SECTION 3

TEST TECHNIQUES AND PRACTICES

3-1 COMMUNICATIONS RECEIVER TESTING

Communications receivers are generally composed of a series of selective RF and AF circuits, each stage of which is designed to amplify the output of the preceding stage. The lowered efficiency of any amplifier, or a change in any one circuit parameter, usually results in lowered over-all efficiency of the receiver. The sensitivity of the receiver may also be decreased by the misalignment of the successive circuits, although each of which may function in a suitable manner as a unit. The sole function of a communications receiver is to receive (selectively) a weak signal; therefore, an objective over-all test on sensitivity is the most significant single check that can be made on the condition of a receiver. Some receivers are provided with a built-in output meter; others require an external indicator, such as a dB meter or a spectrum analyzer to facilitate testing. The only other requirement for a sensitivity check is a calibrated signal for the excitation of the receiver on its various bands. Any decrease in sensitivity must be corrected. In addition to sensitivity checks, qualitative checks as outlined in the planned maintenance manual, (PMM) 43P1, or equipment's technical manual, must also be performed. Adjustment and servicing methods relevant to a specific receiver are discussed in detail in its associated technical manual. When attempting to isolate receiver faults, first test the most accessible (or vulnerable) parts. Previous experience and/or the six-step troubleshooting procedure will aid in isolating troubles. Since a receiver could operate for years with reduced sensitivity before trouble was detected or a complete failure occurred, the maintenance prescribed in 43P1 or the equipments' technical manual is a necessity.

3-1.1 RECEIVER SENSITIVITY

Sensitivity measurements provide convenient and quantitative information regarding a receiver's ability to detect small signals in the presence of electronic noise. The sensitivity of a radio receiver can be defined as the input carrier voltage with standard modulation required to develop a standard value of test output.

3-1.1.1 Impedance Matching Considerations

Sensitivity measurements require the application of an accurately calibrated signal to the antenna input terminals of a receiver, through an impedance matching network that approximates the impedance characteristics of the antenna with which the receiver is designed to be used. The impedance network simulates normal operating conditions, ensures the receiver has the proper match, and ensures the signal current during testing is equivalent to the signal current obtained from an actual signal of equivalent magnitude. In the 15 to 30,000 kHz range, a typical standard matching network consists of a 20-microhenry inductor shunted by a series-connected 400-picofarad capacitor and a suitable 400-ohm resistor, with the shunt combination in series with a 200-picofarad capacitor. This is illustrated in Figure 3-1. The unit is then enclosed within a properly designed grounded shield and used with a signal generator having a resistive output impedance not exceeding 50 ohms. This matching network acts like a 200-picofarad capacitance at low frequencies; like a complex impedance in the 1 MHz region; and like a 400-ohm resistance at frequencies of 2 to 30 MHz. For the measurement of low-impedance-input receivers of 50- to 70-ohms nominal impedance, a signal generator with a 50-ohm output may be directly connected without the use of an external impedance network. Other generator impedances may require special networks to load the generator and the receiver properly, while allowing the equivalent induced antenna voltage to be accurately known. The loudspeaker (or headset) is replaced by a suitable resistor (with the resistance equal to the normal load impedance at the frequency of modulation). Alternately, an indicating device is connected

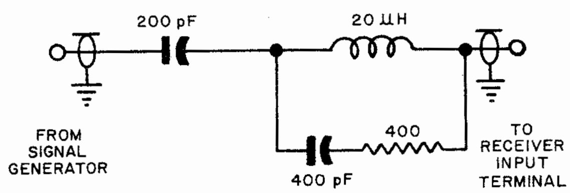


Figure 3-1. Typical Impedance Matching Network

directly to the detector circuit and the power output measured. The sensitivity is then a measurement of the required input signal for a standard power output. The output power may be measured at the receiver output or at the detector, since virtually no noise is generated in the audio stages. Figure 3-2 shows a typical test equipment arrangement for the measurment of receiver sensitivity. In Figure 3-2, 6 dB of attenuation is used to impedance-match the signal generator's output with the receiver. The MX 9407/URC-9 was developed specifically for use with the AN/URC-9 receiver, and has a 600-ohm impedance audio line load incorporated.

3-1.1.2 Am Receiver Sensitivity

Noise-figure measurements are a ratio of the signal-to-noise power ratio of an ideal receiver to the signal-to-noise power ratio of the receiver under test. Sensitivity measurements, however, are relative measurements which are arbitrarily calculated to be the 30 percent, 1000 Hz modulated signal input required to raise the detected audio 10 dB or greater above the receiver's noise level. Sensitivity is measured in microvolts, or in dB below one volt. This arbitrary reference value was selected because a 10 dB change represents a X 10 change in voltage. Therefore, the detected output, as measured across a given resistance with zero signal input to the receiver, can be increased by a factor equal to or greater than 10. This increase can be achieved by increasing the input modulated signal level (in microvolts) to a predetermined level. Note that sensitivity (and selectivity) may be affected by alignment in all types of receivers. As receivers become more complex, alignment becomes more of a problem in amplitude-modulation receivers, improper alignment may result in the loss of weak signals through loss of sensitivity, and through inability to select the desired signal. If the oscillator is shifted off frequency, dial error will be introduced; and tracking error produces a varying intermediate frequency, which results in a loss of signal over portions of the frequency range of the receiver. In frequency-modulation receivers, the discriminator tuning becomes somewhat critical; and in phase-modulated receivers, phasing of the carrier must be correct, adding to the alignment problem. When automatic gain control and automatic frequency control are added to a receiver, proper alignment procedures must be followed, or serious errors may be introduced, further complicating alignment. In equipment employing crystals, as either reference generators, oscillators, or filters, the alignment must center on the crystals since, for all practical purposes, the frequencies of crystals are not variable. As a result, because of the wide variations in circuitry between models of receivers, the actual alignment procedures and specifications provided in the applicable technical manual must be closely followed if the sensitivity check indicates a need for alignment.

3-1.1.3 Single Sideband Sensitivity Measurement Considerations

Sensitivity measurements of single-sideband (SSB) receivers are determined in a manner similar to that used for other amplitude-modulation equipment. However, certain considerations must be taken into account when performing measurements

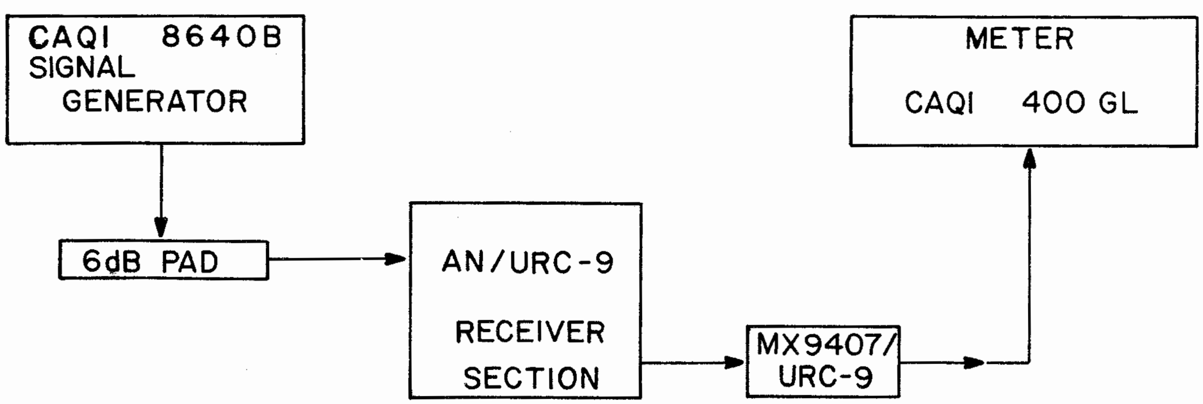


Figure 3-2. Typical Equipment Arrangement for Radio Receiver Testing

on this type of equipment. For example, the oscillator frequencies must be precise because demodulation is directly affected by the oscillator frequency, hence any associated oscillator frequency error will impair reception. The frequency stability requirements of single-sideband equipment operating in the high-frequency region are on the order of 0.2 to parts per million. Frequency errors greater than 30 Hz can cause voice transmissions to be unintelligible with certain signal-to-noise ratios. Intelligibility decreases with an increase in frequency error, even with a high signal-to-noise ratio. When frequency conditions are simulated for the sensitivity measurements, this same high degree of accuracy is still warranted. In addition, filters must also be considered, since these affect the bandpass and rejection of undesired frequencies. Most of the advantages of single-sideband reception depend on the use of these filters. Therefore, proper measurements for single-sideband receivers must include the use of test equipment designed for use with this type of communication equipment, and with an accuracy far greater than the accuracy to be maintained in the receiver. Because of the wide variation in circuitry and layout, actual procedures and specifications must be obtained from the technical manual for the particular equipment involved.

3-1.1.3.1 SSB Receiver Test Equipment

Two very important factors must be borne in mind in the choice of test equipment for maintenance or measurement of SSB receivers: 1) the close tolerance to which the oscillator frequencies must be held; and 2) the sharp edges of bandpass, which must not be defeated. Signal generators specifically designed for single-sideband applications are presently available. Such a generator should include a single-sideband output signal, with provision for either upper or lower sidebands, or both. It should also include a carrier output, with variable (or selectable) frequency and level, both of which should be accurately calibrated. Alternative methods of measurement, using available signal generators, can be used; however, the limitations that may be imposed by the use of these methods should be investigated before a high degree of system analysis or final receiver evaluation is attempted. Signal generators should never be used as reference for oscillator adjustments in either SSB transmitters or receivers, because of the allowable tolerances of such signal generators. Frequency standards with a known accuracy must be used for such adjustments. The reference must have an accuracy greater than the oscillator to be adjusted. Preferably, the accuracy should be greater than ten-to-one; that is, the percentage of tolerance of the reference oscillator should be less than the allowable percentage of tolerance of the oscillator to be adjusted by a ratio of at least ten-to-one.

3-1.1.4 CW(A-1) and Facsimile (A-4) Sensitivity Determination

For determination of CW (A-1) reception sensitivity, some means must be provided to set the output beat note of the receiver to a standard 1000 Hz frequency with reasonable accuracy (about 1000 ± 50). In some receivers, the 1000 Hz "sharp" audio filter provided has a bandwidth narrow enough to allow satisfactory adjustment of the beat note by centering the tone in the pass band. The 1000 Hz internal tone modulation frequency of certain signal generators is also accurate enough, and can be zero-beat against the output beat note. Alternatively, the output of a calibrated audio oscillator and that of the receiver may be fed independently to the deflection amplifiers of an oscilloscope to give the circular or elliptical Lissajous pattern characteristic of identical frequency. For determination of both keyed CW and facsimile (A-1 and A-4) reception sensitivity, set the CW (beat-frequency) oscillator to on, the receiver audio gain at maximum, and the AGC, silencer, noise limiter, and output limiter to off. If not otherwise specified in the receiver technical manual, set audio filters or tone controls for maximum audio range. The antenna trimmer is normally to be peaked at the high-frequency end of each band, and not reset at other frequencies. After the initial adjustments, the following settings are typical: The RF gain control is adjusted to produce 60 microwatts of noise at the receiver output (0.19 volt across 600 ohms) with the receiver tuned to the desired frequency, but with no input signal applied from the signal generator. The signal (carrier only) is then applied, and is tuned as nearly as possible to center on the noise of the overall RF pass-band of the receiver, with the CW oscillator frequency control adjusted to the side of zero beat that produces the higher output with a 1000 Hz beat note. The input-signal voltage is then adjusted to produce a 6-milliwatt output (1.9 volts), resulting in a +20 dB output signal-to-noise ratio. The receiver sensitivity, in terms of input-signal voltage, is then read from the signal-generator voltage calibration. Other reference levels may be used, but the 20 dB voltage relationship should be maintained. In receivers employing frequency synthesis the oscillators are referenced to a frequency standard. The normal preliminary setup would involve setting the function switch to CW (A1), the bandwidth to narrowest position, the AGC to off, BFO (if variable) to midrange, and the audio line terminated in 600 ohms, noninductive. To determine receiver sensitivity, a predetermined signal level is injected into the antenna jack and an indication of receiver sensitivity is obtained from the audio line level meter or on a VTVM connected across the audio load resistor.

3-1.1.5 Voice Modulated (A-3) Sensitivity Determination

The receiver's sensitivity is determined by signal plus noise-to-noise ratio. This is equivalent to a modulation-on/modulation-off ratio. For determination of voice-modulation (A-3) reception sensitivity, a carrier modulated at 30 percent 1000 Hz must be applied. The RF gain control is to be set at maximum, with AGC on and the CW (beat-frequency) oscillator off. (This condition may be automatically established by the reception selector control provided on some receivers.) All other controls, except the AF gain control, are to be set as indicated for CW (A-1) reception. A typical test operation would involve monitoring the audio output line across a 600 ohm load, using a dB meter. A standard voltage input is then applied to the antenna jack from a signal generator through an impedance-matching device. The signal generator is then adjusted to the test frequency. The 1000 Hz, 30 percent modulation signal is switched on and then off for a difference of 10 dB on the dB meter. The minimum signal voltage from the signal generator required to obtain this 10 dB drop is the measure of the receiver's sensitivity. In some applications, the AF gain control is utilized to establish a reference noise level, as indicated on the dB meter. An audio line level meter, incorporated on the receiver, may be employed in lieu of an external dB meter. The drop in dB with the unmodulated signal injected will still be 10 dB, regardless of which method is used.

3-1.1.6 SSB (A-3J) Sensitivity Measurements

Since a single sideband signal is without a carrier, the procedure for measuring SSB receiver sensitivity is similar to that for synthesized CW reception except that no BFO is utilized because the carrier is reinserted by the receiver. The signal is tuned for a peak on the audio line indicator. The output of the signal generator is then set to a predetermined value to obtain a required level on the audio line indicator. Specific values of signal generator settings for audio line indications will vary from receiver to receiver.

3-1.1.7 Tone Modulated (A-2) Sensitivity Measurements

For tone modulation (A-2) sensitivity measurements, the carrier should be 100-percent modulated with a 1000 Hz tone. The receiver RF gain control should be set at maximum with the AGC control off. Then, with the generator modulation off, the receiver RF gain control should be adjusted for a noise output of 60 microwatts (0.19 volt across 600 ohms). The generator modulation should be turned on and the generator output varied until a signal-plus-noise output of 0.6 milliwatts (0.6-volt across 600 ohms) is obtained (20dB output signal-to-noise ratio). If the available signal generators cannot be used at a modulated 100 percent because of excessive frequency modulation or other limitations, an approximate sensitivity measurement may be made by employing 30-percent modulation as used in (A-3) voice modulated procedure. This procedure may give somewhat erroneous results, as detector modulation distortion or modulation clipping by built-in noise limiters may be much less at 30-percent than at 100-percent modulation of the carrier.

3-1.1.8 FSK (F-1) Sensitivity Determination

The receiver, frequency-shift keying (FSK) converter, and the teletypewriter must all operate satisfactorily to produce proper copy in FSK operation. If the receiver checks satisfactorily for CW (A-1) sensitivity, only the additional switching for FSK reception and any special FSK circuitry in the receiver could produce poor FSK operation, at least so far as the receiver proper is concerned. Therefore, the receiver may be checked for FSK sensitivity by initially checking its standard CW sensitivity. If this proves to be normal, switching to FSK operation will allow the output beat frequencies and audio output level to be checked, to ensure they meet the requirements of the specific type of FSK conversion employed.

3-1.1.9 FM (F-3) Sensitivity Measurement

The procedures for measurement of FM (F-3) receiver sensitivity are analogous to those for AM receivers; however, FM signal generator must be used. The modulation signal is a 1000 Hz tone with 2500 Hz deviation. The modulation-on/modulation-off reference is still utilized as in AM and the minimum signal required to obtain a 10 dB drop is a measure of the receiver's sensitivity.

3-1.1.10 Pulse-Modulation Sensitivity Measurement

Continuous-wave generator methods of measuring sensitivity do not provide an accurate indication of the ability of a receiver which is designed for the reception of pulse-modulated signals to receive weak pulse transmissions. A better method of determining the sensitivity of a pulse-modulation receiver involves performing a minimum-discernible signal measurement. This type of measurement consists of measuring the power level to a pulse whose level is just sufficient to produce a visible receiver output. However, because of the relatively wide bandwidths associated with pulse-modulation receivers, a still better performance indication can be obtained by determining quantitatively how much noise is inherent in the receiver, since noise is the limiting factor in the determination of maximum sensitivity. This method of checking sensitivity utilizes a noise generator as the most desirable test equipment for a signal source. The noise in the receiver is related to a calculable noise figure.

3-1.2 DETERMINING I-F BANDWIDTH RESPONSE

A graph showing the bandwidth response of an I-F amplifier can be constructed by plotting frequency horizontally (from left to right) and signal amplitude vertically. This method would be ideal for record retention purposes, but is not necessary for receiver checks and adjustments. For such checks and adjustments, a spectrum analyzer is used in conjunction with a tracking generator. Figure 3-3 illustrates such an arrangement. In Figure 3-3 the voltage control oscillator feeds both the spectrum analyzer's mixer and the tracking generator's mixer. Because of this simultaneous precision tracking, the tracking generator's output frequency acquires the same scan capabilities as the spectrum analyzer. Therefore, the analyzer's calibrated scan widths, which range from broadband to extremely narrow, are acquired by the tracking generator and can be applied to various I-F strips. The configuration in Figure 3-3 makes identification of any point on the display quite easy and unambiguous. The 3 dB point, 60 dB point, center frequency, or any point on the display can be measured by stopping the scan (either electronically or manually) at the point of interest and then reading the indication on the tracking generator.

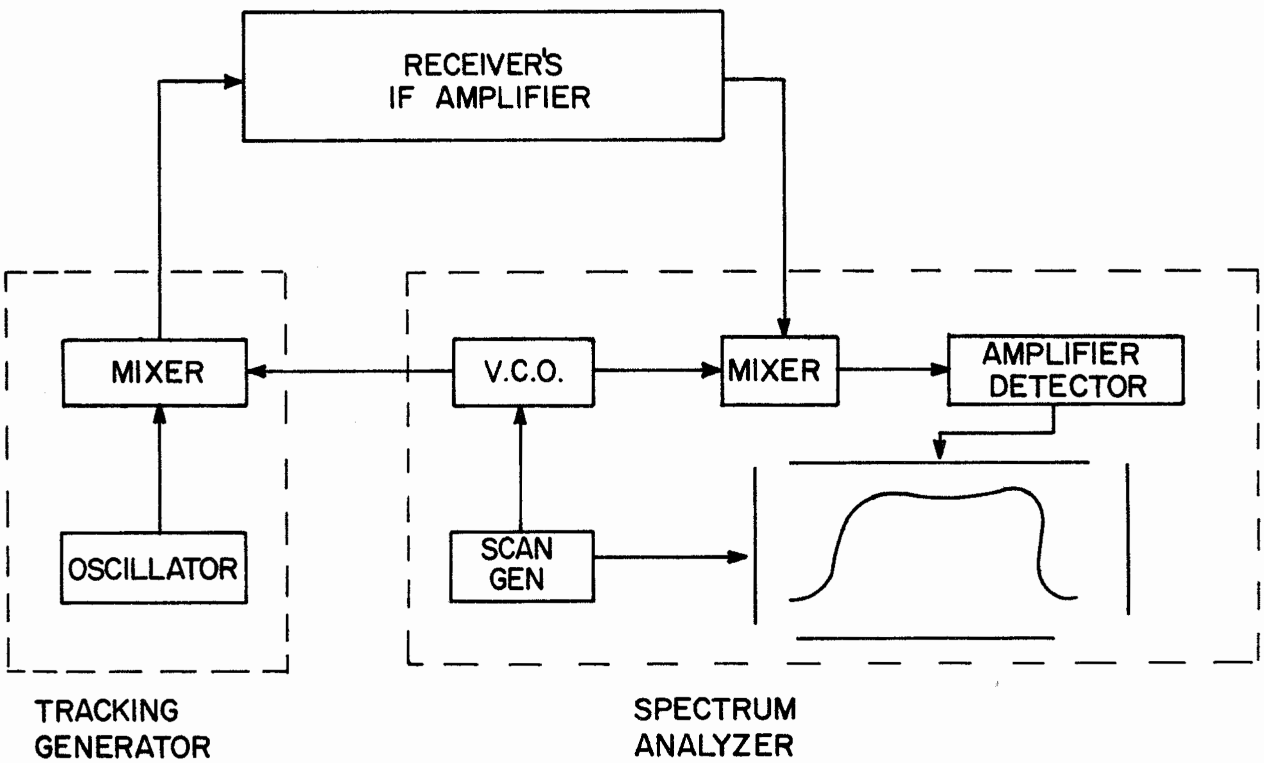


Figure 3-3. Equipment Arranged To Obtain A Visual I-F Bandwidth Response

3-1.3 SELECTIVITY AND BANDWIDTH MEASUREMENTS

Selectivity is the property which enables a receiver to discriminate against transmissions other than the one to which it is tuned. It is usually expressed in the form of a curve obtained from a plot of the strength of a standard modulated-carrier signal required to produce a constant (standard) output, versus off-resonance frequency. Figure 3-4 shows a typical selectivity curve with the carrier signal strength at resonance used as a reference. The bandwidth of a receiver is usually employed to define that portion of the selectivity curve that represents the frequency range over which the amplification is relatively constant. For most receivers, the bandwidth represents the usable portion of the curve, and has a direct relation to the fidelity of the modulated intelligence. Practically, the bandwidth is measured at the half-power (3-dB-down) points, or, for certain applications, at the

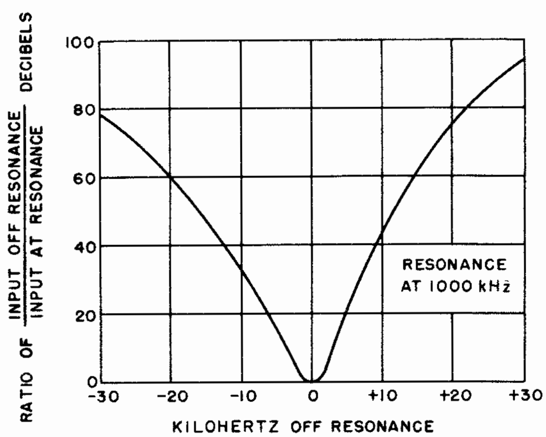


Figure 3-4. Selectivity Curve of Typical AM Receiver

6-dB-down points, and is represented by the frequency range between the two points on a response curve expressed as relative response in dB versus frequency, as shown in Figure 3-5. However, the bandwidth at the 3 dB (or often the 6 dB) points, when compared with the bandwidth at the 60-dB-down points, gives a good indication of the selectivity of the receiver, since the character of the skirts of the curve becomes apparent. This comparison is referred to as the bandwidth, or selectivity ratio. In most receivers, the overall bandwidth is determined by the I-F amplifiers; therefore, bandwidth is sometimes considered as fundamentally an I-F characteristic measurement.

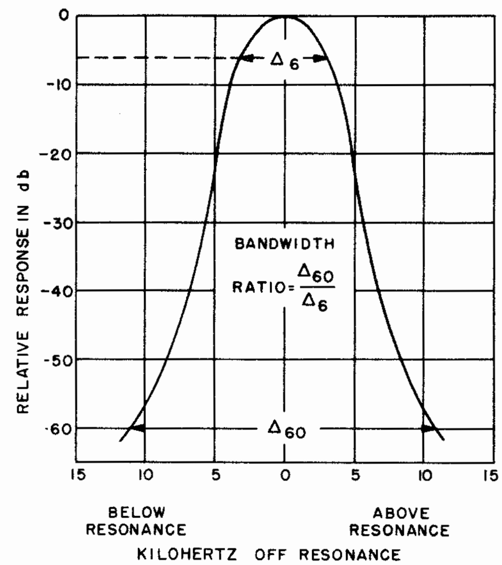


Figure 3-5. Receiver Response Curve

3-1.3.1 Overall Selectivity

Since the RF stages of a receiver are also of some importance in determining the selectivity, and are of fundamental importance in determining the image rejection characteristics, the selectivity factor is most often plotted as overall selectivity. The term "overall selectivity" usually refers to the frequency selectivity of a receiver as measured from (and including) the antenna to the input terminals of the final detector. It does not normally include any elements of the audio system. The overall selectivity of a superheterodyne receiver may be quite difficult to measure accurately with the equipment available in most operating installations, especially at frequencies above 1 MHz. If the lowest signal frequency is at least several times that of the lowest intermediate frequency used in the receiver, the overall selectivity is very likely to be practically the same as the lowest I-F selectivity. Therefore, the lowest I-F selectivity curve may suffice and it is much easier to obtain.

3-1.3.2 Bandwidth

When making bandwidth measurements, the receiver's AGC should be disabled (grounded), connected to a source of fixed bias, or turned off, and the volume control set to maximum. Bandwidth curves can be obtained with the test setup illustrated in Figure 3-3. This procedure can be used for narrow or wide-band receivers employing any type of demodulation. When making I-F bandwidth measurements, the spectrum analyzer is set to the I-F's center frequency. The scan width and scan speed controls are then adjusted to achieve an undistorted display on the CRT. If the scan time is too short with respect to the scan width of the spectrum analyzer, the response curve will appear wider than it actually is and the amplitude will be greatly reduced. This condition is illustrated in Figure 3-6.

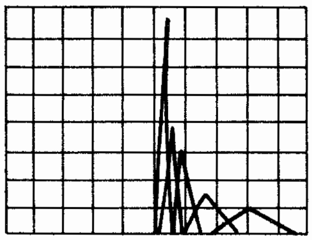


Figure 3-6. Effects of Decreased Scan Speed With Constant Scan Width

3-1.4 AGC MEASUREMENTS

An automatic gain control AGC circuit reduces the effect of signal strength fading by maintaining a constant carrier level at the detector input of an AM receiver, despite variations of the input signal carrier level. To determine the effectiveness of the AGC circuit, its characteristic should be measured at the center frequency of each band covered by the receiver. A curve can then be plotted to compare the change of the receiver output to input signal levels. The standard method of measuring the AGC of an amplitude-modulated receiver is to set the signal generator for 30 percent, 1000 Hz modulation at the carrier frequency. The receiver gain is set to maximum, and the carrier level is then varied over a wide range, such as .4uv to 50uv. The relative output is then plotted in dB as a function of carrier input voltage, which is also presented in terms of dB. Either the output meter incorporated in the receiver or an external indicating device, such as a VTVM or a spectrum analyzer, can be employed to obtain the output measurement. Figure 3-7 illustrates a typical plot of carrier signal vs power output with different percentages of modulation used. In single sideband (SSB) receivers, the procedure transmission is similar except that the carrier is unmodulated.

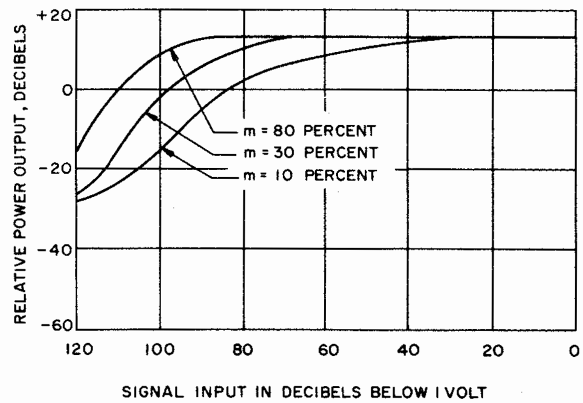


Figure 3-7. Automatic Gain Control, Characteristic Curve

3-1.4.1 Delayed AGC Considerations

Delayed AGC circuits are often incorporated in a receiver because even the weakest signal received in conventional AGC circuits tends to reduce the gain of the receiver somewhat. The delayed AGC adaptation incorporates a separate diode (AGC diode) in addition to the detector diode. Part of the signal fed to the detector diode is coupled to the AGC diode by a small capacitor. The AGC diode is maintained at a suitable bias; this bias keeps the diode until the peak voltage of the amplified signal voltage equals the bias introduced to the diode. For very weak signals which do not produce enough voltage on the anode of the AGC diode to overcome the existing negative potential, no AGC voltage is developed. Thus, the sensitivity of the receiver remains constant, just as if the automatic gain control were not being used. When normal-strength signals are being received, which do not need the maximum sensitivity of the receiver, enough signal voltage will be coupled to the AGC diode to overcome the bias applied. AGC voltage will thus be developed normally for these stronger signals. Measurements on delayed AGC circuits are made in the same manner as described for conventional AGC circuits; however, particular attention should be given at the low-input portions of the curve.

3-1.5 RECEIVER STANDARD MEASUREMENTS

In the measurement of single sideband reception, it is imperative that the standard oscillator used be precise. There are two methods of determining the precision of the frequency standard. One method involves measuring the output of the standard against a frequency counter. The frequency counter must therefore be more accurate than the standard to be measured. Most SSB receivers incorporate a means by which an external primary standard can be compared with the internal receiver standard. The two signals are fed to a difference network whose output is fed to either a meter with zero center swing, or to a light which blinks on and off. In both instances the receiver's standard is compared against the primary standard input for a minimum to zero change. Any necessary adjustments are made only after the receiver has had sufficient time to warm up thoroughly (approximately 3 days), and the adjustments are done in very small increments.

3-1.6 SQUELCH (SILENCER) CIRCUIT MEASUREMENTS

FM and high-frequency receiver circuits inherently have a high noise level when no signal is being received. During communications, where a receiver is tuned to a specific frequency for long stand-by periods in anticipation of signals that may appear at any time, the continuous roar of noise is highly objectionable to anyone in the vicinity of the receiver. It is therefore desirable that a squelch (or silencer) circuit be incorporated to silence the audio output during these periods when no signal is being received. This will eliminate unwanted signal noise and other disturbances that are annoying and fatiguing to the operator. Squelch circuits operate by blocking the input to the audio stage of the receiver whenever the signal voltage is very low or is entirely absent at the detector (in the case of AM receivers), or at the limiter (in the case of FM receivers). The squelch circuit accomplishes this silencing effect by: 1) applying a very large cutoff bias to the first audio amplifier; 2) by actuating a relay to open the audio line; or 3) by gating open the audio line with a Field Effect Transistor (F.E.T.), as illustrated in Figure 3-8. The high pass filter of Figure 3-8 removes all low-frequency signal components and passes the high-frequency noise components. The high-frequency noise increases with a decrease in signal strength, thus providing a gate control signal to the audio output gate F.E.T., cutting it off in low to no signal condition. When cutoff bias squelch is required, it must be sufficiently in excess of cutoff to prevent the noise output from the intermediate amplifiers from causing current to flow in the first audio amplifier stage, even momentarily, on the noise peaks. For the determination of the squelch characteristic, connect

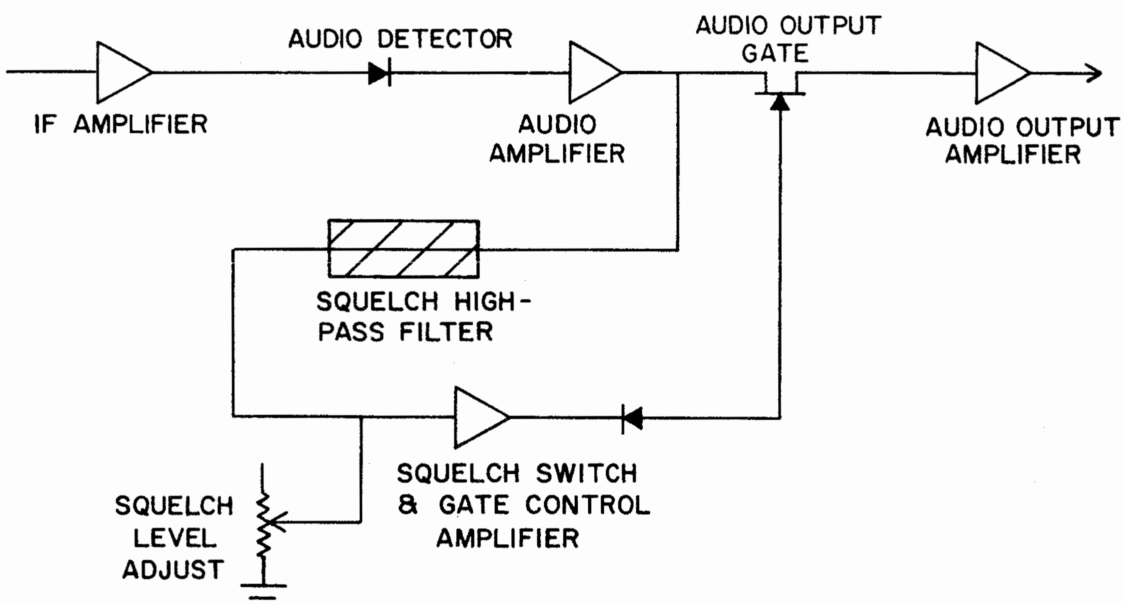


Figure 3-8. Squelch Circuit Employing F.E.T.

the test equipment to the receiver as shown in Figure 3-2. The signal generator should be set to frequency with 1000 Hz, 30 percent modulation. With the signal generator RF output control set for zero output, the receiver output should be noted; it should be essentially zero. Gradually, the signal generator RF output should be increased until the squelch circuit operates. Operation of the squelch circuit is indicated by a sudden increase in the radio receiver output. The signal generator RF output required for the operation of the squelch circuit may be recorded as representing the squelch characteristic.

3-1.7 MODULATION DISTORTION MEASUREMENTS

The distortion produced in the radio-frequency, intermediate-frequency, and detector stages of a receiver can increase significantly as a result of increases in the percentage of modulation. For example, the distortion generated by a square-law detector becomes prohibitive at percentages of modulation greater than about 50 percent, and for this reason such detectors are rarely employed in communications receivers. One method for determining the distortion of a receiver in terms of the percentage of modulation is to connect a signal generator to the receiver input, and to shunt a suitable resistor across the receiver output. A distortion meter should be connected across the resistor. The distortion meter will not respond to the resonant frequency, which is suppressed, but will provide the rms value of the other components of the distorted output signal. It will thereby provide an indication of the amounts of distortion for calibrated percentages of modulation. The signal generator should be modulated at 1000 Hz and set for an output on the order of 50 microvolts. The receiver volume control should be adjusted for a low-level output (on the order of 50 milliwatts) and this level should be maintained throughout the test. Maintaining this low power output level keeps the distortion contributed by the audio section to a low, constant level. The percentage of modulation at the generator is then increased in convenient steps from 10 to 100 percent, and the results are plotted on linear graph paper, with the modulation percentage appearing horizontally and the values of distortion vertically. This test should then be repeated for different RF gain settings to determine whether the RF and the I-F amplifiers affect the modulation.

3-1.8 AFC CHARACTERISTIC MEASUREMENTS

Automatic-frequency-control (AFC) circuits are most often found in frequency-modulation receivers and in very-high-frequency and ultra-high-frequency receivers, because of the high degree of oscillator frequency stability required. Thus, FM receivers incorporate discriminator circuits whose output voltage and polarity are contingent upon the direction and deviation from a center, or mean, value. For purposes of oscillator frequency control, a sampling of this voltage is filtered to remove any AC component. The resulting variation in DC voltage is applied to the local oscillator, which is a VCO (voltage-controlled oscillator). The VCO varies frequency as a function of the direction and magnitude of the applied correction voltage. The voltage-sensitive component of the VCO can be in the form of a reactance tube, a saturable reactor, or a varicap. All three vary the reactance of the oscillator tank circuit, thus changing the oscillator's frequency. The use of this technique can decrease the amount of frequency drift by a factor as high as 100 to 200 percent with respect to an uncontrolled receiver. To determine the locking range of the AFC circuit, connect a signal generator to the input of the receiver at some suitable level at the center frequency. Tune the signal generator both above and below the center frequency and note the break-off points.

3-1.9 RECEIVER ALIGNMENT

Sensitivity and selectivity may be affected in alignment of all types of receivers. As receivers become more complex, alignment becomes more of a problem. In AM receivers using conventional full-carrier signals, improper alignment may result in loss of weak signals, through loss of sensitivity, and the ability to select the desired signals may be impaired. If the oscillator is shifted off frequency, dial error will be introduced, and tracking error produces a varying intermediate frequency, which results in loss of signal over portions of the frequency range of the receiver. In FM receivers the discriminator tuning becomes somewhat critical, and in PM receivers, phasing of the carrier must be correct, adding to the alignment problem. When AGC and AFC are added to a receiver, proper alignment procedures must be followed, or serious errors may be introduced. When multiple conversion is incorporated in the receiver, with two or more herodyne oscillators, additional variables are introduced, further complicating alignment. In equipment employing crystals, as either reference generators, oscillators or filters, the alignment must center on the crystals, since the frequencies of individual crystals are not variable.

3-1.9.1 Alignment Of Crystal Filter Circuits

Crystal filters are incorporated in communications receivers which require an extremely high order of selectivity. These filters are usually located between the receiver-mixer and intermediate-amplifier stages. The crystal is the major component of the filter, and is used because of the extremely high Q that can be obtained from its use. Ordinarily, the filter is externally adjustable so that a variation in bandpass can be obtained. Crystal filters are often used as wave traps, and are extensively used in single-sideband equipment because of the sharp-frequency-cutoff properties required for this type of equipment. However, these are generally of the crystal-lattice type; they are usually hermetically sealed or potted, and should not be tampered with. Some lattice or half-lattice-type crystal filters are provided with adjustable trimmers, accessible as screwdriver adjustments. Some of these are labeled as factory adjustments and should never be disturbed. In any case, the manufacturer's data should always be consulted before any adjustments are attempted. Plug-in filters are readily replaced. Mechanical filters should never be tampered with or repairs attempted under any conditions. Schematic diagrams of crystal filters usually indicate variable capacitors and (often) variable inductances. Such diagrams may be misleading to those unfamiliar with filter circuits. Capacitors in parallel with filter crystals are usually of very small value: on the order of one to several picofarads. These frequently consist only of leads given a slight wrap or twist, or they may be a piece of wire bent near the crystal holder or electrode. Such capacitors are factory-adjusted, and are usually not accessible without dismantling the filter. The schematic symbols for crystal and its equivalent electrical circuit are illustrated in Figure 3-9A. A crystal in its holder is actually a combination of both series-and parallel-resonant circuits, and as such it has two resonant frequencies, as indicated in Figure 3-9B. The series-resonant frequency occurs at the point where the reactance curve crosses the zero-reactance line, and the parallel-resonant (anti-resonant) frequency occurs at the point where the reactance curve rises to a high inductive reactance. It then falls sharply through the zero-reference line to a high capacitive reactance. In most crystals, the two resonant frequency points will

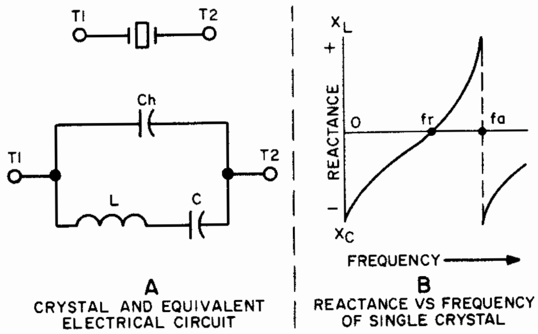


Figure 3-9. Equivalent Electrical Circuit and Reactance Curve of a Quartz Crystal

occur within a few hundred cycles of each other, but the points can be spread (or narrowed) by shunting them with a lump constant so that a suitable filter network can be designed. Phasing controls on interference filters are examples of capacitance introduced into the filter circuit to shift the crystal rejection slot (parallel-resonant frequency) so that specific unwanted signals can be rejected. When aligning filter circuits, the circuit in which it is integrated must be considered. When connected in the intermediate-frequency amplifiers of a communications receiver, either conventional AM or single sideband, the alignment will consist principally of properly tuning the resonant input circuit to the filter, and to the resonant circuit at the output of the filter, as shown in Figure 3-10, for maximum output. The points of parallel resonance must be aligned for sharp cutoff (maximum attenuation) at the design frequency. This is an especially important consideration for equipment containing automatic-frequency-control (AFC) circuits. If sufficient response is not allowed, the carrier may be severely attenuated at a slightly too low (or too high) frequency, thereby causing the AFC circuit to drop control. Thus, the desired limits of AFC operation are also considered in the bandpass of the filter, and vice versa.

3-1.9.2 Alignment Of Wave Traps

The term "wave trap" usually refers to a resonant element used as an auxiliary device to provide additional frequency selectivity in a radio circuit. It may take a distributed form (resonant stub or cavity), or it may consist of a lumped reactor combination (inductor and capacitor), as shown in Figure 3-11. A trap normally provides a means of rejecting (or accepting) signals over only a relatively narrow band of signal frequencies, of a width which depends in part on

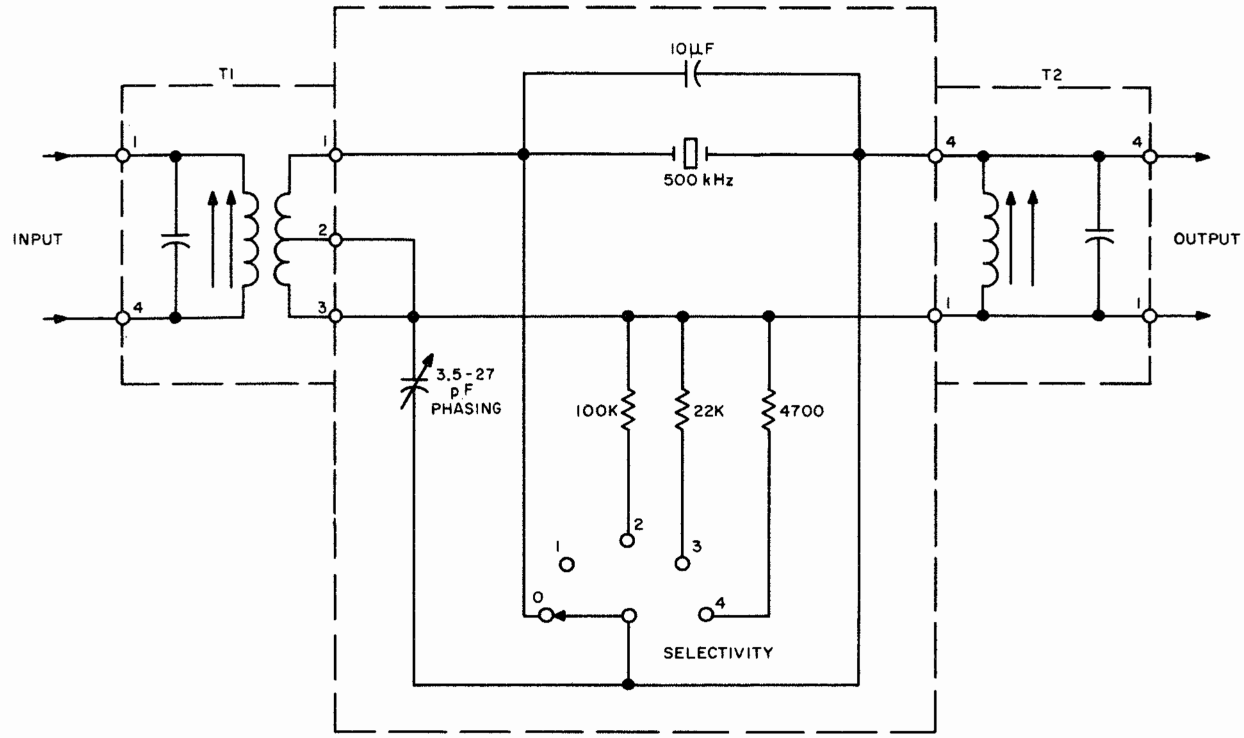


Figure 3-10. Crystal Filter Circuit

the effective Q of the trap circuit. The trapping desired may result from the "shorting" effect of a series-resonant circuit shunted across the signal path; from the selective opposition to the flow of current afforded by a high value of resonant impedance in series with the path; from selective degeneration in an amplifier, produced by using resonant circuits to provide frequency-dependent feedback, etc. Some of the more common lumped-reactor wave-trap applications are shown in Figure 3-11. In addition to those illustrated, many other forms of wave traps may be incorporated in radio equipment. In some applications, a wave trap is used to suppress response at a frequency not desired in one channel. The resulting trap resonance at that frequency is used as a means of supplying the signal to a second channel. Resistance-capacitance (RC), resistance-inductance (RL), and inductance-capacitance (LC) networks, affording high-pass and low-pass characteristics, are also employed to provide band elimination or bandpass effects for wave-trapping purposes. The operating frequencies and apparent effects of wave traps differ from one type of equipment to another. In general, it can be expected that the traps will be left until the last steps in a prescribed alignment procedure, because of their auxiliary or corrective nature. Adjustment of wave-trap trimmers must usually be accomplished at very specific frequencies and under particular conditions, which should be rigidly observed. If adequate instructions for wave-trap alignment are lacking in an equipment technical manual, immediate steps should be taken to obtain further instructions. An incorrectly adjusted trap circuit may produce serious shortcomings in equipment operation, which are not apparent to the operator under ordinary conditions. The signal generator and output indicator commonly employed in the alignment of receiver-tuned circuits will usually serve for wave-trap alignment in receiving equipment. Other forms of radio equipment employing traps, such as field-strength meters and oscilloscopes, may require special instrumentation.

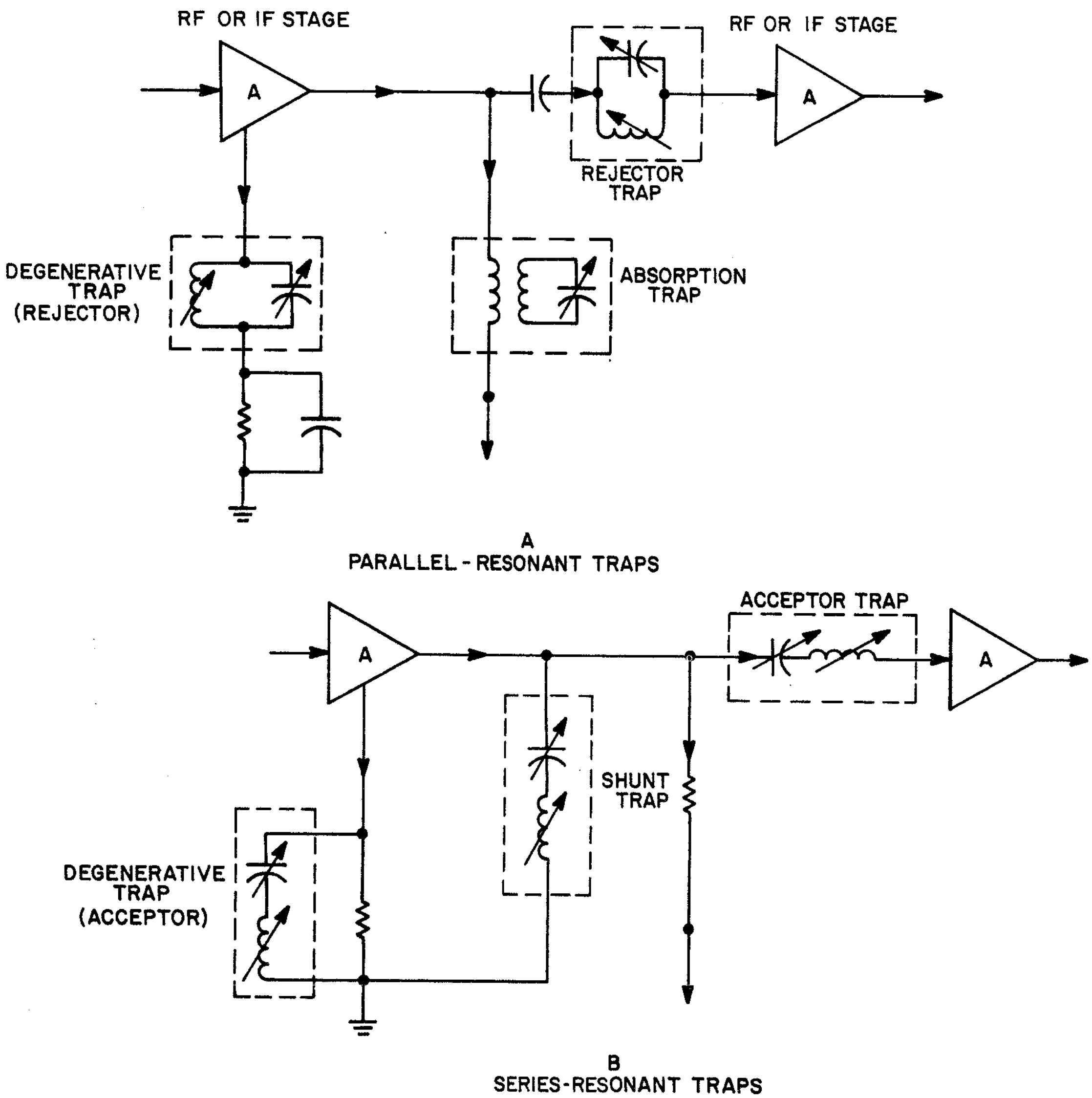


Figure 3-11. Wave-Trap Circuits

3-1.9.3 Alignment of Beat-Frequency Oscillators

Beat-frequency oscillators (BFOs) are incorporated in communications receivers to provide an audible indication of received continuous-wave transmissions. They are also used to calibrate dials in receivers containing internal crystal calibration oscillators. Figure 3-12 shows a schematic diagram of a

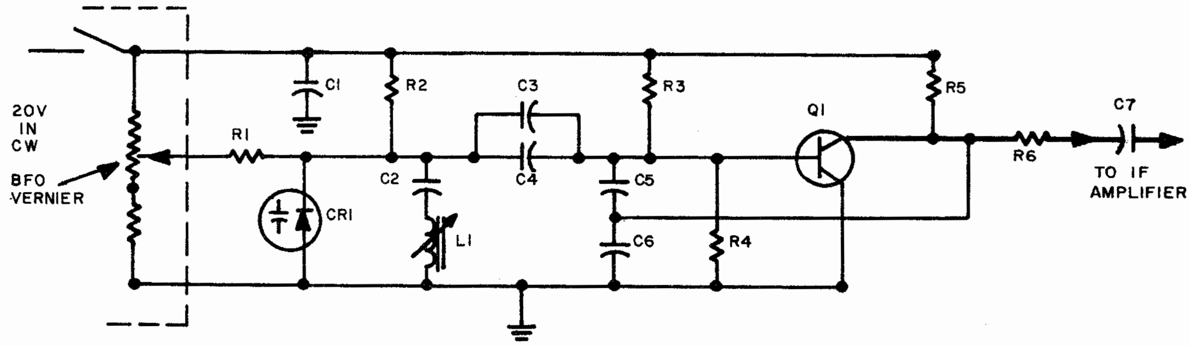


Figure 3-12. Beat-Frequency Oscillator Circuit

typical manually operated BFO which is heterodyned against the intermediate frequency. The procedure for aligning this type of oscillator requires feeding an unmodulated signal into the input of the receiver, with the receiver switched to the CW mode of operation. A frequency counter is connected to the phone jack to record the frequency of the detected signal caused by heterodyning the BFO with the incoming I-F signal. Then L1 is adjusted so that the BFO vernier control will cause a readout on the frequency counter of 6000 Hz above and below the zero beat.

3-1.9.4 AM Receiver Alignment

Prior to the alignment of an AM receiver, the automatic gain control (if possible) should be turned off, and the gain adjusted by means of the manual radio-frequency gain control. The gain level should be set to give the standard 6 milliwatts of audio output with about 100 to 1000 microvolts of signal input at the receiver antenna terminals. This alignment condition is desirable to reduce the detuning effect of receiver gain variations as reflected in changes of overall selectivity. It ensures the circuits will be resonated under average load conditions at approximately the middle working value of amplifier's input reactance, and with freedom from serious regeneration. With most receivers, this condition also reduces receiver noise to a degree which renders it unnecessary to quiet the receiver by removing the amplifier stage preceding the point of alignment-signal injection.

3-1.9.4.1 Disabling Automatic Gain Controls

Amplitude-modulation receiving equipment which operates with automatic gain control as a permanent condition, with no built-in provision for the alternative manual control of RF and I-F gain, may present a problem, especially if considerable regeneration is normally present at full gain. It may not be feasible to disable the automatic gain control in order to add a temporary battery-biased manual gain control potentiometer in its place. In such cases, it will be necessary to align each section of the receiver, with suitable signal-input levels at the various points of signal injection, to produce final detector operation below the threshold of AGC action. The amplifier stage preceding the point of alignment-signal injection should be disabled to preclude the presence of unwanted signals and noise.

3-1.9.4.2 Disabling Local Oscillators

It is preferable not to disable the heterodyne oscillator or oscillators when aligning a receiver, except for the beat-frequency oscillator (BFO) which is used to provide tone output from the final detector in CW reception. The heterodyne oscillator injection voltage is ordinarily a major factor controlling the mixer's operating bias and impedance, with consequent influence on gain and both mixer input and output circuit resonance. In some cases, suitable adjustment of heterodyne oscillator tuning may not be feasible as a means of preventing undesired beats or spurious signals that may result from the interaction of the alignment signal and the heterodyne injection voltage. The oscillator must then be disabled. Stopping an oscillator by short-circuiting its input to ground or by shorting its tank circuit may cause serious damage to the oscillator and to other electronic parts. Therefore, removal of the oscillator output is the safest way to disable the oscillator. The receiver's technical manual will detail the procedure for oscillator removal if it is required for alignment.

3-1.9.4.3 BFO Considerations

The beat-frequency oscillator (BFO) injection voltage in a properly designed CW receiver usually produces a large fixed bias at the final detector, which will mask its rectified voltage changes. This masking is very objectionable when the rectifier signal voltage is employed as an output indication for alignment. Before starting the actual alignment, the technician must disable all those auxiliary functions provided in the receiver which may interfere with proper output indication or circuit resonance. This includes silencer (squelch) circuits and noise limiters.

3-1.9.4-4 I-F Amplifier Alignment

With a few exceptions, such as some trap circuits, I-F resonant circuits are aligned by adjusting their trimmers to produce maximum signal voltage. The I-F trimmers of the typical AM receiver are thus adjusted to produce maximum final-detector signal input voltage, using the input-signal frequency or frequencies prescribed in the technical manual for the equipment. In many cases, this will be the nominal band-center frequency of the particular I-F amplifier. In other instances, usually involving relatively wide I-F passbands, "peaking" of some or all trimmers for maximum response at one or more frequencies off the band-center will be be specified. In general, the last I-F transformer preceding the detector should be aligned first, unless a different order is specified in the equipment technical manual. The input from the signal generator should be adjusted to produce a signal output level which is well above the noise level at the output indicator, but which is also well below the saturation level of the amplifier stages. The signal input should be progressively reduced as needed, as more circuits are brought into proper alignment, with the progression of circuit adjustment moving toward the mixer stage. After the first round of alignment adjustments of the I-F amplifier stages is completed, an overall check of the I-F alignment should be made. A similar procedure should be used for the alignment of the preceding I-F amplifier(s) in receivers employing more than one frequency conversion. The I-F signal input should, in each case, be injected at the input electrode of the mixer preceding that particular I-F amplifier. This ensures inclusion of the transformer located in the output circuit of the mixer. The associated conversion oscillator should be disabled if necessary, as previously discussed.

3-1.9.4.5 RF Stage Alignment

In addition to a suitable signal generator, the dummy antenna specified by the receiver instruction book should be used to simulate an ideal antenna for the receiver. The signal generator (modulated, as required) should first be accurately adjusted to the upper alignment frequency specified for the particular receiver tuning band, using an external frequency standard if necessary. If an antenna trimmer control is provided on the front panel of the receiver, it should be set to the middle of its range. The receiver is then to be carefully tuned to that signal frequency, and the generator output adjusted to produce the desired maximum (or other specified optimum deflection on the receiver output indicator. The tuning-dial frequency indication should coincide closely with the signal frequency being supplied. If it does not, the tuning dial should be reset to indicate the proper frequency. The high-frequency (shunt capacitance) trimmer of the oscillator tank circuit should then be adjusted to produce optimum output from the test signal. Following these adjustments, the interstage and antenna circuit shunt-capacitance trimmers should be adjusted for optimum output, with the test signal input level reset, as needed, to avoid receiver saturation effects. Oscillator shunt trimmers occasionally have an unusually wide range of adjustment. For this and other reasons, it is possible to misalign the circuits so as to place the heterodyne oscillator on the wrong side of the signal frequency. In many instances, this mistake will be revealed in an inability to obtain anything resembling good circuit tracking over the tuning band. Sometimes, however, the mistake will not be so clearly apparent. Therefore, it is always wise to ascertain that the oscillator is being trimmed on the proper side of the desired signal frequency. Determination of the proper relationship from the equipment instruction book, together with careful observation of shunt trimmer positioning (whether its capacitance is increasing or decreasing relative to the two positions of heterodyne response which it produces), will help to prevent error. Next it is necessary to check oscillator alignment at some specified frequency near the low-frequency end of the tuning band. In many military receivers, iron-core or eddy-current trimmers are provided in the RF coils to permit tank inductance adjustments for optimum low-frequency tracking of all RF circuits. The inductance adjustments should be made on all coils except the oscillator coil before the oscillator series padder is checked. The series padder should then be trimmed to produce optimum output while the receiver tuning control is "rocked" back and forth through the region of best signal response. When this process is completed, the shunt trimmer adjustments should be "touched up" for optimum response and correct tuning-dial reading (calibration) at the high-frequency alignment point in the band. The low-end padder adjustments should then be touched up for optimum response, and the tuning-dial reading checked against the test-signal frequency. If oscillator tracing relative to the other RF circuits is poor over the band, as indicated by abnormal variations of gain (and/or output noise) as the tuning control is operated throughout its range, it may be necessary to adjust the oscillator tank inductance trimmer. The correction needed to produce better tracking may be determined by trial readjustment of the oscillator shunt-capacitance trimmer. If the tracking (as checked by tuning from the high-frequency to the low-frequency alignment points) can be improved by increasing the shunt trimmer capacitance, the oscillator tank inductance is low. If the shunt trimmer capacitance must be decreased to obtain improvement, the tank inductance is high. Correction adjustment of oscillator tank induction will necessitate some changes in oscillator series-padder and shunt-trimmer adjustments, and the entire preselector alignment procedure must be repeated. If the tuning dial is still in error over part of the band, it may be possible to correct the calibration to some degree by further slight readjustments, and the entire preselector alignment procedure must be repeated. If the tuning dial is still in error over part of the band, it may be possible to correct the calibration to some degree by further slight readjustments to the oscillator and other trimmers. This realignment should be undertaken only after careful study of the tracking discrepancies and calibration errors over the entire band, and with a full understanding of the superheterodyne tracking problems if adequate directions are not available. In general, it is inadvisable to sacrifice receiver gain and selectivity for the minor convenience of accurate tuning-dial calibration. Receivers that incorporate I-F traps in their RF circuits should be checked by applying a signal, at the intermediate frequency, to the receiver input. The trimmers for such traps are usually adjusted for minimum output at the center frequency of the first I-F amplifier, and may require large input signal amplitude at that frequency.

3-1.9.5 FM Receiver Alignment

The basic difference between receivers used for the reception of frequency or phase-modulated signals and those used for the reception of amplitude-modulated signals lies in the types of demodulator and I-F amplifier circuits employed. In an FM receiver, a frequency-sensitive demodulator is used, and the I-F amplifiers are designed to cause, rather than avoid, amplitude limiting. When testing several amplifier stages that have similar operating functions, such as successive I-F stages, it is possible (but not recommended) to test immediately for an "overall" response curve like that shown in Figure 3-13. Such a curve may be obtained at the output of the last I-F stage or at the grid of the limiter. When using an FM signal generator for testing wide-band equipment, such as an FM receiver, the response curve can be observed directly on the screen of an oscilloscope. Improperly applied, this procedure could consist of varying "at random" the different adjustments in all the stages until the over-all response curve appears to be satisfactory. The apparently good-looking curve results only too often from a compromise. This generally means that a poor alignment in one stage is compensated by overemphasized and shifted alignment in other stages. The reason for this is that one stage may be peaked unsymmetrically, another stage may have a center peak, and the remaining stages may have two response peaks. This has reference to all I-F stages. Hence, no stage by itself satisfies the condition required for linear networks with respect to amplitude and phase, and distortion is bound to result. Regardless of the method of aligning the receiver, the recommended practice is to first align the discriminator (provided that a sufficient signal source is available, the sensitivity of the indicator is high, and the I-F transformers are not excessively detuned). The remainder of the set, up to the discriminator, is then aligned. The correct I-F alignment should always be made by aligning first the I-F stage ahead of the limiter, then the preceding I-F stage, etc. As far as the RF stage or stages before the mixer stage are concerned, ordinary single peaking is generally practiced, since coil and tube damping

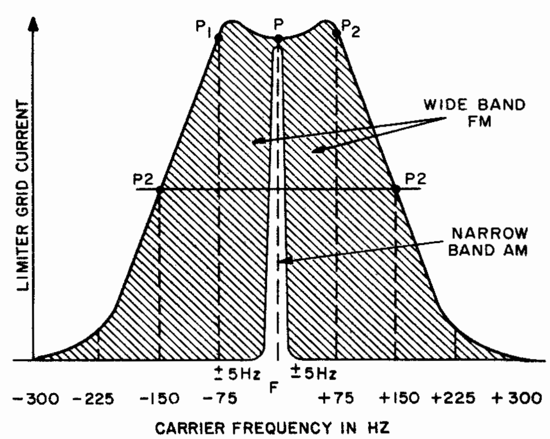


Figure 3-13. FM Versus AM Resonance Curves

nearly always provides the required broadness of the response curve. The receiver can be aligned most conveniently by use of an FM signal generator. However, if such a generator is not available, an AM generator can be used. A meter or an oscilloscope may be employed as an output indicator. Normal receiver alignment consists of the following sequence:

1. Alignment of the demodulator (discriminator) stage.
2. Alignment of the limiter stage.
3. Alignment of the I-F amplifier stages.
4. Alignment of the RF stages.

3-1.9.5.1 Limiter-Type Discriminator Alignment

Figure 3-14 is a schematic diagram of a limiter-type detector. In the double-tuned circuit shown, the primary and secondary are tuned to the carrier frequency. At the carrier frequency, the voltages developed across the diodes are equal to each other, and the diode currents are also equal. Thus, the opposing voltages developed across the output diode resistors are equal and therefore cancel. As a result, no voltage is developed at point A. At frequencies lower than that of the carrier frequency there is more current flow through the lower diode than through the upper diode, thereby causing a negative voltage to be developed at point A. Conversely, at frequencies higher than that of the carrier there is more current flow through the upper diode, causing a positive voltage to be developed at point A. Within the range of carrier-frequency swing, the dc output changes in proportion to the frequency change. There are several ways of measuring the linearity of a discriminator. The most straightforward method requires the use of a high-resistance voltmeter to measure the output voltage between point A and ground (Figure 3-14), while varying the applied intermediate frequency in known steps. A voltage of the center intermediate frequency is applied to the limiter (or to the mixer if the I-F amplifiers are properly aligned and the limiters are properly set). The setting of the secondary capacitor C1 is then varied until zero output is noted. (Note that both terminals of this capacitor are above ground potential; therefore, an insulated screwdriver should be used for this adjustment). The signal generator is then set above and below the intermediate frequency. The voltmeter should indicate equal but opposite direct voltages for equal but opposite frequency deviations. If unequal voltages are obtained, the setting of primary trimmer capacitor C2 is incorrect and must be adjusted until equality is obtained. If necessary, repeat these operations until the proper indications are obtained. The linearity over the entire range can be determined by plotting the values obtained for steps of frequency deviation, as shown in Figure 3-15A. The output of the generator should be constant, or the limiter should be in full operation. A visual method of aligning the discriminator, using an FM signal generator and an oscillscope, has an advantage over the meter method in that the discriminator curve may be observed. Since the effects of the adjustments are visible, no guesswork is involved. There are two methods of setting the discriminator to its proper center frequency. The first method to be described is recommended for its more accurate and more easily observed results. It consists of applying an amplitude-modulated RF signal of the correct center frequency to the discriminator, then adjusting the discriminator secondary for minimum signal output. The output signal will disappear if the discriminator characteristic is symmetrical about the center frequency, since the

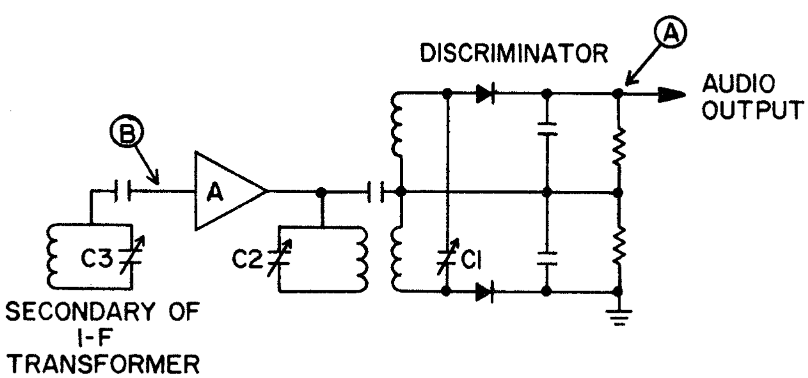
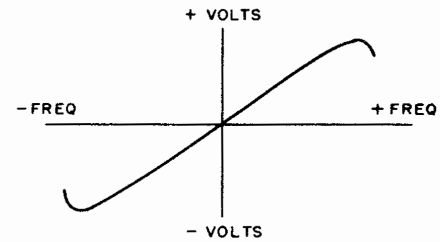


Figure 3-14. Limiter-Type Discriminator Circuit

output will be zero at the center frequency. In the second method, a marker pip is made to appear on the discriminator response curve, and is set for the cross-over point at the center frequency, so that the center frequency is actually determined by noting where the marker disappears and reappears. Because of the difficulty in observing the exact points of appearance of this pip, this method sometimes leads to inaccurate results. To use the AM signal method for aligning the discriminator to its center frequency, connect an AM signal generator to point B (Figure 3-14) and the oscilloscope vertical input to point A. Set the signal generator for 400 Hz amplitude modulation, and adjust the oscilloscope controls for a convenient pattern size. When the discriminator secondary trimmer capacitor, C1, is not adjusted to the current frequency, but is close to it, a pattern similar to that of Figure 3-15B will appear. To align the discriminator to the correct center frequency, adjust the secondary trimmer capacitor slowly in one direction, and then in the other direction, until the 400 Hz signal disappears and then reappears with a further movement of the trimmer. Set the capacitor midway between these points. Then connect an FM signal generator to point B (Figure 3-14). Leave the oscilloscope vertical input at point A, and connect the horizontal input to the modulation circuit of the signal generator. Adjust the signal generator for full frequency deviation, and set the oscilloscope controls for a convenient pattern size. Then adjust the primary trimmer capacitor, C2, for a symmetrical curve similar to the one shown in Figure 3-15A. When using the marker method, connect an FM signal generator and oscilloscope as described above. Then couple a marker generator or wavemeter with the signal generator output to point B (Figure 3-14) so as to produce a pip on the discriminator response curve. With the marker signal generator or wavemeter set at the discriminator center frequency, adjust the secondary trimmer capacitor until the marker disappears at the crossover point in the center of the response curve. Then adjust the primary trimmer capacitor for a symmetrical curve.

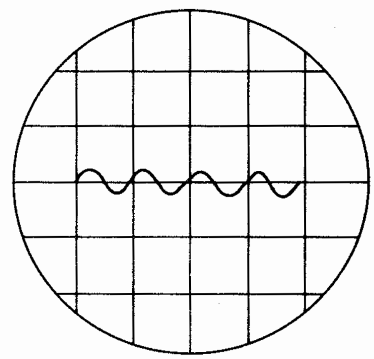
3-1.9.5.2 Ratio Detector Alignment

Another type of FM detector, shown in Figure 3-16, is called a "ratio detector". This circuit is based on changes in the ratio of the voltages across the two diodes, rather than on differences in voltage. A ratio detector is virtually insensitive to amplitude variations. The tuning and coupling provisions are practically the same as in a limiter-type discriminator. As a result, the RF voltage developed across the diodes at any instant depends upon the amount of frequency deviation from the carrier center frequency. Unlike



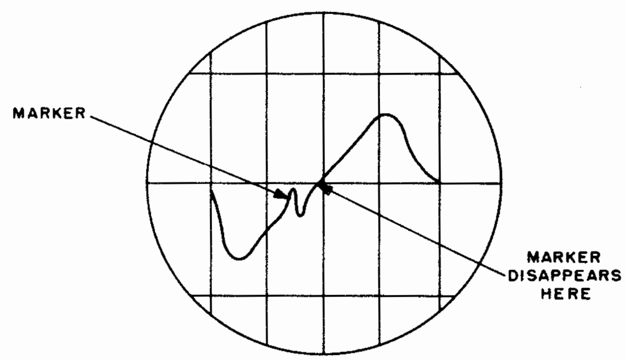
A

TYPICAL DISCRIMINATOR CHARACTERISTIC CURVE



B

DISPLAY FOR DISCRIMINATOR ALIGNMENT USING A-M SIGNAL GENERATOR



C

DISPLAY FOR DISCRIMINATOR ALIGNMENT USING MARKER AND FM SIGNAL GENERATOR

Figure 3-15. Discriminator Characteristic Measurements

the arrangement in the limiter-type discriminator, however, the diodes are connected to conduct simultaneously, so that a negative voltage is developed across the load resistor. A filter capacitor connected across the load resistor has a value sufficient to hold the

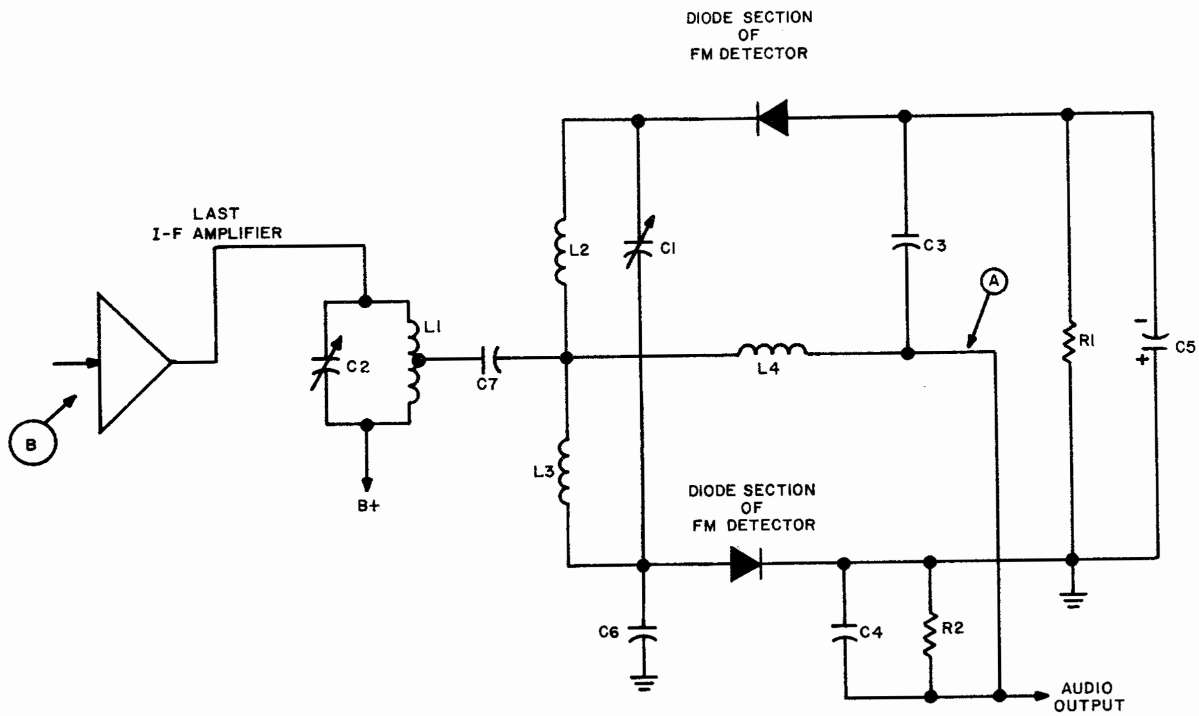


Figure 3-16. Ratio Detector Circuit

voltage constant, even at the lowest audio frequencies to be reproduced. The voltages across the diodes differ according to the instantaneous frequency of the carrier, and the rectified voltages across capacitors C3 and C4 are proportional to the corresponding diode voltages. Although each of these voltages is varying, their sum is held constant by the filter. As a result of this restriction, an audio output is developed across C4. Many modified versions of the ratio detector are in common use. However, the operation and alignment procedures are similar to those described for the ratio detector. To align a ratio detector, connect an FM signal generator between point B and ground, and connect the input to the oscilloscope between point A and ground (Figure 3-16). Set the generator to the center frequency, with maximum frequency deviation. Adjust discriminator, the primary trimmer capacitor, C2, for a curve of maximum amplitude, which will appear somewhat S-shaped if the secondary trimmer capacitor C1 is not excessively detuned. Keep the generator attenuator control set for an output below that to which limiting occurs. Next, adjust the secondary trimmer capacitor until the S-shaped curve is symmetrical (Figure 3-15A), and set to exact center frequency as described previously for discriminators. If necessary, retune the detector primary trimmer for a symmetrical response of maximum amplitude.

3-1.9.5.3 I-F Amplifier Alignment

Specific alignment procedures are generally included in technical manuals for particular receivers. In these cases, specific response curves for each transformer may be given so that a particular over-all response curve may be obtained. However, in general, the intermediate-frequency amplifiers may be aligned by feeding an FM signal from the generator to the grid of the I-F amplifier just preceding the limiter, while observing the discriminator output. The secondary of this I-F amplifier is then adjusted for a symmetrical S-shaped curve, having a proper frequency response of maximum amplitude. The procedure is repeated to tune the primary. If the output can be kept below the threshold of limiting by reducing the generator output, adjust each I-F secondary and primary in sequence, proceeding from the last I-F stage back to the first I-F stage, for a symmetrical response curve of maximum amplitude. Should limiting occur, the response curve of the amplifier can be observed at the grid of the limiter, and the I-F amplifiers tuned for proper bandpass (Figure 3-13).

3-1.9.5.4 RF and Oscillator Stages Alignment

The RF and oscillator stages may also be aligned by means of an FM signal generator and an oscilloscope. To align these stages, connect the output from the generator to the antenna terminals of the receiver through a suitable matching network and connect the oscilloscope input to the discriminator output. Set the generator to a frequency in the approximate center of the band being tested, and set the frequency deviation greater than the receiver bandpass. Observing the receiver output response, tune the shunt (high-frequency trimmer capacitor) for maximum output. Set the signal generator and receiver to the low end of the band. Use a tuning wand and observe the oscilloscope pattern. If the signal amplitude decreases when either end of the wand is inserted into the oscillator coil, the tracking is satisfactory. If the output increases with the brass end of the wand inserted, spread the turns of the oscillator coil; if the output increases with the iron end of the wand inserted, compress the turns of the coil. Do not bend the coil excessively, as only a slight physical change is necessary at the high frequencies at which this type of equipment generally operates. Repeat these steps until no further change is noted. The last adjustment should be that of the shunt (high-frequency) trimmer capacitor. Return the signal generator and receiver to the center frequency, and adjust the shunt trimmer capacitor of the mixer-grid circuit for maximum output. If an RF stage is employed, also adjust the shunt trimmer capacitor of the RF stage for maximum output. Lower the frequency setting of the receiver and generator, and check the tracking of the mixer and RF grid circuits with the tuning wand. If the output increases with the brass end of the wand inserted in the coil, spread the coil turns; if the output increases with the iron end inserted, compress the coil turns. If the output decreases when either end is inserted, the tracking is correct. Do this for both the mixer and RF coils. Repeat these adjustments until no further improvement is noted. An oscilloscope may also be used to check the response of the RF and mixer sections of a receiver. Connect the FM signal generator to the receiver input through a suitable matching network. Then connect the oscilloscope vertical input to the mixer plate decoupling network or, by means of a high-frequency detector probe, to the mixer's output. The first I-F amplifier is disabled during this test to reduce the loading effect of the oscilloscope input. Couple a marker signal generator with the FM signal generator to the receiver input in order to determine the frequency points of the response curve. Set the FM signal generator for the desired frequency deviation; in many communications receivers the front-end response cruve is from 150 to 200 kHz wide. The bandwidth of the front end is largely fixed by the number and Q of the RF circuits, since all of the circuits are usually tuned to identical frequencies. Usually, the bandwidth of the RF circuits is considerably greater than the I-F bandwidth, so that the latter mainly determines the bandwidth of the entire receiver. When making any adjustment of the front end, both the RF and over-all response must be considered, since it is important that the RF response be wide enough to pass all of the important frequency components of the signal. This check may be made by the previous procedure for measuring the RF response.

3-2 COMMUNICATIONS TRANSMITTERS AND TRANSCEIVER TESTING

When testing communications transmitters and/or transceivers, the configuration in which the equipment is installed must be taken into consideration. In the newer installations, system monitoring is used (more so than equipment monitoring). These newer configurations employ built-in test equipment (BITE) to perform system monitoring and a varying degree of fault isolation. In the older installations, front panel meters and dials are relied upon for equipment monitoring and some fault isolation. A number of equipments can be placed in either installation while other equipments are unique to either the system concept or the individual equipment concept. In both instances, however, the same parameters are monitored; only the monitoring method is changed. Another consideration in transmitter/transceiver maintenance involves temperature. High-powered transmitters emit a great deal of heat in their power amplifier and driver stages. If the heat thus generated is inadequately dissipated, premature failures of equipment will occur. Two factors which contribute significantly to inadequate heat dissipation are: 1) failing or failed circulatory systems (either air or water), and 2) dirty filters or heat exchanges. Although the first factor is usually unpredictable, the latter is always avoidable with routine maintenance. In addition, routine maintenance can usually prevent most circulatory system failures. If forced air (blower) circulation is employed to dissipate the heat in a high-powered transmitter, dust settling in the equipment can contribute problems. Because dust forms a film which absorbs moisture, insulation resistance is lowered to a point at which flash-over may occur. Strict adherence to scheduled equipment cleaning is therefore mandatory if the equipment is to be cooled effectively. Insulators must be wiped down, corroded metal parts cleaned, and arc-overs repaired. It is possible to detect poor contacts by inspecting for evidences of local overheating (or arcing). Such contacts must be thoroughly cleaned and tightened.

3-2.1 FREQUENCY GENERATION

On any given day, several thousand different frequencies may be used simultaneously by several transmitting units. Since atmospheric conditions dictate which frequency will best be propagated at any given time, the transmission frequency in use may need to be changed quite often in order that the individual units may maintain efficient and reliable communication. Regulations require each transmitted frequency to be precise (to very close tolerances), and that it comply with harmonic frequency emission limits and sideband emission limits. To meet these rigid standards, crystal oscillators and frequency synthesis are employed to generate the required transmission frequency. When frequency synthesis is used, a calibrated frequency standard with a very high degree of accuracy (on the order of 1 part in 10^9 per day) serves as the basic oscillator. This standard is external to the transmitter or transceiver, and its output is fed to each transmitter/transceiver via an amplifier distribution system, as illustrated in Figure 3-17. A back-up internal standard is incorporated in each transmitter/transceiver for use in event of primary standard malfunction. The output of the primary standards is multiplied and/or divided in the transmitter to obtain the desired frequencies for use in the frequency synthesis process. Figure 3-18 shows an example of an HF transmitter's synthesizer. In the example shown in Figure 3-18, all of the output operating frequencies are derived from the 5MHz primary standard input frequency. This ensures each operating frequency is as accurate as the primary standard.

3-2.2 FREQUENCY MEASUREMENT

The primary standard used in the frequency standard distribution system requires routine calibration by a Class A calibration facility. This standard must be more accurate than any other general-purpose test equipment maintained aboard ship. Primary power must be applied to the standard at all times to ensure stable operation. As a provision against a loss of ac power, a battery source of alternative power is incorporated in the standard. The battery becomes switched in automatically when ac power is lost, but its staying time is limited to approximately eight hours on a full charge. The primary power system is alarmed and must be checked frequently to ensure battery-power readiness. This is accomplished by manually switching to the battery power source. The alarm should then sound and the standard should operate automatically from the battery source and without interruption on the batteries. Before switching manually to battery operation, make sure the standard's front panel meter indicates a battery charge exists. If all power to the primary standard becomes lost, the transmitters and transceivers in the frequency distribution system must then be shifted over to the internal frequency standard. The internal frequency

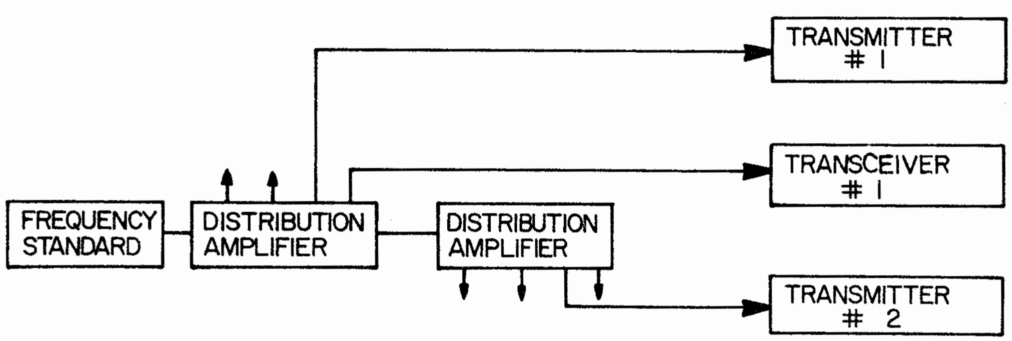


Figure 3-17. Frequency Standard Distribution System

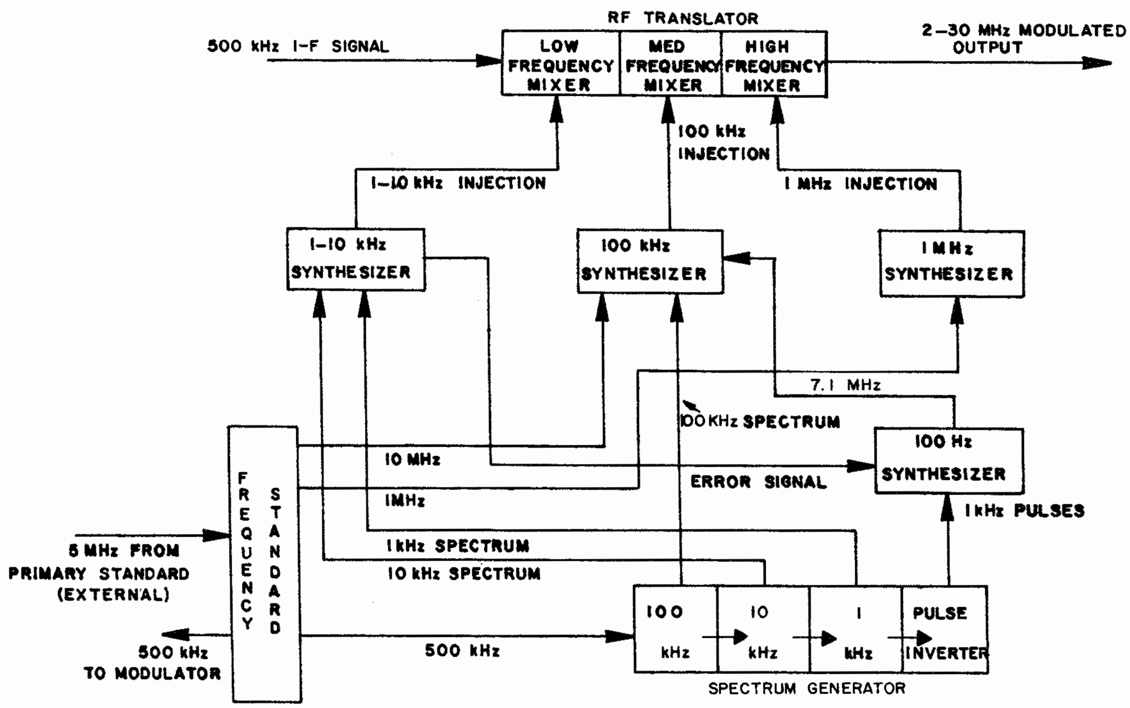


Figure 3-18. Frequency Synthesizer

standard should normally be allowed twenty-four hours to become stable; in emergency operations the internal standard requires at least four hours of warm-up time. Although the frequency accuracy of the primary standard cannot be checked realistically, a check with a calibrated frequency counter will indicate whether or not the standard is grossly off frequency. Harmonic distortion of the primary standard's frequency must be guarded against because of the unwanted changes this frequency will undergo during the frequency synthesis process in the transmitters and transceivers. Spectrum analysis is used to measure harmonic distortion because low-level distortion will not appear in the time-domain display of an oscilloscope. Such distortion will show up, however, in the frequency domain display of the spectrum analyzer, as illustrated in Figure 3-19. High-level distortion will show up in either display. If more than 0.2% of distortion is encountered in the standard's output, the standard must be sent to a shore facility for repair and/or calibration. The two best means of measuring a transmitter's output frequency are: 1) the frequency counter; and 2) the spectrum analyzer. The frequency counter is by far the more accurate of the two because interpolation is not required. The spectrum analyzer, on the other hand, can measure second- and third-order harmonic emissions, sideband emissions, and intermodulation distortion, as well as output frequency. In addition, when the spectrum analyzer is used in conjuction with a directional coupler and calibrated attentuators, it can also measure power output and reflected power. The spectrum analyzer can thus provide a better indication of transmission quality than can a frequency counter.

3-2.3 AMPLITUDE MODULATION MEASUREMENTS

If an AM signal is applied to the vertical input of an oscilloscope, the oscilloscope will display a wave-envelope pattern of the AM signal. Figure 3-20 illustrates such an oscilloscope display. When the modulation voltage is used as an external sweep

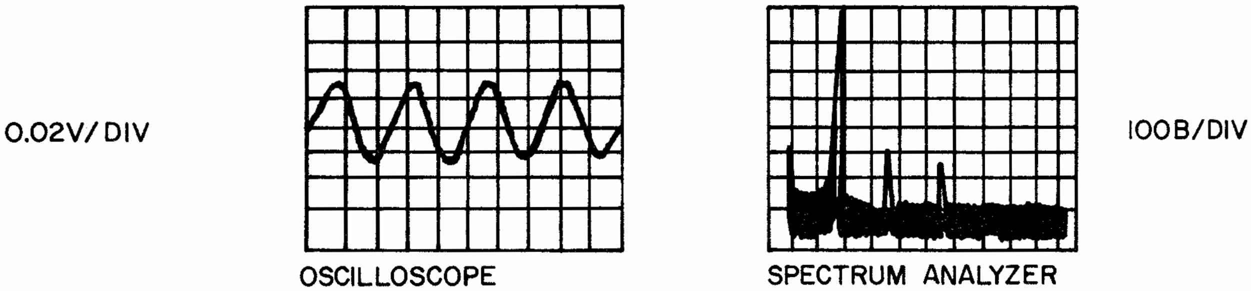


Figure 3-19. Harmonic Distortion Displays

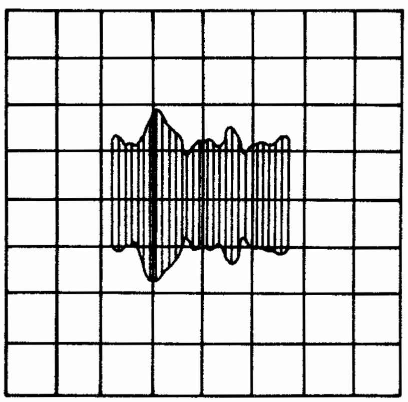


Figure 3-20. Amplitude-Modulated Carrier

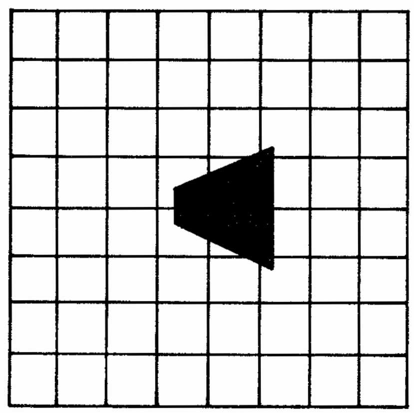


Figure 3-21. Trapezoidal AM Carrier Pattern

voltage, a trapezoidal pattern, as illustrated in Figure 3-21, becomes displayed. In this display, the amplitude-modulated carrier is plotted as a function of modulation, rather than of time. The resultant pattern in Figure 3-21 will remain stationary and its shape is determined by the percent of modulation. The presence of an abnormal pattern does not necessarily mean that there is a defect in the operation of the transmitter. A pattern with elliptical sides merely implies that, with respect to the signals applied to the oscilloscope, there is a phase difference between the modulating signal and the modulated radio-frequency signal. The curved appearance of the sides is simply a Lissajous effect denoting phase difference. In fact, the straight sides of a properly formed trapezoid simply represent the characteristic Lissajous pattern of a zero phase difference. When the modulating signal is not a sine-wave, a trapezoid is still formed, but a series of bright vertical bands appear in the pattern. These bands simply indicate points at which the lateral motion of the electron beam is relatively slow. Distorted trapezoid waveforms due to certain circuit malfunctions are shown in Figure 3-22. Part (A) of Figure 3-22 shows an amplitude-modulated waveform and its corresponding trapezoidal display, resulting from excessive plate voltage applied to the radio-frequency power amplifier in a radio transmitter. Part (B) shows the reverse situation, where there was insufficient voltage applied to the plate of the transmitter power amplifier. Parts (C) and (D) of Figure 3-22 show the effects obtained from imperfect neutralization in a radio transmitter power amplifier. There is nonuniform density in the amplitude-modulated waveform; light and dark bars, such as

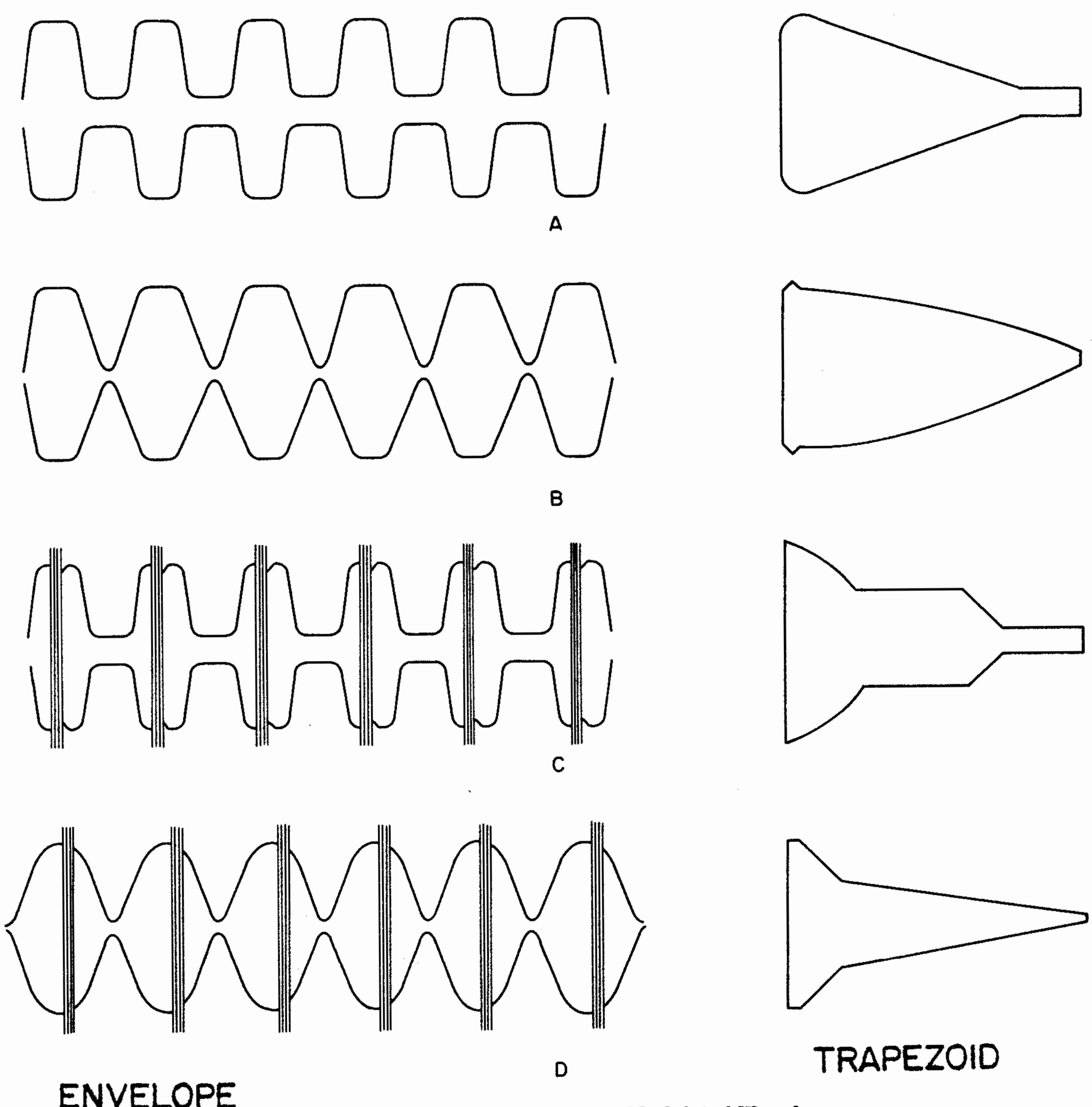


Figure 3-22. Distorted Amplitude-Modulated Waveforms

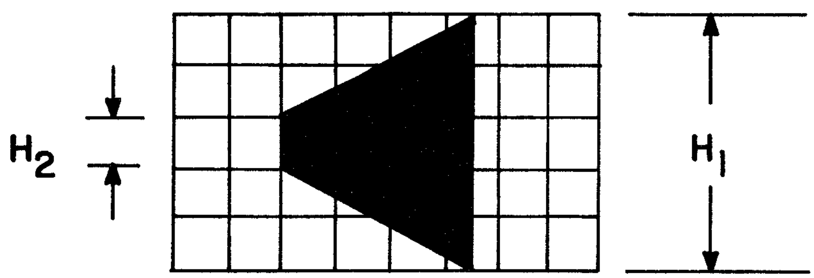
those shown in the figure, usually indicate spurious oscillation. To calculate the percent of modulation, the trapezoid pattern provides the most convenient form to use. The horizontal and vertical gain controls are adjusted for a suitable display on the screen, such as shown in Figure 3-23. The modulation percent is then calculated, using the formula:

$$\frac{H1 - H2}{H1 + H2} \times 100$$

where H1 is the greatest vertical height (amplitude) and H2 is the lesser vertical height. Thus, using Figure 3-23 as an example, the precent of modulation would be:

$$\frac{5 - 1}{5 + 1} \times 100 = \frac{4}{6} \times 100 = 66.6\%$$

The longer side of the trapezoidal pattern represents modulation peaks, or crests; the shorter side indicates modulation troughs, or low points. At 100-percent modulation, the wedge-shaped pattern assumes a point on the shorter side; modulation over 100 percent causes this point to extend and form a horizontal



$$\frac{H1 - H2}{H1 + H2} \times 100$$

Figure 3-23. Trapezoid Method of Determining % of Modulation

line, or tail, as illustrated in Figure 3-24. Because the trapezoid type of display retains its triangular characteristics even with varying degrees of modulation, it provides a more easily discernible indication of overmodulation as well as the modulation percentage. To obtain correct results, care should be taken to avoid stray radio-frequency pickup which may distort the oscilloscope presentation.

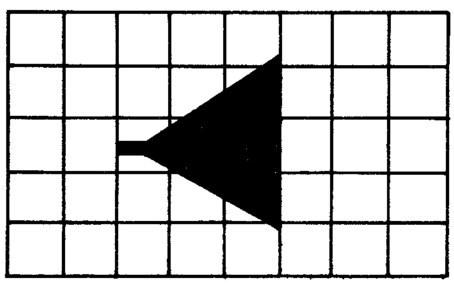


Figure 3-24. Overmodulated Carrier

3-2.4 FREQUENCY MODULATION MEASUREMENTS

The concept of percentage of modulation as discussed in connection with amplitude modulation does not apply to frequency modulation. The amplitude of the FM wave is constant, and the extent of modulation must be described in other terms than those of the amplitude-modulated wave. When referring to a class of stations, a certain maximum frequency swing is established as representing 100 percent modulation. For example, in the case of FM broadcast stations, a frequency swing of ± 75 kHz from the unmodulated center frequency (frequency deviation) is commonly considered as being the equivalent of 100 percent modulation. However, the more widely accepted method of describing the extent of modulation is to state the value of the modulation index. This index (m) is the ratio of the amount by which the transmitted frequency swings from its average frequency (frequency deviation) to the frequency of the modulating signal. The relationship of these quantities is shown by the following equation:

$$m = \frac{Fd}{Fm}$$

where:

m = modulation index
Fd = frequency deviation
Fm = frequency of modulating signal

By means of this basic relationship, it is possible to determine the frequency deviation when the modulation index and the modulating frequency are known. It should be carefully noted, in describing the extent of frequency modulation, that the modulation percentage and the modulation index are defined in a different manner. The percentage is proportional to the frequency swing. The modulation index is also directly proportional to the frequency swing, but in addition, is inversely proportional to the highest modulating frequency. Thus, in contrast to amplitude modulation, the modulation index of a frequency-modulated wave is not the decimal equivalent of the modulation percentage. The modulation index of a frequency-modulated wave, for example, will exceed 1 (unity) by many times when the frequency swing is large and the modulating frequency is low. The frequency-modulated output is the sum of a center frequency component and numerous pairs of sideband frequency components. The center frequency component has the same frequency as the unmodulated carrier. The two components of the first sideband pair have frequencies respectively higher and lower than the center frequency by the amount of the modulating frequency, just as in amplitude modulation. In frequency modulation, however, there are additional pairs of sideband components which can have appreciable amplitude. For example, the second pair of sidebands, having frequencies that are higher and lower than the center frequency by twice the amount of the modulating frequency, can also be important. The same can be true of the third pair of sidebands, which are removed from the center frequency by three times the modulating frequency, and of even higher orders of sideband pairs, whose frequencies differ from the center frequency by correspondingly greater amounts. When the modulation is only slight, only the pair of sidebands nearest in frequency to the carrier frequency component will have sufficient amplitude to be important. Under this condition, the bandwidth required is no greater than that for an amplitude-modulated wave. As the frequency modulation is increased, however, more pairs of sidebands acquire appreciable amplitude, and the bandwidth requirements are greater than the amplitude modulation. The actual amplitudes of the frequency-modulated-wave sidebands and carrier, as compared with an unmodulated carrier amplitude of 1, may be read directly from Table 3-1 for modulation indices up to 6. To find the amplitude of any sideband pair, determine the modulation index (m), read the corresponding amplitude factor for the sideband pair, and multiply this factor by the amplitude of the unmodulated carrier. The amplitude of the carrier during modulation is found in the same manner, taking the amplitude factor from the $J_0(m)$ column. Where no value is given in a column, the amplitude factor is less than 0.005, and the sideband pair will not be important for normal considerations. The values of $J_0(m)$, $J_1(m)$, and $J_2(m)$ over the range m = 0 to m = 16 are shown plotted in Figure 3-25. A study of these curves reveals some interesting facts about the composition of frequency-modulated waves. $J_0(m)$ is less than 1 for all values of m greater than zero. This indicates that as sideband components appear with modulation, the amplitude of the center frequency component is less than its amplitude in the absence of modulation. This fact is evident when it is remembered that the amplitude of the frequency-modulated wave is constant, so that the average power during each radio frequency cycle is the same as that during any other radio frequency cycle. In order that the power in the wave will not change when frequency modulation causes sideband currents to appear, the amplitude of the center frequency component must decrease sufficiently to keep the total of the I^2R products of all the components equal to the power of the unmodulated wave.

Table 3-1. Bessel Factors for Finding Amplitudes of Center and Sideband Frequency Components

m	$J_0(m)$ F	$J_1(m)$ F±Fm	$J_2(m)$ F±2Fm	$J_3(m)$ F±3Fm	$J_4(m)$ F±4Fm	$J_5(m)$ F±5Fm	$J_6(m)$ F±6FFm	$J_7(m)$ F±7Fm	$J_8(m)$ F±8Fm	$J_9(m)$ F±9Fm
0.0	1.000									
0.1	0.9975	0.0499								
0.2	0.99	0.0995								
0.3	0.9776	0.1483	0.0112							
0.4	0.9604	0.196	0.0197							
0.5	0.9385	0.2423	0.0306							
0.6	0.912	0.2867	0.0437							
0.7	0.8812	0.329	0.0589	0.0069						
0.8	0.8463	0.3688	0.0758	0.0102						
0.9	0.8075	0.4059	0.0946	0.0144						
1.0	0.7652	0.4401	0.1149	0.0196						
1.2	0.6711	0.4983	0.1593	0.0329	0.005					
1.4	0.5669	0.5419	0.2073	0.0505	0.0001					
1.6	0.4554	0.5699	0.257	0.0725	0.0150					
1.8	0.3400	0.5815	0.3061	0.0988	0.0232					
2.0	0.2239	0.5767	0.3528	0.1289	0.034	0.007				
3.0	–0.2601	0.3391	0.4861	0.3091	0.1320	0.0430	0.0114			
4.0	–0.3971	–0.066	0.3641	0.4302	0.2811	0.1321	0.0491	0.0152		
5.0	–0.1776	–0.3276	0.0466	0.3648	0.3912	0.2611	0.131	0.0534	0.0184	
6.0	0.1506	–0.2767	–0.2429	0.1148	0.3576	0.3621	0.2458	0.1296	0.0565	0.0212

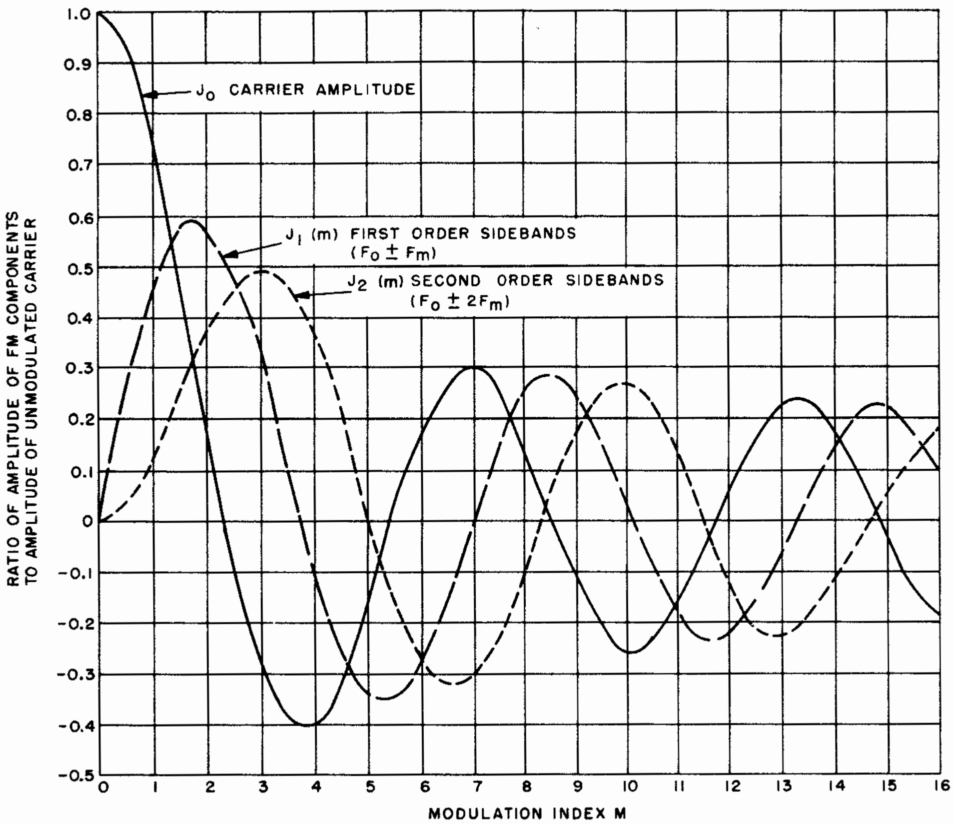


Figure 3-25. Variation of FM Wave Component with Degree of Modulation

3-2.5 FREQUENCY DEVIATION MEASUREMENTS

Regardless of the differences between amplitude modulation and frequency modulation, it is possible to make an analogy between percentage of amplitude modulation and frequency deviation. Specifically, frequency deviation is proportional to the amplitude of the modulating signal, as is the percentage of amplitude modulation. Because of this analogy, it is convenient to extend the concept of percentage of modulation to frequency modulation by arbitrarily designating the maximum allowable frequency deviation of a class of operation as 100 percent modulation. An important distinction to remember is that no distortion results from modulation percentages greater than 100 in FM transmission. However, any percentage larger than the figure sanctioned by the proper authorities will produce excessive channel width, making interference with other stations possible. For example, the maximum frequency deviation for commerical FM stations is limited to 75 kHz; for military applications the maximum deviation is limited to 40 kHz (and is classed as narrow-band FM transmission), and the sound transmission of television stations is restricted to a deviation of 25 kHz. It was stated earlier that modulation index (m) determines the relative amplitude of the carrier and sideband frequencies emitted by an FM transmitter. The modulation index may be measured by utilizing the fact that the carrier amplitude becomes zero whenever the modulation index is such that $J_0(m) = 0$, where J_0 is a Bessel function of the zero order. The values of the modulation index for these conditions are given in Table 3-2. Specifically, the carrier component disappears completely for certain values of m; i.e., m = 2.405, 5.52, 8.654, etc. (Note that $J_0(m) = 0$ in Figure 3-25 for these values of m.) For these specific values of m, all of the transmitter power is contained in the sidebands. This fact allows the measurement of specific values of the modulation index by measuring the amplitude of the carrier component only. The level of modulation on the FM carrier is increased from zero to the first point at which the detected carrier disappears. The point at which the carrier first

Table 3-2. Values of Modulation Index for Which a Carrier Wave Has Zero Amplitude

ORDER OF CARRIER ZERO	MODULATION INDEX
1	2.405
2	5.52
3	8.654
4	11.79
5	14.93
6	18.07
m(m 6)	18.07 + π(m-6)

disappears corresponds to m = 2.405. Upon increasing the modulation further, a carrier reappears and then disappears a second time. The second vanishing of the carrier corresponds to m = 5.52. Further increases in modulation will produce the higher carrier zeros (or null points). The frequency deviation at the first null point, for example, is: $F_d = 2.405\ F_m$. This means of determining frequency deviation is generally known as the "Bessel zero method". Modulation indices between carrier zero-points would involve considerable interpolation, leaving room for large error if the Bessel zero method is used. In addition, the modulating frequency must be started at zero amplitude in order to determine which zero point is being displayed. A more accurate method involves comparing the carrier amplitude with respect to the sideband amplitudes. Figure 3-26 illustrates a spectrum analyzer display of an FM signal where deviation has caused the carrier to be at thirty percent of its unmodulated level. In parts B and C of Figure 3-26 the carrier is in the negative region of its curve, as shown in Figure 3-25. The spectrum analyzer will still display the carrier level as 30%, but by measuring sideband levels (two or three should prove sufficient), the correct index can be determined, as illustrated in Figure 3-26. From this procedure, the precise frequency deviation can be readily determined.

3-2.6 SINGLE-SIDEBAND (SSB) MEASUREMENT

Unlike AM or FM, where the carrier is transmitted along with the intelligence, single-sideband transmitters and transceivers transmit only intelligence. Instead of measuring the degree of modulation, as in AM and FM, the degree of carrier suppression and the amount of signal distortion are the characteristic measurements in an SSB transmission. The characteristic single-sideband signal is extracted with balanced modulations, sharp filters, and linear amplifiers.

3-2.6.1 Balanced Modulators

The primary purpose of any balanced-modulator circuit is to produce the sidebands of an amplitude-modulated RF carrier and to suppress or reject the carrier. The amount of carrier suppression depends upon the degree of balance between the two legs of the balanced circuit. In such circuits using vacuum-tubes, two tubes the same type will generally balance close enough to suppress the carrier approximately 10 to 15 dB without any external adjustments. Since a carrier suppression of at least 35 to 40 dB is desirable in SSB, separate bias supplies and R-C balance adjustments are provided in the modulator to ensure correct balance of these circuits and thus provide adequate suppression of the carrier. In diode-rectifier balanced modulators (balanced-bridge, ring, or lattice-type modulators), an important factor that must be considered is the proper ratio of the RF carrier voltage to the modulating-signal voltage. If the distortion products generated in such nonlinear devices are to be kept to a practical minimum, this ratio should be kept high. For circuits employing germanium crystals or copper-oxide rectifiers as the diode-modulator elements, the level of the RF voltage should be on the order of 3 to 6 volts, which is at least eight to ten times the level of the peak modulating-signal voltage. In fact, approximately the same voltage ratio should be observed in all balanced-modulator circuits, regardless of the type of device used as the modulator element. If a pilot carrier is transmitted with the single-sideband signal, the level of the reduced carrier is normally 10 to 20 dB below the level of the sideband signal. In such systems it is desirable that all of the reduced carrier signal be applied through the pilot-carrier re-insertion circuits, and that the amount of carrier due to imbalance of the balanced modulator be kept to a minimum. Attainment of this condition ensures more accurate control of the pilot-carrier level by the carrier-level control. The first step in adjusting a balanced

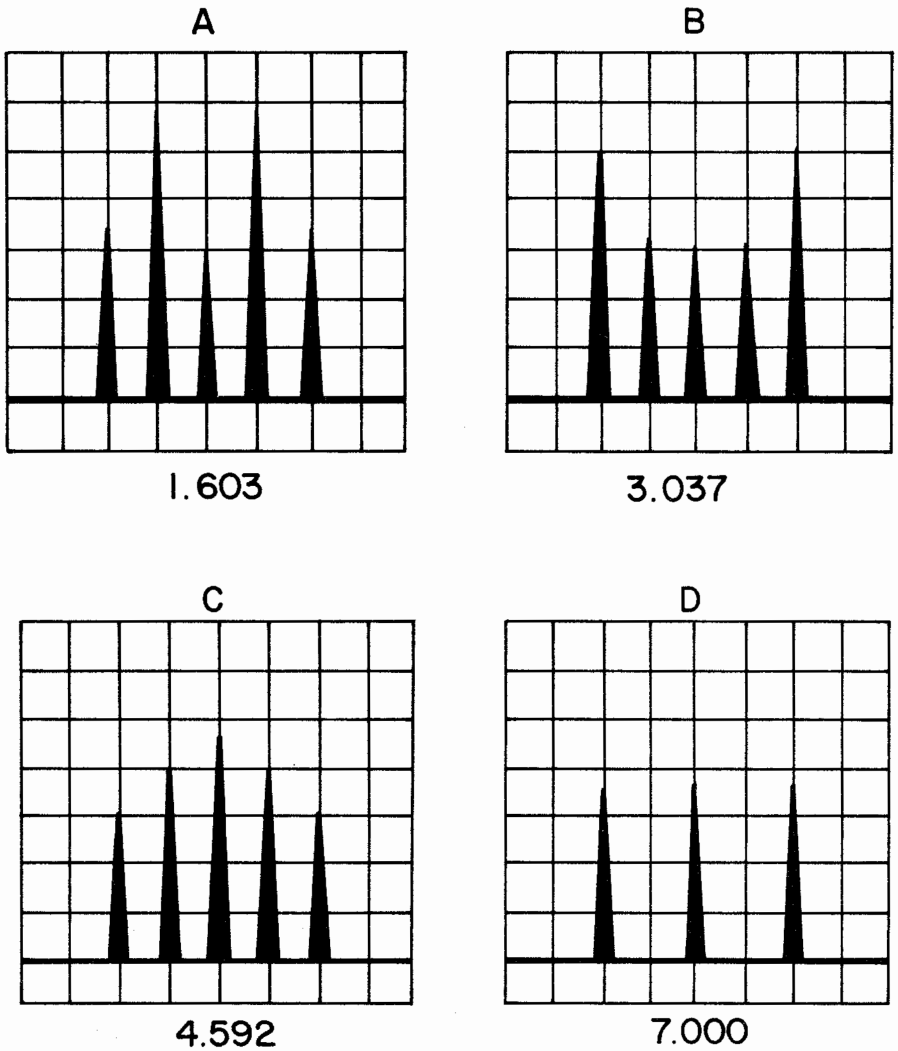


Figure 3-26. FM Spectral Display at Indices of 1.603, 3.037, 4.592, and 7.000, respectively (Carrier @ 30%)

modulator is to check the circuit operating voltages, and to make the necessary corrections (parts replacement, bias supply adjustments, etc) for any errors thus found. When a change of parts is required, caution must be observed to ensure the replacement parts comply with the value and tolerance specified for the circuit. If it is found necessary to change one tube in a circuit using separate balanced-modulator tubes, care should be exercised in determining that the replacement tube matches the tube remaining in the circuit. This can be done by checking the mutual conductance of both tubes (the remaining tube and the replacement tube) on a reliable tube checker. After it is ascertained that the tubes are properly matched (by double-checking their operation in the circuit) and that all operating voltages are in compliance with the equipment specifications, the modulator balance controls can be properly adjusted. The carrier balance can be adjusted by applying a test tone of constant amplitude and frequency (usually around 1000 hertz) to the input of the SSB's audio input. A spectrum analyzer is connected to the output of the balanced modulator and the balance controls are adjusted until a signal similar to that shown in Figure 3-27 is obtained. In Figure 3-27 the carrier is greater than 40 dB below the sidebands (scale is 10 dB per division). Distortion products are down greater than 50 dB for this test.

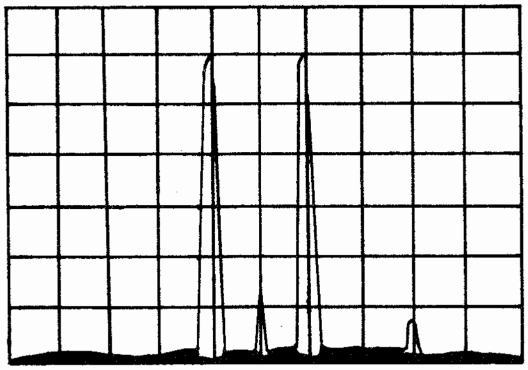


Figure 3-27. Typical Single-Tone Output of a Balanced Modulator

3-2.6.2 Sideband Filters

The purpose of the sideband filters in an SSB transmitter is to pass only the desired band of frequencies with minimum distortion and loss. The filter characteristics are primarily determined by the filter design engineer and are usually permanent once the initial design is completed. Not only is it desirable to pass a certain band of frequencies with minimum loss and distortion, but it is equally important that all frequencies outside this desired band be attenuated sufficiently so that they do not appear in the output of the transmitter. This fact is most important in dual-channel SSB systems where information from two unrelated sources is transmitted on opposite sides of a suppressed, or reduced, carrier. Sideband filters are not field-adjustable. The necessary adjustments are made and sealed-in during the construction of the filter unit. This is especially true of mechanical filters; some types of these filters are hermetically sealed and no external adjustments are possible. Attempts to adjust mechanical filters on board ship are not to be undertaken under any conditions. The trimmer adjustments associated with certain types of crystal-lattice filters must not be altered by maintenance personnel in the field. If all other attempts at alignment of a transmitter fail, and it is assumed that the filter is out of adjustment, replace it. The importance of the sideband filters with regard to the overall frequency response and proper performance of a single-sideband transmitter cannot be stressed too strongly. Therefore, strict adherence to the manufacturer's instructions concerning the sideband filters used in this equipment is essential.

3-2.6.3 Sideband Suppression Testing

The procedure used to test for carrier suppression can also be employed for checking sideband suppression. Instead of connecting the spectrum analyzer to the output of the balance modulator, it is connected to the output of the sideband filter. Either the upper or the lower sideband signal will be attenuated, depending on which sideband's filter is being checked. As in carrier suppression, the desired sideband suppression for proper operation is 35 to 40 dB. Figure 3-28 presents a graphic illustration of the output of the sideband filters as they would appear on the spectrum analyzer. (Scale is 10 dB per division.)

3-2.6.4 Distortion in SSB System

An important requirement of SSB systems is low intermodulation distortion. In such systems the linear power amplifiers are the main source of this form of distortion. Usually the distortion caused by the even-order products (2nd, 4th, etc.) are sufficiently removed from the desired signal that normal tuning will eliminate them. Most of the intermodulation distortion in the linear amplifiers is caused by the odd-order products (3rd, 5th, etc.) that fall in or near the desired frequencies. Of equal importance are the distortion products that fall outside the assigned SSB channel, because these products may cause interference in the reception of weak signals in equipment operating on the adjacent channel. A form of distortion peculiar to phase-shift SSB transmitters is called "post-phasing distortion". This form of distortion is caused by harmonics generated in the audio amplifiers following the audio phase-difference networks, and by improper adjustment of the balanced modulators. When the signals (including the harmonics) from the two audio amplifiers are applied to the balanced modulators, the distortion products will not be in the correct phase to be cancelled in the output circuit of the modulators. Specifically, the third-order products will cause a single-sideband signal to be present on the

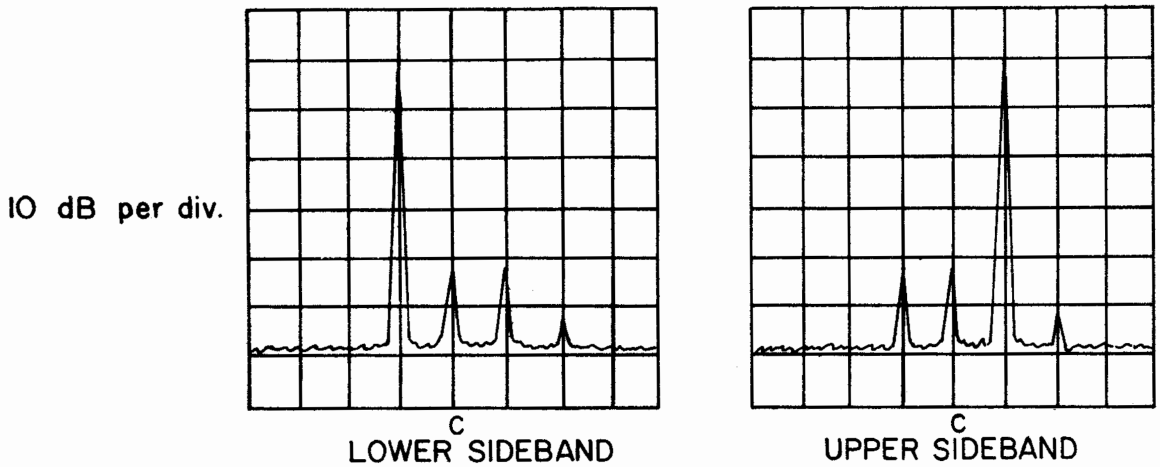


Figure 3-28. Sideband Filter Outputs

undesired side of the suppressed carrier; the fifth-order products will cause distortion to be present in the desired sideband; and all even-order products will be transmitted as double-sideband signals with no carrier. If the carrier is not completely balanced and appears in the output, phase modulation, as well as amplitude modulation, will result.

3-2.6.5 Two-Tone Testing Procedure

The two-tone test is the most widely used method of testing SSB transmitters. This test involves the application of two separate tone-signals to the input of a system or circuit, and observing the results on an oscilloscope, spectrum analyzer, or some other indicating device. The two tones should be equal in amplitude and have a difference in frequency of about 1000 Hz, in order to achieve the results desired from the test. Typical examples of two-tone test waveforms are illustrated in Figure 3-29. The sources of the test tones can be either RF signal generators or audio oscillators, whichever are applicable to the test being performed. In filter-system transmitters, the two-tone test signals can sometimes be obtained by applying a 1000 Hz audio tone to the transmitter input and slightly unbalancing the I-F balanced-modulator to allow a portion of the carrier to "feedthrough". A similar method of obtaining the two tones can be used in phase-shift system SSB transmitters. In these applications it is only necessary to disable one of the

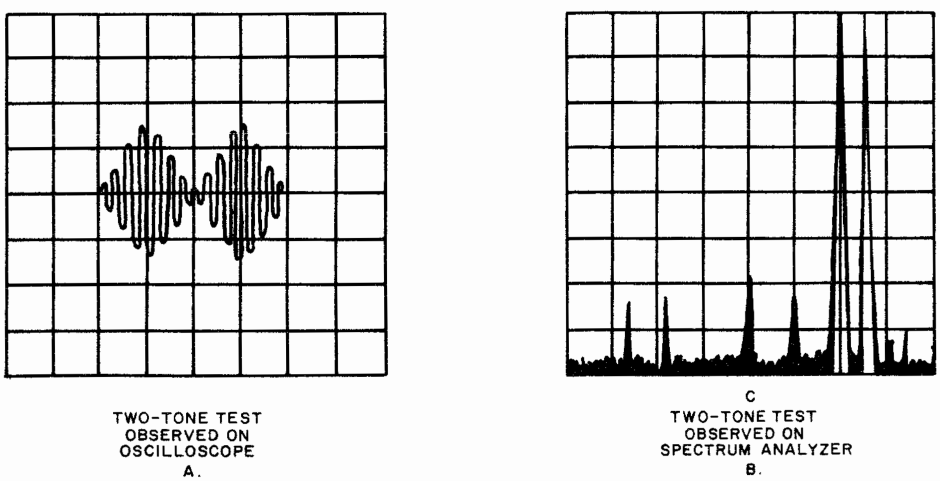


Figure 3-29. Examples of Ideal Two-Tone Test Waveforms

balanced modulators, instead of unbalancing the circuit. When these methods are used, care must be exercised to ensure that the amplitudes of the two tones are maintained constant. Upon completion of the two-tone test (using these methods), the carrier balance should be reset and checked to ensure that it is at the proper point for correct functioning of the equipment.

3-2.6.6 Signal-To-Distortion Ratio Measurement

The signal-to-distortion ratio, stated in dB, is the ratio of the amplitude of one test tone to the amplitude of the third-order product, and is usually determined by using the two-tone test. Although present designs of linear amplifiers can produce signal-to-distortion ratios of approximately 35 to 40 dB, lower overall system distortion values can be obtained by using some form of distortion cancellation, such as RF feedback. Since the principal causes of distortion in linear amplifiers are non-linearity and grid-current loading, care must be exercised to ensure that the linear amplifiers are not overdriven in SSB systems. The signal-to-distortion measurement can be made by applying two tones (at the same level) to the input of the transmitter and then measuring the transmitter's output. The output of the higher power transmitters, however, must be attenuated to achieve proper measurement. Since the method of connecting the test equipment will vary for different system applications, the equipment technical manual should be consulted for each individual system. However, the basic setup will be as shown in Figure 3-30. Figure 3-29B is an example of the results derived by the test setup shown in Figure 3-30. The two-tone test has proven to be the best overall test of SSB transmitter operation. When two audio tones of the same level are introduced at the input of the transmitter and the output of the transmitter is then monitored on the spectrum analyzer, carrier suppression, sideband suppression and intermodulation-distortion can all be monitored simultaneously, as illustrated in Figure 3-29B. (Scale is 10 dB per division.)

3-2.7 I-F AND RF AMPLIFIERS

There are three basic methods used to obtain a transmitter's operating frequency. One method mixes the output of various crystal oscillators of different frequencies; another method involves multiplying the basic crystal frequency by certain factors; and a third method employs frequency synthesis, whereby the basic oscillator's frequency is used to generate harmonics which are then amplified and mixed. On occasion, more than one method is

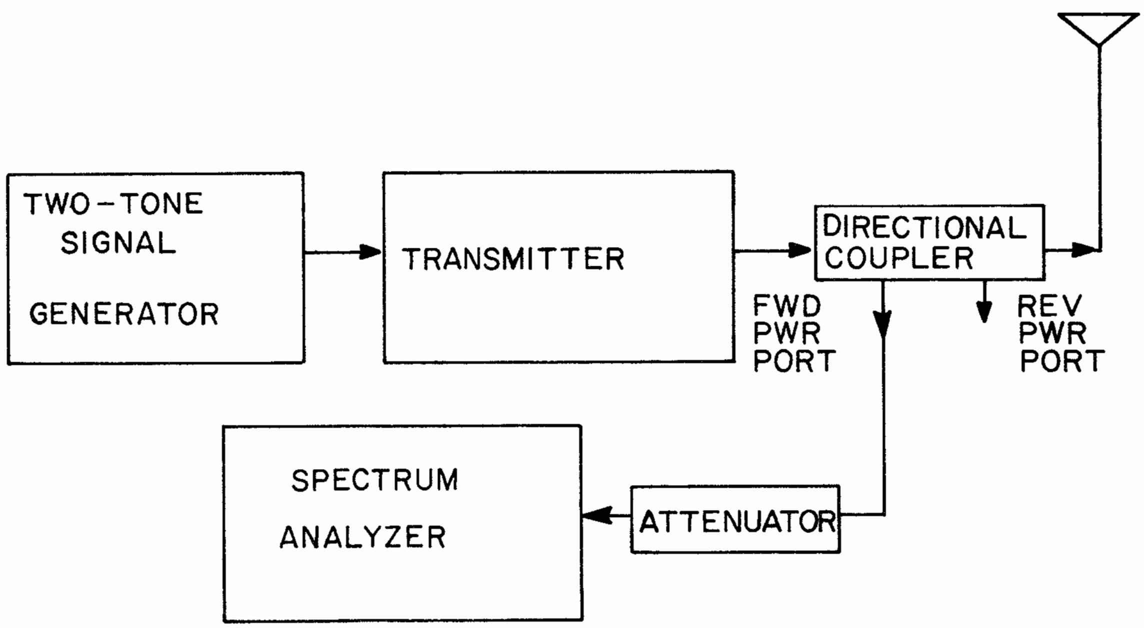


Figure 3-30. Equipment Setup for Measurement of Signal-To-Distortion Ratio, Using the Two-Tone Test

used; however, in each method, intermediate (I-F) and output (RF) frequencies are developed which require amplification to attain the rated output power of the transmitter. Both I-F and RF-type amplifiers can distort a modulated signal if the amplifiers are not operated in the linear portion of their characteristic curve. However, linear operation is not always the most efficient or desirable method of operating an amplifier. The output power requirements and the type of modulation used will be the determining factors in the design of I-F and RF amplifiers.

3-2.7.1 I-F Gain Measurement

When the gain factor of an I-F signal is the main consideration and the I-F stage is not amplifying an AM or SSB signal, nonlinear amplification is used for maximum efficiency. Since the I-F is not amplitude-modulated, distortion products can be eliminated by installing fixed or tunable filters in the output stage of each amplifier. There are two prime considerations to keep in mind when testing this type of I-F amplifier: 1) the gain of the amplifier; and 2) the selectivity (response) of the filter. In general, the undesired, out-of-band signals should be reduced by more than 40 dB. Because I-F gain will vary by equipments, the individual equipment's technical manual must be consulted in determining what the I-F gain factor should be. Figure 3-31 illustrates a basic setup for measuring I-F gain.

3-2.7.2 FM and SSB Requirements

In FM transmissions, the basic frequency is modulated, and it is this frequency which is mixed, translated or multiplied to obtain the transmitter's output frequency. Since the distortion generated in nonlinear amplification primarily affects the amplitude of the carrier more than the frequency, FM is less susceptible than AM to the distortion created in nonlinear amplification. Therefore, the I-F and RF amplifiers of an FM transmitter can be operated for maximum efficiency. The primary consideration in FM is that the I-F filter's response be broad enough to pass the required frequency deviation of the FM signal with sufficient amplitude. The test setup for checking and FM transmitter's I-F stages is similar to that for testing an FM transmitter's I-F stages except that a sweep generator is used as a signal source, instead of a CW signal generator. In single-sideband operation, the audio is translated onto the I-F frequency in the balanced modulator. The output of the balanced modulator is the single-sideband I-F signal. For maximum effectiveness, this SSB signal must remain virtually free of intermodulation distortion. Distortion in excess of 40 dB can affect the SSB signal to the point of unintelligibility. To prevent distortion to the SSB signal, linear amplification is used in the I-F and RF stages. Again, testing the I-F stages of a single-sideband transmitter is similar to that for testing the I-F stages in AM transmitters. Here a two-tone signal generator is used in lieu of a CW signal generator.

3-2.7.3 Variable Tuned Filters

The output circuit of each I-F and RF amplifier is equipped with a tuned circuit (filter) designed to eliminate distortion and/or to pass the desired band of frequencies. In fixed frequency transmitters such filters are of a fixed value, but in variable transmitters, the filters can also be varied as to what they will pass. This is done either by varying the spacing between the capacitor plates or by shorting out appropriate turns of the coil, or both. In the VHF and SHF ranges, the size of the filter's resonant cavity is changed to accomplish the same effect. In each case, the filter must be able to track the range of frequencies in which it will operate. Trimmer components, designed to set the end ranges of the filter, are incorporated in the circuitry to permit alignment for proper tracking. Where large frequency ranges must be tracked, additional compensation, such as slugs in a resonant cavity or sectional rotor plates on the main tuning capacitor, may be incorporated, as illustrated in Figure 3-32. Detailed adjustment and alignment procedures for the trimmer and tracking components are described in the specific equipment's technical manual. As a general rule, it is best to start aligning at

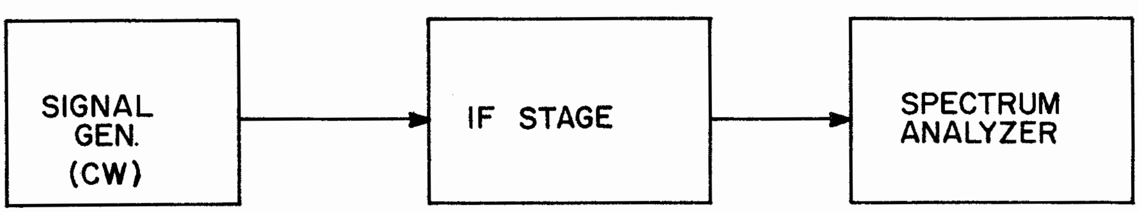


Figure 3-31. I-F Gain and Distortion Measurement, Test Equipment Arrangement

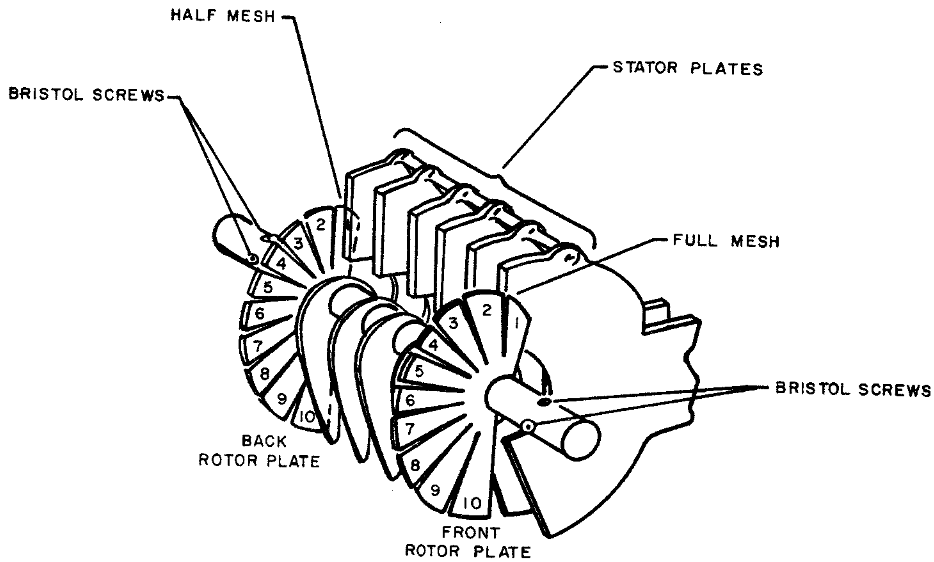


Figure 3-32. VHF Transceiver Main Tuning Capacitor

the high frequency end. The capacitor is adjusted first, then low end and the coils are adjusted. If tracking is then needed, the best procedure is to start again at the high end, making incremental tracking adjustments as the equipment is tuned to the low end. Power output monitoring provides the best indication of correct adjustment, and can be done with a wattmeter or power meter connected to the output of the transmitter. If an I-F stage is being adjusted, a spectrum analyzer at the output of the last I-F stage is used to monitor the signal level.

3-2.7.4 RF Power

Because even a substantial reduction of transmitted power does not necessarily decrease the operating range significantly, the precise measurement of RF power may appear to be unnecessary. This is a false conclusion. A change in the power output may result from altered operating conditions that are capable of causing equipment breakdown unless remedied. In addition, power measurements are often the surest way of determining whether the over-all performance of a transmitter is normal and, in general, is consistent with its designed specifications. Tests should therefore be made periodically, either with the same test equipment or with different instruments of equal accuracy. If this practice is observed, a change in the test indication will reliably signify trouble rather than a discrepancy between the instruments. There is also a direct relationship between the maximum power output of a transmitter and the amount of distortion permissible in the system. For this reason, the power output rating of a transmitter is usually given in terms of the maximum power that can be delivered with respect to a specified amount of distortion that can be tolerated. If no value of distortion is specified, it is understood that the distortion will be kept within limits considered to be acceptable for the system. The direct relationship, as well as the importance of this relationship, between the transmitter power output and permissible distortion should not be overlooked when considering the power output of a transmitter.

3-2.7.5 Peak-Envelope-Power (PEP) Output Measurement

The power output of an SSB transmitter is usually given in terms of peak-envelope power (PEP). This term is defined as the rms power developed during the peak RF cycle. Peak-envelope power is equal to the sum of the amplitudes of the sideband components and pilot carrier. The measurement of PEP usually employs the two-tone test procedure with the carrier turned "ON". During the two-tone test, the PEP occurs when the peaks of the two-tones are in coincidence. Another rating sometimes given to SSB transmitters is called "peak-sideband power" (PSP). Peak-sideband power is similar to peak-envelope power except that the measurement is made with the carrier turned "OFF"; i.e., all the transmitter power is applied to the sidebands and none is applied to the carrier. "Talking power" is a term often used in reference to SSB transmitters. Talking power is defined as that portion of the transmitter output power that carries the intelligence of the message. Since only the desired sideband is radiated in SSB systems, the talking power and peak-sideband power of such systems are closely related. As carrier suppression increases, the amount of available sideband power (and the talking power) will increase. If the carrier is suppressed completely, the total output energy of the transmitter will then be applied to the sideband signal.

3-2.7.6 AM/FM Considerations

Both AM and FM transmitters are rated in terms of average power. In AM transmitters the power contained in the carrier does not change with an increase or decrease in the percent of modulation. Sideband power can increase the power output of a transmitter by as much as fifty percent at 100% modulation. In FM transmitters, the dispersement of power is averaged throughout the sidebands and the carrier, therefore, there is neither increase nor decrease in power as modulation becomes changed. Automatic drive control incorporated in a SSB transmitter prevents the transmitter from exceeding its peak envelope power rating. The control reduces the drive to the linear power amplifier, based on the output of that amplifier. As a result, the rated average power output of the transmitter can be monitored as a quality indicator instead of monitoring the peak envelope power.

3-2.8 TRANSMITTER POWER MEASUREMENT

To measure the average power output of a transmitter, the output line is terminated into a characteristic load (usually resistive) through an in-line wattmeter or via a directional coupler, as illustrated in Figure 3-33, A and B. The configuration in Figure 3-33A is practical up to 1000 watts. The configuration in Figure 3-33B is limited by the power rating of the directional coupler. When taking power measurements using the directional coupler method (Figure 3-33B) the attenuation at the measuring port must be added to the meter indication. Most directional couplers also include a reverse power port, which is convenient for measuring reflected power. Transmitters incorporating power meters use the directional coupler method in checking outputs. This allows constant power monitoring during equipment operation. The circuitry is similar to that illustrated in Figure 3-34.

3-2.9 NEUTRALIZATION PROCEDURES

Neutralization is the process of balancing out the voltage that feeds through an RF amplifier by means of the interelectrode capacity of the electron

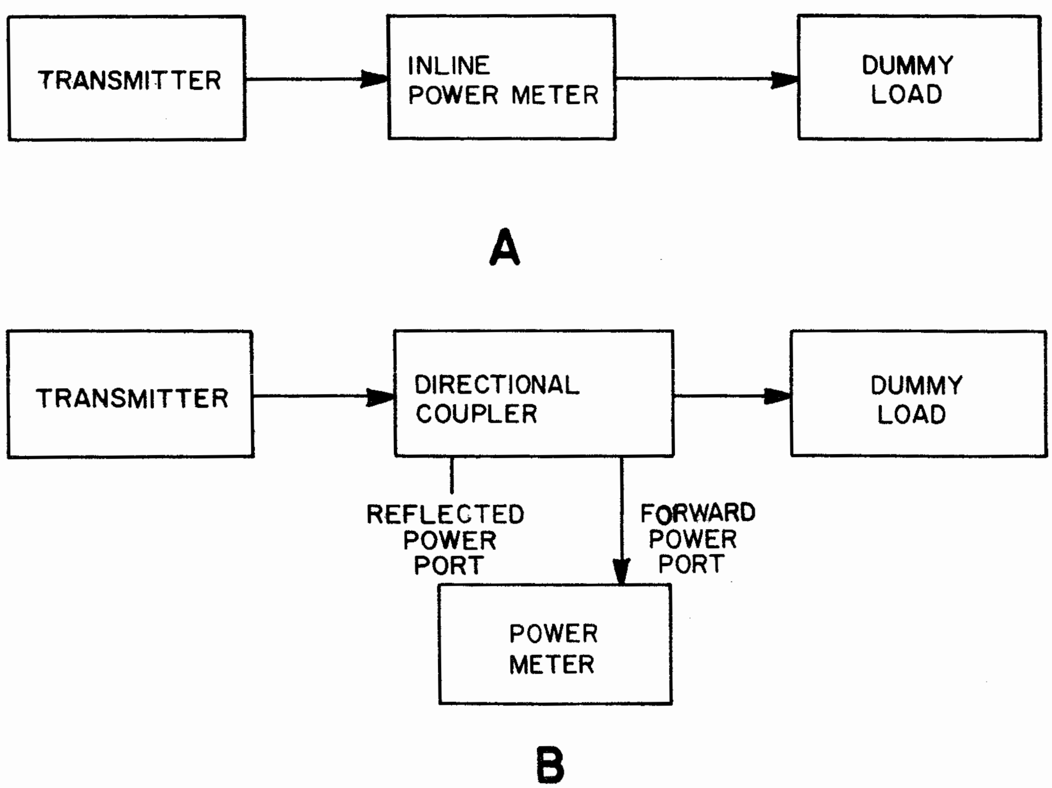


Figure 3-33. Transmitter Average Power Measurement Configurations

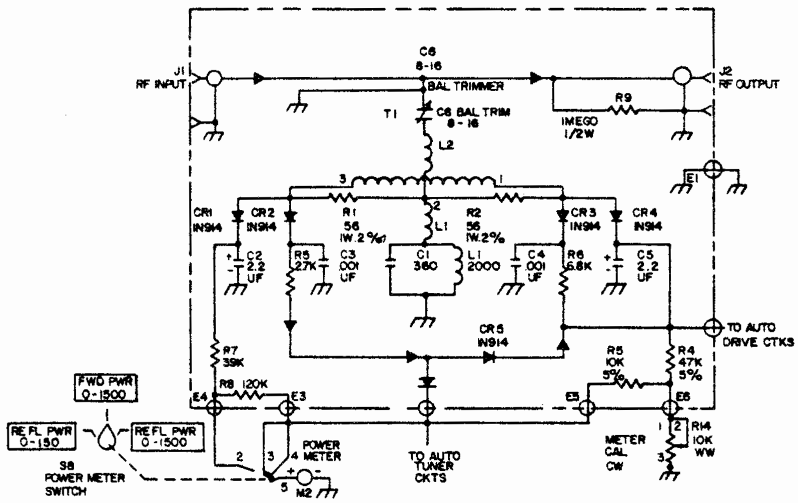


Figure 3-34. Directional Coupler With Automatic Drive Control Signal

tube. Since the usual amplifier operates with its input and output circuits tuned to the same frequency, the amplifier will break into oscillation as a tuned-plate, tuned-grid oscillator if the RF feedback between the grid and plate is not brought to the necessary minimum. Triodes have the greatest interelectrode capacitance, and, therefore, require maximum stabilization. Although the plate-grid capacitance of the screen-grid tubes normally used is reduced to only a fraction of a picofarad by the screen grid, the power sensitivity of these tubes is so great that only a small amount of feedback is necessary to induce oscillation. Therefore, to ensure stabilization of the amplifier, it is usually necessary to either load the grid circuit or to use a neutralizing circuit external to the tube in order to balance the voltage feedback through the grid-plate capacitance by another voltage of opposite phase. For proper neutralization, the neutralizing voltage must be opposite in phase and equal in amplitude with respect to the feedback voltage between the grid and plate. When the plate circuit is divided so that the neutralization voltage is developed across part of it, the amplifier is said to have grid neutralization. There are many variations in the methods employed to provide neutralization. As the operating frequency of an RF amplifier is raised, the output energy coupled back to the input circuit by the grid-late interelectrode capacitance eventually becomes large enough to cause sustained oscillations of the tuned-plate tuned-grid amplifier. This kind of oscillation differs from parasitic oscillations in that the oscillations occur at the resonant frequency of the tank circuits. In the neutralization process, the regenerative feedback is either cancelled out (neutralized) by a voltage of opposite polarity, or prevented from reaching the input circuit by a high series impedance in the feedback path. The first method utilizes various bridge circuits to provide the neutralizing signal, while the second method makes use of a suitable inductor shunted across the feedback capacitance, to develop a high series impedance at the oscillating frequency. As a general rule, when triode RF stages are employed, regeneration occurs above 100 kHz, but when screen-grid tubes such as tetrodes, pentodes, and beam tetrodes are used, regeneration is seldom troublesome below 30 MHz. The superiority of screen-grid tubes in this regard is attributable to the shielding action of the screen and suppressor grid, which are placed at RF ground potential. Because of this design the grid-plate interelectrode capacitance is greatly reduced in magnitude. The low value of grid-plate capacitance precludes oscillation at all but extremely high frequencies. No neutralization is required in frequency-multiplier circuits, because the grid and plate circuits are tuned to different frequencies. Another method of reducing feedback is the use of a grounded-grid amplifier, which is particularly effective in VHF and UHF receivers. With the grid of the tube connected to ground so that the input signal is applied to the cathode, any energy coupled from the plate to the grid by the grid-plate capacitcance is returned to ground directly. The amplifier is characterized by low input impedance, which leads the exciting circuit, and by high output impedance Neutralization Circuits: bridge methods are shown in parts (A) through (D) of Figure 3-35. In part (A), a tapped inductor is used in plate tank circuit of a single-end triode RF amplifier. Since the RF voltages at the ends of the tank are 180 degrees out of phase, proper adjustment of the neutralizing capacitor will result in a null across the grid circuit at the frequency of oscillation. This method is successful only at frequencies below 7 MHz. The split-stator method, shown in part (B) of Figure 3-35, is used more widely. This arrangement makes the electrical balance virtually independent of the mutual coupling within the coil, and also of the point where the coil is tapped. If adjustment is made at a relatively high frequency, such as 15 MHz, the stage can usually be operated at lower frequencies without requiring further adjustment. Sometimes an additional balancing capacitor is inserted between ground and the junction of the split-stator and balancing capacitors. Its purpose is merely to maintain equal capacitance to ground on each side of the balanced plate tank circuit. Part (C) of Figure 3-35 shows a plate-neutralizing circuit similar to the one in part (A), but lacking its limitations. A separate neutralizing coil is inductively coupled to the plate tank inductor. Part (D) shows a similar arrangement in which a coil is inductively coupled to the grid tank. Note that there is no flow of tank current through the neutralizing coil in either case. The size of the neutralizing capacitor, NC, depends on the coefficient of coupling between the tank and the neutralizing coils, and upon the relative values of the inductances. By proper proportioning of the neutralizing coil used on each band of operation, it is possible for one value of neutralizing capacitance to be used on all bands. In another form of grid-circuit neutralization, the tank coil is center-tapped in the same way as the plate tank coil shown part (A). In push-pull stages the symmetry of the circuit makes neutralization especially simple, as shown in part (E) of Figure 3-35. Push-pull neutralization has the advantage of making balance possible more readily than in single-ended amplifiers: this is an especially desirable property at very high frequencies. In addition, neutralization is usually preserved throughout the various bands of operation. The coil-neutralizing method is shown in part (F) of Figure 3-35. Inductor L_N is resonated with the grid-plate capacitance of the tube at the frequency of oscillation. This causes the grid-plate impedance at that frequency to be high enough that regeneration due to interelectrode capacitance is unable to occur. The principal advantage of this method is that single-ended tank circuits can be used with a single-ended amplifier. However, there is also the disadvantage of restricting the neutralization to a limited range of frequencies. This limitation can be offset somewhat by shunting a trimmer capacitor across the neutralizing coil. The stage can then be neutralized at any desired frequency within a band of operation, provided that the trimmer is tuned appropriately. Normally adjustment of the neutralizing circuitry is unnecessary. Component aging or deterioration, and replacements may result in the need to adjust the neutralizing circuitry. The specific transmitter's technical manual must be consulted before adjustments of the neutralizing components are made, because procedures will vary due to the design characteristics of the transmitters.

3-3 FACSIMILE SYSTEMS TESTING

Facsimile is a process of transmitting a picture map, or other graphic material (page copy) from one terminal to another. At the transmitting equipment, the picture is mounted on a circular drum or cylinder. The picture is then scanned by an electo-mechanical device. Picture variations along the scanning lines are converted to light variations, and then into electrical impulses which are used to modulate a transmitter signal. At the receiving end, the demodulated electrical impulses are fed to a facsimile recorder which reproduces the picture, using an electro-mechanical scanning device similar to that used at the transmitting end except that it reverses the process. Received copy can be recorded either directly on chemically coated paper, using electrical impulses; or photographically in positive or negative form by converting the electrical impulses back into light variations. Facsimile is used with either wire or radio communication circuits. When used with wire line services, the facsimile transceiver can be connected directly to the line or through a coupling coil; a conventional telephone headset may be coupled to the line for signal monitoring purposes. When used with radio communication circuits the facsimile transceiver can be connected in several ways, by use of auxiliary equipment, to produce different types of radio signals. Facsimile signals may be transmitted by radio by the audio frequency tone shift (AFTS) or the radio frequency carrier shift (RFCS) method. Conventional superheterodyne receivers are used to receive either type of transmission. The output of the receivers is in the form of AFTS signals in which 1500 Hz represents the maximum (black) signal, and 2300 Hz represents the minimum (white) signal. Another system employs AFTS signals in which 2300 Hz represents the maximum signal, and 3100 Hz represents the minimum signal. In either case, an 800 Hz shift is maintained between the maximum and minimum signal outputs from the facsimile transmitter at the sending station.

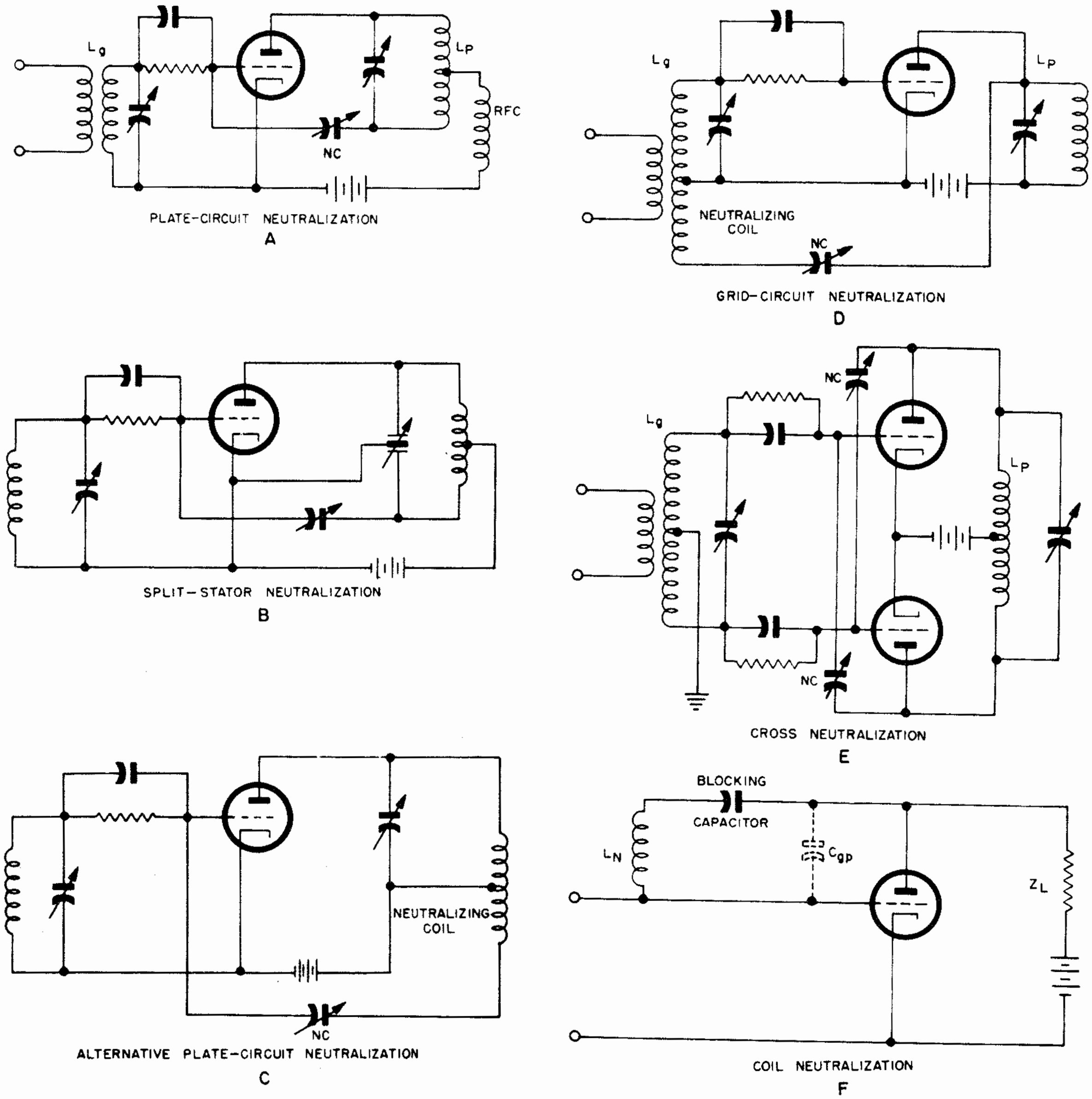


Figure 3-35. Neutralization Circuits

A simplified block diagram of a simplex facsimile system is shown in Figure 3-36. On the transmit side, a 2400 Hz carrier (amplitude-modulated corresponding to the various shades of the copy to be transmitted) is produced at the output of the transmitter section of Facsimile Receiver-Transmitter TT-321U/UX. The MD-168A/UX converts the AM signals to AFTS signals which modulate the RF carrier generated by the transmitter. The transmitter carrier is modulated plus or minus 400 Hz. If the RFCS method of transmission is to be used, the facsimile receiver-transmitter output is fed to Keyer Adapter KY-44C/FX. Keyer Adapter KY-44C/FX converts the TT-321U/UX AM signal output into dc keying

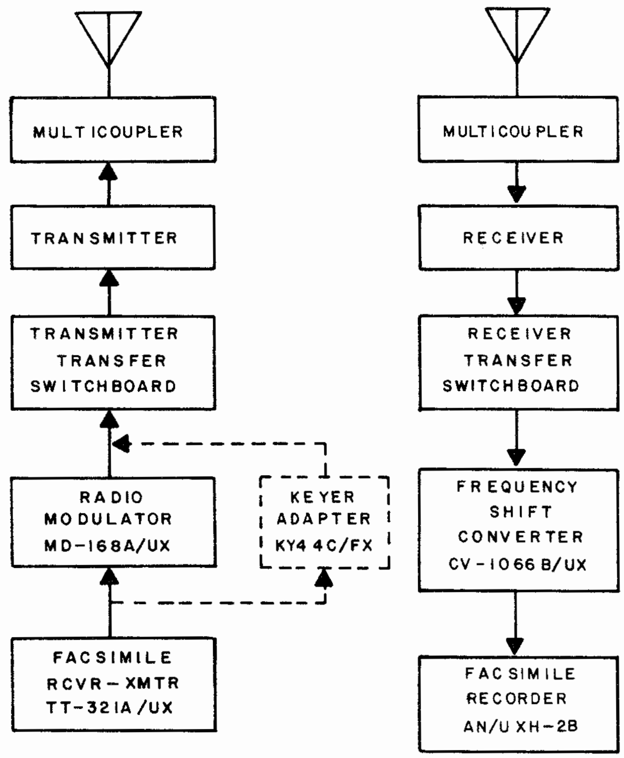


Figure 3-36. Facsimile System, AFTS and RFCS Modes.

signals suitable for use with a frequency-shift exciter unit. The receiver output (see Figure 3-36) is in the form of AFTS signals from 1500 Hz to 2300 Hz, or 2300 to 3100 Hz, depending upon the setting of the receiver BFO. These AFTS signals are fed to Frequency-Shift Converter CV-1066B/UX where they are converted into equivalent AM signals suitable for operating the facsimile recorder. The Facsimile Recorder AN/UXH-2C, Facsimile Receiver-Transmitter TT-321U/UX, Radio Modulator MD-168/UX, Keyer Adapter KY-44C/FX, and Frequency Shift Converter CV-1066B/UX are described later in this chapter. The data mode is also used for the transmission of facsimile signals as shown by Figure 3-37. The AM signals from the TT-321U/UX are fed to the A/D (analog/digital) converter, where they are converted into equivalent digital data signals and fed to the "modem" unit. The modem unit converts the digital data signals into the AFTS signals that modulate the transmitter carrier. On the receiver side, the receiver output signals are fed to the modem unit, where they are converted to digital data signals and fed to the D/A (digital/analog) converter. The D/A converter converts the digital data signals back to AM signals suitable for operating the facsimile recorder. Secure facsimile systems are produced by connecting security equipment between the modem units and the A/D and D/A converters.

3-3.1 FUNDAMENTALS OF FACSIMILE

A facsimile system consists of a transmitter and recorder. These are used to convey graphic matter from its source to a distant point. The graphic matter (subject copy) may be photographs, sketches, and typewritten, printed or handwritten text, as well as weather maps or charts. Transmission is by a direct wire link or by a radio communications system.

3-3.1.1 Facsimile Transmitter

The facsimile transmitter must resolve the copy to be transmitted into very small incremental areas and must transmit the average density of each area separately. These incremental areas must be considerably smaller than the smallest intelligence that is to be transmitted since the received image is comprised of dots the size of the incremental area. An exciter lamp is used as a source of uniform illumination of the copy and a phototube is used to determine the brightness of each incremental area. The copy to be transmitted is clamped securely on a drum of the facsimile transmitter. An image of the copy to be

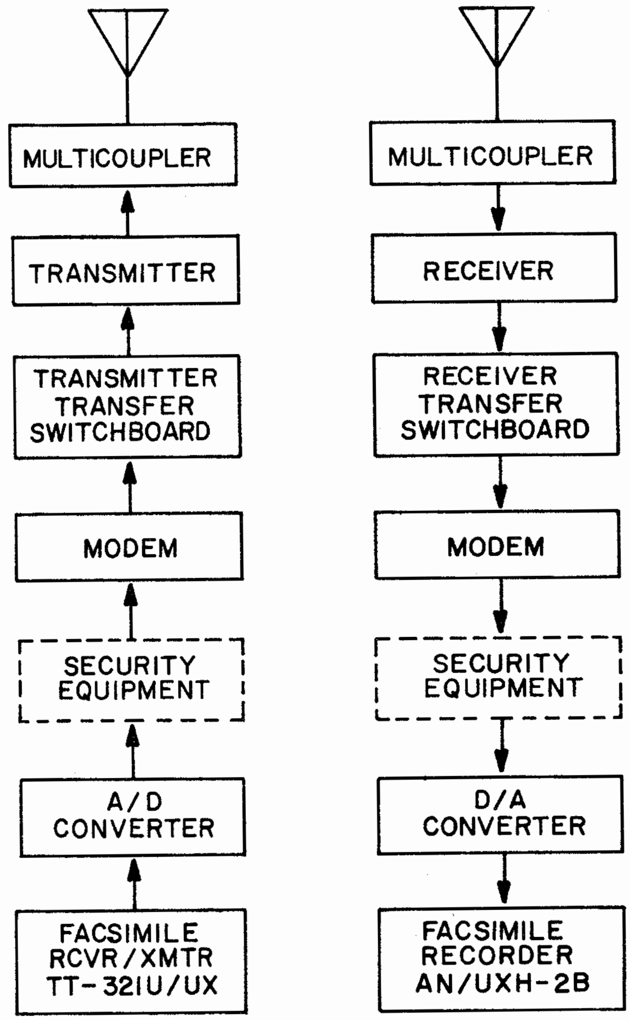


Figure 3-37. Facsimile System, Data Mode

transmitted is focused upon an aperture plate located in front of the phototube. The size of the hole in the aperture plate determines the area of the copy image (incremental area) which passes through the phototube. The phototube acts as a variable resistor in a modulator circuit to control the amplitude of a carrier frequency. When a dark area of the copy is seen by the phototube through the aperture, the phototube allows a maximum signal to pass. A white area seen by the phototube results in a minimum signal, and a grey area produces an intermediate-amplitude signal. These phototube modulator outputs represent the facsimile signal which is transmitted to the facsimile recorder, although the grey area signals are normally not true representations of the original copy. In order to scan each incremental area of the copy, the drum upon which the copy is clamped rotates and, at the same time, moves laterally from right to left. The relative position of the optical system thus shifts by one incremental area for each drum revolution. When the optical system has completely scanned the copy, all the incremental areas on the copy have been seen by the phototube, and a signal of corresponding amplitude has been transmitted for each incremental area.

3-3.1.2 Facsimile Recorder Set

The facsimile recorder set amplifies facsimile signals originating at the facsimile transmitter, and converts the result into appropriate density variations on the recording paper. The received facsimile signal is amplified and applied to three small styli, which are driven across the recording paper. These styli are driven by a belt, and only one stylus can contact the recording paper at a given time. The recording speed of the stylus at the facsimile recorder is the same as the peripheral speed of scanning at the facsimile transmitter. A stylus makes a traverse of the width of the recording paper in the same interval that the drum of the facsimile transmitter makes one revolution. Thus, if the recorder is properly phased at the start of a transmission, the recording stylus is maintained in the same relative position as the scanning spot at the transmitter. When a black area of the copy is scanned by the transmitter, a maximum signal is transmitted, amplified by the recorder, and then impressed upon the recording paper by the stylus to print a corresponding incremental area of maximum density (black). When a white area is scanned, the received signal is so low in value that nothing is recorded in the corresponding incremental area on the recording paper. When the average density of the incremental area is grey, the stylus is driven with less magnitude than that of the black signal, therefore a grey variation is reproduced.

3-3.1.3 Signal Requirements

The facsimile recorder set is capable of recording double-sideband, amplitude-modulated carriers such as the output of Facsimile Set Series TT-321U/UX or a vestigial-sideband having the following characteristics: An amplitude modulated carrier of 2400 hertz with the upper sideband suppressed at the facsimile transmitter and with a carrier modulation no less than 75 percent. The facsimile recorder set responds only to an amplitude modulated signal. This signal may be transmitted directly from the facsimile transmitter to the facsimile recorder set via land lines or may be transmitted via a radio link.

3-3.1.3.1 Land Lines

The carrier that transmits intelligence via land lines has a frequency of 1800 Hz to 2400 Hz. The signal is superimposed upon the carrier so that the black level is 12dB above the white level and signal amplitudes between the black and the white levels correspond to gradations in the density of the transmitted copy.

3-3.1.3.2 Radio

In a radio transmission, intelligence is conveyed by a frequency shift carrier with an effective center frequency of 2700 Hz, a lower frequency limit of 2300 Hz, and an upper frequency limit of 3100 Hz. The 2300 Hz signal corresponds to the black level; the 3100 Hz signal corresponds to the white level; and the signal frequencies between the black and white signal frequencies correspond to the gradations in the density of the transmitted copy. The frequency shift carrier, which is insensitive to noise variations, is imposed upon the carrier of a radio transmitter and is subsequently received, detected and amplified by a radio receiver. The received signal, at the output of the associate receiver, is a 2300 to 3100 Hz audio frequency shift signal that must be converted to the amplitude modulated signal required by the facsimile recorder set. This conversion is provided by a frequency shift converter such as the CV-1066/UX Series.

3-3.1.4 Control Signals

The facsimile recorder set is designed for automatic, unattended operation, which is enabled by the transmission of automatic control signals. These control signals are in the form of audio tones amplitude modulated upon the carrier. The control signals required for automatic, unattended operation are:

1. A start signal which is a 300 Hz transmitted for three to five seconds.
2. A phasing signal, which is a tone corresponding to the maximum density (black) level that periodically changes to the white level. When operating at a scan speed of 60 lines per minute, a phasing pulse recurs every second, during which the white level recurs for 25 msec. The phasing signal is transmitted for 15 seconds, during which interval the facsimile recorder set is phased and, when manual gain is used, the recording level is set by the operators.
3. A start-record signal which is a 60 Hz square-wave modulation of the carrier that is transmitted for one second; and
4. A stop signal, transmitted after the facsimile copy signals. This is a 450 hertz, square-wave modulation of the carrier, transmitted from three to five seconds duration.

NOTE

After transmission of the stop signal, the carrier frequency is removed from the line for three to five seconds.

When the transmitter does not have the capability of generating automatic control signals, a phasing signal is optically generated by scanning black and white bar patterns. The operator, in response to an audible presentation of this signal, manually phases the facsimile recorder set.

3-3.2 OVERALL FUNCTIONAL DESCRIPTION

The facsimile recorder set (Figure 3-38 and 3-39) starts automatically at the beginning of a transmission, prints copy the proper density and in the proper position on the recording paper, and stops when transmission is complete. All control and copy signals

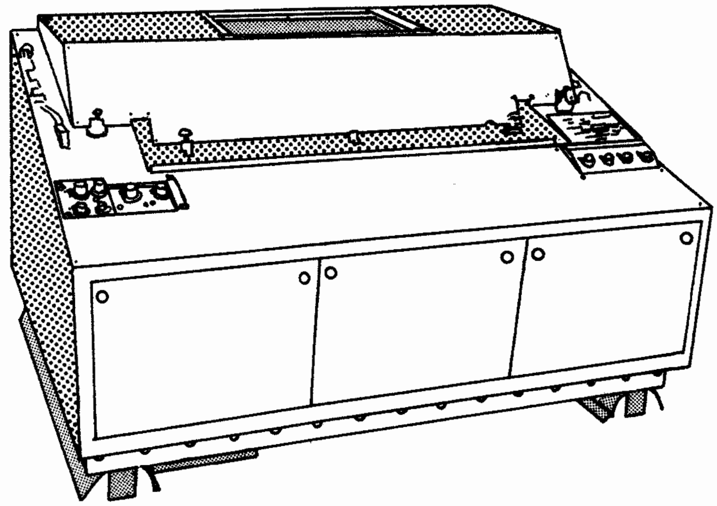


Figure 3-38. Facsimile Recorder Set AN/UXH-2C

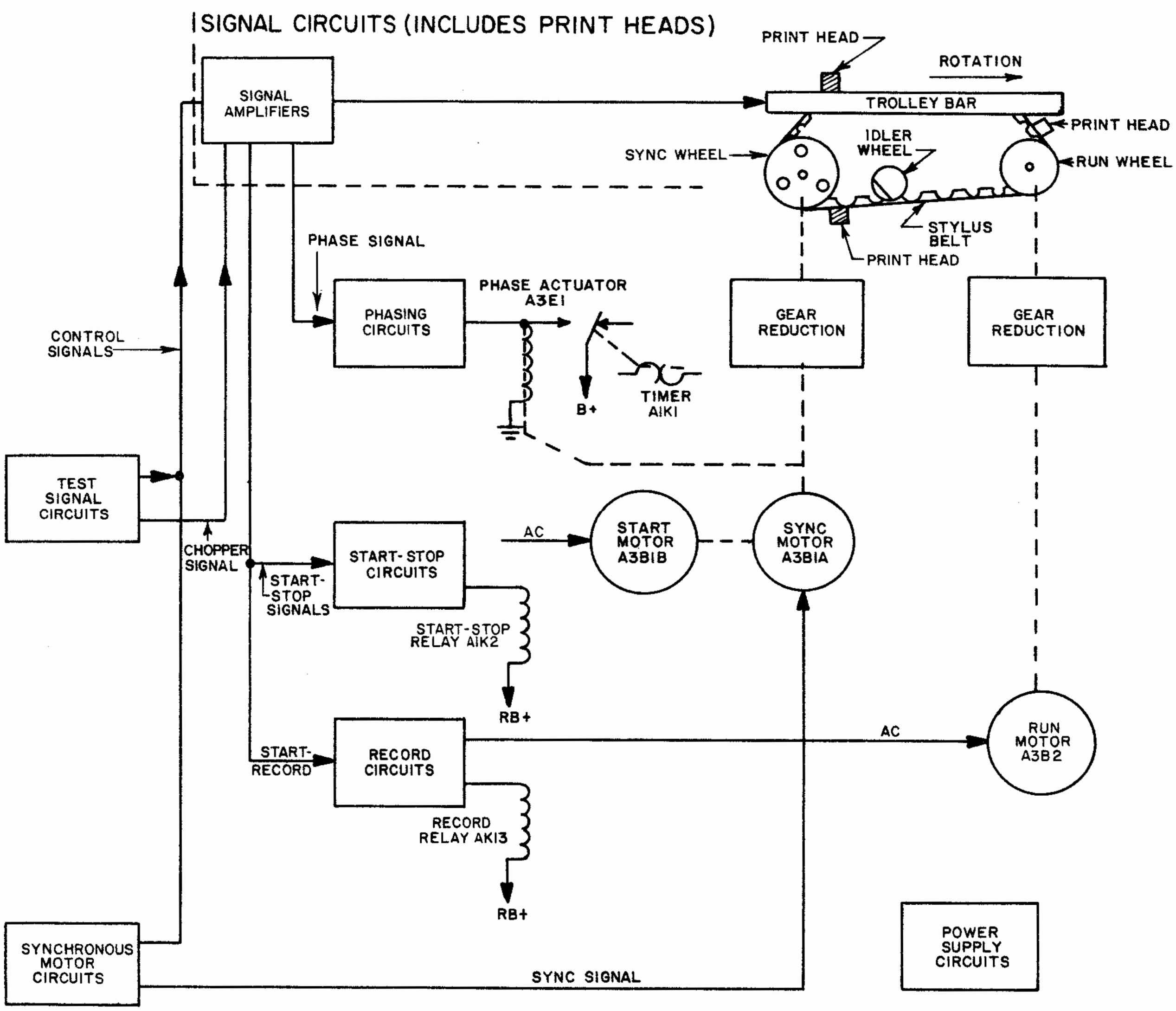


Figure 3-39. Facsimile Recorder Set, Functional Block Diagram

are applied to the speaker amplifier, for monitoring purposes, and are amplified within the signal circuits. Figure 3-39 depicts the functional block diagram.

3-3.2.1 Stand-by Condition

With power applied but prior to the reception of any control signal, the facsimile recorder set is at stand-by. In this condition, the print output of the signal circuits is cutoff and power to the run motor is interrupted by the record circuits. In turn, the record circuits are maintained at this inoperative condition by start-stop relay A1K2 in the energized condition.

3-3.2.2 Start Cycle

The start cycle is initiated by the 300 Hz start tone which occurs during a three to five second transmission interval. This signal is applied to the start-stop and record circuits, but only the start-stop circuit responds. Relay A1K2 releases, which, in turn:

1. Releases phase actuator A3E1 so that the equipment will phase in response to the subsequent phase control signal.
2. Energizes timer A1K1, the contacts of which close after a 7-second delay to energize the phase actuator. During these seven seconds, phasing is accomplished.
3. Makes a holding voltage available so that the record circuits can respond to subsequent start-stop signals.

3-3.2.3 Phase Cycle

The phase cycle is initiated by the 15-second phase signal that consists of the maximum intensity (black level) signal periodically interrupted by the white level. The phasing signal appears immediately after the start signal which had released the phase actuator. The released actuator responds to the phasing pulse in such a manner that a correct phase relation is established between the remote transmitter and the sync drive circuits of the recorder set.

3-3.2.4 Start-Record Cycle

The start-record cycle is initiated by the 60 Hz start-record signal applied to the start-stop and record circuits from the signal amplifiers. Only the record circuits respond to this signal. Relay A1K3 energizes which, in turn:

1. Removes the cutoff voltage from the print output of the signal amplifiers.
2. Starts the run motor to drive the stylus belt and paper-advance mechnism.
3. Applies a holding voltage to phase actuator A3E1.

3-3.2.5 Copy Cycle

The incoming copy signal is amplified by the signal amplifier and applied through the trolley bar to the print heads. Simultaneously, the mechanical systems provide the following functions:

1. The run system drives the paper feed mechanism which advances the recording paper relative to the copy being recorded.
2. The run system drives the stylus belt, but the sync system restrains the belt such that the speed of the recording stylus exactly matches the speed of the facsimile transmitter scanner.

3-3.2.6 Stop Cycle

The stop cycle is initiated by the 450 Hz stop signal that causes the start-stop circuits to re-establish the start condition.

3-3.2.7 Test Circuits

The test circuits simulate control and test signals which are used in maintenance of the equipment.

3-3.2.8 Power Supply Circuits

The power supply circuits provide a regulated output of 225 volts, an unregulated output of 400 volts, four separate bias supplies of -55 through -24 vdc (bias no. 1), -24 vdc (bias no. 2), -1.5 vdc (bias no. 3), and -1.4 vdc (bias no. 4) and separate filament voltages of 6.3 vac at 9 amperes and 6.3 vac at 2 amperes.

3-3.2.9 Automatic Operation

Automatic operation, in response to control signals, provides the following relay operating states:

Condition	A1K1	A1K2	A1K3	A1E1
Standby Cycle	R	E	R	E
Start Cycle	E	R	E	R
Phase Cycle	E	R	E	R
Start-Record Cycle	R	R	E	E
Recording Cycle	R	R	E	E
Stop-Record Cycle	R	E	R	E

R = Released E = Energized

NOTE

Note that relays A1K2 and A1K3 are always in alternate conducting states.

3-3.2.10 Manual Operation

In manual operation, instead of the control signals, operation is initiated by manual switching. These switching procedures provide the following relay operating states:

Condition	A1K1	A1K2	A1K3	A3E1
Standby Cycle	R	E	R	R
Phase Cycle	R	E	R	R
Recording Cycle	R	R	E	E
End of Copy Cycle	R	E	R	E

R = Released E = Energized

3-3.2.11 Normal Operation

Normal operation of the facsimile recorder set may be assessed by the quality of the recorded copy. If the copy is clear, intelligible, and has sufficient contrast, it may be assumed that the system is operating normally. In event of abnormal operation, check that fuses are functional and that during the recording cycle the POWER ON and RECORD indicator lamps are glowing. Use the following checklist to verify that control settings are correct.

Control	Position
AUTO/MANUAL Switch	Is set at AUTO when associate facsimile transmitter is capable of transmitting automatic control signals; if not, set the switch at MANUAL.
SCAN/MIN Switch	Set to 60, 90, or 120, as required by the facsimile transmitter.
AUTO/MANUAL GAIN Control	Set at AUTO when using a direct-wire or compatible radio link, and PRINT LEVEL meter reads between 50 and 150; set at MANUAL when the PRINT LEVEL meter reads less than 50. In MANUAL, set the GAIN during the phasing cycle for a meter reading of 100.
DENSITY CONTROL	Set for acceptable printing level.

Any deviation from "normal operation" need not be restricted to a faulty facsimile recorder set. Faulty operation may also be caused by failure of the operator to perform a required function, by spurious signals on the transmission line, or by a malfunction of either the associate frequency shift converter or the facsimile transmitter. For example, if the background prints a light grey or if there is a partial tone reversal in the recorder copy, either the transmitter contrast is incorrectly set or extraneous signals are appearing in the transmitter circuit. Similarly, if the recorded copy displays a group of vertical lines, the transmitter drum is not feeding. Use the STEADY, START, RECORD, and STOP test switches (Figure 3-40) to simulate the control signals normally sent by the transmitter; use the STEADY or CHOPPER switches to simulate the copy signal.

3-3.2.12 Additional Checks

Perform a visual check on the facsimile recorder set. Make certain there are no loose cables or signal input leads; check for charred or discolored insulation, improper control settings, or other evidence of equipment malfunction. Check that the line voltage switch on the power supply is set for the available line voltage, and that the proper rated fuses are used. Check for the illumination of blown-fuse indicators when power is applied to the facsimile recorder set. Check that primary power is available for operation of the equipment. Mechanical failure may cause trouble as well as troubles due to defective or incorrect electronic components. When mechancial faults are present, they must be eliminated before electronic circuits can be adjusted properly. The initial step in troubleshooting the facsimile recorder set is to recognize that a trouble exists and, having determined that operation is abnormal, to localize the trouble either to the facsimile recorder or to peripheral equipment. For more detailed troubleshooting information, refer to NAVSHIPS 0967-287-9010 technical manual for Facsimile Recorder Set AN/UXH-2C.

3-4 TELETYPEWRITER TESTING

Because of the increasing variety of teletypewriter equipment installed aboard ship it is impractical to describe every piece of equipment likely to be encountered. The equipment listed in Tables 3-3 through 3-10 is representative of the types commonly employed in shipboard installations. In some instances this equipment may be designated by nomenclature different from that shown in the tables; but in most instances, this variance in nomenclature merely indicates a modification of the basic equipment described herein.

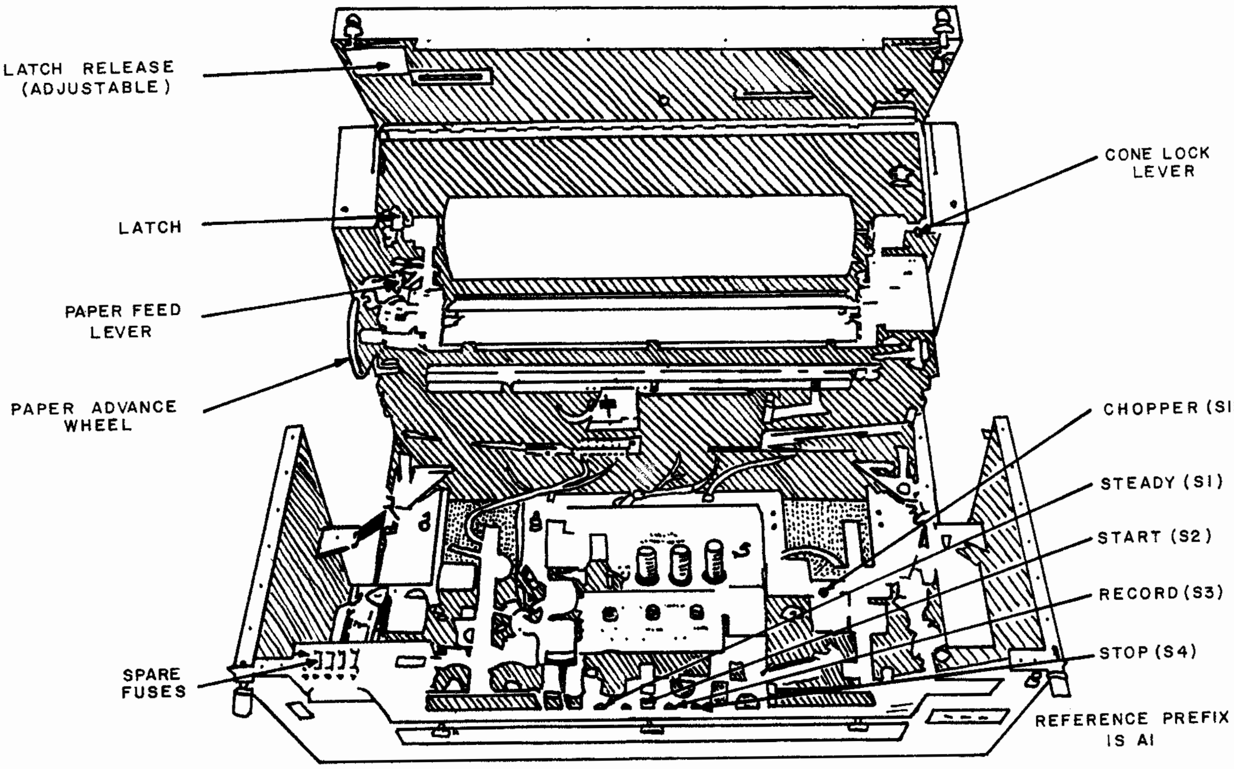


Figure 3-40. Location of Test Switches

3-4.1 TELETYPEWRITER EQUIPMENT

The function of the teletypewriter is to provide a record copy of message traffic (Page Printers), tape for transmission purposes (Perforators and Reperforators), and message transmission equipment (Transmitter Distributors, Keyboards).

3-4.2 RANGE ORIENTATION

The range finder on a teletypewriter printer is used to determine the time at which character elements are sampled in order to compensate for biased signal. Range finders are numbered from 0-120. A well adjusted teletypewriter should be able to produce errorless copy over 80 points of this range, when a "no-bias" test signal is provided for input. The numerical value of the range limits are relatively unimportant as long as the total range is 80 points. Thus, ranges may be 40-120, 25-105, 30-110 and even 35-115. These ranges will vary from teletypewriter to teletypewriter. Regardless of range, the range finder should be centered on the optimum setting. If the range is 20-100, then the optimum setting would be 60. This figure is obtained by adding the two extremes and dividing by two.

3-4.2.1 Page Printer Malfunction

If a circuit is printing garbled traffic and the cause has been isolated to a page printer, the first step is to replace it with one known to be good. If this is not possible, patch a local test to the page printer to check if it has the proper range.

3-4.2.2 Range Finding

To find the range while the test is running, turn the range finder up until it garbles, then turn the range finder down until it again garbles. These two points are the minimum and maximum of the range. The center point between this low and high range is the optimum setting. If this range is 80 points or more, the printer is working properly. At this point the range finder should be placed on the optimum setting, but in some circumstances it may be necessary to work up and down the range a little bit. Do not make this a frequent practice, for only during very unusual circumstances will teletypewriter performance be improved on a range higher or lower than the optimum setting.

3-4.3 MAINTENANCE AND ADJUSTMENTS

Lubrication, preventative maintenance, and adjustments of teletypewriter equipment must be conducted in accordance with the appropriate PMS schedule. Because of the large amount of information involved in PMS, this subject is not addressed herein.

Table 3-3. Automatic Send-Receive Page Printer Sets

NOMENCLATURE	DESCRIPTION
AN/UGC-5( )	Console type (high-level) keyboard, page printer, typing reperforator, transmitter distributor. 7.42 unit code.
AN/UGC-6( )	Console type (high-level) keyboard, page printer, typing reperforator, transmitter distributor. 7.42 unit code.
AN/UGC-13( )	Console type (high-level) keyboard, printer, typing reperforator, transmitter distributor, NTDS use with special computer interface electronics. 7.00 unit code.
AN/UGC-48( )	Console type (low-level version of AN/UGC-6) meterological use. 7.42 unit code.
AN/UGC-49( )	Console type (low-level version of AN/UGC-13). 7.00 unit code.
AN/UGC-59( )	Console type (high-level) keyboard, page printer, keyboard non-typing reperforator, type reader. 11.0 unit 8 level code. AN/UYA-5 or AN/UYK-7(v) computer interface equipment.

Table 3-4. Keyboard Send-Receive Page Printer Sets

NOMENCLATURE	DESCRIPTION
AN/UGC-20( )	Compact cabinet (high-level) keyboard, page printer. 7.42 unit code.
AN/UGC-47( )	Floor console type (low-level version of TT-47) keyboard, page printer. 7.00 unit code.
AN/UGC-60( )	Floor console type (low-level) keyboard, page printer, similar to TT-47L/UG. 7.42 unit code.
AN/UGC-77( )	Compact cabinet type (low-level version of AN/UGC-20) keyboard, page printer. 7.42 unit code.
AN/UGC-79( )	Floor console type (high-level) keyboard, page printer. 11.0 unit 8 level code. Model 35 modified for computer interface with missile and weapons control systems consoles.
AN/UGC-91( )	Compact cabinet type (low-level) keyboard, page printer. Meteorological use. 7.42 unit code.
TT-47( )/UG	Floor console type (high-level) keyboard, page printer. 7.42 unit code.
TT-69( )/UG	Cabinet type (high-level) keyboard, page printer. 7.42 unit code.
TT-70( )/UG	Cabinet type (high-level) keyboard, page printer, series governed motor. 7.42 unit code.
TT-128( )/UG	Cabinet type (high-level) keyboard, page printer. Meteorological use. 7.42 unit code.
TT-130( )/UG	Cabinet type (high-level) keyboard, page printer. Meteorological use. 7.42 unit code.
TT-176( )/UG	Rack mount type (high-level) keyboard, page printer. 7.42 unit code.

Table 3-5. Receive Only Page Printer Sets

NOMENCLATURE	DESCRIPTION
AN/UGC-25( )	Compact cabinet type (high-level) page printer. 7.42 unit code.
AN/UGR-9( )	Compact cabinet type (low-level version of AN/UGC-25) page printer. 7.00 unit code.
AN/UGR-16( )	Compact cabinet type (low-level) page printer. Meteorological use. 7.00 unit code.

Table 3-6. Multiple Page Printer Console Sets

NOMENCLATURE	DESCRIPTION
AN/FGC-79( )	Floor console type (high-level) 3 page printers, keyboard. 7.42 unit code.
AN/FGC-100( )	Floor console type (high-level) 4 page printers, no keyboard. 7.00 unit code.
AN/UGC-61( )	Floor console type (low-level version of AN/FGC-79( ) 3 page printers, keyboard. 7.42 unit code.
AN/UGC-61(A)	Same as AN/UGC-61 with structurally reinforced cabinet for shipboard use.
AN/UGR-10( )	Floor console type (low-level version of AN/FGC-100) 4 page printers, no keyboard. 7.00 unit code.
AN/UGR-10(A)	Same as AN/UGR-10 with structurally reinforced cabinet for shipboard use.

Table 3-7. Send-Receive Typing Reperforator

NOMENCLATURE	DESCRIPTION
TT-253( )/UG	Cabinet type (high-level) keyboard, typing reperforator. 7.42 unit code.
AN/UGC-78( )	Cabinet type (low-level version of TT-253( )/UG) keyboard, typing reperforator. 7.42 unit code.

Table 3-8. Receive Only Typing Reperforators

NOMENCLATURE	DESCRIPTION
TT-192( )/UG	Compact cabinet type (high-level) no keyboard, typing reperforator. 7.00 unit code.
TT-571( )/UG	Cabinet type (low-level version of TT-192( )/UG) no keyboard, typing reperforator. 7.00 unit code.
TT-605( )/UG	Compact cabinet type (low-level version of TT-192/UG) no keyboard, typing reperforator. 7.00 unit code.

Table 3-9. Tape Transmitter Distributor Sets

NOMENCLATURE	DESCRIPTION
TT-187( )/UG	Standard cabinet (high-level) 7.42 unit code.
TT-570( )/UG	Compact cabinet (low-level) version of TT-187( )/UG). 7.00 unit code.
TT-603( )/UG	Miniaturized cabinet (low-level) 7.42 unit code.

Table 3-10. Signal Distortion Test Sets

NOMENCLATURE	DESCRIPTION
TS-2/TG	Chest type, portable ED-57GC test set. Code disk operated.
AN/UGM-8B(V)	Teletypewriter Test Set which generates a wide variety of start-stop and synchronous type teletypewriter signals used for testing teletypewriter equipment and systems.

3-4.4 FACTORS AFFECTING QUALITY OF COMMUNICATIONS

An ideal teletypewriter circuit reproduces signals at the receiving end exactly as they were produced at the sending end. Unfortunately, this ideal signal seldom occurs under actual operating conditions. This is because signal units have a way of lengthening and shortening as they travel along the circuit. This lengthening and shortening of MARKS and SPACES occurring during transmission reduces the quality of the signal and is called distortion. Four fundamental types of distortion adversely affect fidelity of teletypewriter signals; Bias, Fortuitous, End and Characteristic. Start-stop and Synchronous are the two modes of teletypewriter operation, and both are explained in order to better understand distortion.

3.4.4.1 Start-Stop Mode

If a teletypewriter signal could be drawn on paper, it would resemble Figure 3-41. Shaded areas show intervals during which the circuit is closed (MARKING). Blank areas show the intervals during which the circuit is open (SPACING). In the code most frequently used in Navy teletypewriter communications, 7.42 units represent one teletypewriter character. The start element precedes the first code element and one teletypewriter character. The start element precedes the first code element and is always a SPACE signal. Its purpose is to start the receiving machine. The stop element follows the last code element and is always a MARK signal. Its purpose is to stop the receiving machine in preparation for receiving the next character. The start element is always equal in duration to the code elements. The stop element must be equal in duration to at least one element of the code. The most common mode uses a stop element 1.42 times the length of one element. It is common practice to refer to a code element as a unit and the duration of a unit as the unit interval. The code just described is technically referred to as a "five-unit code, start-stop, with a 1.42 unit stop." This code is more commonly referred to as the 7.42 unit code. This method of teletypewriter communication described is called the start-stop method and gets its name from the start-stop units. The start-stop method keeps teletypewriter machines and signals in synchronization

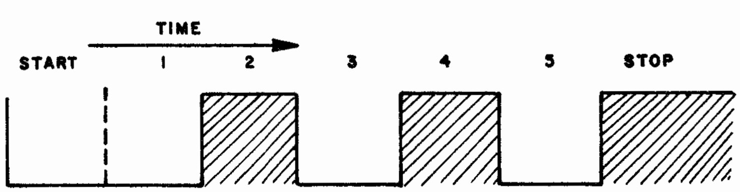


Figure 3-41. Mark and Space Signals in the Teletype Character R

with each other. With this method the selecting mechanism in the receiving machine comes to a complete stop after each character.

3-4.4.2 Character Transmission

Different characters are transmitted from the keyboard by an automatic process that selects various combinations of MARKING and SPACING in the five intelligence units (Figure 3-42). When reading tape, holes in the tape represent MARKS and solid areas represent SPACES. The reason is that holes in the tape allow the transmitter-distributor pins to rise, sending a current pulse. No holes in the tape prevent sensing pins from rising, thereby preventing current pulses. Thus we have spacing intervals. The machine, without benefit of tape perforations, automatically generates the start and stop elements.

3-4.4.3 Character Combinations

A total of 32 combinations can be obtained from the five intelligence units, but by using uppercase and lowercase, the number of characters obtainable is greatly increased. When a teletypewriter printing mechanism is shifted to uppercase as a result of receiving a FIGS shift character, all succeeding characters received before a LTRS shift character, print in uppercase as numerals and punctuation marks. The machine does not, however, make such double use of all 32 possible combinations, because six are used for the functions of carriage return, line feed, figures shift, letters shift, space and for one normally unused blank key. This leaves 26 of the 32 that can be employed in both uppercase and lowercase. When the six special functions are added, the total is 58 characters and functions that can be sent from a teletypewriter keyboard.

3-4.4.4 Signal Quality

Theoretically, Figure 3-41 represents a perfect signal. The quality of each element remains the same during its transmission, and the shift from MARKING to SPACING (and vice versa) is instantaneous. These changes are called "transitions". They occur at the beginning and end of each of the solid blocks. Some are MARK-TO-SPACE transitions, and others are SPACE-TO-MARK transitions. For some other character combination a transition may occur between "start" and intelligence units, but in any transmitted character there can be only 2, 4, or 6 transitions.

3-4.4.5 Unit Lengths

As previously discussed, the first 6 units of the signal are the same length, but the 7th (stop) unit is longer. Each of the first 6 units requires 13.5 milliseconds of the circuit time for transmission. This timing is based on a transmission speed of 100 words per minute. The stop unit requires 19.0 milliseconds. If a value of 1 is assigned to each of the first 6 units, then the stop unit has a value of 1.42. Thus, the total number of units in letter R (for example) is 7.42, requiring a transmission time of 100 milliseconds. No allowance is made for transition time, because a transition has negligible time duration. (See Figure 3-43).

3-4.4.6 Transmitter Contacts

Transmitter contacts are actually a set of mechanically-controlled switches that can produce a different combination of 7.42 unit signal for any letter or function lever depressed. The selector magnet of the receiving teletypewriter mechanically releases a start lever when the start pulse is received, thus allowing the selector cam clutch to rotate through one revolution. During this revolution, five selector levers in the selector unit are positioned by the operation or release (MARKING or SPACING) of the selector magnet armature as determined by each intelligence pulse received. The time required to position each selector lever is approximately 20 percent of the time of one intelligence pulse, or 2.7 milliseconds. This time, again, is based on a teletypewriter running at 100 WPM. Cams on the selector cam-clutch are so located that the time between each selector lever operation is fixed at 13.5 milliseconds.

FIGURES 1	↑	⊕	o	↗	3	→	↘	↓	8	↙	←	↖	·	⦶	9	∅	1	4	BELL	5	7	●	2	/	6	+	—	LETTERS	FIGURES	SPACE	CARRIAGE RETURN	LINE FEED
FIGURES 2	—	5/8	1/8	$	3	1/4	&		8		1/2	3/4	,	7/8	9	0	1	4	BELL	5	7	3/8	2	/	6	"						
FIGURES 3	—	?	:	$	3	!	&	#	8		(	)	.	,	9	0	1	4	BELL	5	7	;	2	/	6	"	//	▼	▲	□	<	≡
LETTERS	A	B	C	D	E	F	G	H	I	J	K	L	M	N	O	P	Q	R	S	T	U	V	W	X	Y	Z	SYMBOLS ABOVE ON TAPE ONLY					
	1	1		1	1	1				1	1						1		1		1		1	1	1	1		1	1			
	2		2				2		2	2	2	2				2	2	2			2	2	2					2	2			2
			3			3		3	3		3		3	3		3	3		3		3	3		3	3			3		3		
		4	4	4		4	4			4	4		4	4	4			4				4		4				4	4		4	
		5					5	5				5	5		5	5	5			5		5	5	5	5	5		5	5			

1 WEATHER 2 FRACTIONS 3 COMMUNICATIONS

Figure 3-42. Five-Level Code Chart

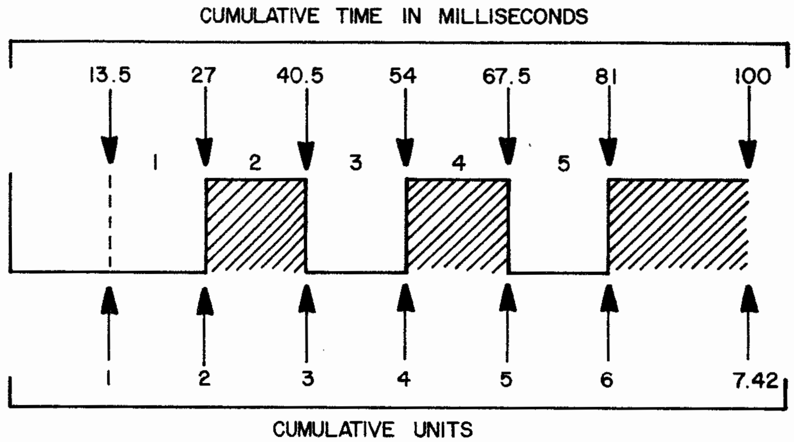


Figure 3-43. The 7.42-Unit Teletype Signal

3-4.4.7 Lever Positioning

During 2.7 milliseconds of the first pulse the first selector lever is positioned; during 2.7 milliseconds of the second pulse the second selector lever is positoned, and so forth until all 5 selector levers are positioned. (See Figure 3-44). These selector levers control the internal mechanism of the teletypewriter so as to select and at the proper time print the correct character.

3-4.4.8 Transmission Speed Factors

In the start-stop mode, the receiving machine is allowed to run for only one character and is then stopped to await the reception of a start signal indicating that the next character is about to start. In this manner any difference in speed between the transmitting and receiving machines can accumulate only during the duration of one character. There is a penalty to pay for this advantage. The length of each character must be increased to include an element to start the receiving machine and another added to stop it. Also, the receiver must be slightly faster than the transmitter, to ensure that the receiver will complete its cycle and be ready to receive another character before the transmitter sends it. Thus, speed of transmission and operating margins are sacrificed for synchronization.

3-4.5 SYNCHRONOUS MODE

Synchronous teletypewriter operation, as opposed to start-stop operation, does not in all cases have to rely upon elements of the transmitter character to maintain proper position in relation to the receiving device. External timing signals may be used, allowing the start and stop elements to be discarded. Then only the elements necessary to convey a character (and in some cases, a reference element) need to be transmitted. This results in more information being passed in a given period of time and all elements are of equal length. Synchronous operation is employed in AUTODIN Digital Subscriber Terminal Equipments (DSTEs) and in most forms of digital data transmission other than teletypewriter. Due to its current limited use in ships, it will not be discussed here in detail.

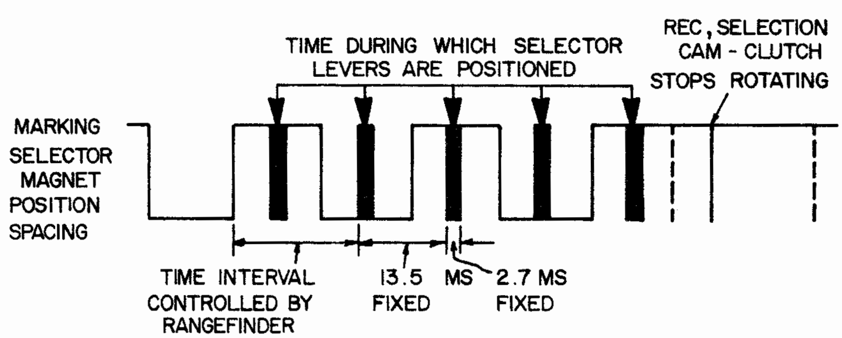


Figure 3-44. Selecting Intervals for Letter Y

3-4.6 BIAS DISTORTION

Bias distortion is the uniform lengthening or shortening of the MARK or SPACE elements, one at the expense of the other. This process means that the total time for one MARK plus one SPACE never changes; only the length of the MARK or SPACE element changes. Bias distortion may be caused by maladjusted teletypewriter line relays, detuned receivers, or a drift in frequency of either the transmitter or receiver. Throughout the overall teletypewriter circuit there are many things that could cause bias distortion. (See Figure 3-45 B and C as contrasted with an unimpared signal, A.)

3-4.6.1 Relays

Maladjusted relays can cause bias distortion by the armature's remaining longer on one contact than it does on the other. Bias distortion can also be caused by an improper balance between line and bias currents in a neutral relay.

3-4.6.2 FSK-Caused Distortion

FSK (Frequency Shift Keyed) equipment will cause bias distortion when the time required to change from MARK-to-SPACE and the time required to change from SPACE-to-MARK are not identical, even though the MARK and SPACE frequencies are correct. This fault can usually be eliminated by proper setting of the center frequency while transmitting unbiased reversals (alternate MARKS and SPACES of equal length).

3-4.6.3 Receiver Tuning

In certain FSK equipment, improper receiver tuning or transmitter drifting will cause bias distortion. This equipment uses some type of balanced FM (Frequency Modulated) discriminator. When the MARK and SPACE frequencies are equally centered on the discriminator response slope, and equipment is properly tuned and aligned, the result is a signal without bias.

3-4.6.4 Monitor Distortion

Bias distortion may be introduced into dc circuits by monitor printers inserted for checking purposes in series with the circuit. This is the case on neutral circuits where the added monitor printer inductance changes the shape of the waveform. The use of monitor printers that are not equipped with line relays to isolate the selector magnet will cause more bias distortion than those equipped with relays. High impedance shunt monitor devices should be used where possible, to reduce the effects normally encountered with series monitoring.

3-4.6.5 Speed Distortion.

Differences in speed between the transmitting and receiving devices will appear as signal distortion to the receiver. An analyzing device capable of sensing each signal element will show a proportional increase in distortion with each element sampled if the speed error is in the transmitter.

3-4.7 FORTUITOUS DISTORTION

Fortuitous distortion is the random displacement, splitting, or breaking up of the MARK and SPACE elements. It is caused by such things as cross-talk interference between circuits, atmospheric noise, powerline induction, poorly soldered connections, lightning storms, and dirty keying contacts. (See Figure 3-45F). On radio circuits, fortuitous distortion is caused by noise, interference, and multipath conditions. The results vary from an occasional hit to a complete disruption of the signal. Interchannel or adjacent channel interference on multichannel radio or carrier equipment may cause fortuitous distortion. Any minor breaks in signal elements are properly called fortuitous distortion, but the duration or length of such breaks may be so short as to go unnoticed except in some sensitive distortion measuring devices. If a break of 1 millisecond occurs at the instant of sampling, and the receiving device has sampling interval of 4 milliseconds, then the loss will not be noticed. If the same conditions exist, but the sampling interval is only one-half millisecond, then the element may be lost completely. To properly interpret the effect of fortuitious distortion on a traffic circuit, extensive knowledge of the receiving devices and distortion-measuring equipment characteristics is required.

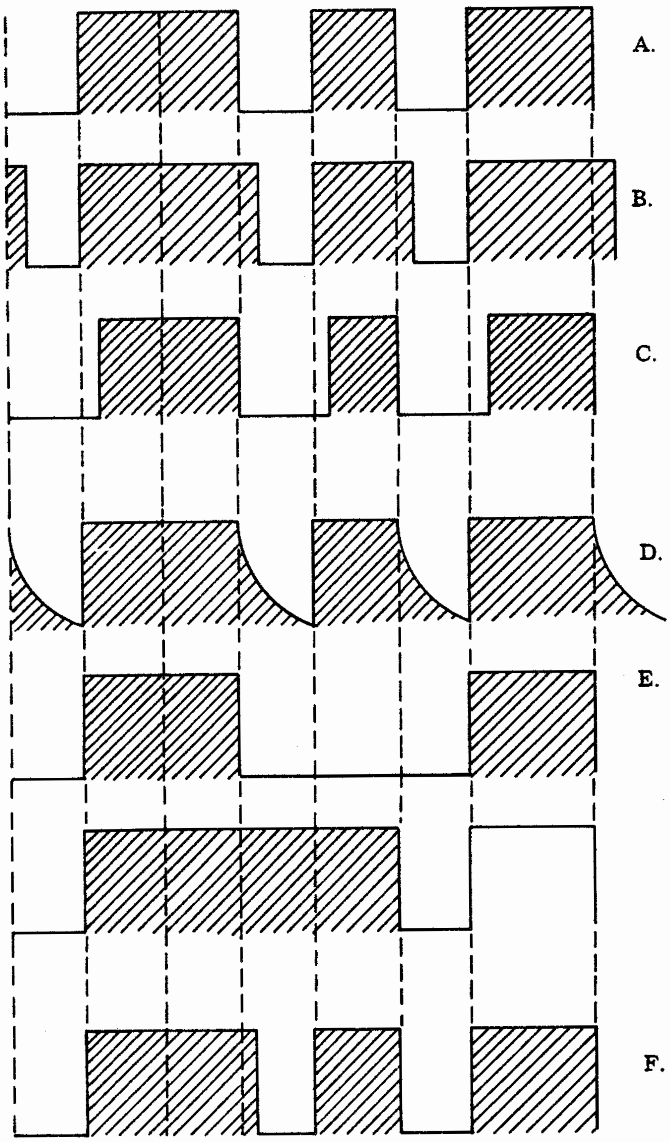
3-4.8 END DISTORTION

End distortion is the uniform displacement of MARK-to-SPACE signal transitions with no significant effect on SPACE-to-MARK transitions. It is caused by the combination of resistance, inductance, and capacitance in the circuit. (See Figure 45D).

3-4.9 CHARACTERISTIC DISTORTION

Characteristic distortion is a repetitive displacement of disruption peculiar to specific portions of the signal. It normally is caused by mechanical maladjustment or by broken parts of the sending equipment, and differs from fortuitous distortion in that it is repetitive instead of random. An example would be the repeated splitting of the third code element of a teletypewriter signal, resulting in the transmission of a MARK vice a SPACE. (See Figure 3-45E).

TELETYPEWRITER DISTORTION



A. Perfect signal, 7.42 unit interval, 74.2 BAUD (character "J")

B. Marking Bias - approx. 25%

C. Spacing Bias - approx. 25%

D. End distortion

E. (1) Characteristic Distortion. 4th intelligence unit stays on space. Effect is to make letter "J" print as letter "A".

(2) Characteristic Distortion. 3rd intelligence unit stays on a mark. Effect is to make letter "J" print letter "K".

F. Fortuitous Distortion.

Figure 3-45. Teletypewriter Distortion

3-4.10 DISTORTION MEASUREMENTS

Correct analysis of received teletype signals depends on the skill of the individual, understanding of the nature of the receiving devices, and the capabilities of distortion-measuring devices used. The two modes of transmission, start-stop and synchronous, use different methods of establishing a reference point from which to obtain a sampling interval. Therefore, distortion-measuring equipment should be capable of "sensing" the received signal in the same manner as the receiving device used on the circuit. The term "maximum distortion", used in a table of allowable distortion, is a percentage calculated from the ratio of the difference between an observed unit interval and an ideal unit interval. For example, if the correct unit interval is 22 milliseconds, and the maximum displacement, early or late, of a transition is 8 milliseconds, the ratio of maximum distortion is 8:22, or approximately 36 percent. The degree of distortion present on start-stop signals is the same as the maximum distortion explained above except that the timing (which establishes the relationship or ratio between the observed signal and the ideal signal) is based upon the beginning of the start element in the signal being measured. The ideal signal timing is established internally within the distortion-measuring device, and is triggered by the beginning of the start impulse. Distortion measurements of synchronous signals can be made with distortion measuring devices set to the start-stop mode. This method of measurement uses the mark-to-space transition as the reference point from which to establish an ideal and compare the deviation of transitions. It also counts a block of elements equivalent to the unit code setting of the measuring device. The next mark-to-space transition which occurs is used as the reference point for the next block of elements.

3-4.11 DC MEASUREMENTS

The following paragraphs describe the test procedures for the various performance-monitoring functions associated with Quality Monitoring of Communications Systems. Each subsection outlines the purpose of the test, equipment required, test procedures, test results, and the necessary tables and pictures for comparing the actual results.

3-4.11.1 Loop Current Measurements

Loop current measurement tests are designed to minimize the effects of distortion, feedover, and other undesired effects. Low loop-current will cause distortion. Excessive loop current may burn out switching transistors, shorten the life of other electronic components within the circuit, and cause undesired feedover.

3.4.11.1.1 Equipment Requirements

No special test equipment is required for this test other than utilizing the equipment's built-in current meters (front panel meters of converters, keyers, dc patching panels, power supplies, etc.) for measurement of dc loop currents.

3-4.11.1.2 Connections

Loop current tests can be performed at the RED Patch Panel (SB-1210)and the BLACK Patch Panel (SB-1203). Each patch panel provides six channels, each comprising a circuit of two or three looping jacks, at least one set jack, and a rheostat for adjusting loop current. Loop current is monitored by a milliammeter which can be switched to any one channel by a rotary selector switch. When loop current is not being adjusted, this switch should be in the 0 (zero) position to avoid damage to the meter movement.

3-4.11.1.3 Control Settings

None required.

3-4.11.1.4 Test Procedures

The method of checking loop current requires that a MARK be placed on the loop being tested. When this is accomplished, rotate the meter switch to the position corresponding to the loop under test. Adjust the RHEOSTAT for that loop to 60 ma ± 1 ma on the meter. Loop current is now properly set. Loop current should be adjusted each time equipment is added, deleted, or substituted. When patching is necessary on these patch panels, always patch "SET to LOOP" to eliminate shock hazards.

3-4.11.1.5 Test Results

If observations indicate operation within the recommended standards, quality has been assured within the system. Operations not meeting these standards must be corrected before making further checks.

3-4.11.2 Low Level Loop Voltage Measurement

Low Level Loop Voltage Measurement tests are designed to aid in minimizing the effects of distortion and other undesired effects.

3-4.11.2.1 Equipment Required.

No special test equipment is required for this test other than utilizing the quipment meters (front panel meters of converters, keyers, dc patch panels, power supplies, etc.) for measurement of dc loop voltage.

3-4.11.2.2 Connections

None required.

3-4.11.2.3 Control Settings

None required.

3-4.11.2.4 Test Procedure (General)

Loop Voltage measurements can be made at the front panel voltmeter on the Low Level patch panel. Both MARK and SPACE voltages can be measured by using the panel voltmeter.

3-4.11.2.5 Test Results

If measurements indicate operation is within the recommended standards, quality has been assured within the system. Operations not meeting these standards must be corrected before making further checks.

3-4.11.3 DC Distortion Measurements

The test procedures described herein provide operators with a primary and alternate method of checking dc distortion. An increase in dc distortion frequently warns the operator of possible radio path deterioration or equipment failure, and greatly aids in determining what corrective action is required.

3-4.11.3.1 Equipment Requirements

TS-2446/UG Distortion Analyzer or equivalent.

3-4.11.3.2 Connections

LOW Z INPUT - Insert one end of patch-cord into the low Z input and the other end into loop being checked. CAUTION: ALWAYS plug patchcord into the TS-2446/UG before patching into looping jack to prevent electrical shock to operator or a possible blown fuse.

3-4.11.3.3 Control Settings

1. CAL CHECK - With power switch "ON", turn CAL SWITCH to ON position and set meter on centerscaled red line by means of CAL-ADJUST SWITCH. Turn CAL SWITCH off.

2. START-STOP/SYNC SWITCH - Set switch to START-STOP.

3. AUTO-RESET SWITCH - Set to ON position.

4. MA SELECTOR

a. 100 ma - Check line current for 60 ma.

b. Turn ma meter switch to "OFF" position after 60 ma adjusted.

5. SPEED/BAUD SWITCH - Set to correct position as indicated in the chart below.

TYPE EQUIP	RED BAUD SPEED	BLACK BAUD SPEED
TSEC/KW-37	75.0	75.0
TSEC/KG-14	75.0	75.0
TSEC/KW-7	75.0 (RECV)	75.0
	74.2 (SEND) *	75.0
TTY EQUIP	74.2 (SEND) *	

*NOTE: Most commonly used. 75 BAUD may also be encountered.

6. DISTORTION SELECT - Average Bias.

7. TRANSITION MARKERS - OFF.

8. INPUT FILTER - OFF.

9. INPUT SELECT - Set according to dc loop being checked (60 ma neutral, 20 ma neutral, 20/20 ma polar.)

10. LOW Z INPUT POLARITY - Set to polarity which causes signal indicator light to illuminate.

11. TRANSITION SELECTOR - Set to "ALL" position.

NOTE: Low level circuitry cannot be measured with a TS-2446/UG without prior modification.

3-4.11.3.4 Test Procedures

1. In order to maintain optimum performance of the equipment, average bias should never exceed the parameters outlined in Recommended Standards. Both the MARK/SPACE indicator lights and the percentage of distortion meter provide the operator with a firm indication of circuit conditions. If the MARK light is predominantly lit, the signal is being distorted by marking bias. If the SPACE light is predominantly lit, the signal is being distorted by spacing bias. In either case, the fault is usually caused by maladjusted equipment or equipment failure. In those cases where the bias exceeds the parameters in Recommended Standards and the MARK/SPACE lights are being alternately lit without one or the other being predominant, a deterioration in radio path is usually indicated. If so, operators should consider shifting to a more suitable frequency.

3-4.11.3.5 Test Results

1. If observations indicate operation is within the recommended standards, quality has been assured within the system. Operations not meeting these standards must be corrected before making further checks.

3-4.11.4 DC Distortion Measurements - Method

In the event of failure of the TS-2446/UG Distortion Analyzer or equivalent, an alternate method of measuring DC distortion is available and is described below.

3-4.11.4.1 Equipment Requirements

AN/USM-117 Oscilloscope or equivalent.

3-4.11.4.2 Connections

Fabricate a modified patchcord as shown in Figure 3-46 and connect as shown.

3-4.11.4.3 Control Settings

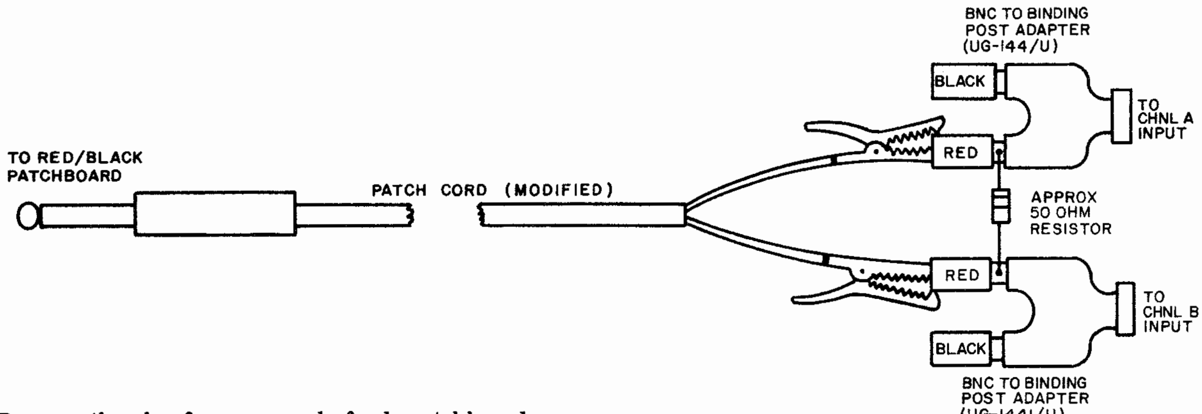
1. Adjust INTENSITY control to obtain a visible trace.

2. Set the SWEEP STABILITY control to "trigger".

3. Set the channel A and B polarity selector switches to +DC.

4. Set the INPUT SELECTOR switch to the "added" position when utilizing the AN/USM-117, and to the "A-B" position when using the AN/USM-281 oscilloscope.

5. Set the CHANNEL switch on both channel A and B to 1 volt per cm (IV/cm).



1. Remove the plug from one end of a dc patchboard.

2. Solder an alligator clip to each of the leads removed from the plug.

3. Connect an alligator clip to the red post of the BNC to binding post adaptor as shown above.

4. Connect a 47 to 50-Ohm resistor to the two red binding posts (the signal is then developed across the resistor).

Figure 3-46. Test Configuration for DC Distortion Tests

6. Adjust the channel A VOLT/DIV variable switch (concentric red knob) until the signal expands to approximately 1 inch vertical height.

7. Set the TRIGGER SLOPE LEVEL switch to int ±.

8. Set the TIME/DIV switch to 0.5 milliseconds position.

9. Adjust the TIME/DIV VARIABLE control until a crossover pattern is displayed on the scope. (See Figures 3-47 & 3-48).

NOTE: Ignore the large positive spike present in circuits using KG-14. It will not deteriorate the crossover pattern.

10. Adjust the SWEEP STABILITY control until the presentation remains stationary on the scope.

3-4.11.4.4 Test Procedures.

1. A bias-free signal or a signal with very low bias will resemble that shown in Figure 3-47. A signal containing bias or fortuitous distortion will resemble the Figure 3-48. Signals of this type often provide an indication that the incoming radio path is deteriorating, or that equipment in a given channel is failing and maintenance is required.

3-4.11.4.5 Test Results

1. If observations indicate operation is within the recommended standards, quality has been assured within the system. Operations not meeting these standards must be corrected before making further checks.

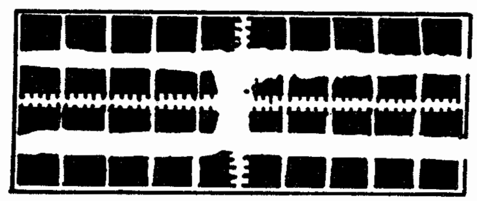


Figure 3-47. Crossover Pattern Seen on a Bias-Free Signal

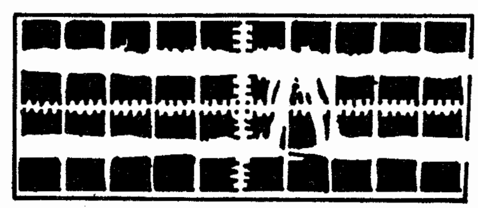


Figure 3-48. Crossover Pattern Seen on Active Traffic Channel with Approximately 25% Distortion

3-5 INFRARED EQUIPMENT AND TESTS

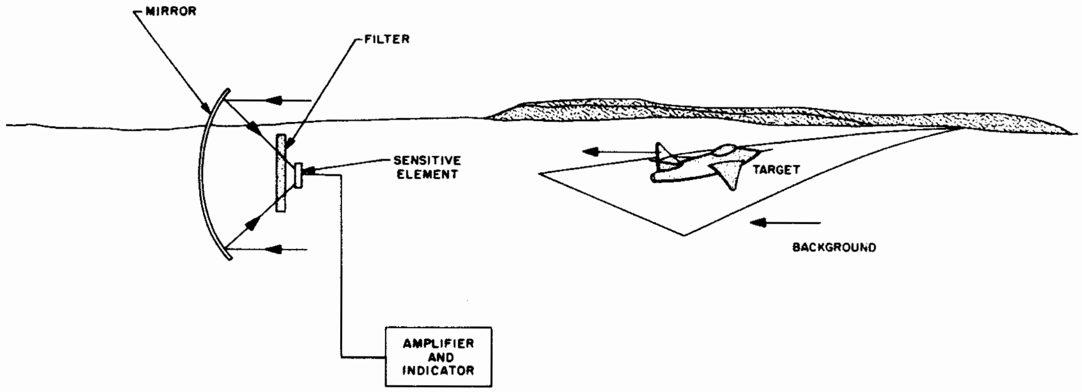
Infrared equipment and tests between visible light and microwaves encompass the band of radiation called infrared. Infrared radiations extend from the limit of the visible red region of the electromagnetic spectrum to the upper edge of the radio microwave region. Since infrared waves are invisible to the human eye, specially designed equipment is necessary to detect these extremely useful radiations and to provide a visible presentation for their analysis. Infrared radiations have characteristics similar to those of visible light rays; they can be reflected by mirrors and refracted by lenses. In addition, these radiations can be transmitted through substances such as silicon and germanium, which are opaque to visible light. Therefore, they can also be compared to RF radiations. Infrared equipment is designed to create, control or detect invisible infrared radiation. The infrared transmitter (source) equipment is designed to produce and direct the radiations, whereas receivers are designed to detect and convert the radiations into either visible light, for viewing purposes, or into voice or code signals, for audible presentation. Infrared devices can be used for weapon guidance, detection of enemy equipment and personnel, navigation, recognition, aircraft proximity warning, and communications. Depending upon the application, this equipment is either passive or active. The active method employs both infrared transmitter and receiver equipment, whereas the passive method requires only receiver equipment. Figure 3-49 shows the basic components required for both active and passive methods. If the human eyes were sensitive to infrared rather than to visible radiation, the eye would perceive a very different visual world. Objects are normally seen by the light they reflect, because most substances emit little, if any, light of their own. In the infrared region, the situation is almost reversed; most objects reflect very little but emit copiously. The infrared world is a radiant world, a world almost without shadow. Warmer objects would appear brighter to the eye, and the band of "colors" visible to the eye might be much wider. Yet, although purists maintain a distinction between infrared and visible light, the differences are mainly biological. Infrared radiation is sensed as heat rather than light. To those working in optical technology, both frequency bands are, along with the ultraviolet, part of the optical spectrum. Infrared equipment uses lenses, mirrors, and prisms, and, like visible light, infrared radiation can be detected by its photo-effects. The infrared spectrum was discovered in 1800 by the elder Herschel, Sir Frederick William, who used a thermometer to detect the energy beyond the end of a visible prismatic spectrum. In 1861, Richard Bunsen and Gustav Kirchhoff firmly established the underlying principles for infrared spectroscopy. After a century of steady advances, the utility of infrared technology is no longer confined to spectroscopy, nor is it used just for devices for night vision and rifle sighting, or for infrared heating. There is developing a fascinating variety of applications of electromagnetic radiation constituting the spectrum from about 1 micron to about 1 mm - the three decades between visible light and very-high-frequency radars shown in Figure 3-50. The list of applications is long, ranging from missile defense and fire detection to meteorology and medical diagnostics. However varied the applications, infrared systems can be considered in four basic units: an infrared source, an optical system, a radiation detector, and an arrangement for presenting the data. Passive systems are used in most applications; the object of interest is in this case the source, which is seen by its self-emisssion. In active systems, a controlled source is used to illuminate the object under observation.

3-5.1 INFRARED RADIATION THEORY

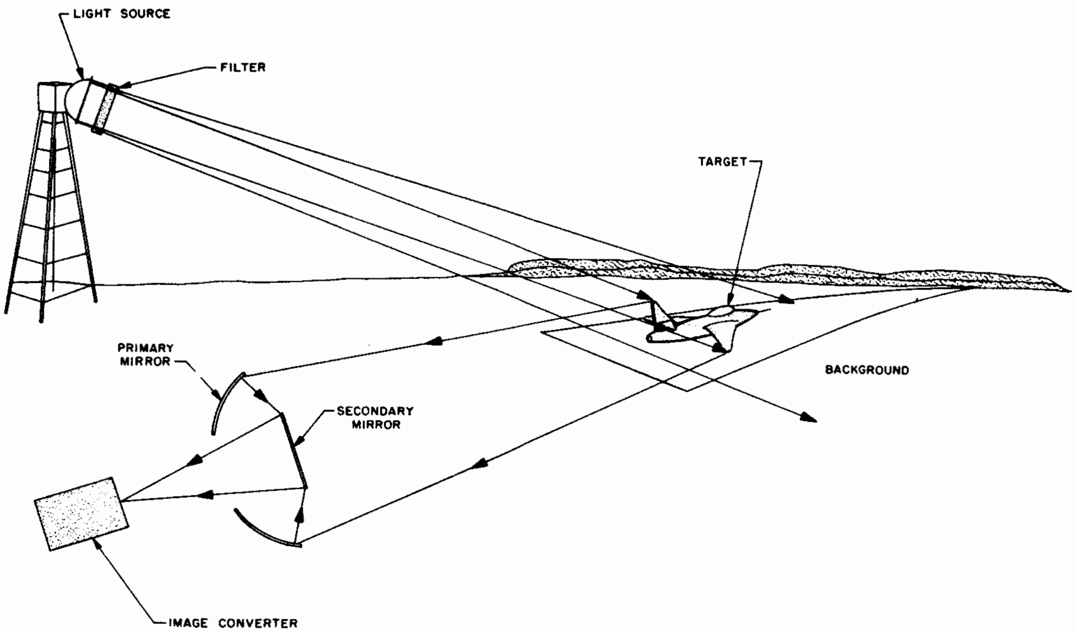
All substances emit electromagnetic energy. As stated, most objects on earth radiate abundantly in the infrared portion of the spectrum. The simplest emitters to describe are theoretically perfect radiators, or black bodies which emit radiation with a spectral distribution shown in the marginal curves. Max Planck formulated the complicated law describing this distribution back at the turn of the century.

3-5.2 INFRARED SOURCES AND ATMOSPHERIC EFFECTS

The intensity and spectral distribution of radiation emitted from an infrared source are determined by its physical characteristics and by its temperature. This radiation usually passes through some medium, like the atmosphere, and the radiation impinging on any receiving device is thus a function of both the emission characteristics of the source and the absorption (or transmission) characteristics of the medium. Figure 3-51 illustrates the overall absorption spectrum of the atmosphere, made up of the superimposed spectra of all the atmospheric constituents and, as the blownup section from 3.4 to 3.5 microns shows, the spectrum is actually quite complex. The complete transmission curve has comparable structure. In addition, the atmosphere is not constant; it changes with season, altitude, time of day, viewing angle, etc. A calculation of transmission over any path must take all of these factors into account. For most system designs, the engineer requires a knowledge of the quantity and spectral distribution of radiation arriving at the collecting aperture (entrance pupil) or "dish" of the instrument. This spectral irradiance $\underline{H}\lambda$ is usually specified



A
PASSIVE INFRARED METHOD



B
ACTIVE INFRARED METHOD

Figure 3-49. Basic Components of an Infrared Facility

as the number of watts per square centimeter per micron. For some problems it is known with great accuracy; for others an uncertainty of 200% is not unusual. If this figure seems surprising, consider the nature of some sources, particularly exhaust gases from rocket plumes. Energy is emitted in a series of narrow spectral lines, bounded by a black-body envelope; the atmosphere operates like a comb filter

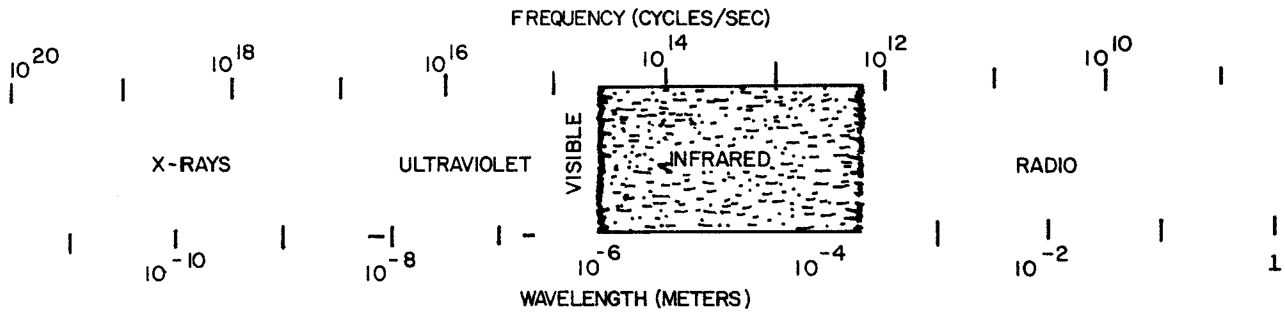
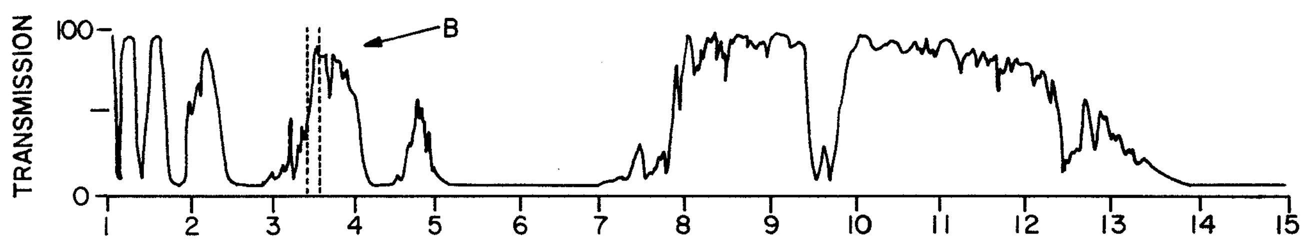
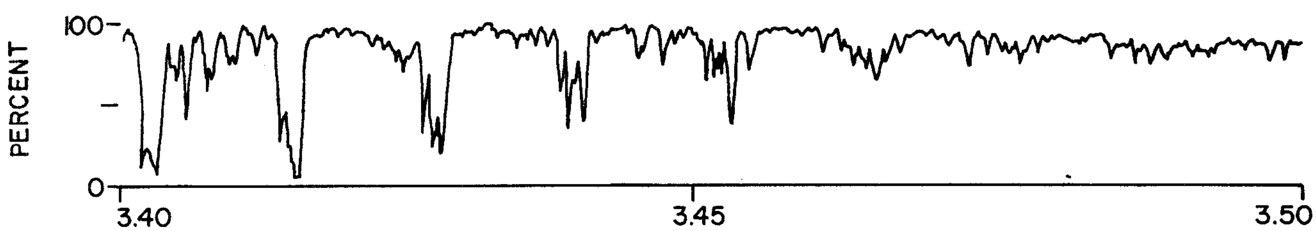


Figure 3-50. Infrared Band



Wavelength (microns)
A 1-15 Microns



Wavelength (microns)
B Section of A

Figure 3-51. The Atmospheric Transmission Spectra

on these lines, absorbing some frequencies, transmitting others. In addition, the atmosphere reradiates at the frequency of the absorbed power. Thus, the quantity and spectral distribution of the collected radiation is a complex combination of the filtered target radiation and the self-emitted atmospheric radiation.

3-5.3 SOURCES FOR ACTIVE SYSTEMS

Passive infrared systems depend upon sensing the self-emission of the objects of interest. Conversely, active systems depend upon sensing source radiation that is reflected by the object of interest. Consequently, the source problem is quite different. The nature of the source is known and controlled; the designer is plagued rather by a lack of good sources. The need for sources is acute because, although active systems sacrifice the advantages of secrecy so important to the military, they can perform functions for which passive systems are just not naturally suited, particularly ranging and communication. Active systems have been restricted in range by power limitation because sources have not had sufficient spectral and spatial power densities. For example, only a small amount of the source power lies in the spectral region of interest. Sources must be relatively large to generate appreciable amounts of power, must be modulated externally, and are presently incoherent so that no use can be made of the information contained in the shape and phase of the waveform. Laser sources may change the situation radically. Infrared laser beams share the advantages of visible coherent light for use as an information carrier for communication, ranging and navigation systems. Infrared lasers provide good sources for night-vision devices because of the high spectral and spatial densities.

3-5.4 OPTICS FOR INFRARED SYSTEM

Whereas the detector is often called the heart of the infrared system, the optical device must be considered analogous to the human eye. It collects the radiation from some region of space and then filters and focuses it onto the surface of the detector. Optical systems for infrared use are, for the most part, telescopes with resolutions of a fraction of a milliradian, and must often perform well over the broad infrared spectrum from 1 to 15 microns. In size, the collecting mirrors or lenses range from about one inch in diameter, for small laboratory equipments and some missile systems, to nearly a yard across for long-range detection and astronomical applications. Another important optical element is the spectral filter. A given detector will usually be sensitive over a broader wavelength band than that in which the target radiates. Hence, background can be cut down considerably by filtering out radiation from all regions except the one of interest. Although there exists a wide variety of filter types for use in the infrared part of the spectrum, the simple absorption filter and the all-dielectric interference filter presently reign supreme. The first of these is conceptually simple: every material is transparent in parts of the electromagnetic spectrum and opaque elsewhere. By selecting materials properly, only the desired band of wavelengths can be allowed into an instrument. Figure 3-52 indicates the regions of transparency for a wide variety of optical materials. Cutoff wavelengths are those at which a 2-mm thick sample has 10% transmission. Those marked with asterisks have less than 10% transmission throughout. The all-dielectric filter, currently the most versatile and useful filter, is based upon the principle of interference rather than absorption. The transmission of optical elements can be improved by applying on their surface a thin coating of transparent material; the coating thickness and refractive index are chosen so that the waves reflected from

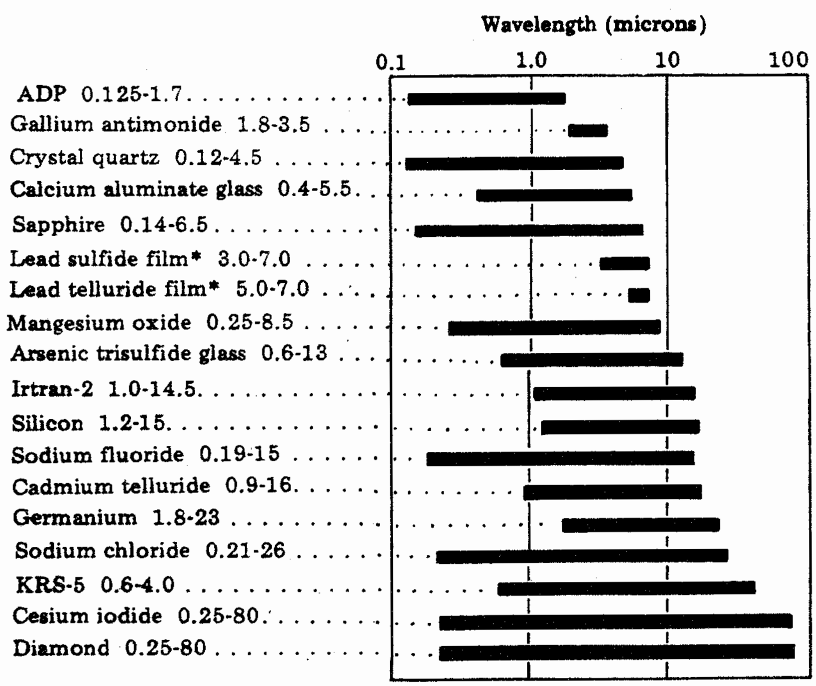


Figure 3-52. Transparent Region of Various Materials

the second surface interfere destructively with those from the first. The amplitude of the transmitted wave is larger. Because an arrangement like this can only be satisfied for a restricted, but not sharply defined, band of wavelengths, such a thin film is a filter, although a very unsatisfactory one. However, by combining a series or stack of such layers together in different arrangements, a great variety of filter characteristics can be obtained.

3-5.5 PROCESSING INCOMING RADIATION

To process incoming radiation, the reticle, a small, patterned rotating disk comprised of sections which are completely transparent, completely opaque, or partially transparent, is often placed in a primary or secondary focus of the optical system. This remarkable device can be less than 1 cm in diameter, yet can perform as many as three different operations simultaneously. The first of these functions is the easiest to understand, were the reticle not there. The radiation would be focused onto the detector at all times, and a steady source signal would generate a steady DC signal. With the rotating reticle in position, however, this steady light intensity is interrupted and a more manageable AC signal is generated. In addition, particular types of modulation are accomplished as desired by the size and shape of the bars of the reticle. A target spot can generally be located circumferentially by measuring the time lag between a reference pulse and the first pulse received from a point target. The reticle with nonradial spokes can provide radial position information by the pulse width. Many other coding schemes are possible. These two bits of information can provide the two angles of a spherical coordinate system that are necessary for pointing. The third function is independent of the first two, but naturally the reticle design is not. Most military targets (jet exhausts and rocket plumes) and many commercial ones are small, while backgrounds like clouds, lakes, and terrain are relatively large. Therefore, if the spacings of the reticle are made small, the output waveform for a target remains well modulated, while that of a background becomes smoothed. In this way, large objects can be discriminated from small ones.

3-5.6 INFRARED BANDS

The devices used in the near and middle infrared bands are employed for ranging, recognition, and communications. The usable distance range of near infrared equipment is normally between 6.5 and 10 miles. Equipment which operates in the far infrared band is used for ranging, missile guidance, and the detection and location of personnel, tanks, aircraft, and ships. The usable distance range is between 100 yards and 12 miles. The bulk of target radiation is at the lower wavelengths, but present infrared detectors and optical components do not operate efficiently in the far infrared bands. Personnel must avoid being exposed to the beam of an infrared transmitter. Usually, the danger is not great because the heating effects of the beam will be felt before damage occurs. Unfortunately, however, the eyes can be damaged before the physical heating effect of the radiation provides sufficient warning. Therefore, particular caution should be observed so as not to look into a source of intense infrared radiation. Actually, infrared radiation does not consist of heat waves, but is the result of molecular agitation on the surface of a heated object. The greater the heat of an object, the more the radiation emitted. These radiations are converted into thermal energy whenever they strike an object. From this discussion you would assume that the infrared energy emitted from an object is a direct function of the absolute temperature of that object. However, actual radiation depends on the smoothness of the radiation surface, as well as the surface heat. The smoother the object surface finish, the less the radiated energy provided for a given temperature. The emitted infrared energy is attenuated and scattered over some portions of its spectrum, and not attenuated very much nor scattered over other portions of its spectrum. The atmosphere filters some parts of the infrared spectrum and permits other parts to pass freely. Those parts of the infrared spectrum which are not affected by the atmosphere are termed "windows." These windows are more narrow, and, consequently, more pronounced at sea level; they become broader as height is obtained until, at approximately 30,000 feet above sea level, only two narrow attentuation hands remain. All bodies, either hot or cold (above absolute zero) radiate some heat. Even ice, at its melting point, is approximately 273 degrees Kelvin above absolute zero, and will radiate a certain amount of heat. Much of this radiation is in the form of infrared, and can be detected by infrared devices. A typical experiment can be performed in the field to measure the infrared with an ammeter, as is illustrated in Figure 3-53. As shown, a prism is used to separate the beam of light into specific wavelengths. Normally, this separation is accomplished by means of a spectroscope or spectrometer. If you place a camera on the spectroscope or spectrometer to obtain an image of the spectrum on photographic film, you have an instrument called a spectrograph, and the photographic record is called a spectrogram.

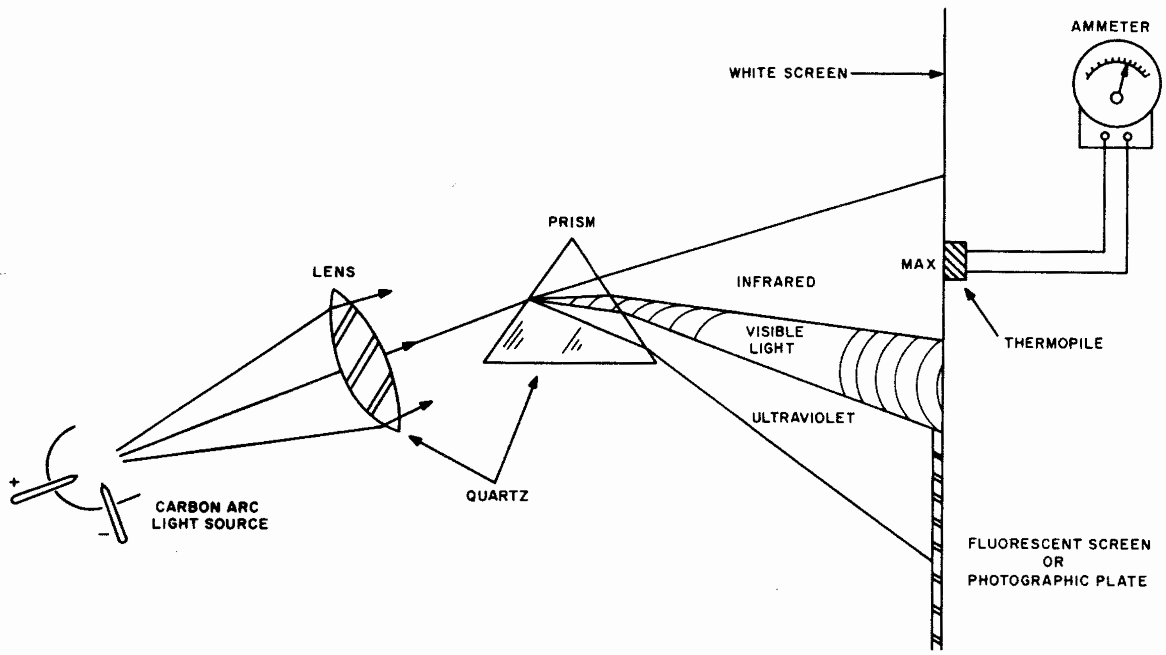


Figure 3-53. Spectrum Analysis Experiment

3-5.7 NANCY GEAR

The sending and receiving of messages or information by means of infrared energy in the Navy has been assigned the program name NANCY. By way of definition, NANCY gear is Navy equipment which creates, controls, or detects invisible infrared radiation for various military purposes. NANCY sources are devices to produce and direct infrared radiation; NANCY receivers are instruments for detecting infrared radiation and for converting it into light that can be seen or into voice or code signals that can be heard. Although research and development is being conducted in all areas of the electromagnetic spectrum, most Navy infrared equipment operates in the near infrared region, from 0.8 to 1.5 microns in wavelength (10,000 microns in one centimeter). This region is used because it is relatively easy to produce infrared radiation since it is present in quantity in the light emitted by ordinary incandescent electric lamps. Consequently, a good source (or transmitter) can be had just by filtering out all the visible light from standard Navy signaling lights already in use. Wherever possible, NANCY sources now in fleet use are adaptations of lights already in use by the Navy, and require little more than the addition of a NANCY filter to screen out all undesired light waves. Unfortunately, no NANCY receiver yet developed is as sensitive to infrared radiation as the human eye is to visible light. Consequently, for NANCY broadcast signaling, it has been necessary to devise yardarm equipment that is much more powerful than the blinkers used for visual signaling. At present there are NANCY sources in fleet use that fulfill nearly all the purposes to which visible signaling lights are put. The primary system employed, however, has been narrow beam directional signaling, using the standard Navy 12-in. searchlight with NANCY filter and hood, plus an image-forming receiver. With equipment in good order, and in the hands of trained personnel, this system is capable of sending messages between ships at ordinary daytime signaling speeds and up to horizon distance, even during blackout conditions or conditions of radio silence. NANCY image-forming receivers are relatively small, portable, "telescope-like" optical devices weighing from one to about 40 pounds. Their purpose is to convert invisible infrared radiation into visible images of the source at the receiving end of a "system." Non-image-forming receivers consist of larger, more powerful equipment, permanently installed, and used for code and voice communications.

3-5.8 INFRARED RECEIVERS

Infrared receivers are separated into two distinct groups. The phosphor button type is the simplest, containing a small disk composed of infrared radiation into visible light. The second group, termed the electronic type, contains an image tube rather than a phosphor button to convert the radiation into light.

3-5.8.1 Detectors

The detector is the fundamental sensing device of infrared equipment, and, as such, plays a critical role in infrared equipment operation. The detector generally functions as a transducer and converts radiant energy into an electrical signal. This electrical signal may be used in other communications-electronics devices, or for observation purposes. Detectors may be separated into two primary classes, depending on their working principle of operation. The two basic types are thermal detectors and photo detectors. Thermal detectors are hot wavelength-sensitive, but respond to radiation throughout the infrared spectrum. In terms of image-forming detectors, photo detectors vary with the frequency or wavelength of the incident radiation. The terminology of infrared detectors is listed for ready references:

1. Sensitivity. This term is a measure of the minimum detectable part of the quantity of infrared radiation.
2. Response. This term is a quantitative expression of output versus input.
3. Spectral Sensitivity. This term refers to the operating range of the detector, and is measured in terms of the band of the infrared spectrum to which the detector will respond. Measured in microns, it is expressed in terms of wavelength.
4. Minimum Detectable Power. The limit of usefulness of a detector is the minimum amount of radiant power expressed, in watts, to which it will respond. This minimum is that point where the detector output voltage is no greater than the internal noise voltage of the detector. Actually, the minimum detectable power is the reciprocal of the sensitivity of a given detector.
5. Time Constant. This is a measure of the time required for the detector to respond to a given radiation.

3-5.8.2 Limitations

The limitations imposed on infrared detectors are severe in that the function of a detector is critical. The ideal detector should have a broad spectral response, a short response time, low minimum detectable power, a high output/input ratio, and the ability to discriminate between targets and background noise. Unfortunately, present-day detectors cannot cope with all of these requirements. Thermal detectors have broad spectral response, but they have a long time-constant and low sensitivity. Photo detectors have a short time-constant and are approximately a thousand times more sensitive than thermal detectors, but their spectral response is limited to the near-infrared region. Noise constitutes one of the greatest limitations to both detector performances, because the noise component restricts the interpretation of target data from a given output signal. Noise is the result of the random time of target emission, background noise from the atmosphere, Johnson noise or current noise within the detector, and various other minor background noises. Johnson noise is a type of thermal shot effect caused by internal temperature fluctuations within the sensitive element. Johnson noise can be reduced by lowering the operating temperature of the detector and by using as high a chopping frequency as possible. Current noise is most prevalent in photoconductive detectors, and is the result of internal resistance fluctuations caused by random contacts between the semiconductor microcrystals and the thermally generated current carriers. Cooling the detector will reduce this type of noise. When discussing noise level and detector sensitivity, the term noise equivalent power (nep) is used. Actually, this factor is comparable to the minimum detectable power, discussed previously. The nep is the infrared power, expressed in watts, which must be received by the detector to generate a signal equal to its rms noise level over a given bandwidth. Infrared detector characteristics are listed in Table 3-11.

3-5.9 THERMAL DETECTORS

Any temperature change caused by incident radiation can be made to produce the following effects in the detector:

1. Generation of an electromotive force (thermocouple).
2. Resistance change (bolometer).
3. Gas volume change (Golay cell pneumatic detector).

The thermal detector is used primarily because of its wide spectral response. It depends solely on the heating effect of the infrared radiation rather than on its wavelength. The choice of which one of the several types of thermal detectors should be used for a particular application is chiefly dependent on the chopping frequency requirements. The thermocouple detector operates best at low chopping frequencies (over 30 Hz). The basic concepts, principles of operation, and illustrative requirements of both thermocouples and bolometers were covered elsewhere within this publication.

3-5.10 PNEUMATIC DETECTORS

The pneumatic, or Golay, cell consists of a hermetically sealed cell filled with air or other gas which does not absorb radiation easily, a

Table 3-11. Infrared Detector Characteristics

DETECTOR	TIME CONSTANT (SECONDS)	SPECTRAL SENSITIVITY (MICRONS)	FORM OF OUTPUT
Bolometer	4×10^{-3}	0.1–16	Change in resistance
Thermistor bolometer	3×10^{-3}	1–15	Change in resistance
Ceslum	1×10^{-6}	0.1–1.3	External photoeffect
Evaporagraph		0.1–12	Latent image
Germanimum	1×10^{-3}	0.3–1.7	
Golay cell	3×10^{-3}	0.5–15	Pneumatic
Lead selenide	1×10^{-6}	0.1–4.5 0.1–6(cooled)	Internal photoeffect
Lead sulfide	$1\text{-}4 \times 10^{-6}$	0.7–3.0 0.5–4(cooled)	Internal photoeffect
Lead telluride	1×10^{-6}	1–5.5(cooled)	Internal photoeffect
Photographic, type 2		0.1–1.2	Latent image
Thermocouple	50×10^{-3}	5–15	Electromotive force
Thermopile	10.50×10^{-3}		Electromotive force

transmitting window, and an absorbing membrane located in the center of the gas-filled cell. The optical photoelectric components consist of a glass within 100 ruled lines per inch, a meniscus lens, condensing lenses, a photocell-exciting lamp, and a photocell. A typical Golay cell arrangement is illustrated in Figure 3-54. The Golay cell detector is actually a type of thermal detector in that its output depends on the detected heat radiation. However, rather than undergoing a resistive change within the sensitive element, the Golay cell depends on a volume of gas change which is proportional to temperature change. As the Golay cell uses an all-metal film as a radiation absorber, it is not selective over the entire infrared spectrum. The change in indication is the result of a photoelectric circuit which is affected by physical displacement of one surface of the gas container. This moving surface is the criterion used to determine the limits of sensitivity. The actual operation of the pneumatic, or Golay cell is simple. Radiation entering the window heats the absorbing membrane, which heats the surrounding gas which, in turn, distorts the diaphragm. The diaphragm reflects light to the photocell in an amount depending on the diaphragm distortion. The spectrophone is another type of pneumatic detector which operates on the gas-expansion principle. However, sound, rather than an electrical photocell signal to an indicating device, is produced for measurement with a microphone hookup.

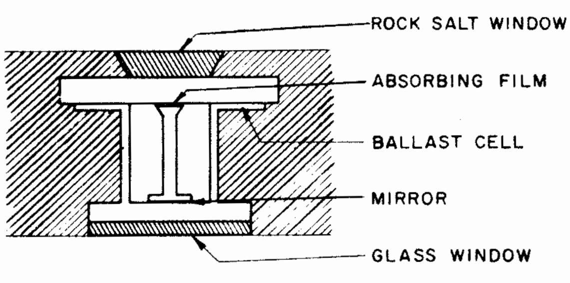


Figure 3-54. Golay Cell Structural Arrangement

3-5.11 PHOTO DETECTOR

The photo detector is sensitive to wavelength change rather than to heat. It is a selective device with low-frequency cutoff. Typical photo detector devices, which change electrically when exposed to infrared radiation, are photoemissive, -conductive, or -voltaic. Actually, detectors which depend on luminescent properties are also considered as photoelectric devices, as are photographic photo detectors. In photoemissive type detectors, electrons are emitted from the surface of some substance when that substance becomes exposed to infrared radiation. The photoemissive cell is chiefly composed of a cathode which emits electrons when exposed to infrared radiation, and an anode which receives these electrons; the resultant plate current is sent to an amplifier, as shown in Figure 3-55. The fundamental relationship governing the operation is explained by the following photoelectric equation:

$$E = hv - e\phi$$

where:

E = kinetic energy
h = Planck's constant
v = radiation frequency
eϕ = energy required from the surface of an exposed material (work function)

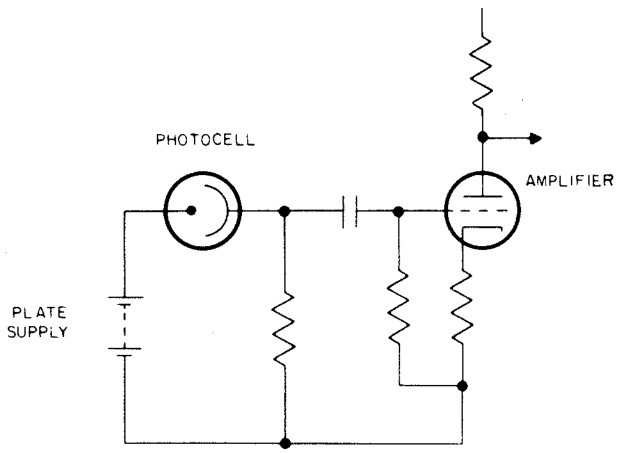


Figure 3-55. Photoemissive Cell Diagram

The spectral sensitivity of photoemissive detectors extends from the red end of the visible spectrum into the near-infrared range. Therefore, beyond 1.2 microns, the quantum energy level of the radiation is not sufficiently high enough to cause photoemission. One limitation to detector sensitivity is a dark current. Dark current is that which flows in either the photoemissive or photoconductive detector during total darkness (no applied radiant energy), and consists primarily of leakage current and thermionic emission. Since these effects are primarily due to temperature rise, operating the detector at a reduced temperature will tend to reduce both leakage currents and thermionic emission currents. The construction of the photoemissive detector is rather simple; it generally consists of only a cathode of silver- oxygen or antimony-cesium and an anode (plate) both in a vacuum cell. The vacuum cell suffers less from fatigue than does a gas-filled cell. The gas-filled cell permits a multiplication of up to 100 times the original photocurrent because the electrons emitted by the cathode ionize the gas molecules. However, as was mentioned above, the gas-filled cell is fatigued easily, and, it is unstable for low-level measurements; in addition, the presence of positive ions adversely affects the frequency response, which permits a sharp chopping frequency cutoff for frequencies beyond 1000 Hz. The photomultiplier type of detector has the advantage of internal radiation current multiplication and high sensitivity to radiation. The current multiplication occurs because electrons emitted from the cathode strike a surface that releases secondary electrons, which, in turn, strike another surface that releases electrons. The electrons continue to build up until they strike the final anode (plate). However, the photomultiplier detector is rarely used for infrared detection. The photoconductive detector obtains the photoelectric effect by virtue of the fact that some materials increase their electrical conductivity as a result of infrared illumination. This effect is not due to heat; it is the result of electron absorption in the conductive bands when exposed to a sufficient quanta of radiation. Since the photoconductor has a definite long-wave limit (it is wavelength-sensitive), only quanta of minimum magnitudes will activate the photoconductor. Semiconducting materials such as lead salts, silicon, and germanium are used in photoconductive detectors, because these materials will reflect or absorb high-energy light rays while they transmit low-energy infrared rays. These materials are normally gold-doped to instill the desired properties. The photoconductive effect is due to the fact that two kinds of charged carriers are contained in the semiconductor: negative electrons and positive holes. When the semiconductive material has been adequately charged, electrons in the valence band will shift over to the conduction band, illustrated in Figure 3-56. As the electrons in the conduction band circulate more freely, more current will result. Photoconductive detectors are constructed by either evaporating the semiconducting material on a glass slide or placing it there by chemical process and vacuum-sealing

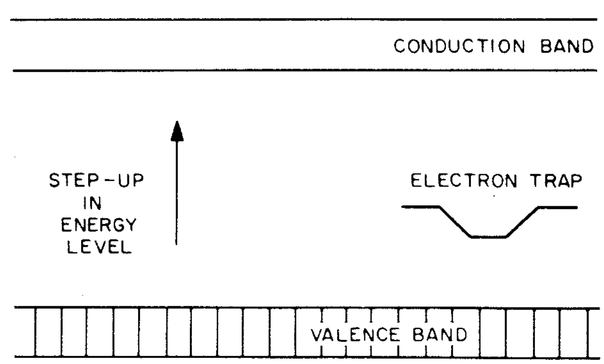


Figure 3-56. Structure of a Semiconductor

the glass between two electrodes. The material is then sensitized by introducing oxygen at a high temperature. The combination of elements within the cell will determine its physical size. Normally, photoconductive detectors are operated without special cooling. However, special applications may require that the cell be artifically cooled. The coolant (liquid helium hydrogen) is injected, between the specially constructed glass walls of the cell, by capillary action. The coolant will boil at reduced pressure in its vacuum enclosure, and thus provide adequate cooling. The sensitivity of the photoconductive detector is many times greater than that of a thermal detector. The spectral sensitivity of a photoconductor is normally around 9 microns without artificial cooling, and up to 120 microns (reduced sensitivity) with artificial cooling. Both the temperature and the illumination determine the time constant function. The time constant is not as fast as for photoemissive detectors, but is faster than for thermal detectors. This permits high chopping frequencies (1000 Hz) without loss in sensitivity.

3-5.12 PHOTOVOLTAIC DETECTORS

The photovoltaic and thermocouple detectors are similar in that both depend on irradiation at the junction of dissimilar metals. However, the photovoltaic detector depends on the photoelectric properties of the material rather than the thermoelectric properties. Three types of photovoltaic cells are mosaic, thallium, and lead sulfide. Mosaic cells are too critical for normal use, and the lead sulfide is not sensitive enough for normal use, The thallium sulfide cells are, therefore, the only cells of interest. These are produced by depositing thallium on a metal base, converting it to sulfide, and covering it with a gold or platinum layer.

3-5.13 LUMINESCENT DETECTORS

Some substances become luminescent when stimulated by infrared radiation. The emission of light is not caused by heat or incandescense, because it occurs at low temperatures. Phosphorus is the normal substance used as the luminescent material, because it stores energy when exposed to visible light, and later emits a bright visible light when illuminated by infrared. The use is normally limited to detecting (receiving) infrared transmissions at night. The phosphorus is exposed to sunlight for energy storage, and is then used as a detector after dark.

3-5.14 IMAGE-FORMING DETECTORS

Actually, the image-forming detector is a type of photo detector. It is treated separately because it provides a pictorial display. The image-forming detector requires optical devices to focus the image, but does not require chopping devices as does the non-image-forming detector. The image-forming detector is not very sensitive, and its performance is low. The target must be illuminated with infrared radiation; therefore, because of security considerations, use of the image-forming detector by the Navy is limited. As shown in Figure 3-57 the photoemitter type of image-forming tube employs a photosensitive cathode element which emits electrons when exposed to infrared radiation. These electrons are attracted to an anode (fluorescent screen), to provide a fluorescent display having a point-for-point light-to-radiation correlation. The picture is actually formed by modulating the electron beam sent from the cathode to the anode; the signal can be used in standard video circuits. Figure 3-58 illustrates the basic types of the many developed image tubes.

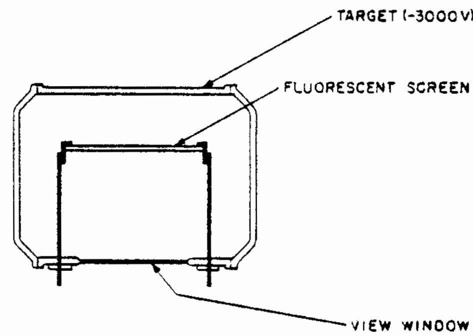


Figure 3-57. Electron Image Converter

3-5.15 TARGETS AND BACKGROUND

The target is considered to be whatever object is sought with infrared receiving or detecting equipments. The background is considered to be the remainder of the picture, which contains the desired object. Actually, the problem is relative, in that the background of one image may be considered as the target under other circumstances.

3-5.16 INFRARED TRANSMITTERS

The infrared transmitter is composed primarily of an energy source and a visible light filter.

3-5.17 RADIATION SOURCE

The radiation source transmits over a wide frequency band, including visible light, and the filter removes the visible light for security reasons. The source could be a tungsten lamp if its response speed were not too slow as to restrict Morse code operations to less than 8 words per minute. However, a high-speed mechanical shutter will increase the signaling speed, and the development of rugged, shockproof tungsten lamps

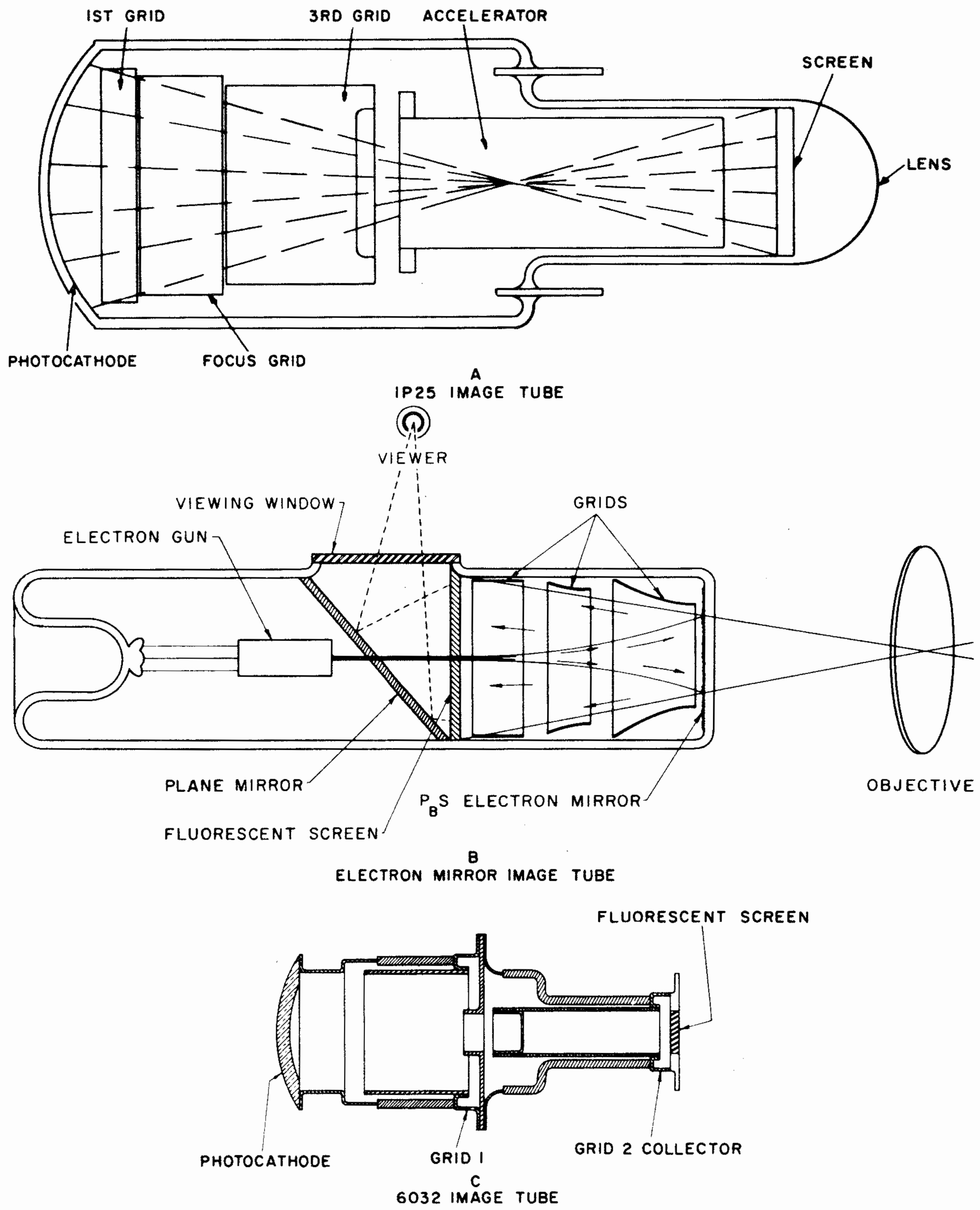


Figure 3-58. Typical Infrared Image Tubes

was initiated to eliminate both shock and vibration effects. For some applications, where the source energy must be modulated, a slotted cylinder is rotated to create a chopping or modulation effect. Cesium lamps have been developed for use instead of tungsten lamps because they have less distortion, a higher modulation rate, increased power output (5 times that of tungsten), and narrow-band selectivity.

3-5.18 MODULATION

Modulation may be accomplished by passing the radiation through a filament lamp or gas discharge source, at the radiation source. Modulation may also be accomplished with the aid of an electro-optical device (Kerr cell) or some form of mechanical modulation of the outgoing beam. Vibrating mirrors or any beam-chopping device may be employed at the output for mechanical modulation purposes. In RF applications an electrical modulation method using supersonic vibration of an optical element is employed. Electrical modulation is the most efficient method of modulation. However, the mechanical method of modulation, while not as efficient, is lighter in weight. The modulation method to be employed will determine the size, weight, power, efficiency, and security of the radiation source. In addition, it may partially affect the design of the receiving equipment.

3-5.19 MEASUREMENTS

Measurements and tests are not normally performed by shipboard personnel. Malfunctioning night vision equipment is usually turned in to a designated repair facility on a one-to-one basis.

3-5.20 OPTICAL TEST EQUIPMENT

Infrared optical assemblies normally do not require alignment unless optical components have been replaced. The alignment procedures for elaborate optical assemblies on a large infrared equipment may require special training prior to alignment performance.

3-5.21 BASIC LENS CHARACTERISTICS

The focal length of a lens is the physical distance from the optical center of a lens to a point where it will bring parallel light rays to a focus. Lenses that are thin have long focal lengths, and are said to be weak. Thick lenses have short focal lengths, and are said to be strong. The terms weak and strong are references to the converging power of a lens, that is, the ability of a lens to refract rays and to bring them to a focus within a certain distance. It is often necessary to measure the convergence, or optical power, of a lens. The unit of measure is the diopter. Weak, long-focus lenses have small diopter values, while strong, short-focus lenses have large diopter values; i.e., the lens power is inversely proportional to its focal length. The mathematical definition of a diopter is:

$$\text{Diopter} = \frac{1}{\text{focal length (in meters)}}$$

Thus, a lens with a focal length of 1 meter has a power of 1 diopter, and a lens with a focal length of 1/2 meter has a power of 2 diopters. In any lens assembly, the object distance and the image distance bear a definite relationship to each other. This relationship may be expressed by the following simple lens formula:

$$\frac{1}{\text{focal length}} = \frac{1}{\text{object distance}} + \frac{1}{\text{image distance}}$$

The optical measurements used to test, align, calibrate, or adjust infrared receivers are similar to those used for telescopes. In addition, tests have been devised to measure the sensitivity of the phosphors or image tubes associated with infrared receivers.

3-5.22 COLLIMATION

All infrared receivers have an optical assembly, usually of the reflector type, which collects radiation from the target and projects it in the form of a small, inverted image on the phosphor button of the receiver, or on the cathode of the image tube. When the collimator lens is moving toward the target, the rays become more divergent, and when the lens is moved away from the target, the rays become more convergent. Therefore, when a target is viewed through the collimator lens, it is possible to establish whether the infrared receiver is accurately focused for infinity simply by moving the collimator lens and observing the effect of the focus through the receiver. The clearest image should be obtained when the collimator is in its normal position, for only then is the receiver objective in correct focus for infinity. Any deviation from correct focus can be determined by noting the distance the collimator lens is moved to bring the image into focus. Deviations from correct focus can be obtained if the position of the collimator is indicated on a linear scale calibrated in diopters.

3-5.23 THE DIOPTOMETER

The image formed on the phosphor button or the screen of the electron image tube of an infrared receiver is too small and too close to the eye to be seen. An eyepiece (or ocular) is provided to produce a vertical image magnified 10 to 15 times, and sufficiently far from the eye of the observer to be clearly seen. In case of phosphor-button receivers, where the image on the phosphor is upside down, the erector lens serves to turn the image right side up. Signals viewed through infrared receivers are usually of low brightness, and can be seen only by an observer who has normal eyesight and is partially or completely dark-adapted. Under these conditions, the visual image in the receiver eyepiece is most clearly defined and most easily seen when the eyepiece is adjusted to a power of from minus 1 to minus 2 diopters. The power of the eyepiece can be measured by means of a dioptometer. In principle, the dioptometer is a tiny telescope whose eyepiece can be adjusted until cross-wires located at the principal focus of the eyepieces become sharply defined. When this has been done, light waves through the eyepiece of the dioptometer enter the eye of the observer as parallel waves. In using this instrument, first set up the infrared receiver with the eyepiece to be checked, so that is it illuminated by a collimated signal. Then take the dioptometer, holding the knurled end ring next to the eye. Peer into the eyepiece and turn the knurled ring until the crossline reticule comes into sharp focus. This operation sets the eyepiece for the individual using the instrument; any other person using the instrument will have to make this same adjustment. Now unscrew the diopter ring to the extreme negative position of -5. Place the dioptometer against the eyepiece of the receiver, and screw the diopter ring in until the image comes into sharp focus. Take the reading from the diopter ring. The reading obtained in this manner is a measure of the eyepiece adjustment of the receiver, and should be -1 or -2 diopters. The dioptometer can also be used to check the collimating lens. When the collimator is in the proper position for checking a receiver, a reading of zero should appear on the diopter scale when light coming through the collimating lens is observed through the dioptometer.

3-5.24 THRESHOLD

The threshold of an infrared receiver is the smallest amount of infrared radiation that can be seen through the receiver by a dark-adapted observer, using for the infrared source a lamp having a specified color temperature, and using a filter of specified spectral transmission. The threshold figure, therefore, is a measure of the sensitivity of an infrared receiver. The unit of measurement used in threshold tests is the nautical-mile-candle (nmc). It is a very small unit (there are approximately 36,000,000 nautical mile candles in 1 foot-candle). One nmc is the illumination received at a distance of 1 nautical mile from a source of 1 candle power. If an instrument can detect a source of 1 candle power at a distance of 1 nautical mile, then the threshold of the instrument is 1 nmc. The numerical value of this unit is inversely proportional to the actual sensitivity of the instrument. In other words, an instrument which requires a source of 4 candle power to operate at a distance of 1 nautical mile is a more sensitive instrument than one rated at 5 nmc. The finest instruments, therefore, are those of low nmc value; the least sensitive instruments are those of high nmc value. A poor threshold figure in a receiver may indicate a defect in the optical assembly, a faulty phosphor, or a defective charging system; it may also indicate a faulty image tube or a defective electrical system in the case of electronic-type instruments.

3-5.25 RESOLUTION

The resolution of an infrared receiver is a measure of its ability to distinguish between two point sources separated by a small angle. The unit of measurement used in resolution tests is the angle of separation of the two points, given in minutes (60 minutes in 1 degree) of arc. Unsatisfactory resolution in a receiver usually indicates poor optical alignment or focus. Threshold and resolution tests can be made on a specially constructed infrared optical tester.

3-5.26 THE INFRARED OPTICAL TESTER

The infrared optical tester is essentially a point light source which is placed at infinity by means of a collimator lens. This light source consists of two pieces of colorless glass ground on both sides and illuminated by a coiled filament tungsten lamp. The radiation from this source is given the proper spectral characteristics by passage through a standard infrared polaroid filter. The intensity of the light may be varied by known increments by interposing apertures of different sizes over the ground glass, and by moving the tungsten lamp measured distances from the ground glass. Since low light intensities are desired at the receiver for making threshold tests, a lens is placed between the light source and the collimator to reduce the light intensity and permit the use of a short-focal-length-collimator. Without this lens, a collimator lens of very long focal length (about 30 feet) would be required to retain the same effect. A hole plate in the optical tester contains 10 threshold apertures and 6 resolution patterns. The threshold holes vary in diameter from 0.027 to 0.177-inch to provide threshold values from 0.25 to 10 nautical-mile-candles. The resolution patterns are pairs of holes separated by various distances to correspond to resolution readings of 4, 6, 10, 15, 20, and 25 minutes of arc. (The separation of the holes in tenths of an inch exactly equals their separation in minutes of arc as seen through the optical assembly.) It is important that the threshold holes be kept clear of dust and dirt. The microscope objective and the collimating lens are mounted in threaded sleeves to facilitate adjustment. These are set and locked at the time the tester is calibrated, and should not be moved. The microscope objective can be removed for cleaning, storage, or shipment. When replaced, it should be screwed all the way into the sleeve. The hole plate is connected through gears to a shaft which is rotated by means of the knob on the front panel of the tester. The holes are arranged so that threshold and resolution values increase as the knob is rotated in a clockwise direction. Connected to this shaft is the detent plate which locates the holes as the hole plate is rotated. If it is desired to work between two definite values, as in inspection work, the detent is pulled out to its first slot, rotated, and locked. The detent stops are then screwed to the detent plate at the positions where the desired holes are in place when the stop strikes the detent. The detent can be pulled out further and locked again in such a postion that the stops will clear the detent and the shaft can be rotated its full 360 degreees. The values in nautical-mile-candles and minutes of arc are indicated through a hole in the front panel by self-luminous numbers on the detent plate. These numbers can be read in total darkness without the aid of a flashlight or auxiliary lamps. Before threshold readings can be made, the observer must be completely dark-adapted. This is accomplished in one of two ways. The first method is to remain in total darkness for 30 minutes, the second is to wear approved dark-adaptation goggles for 20 minutes and then spend 5 or 10 minutes in the dark before starting to make the readings. Either method accomplishes the same degree of dark-adaptation, the latter permitting the observer to work under normal lighting conditions. Once dark-adapted, the eyes should by no means be exposed to white light, and only to low levels of red light. Reading should be made at the dim threshold after the observer is dark adapted. The dim threshold is defined as the lowest level of illumination at which the point source is still continuously visible, but does not tend to appear and disappear as at the absolute threshold. In this text the dim threshold is referred to simply as the threshold. To make the readings, the instrument to be tested is placed in the jig in front of the collimator, and the light value turned up to about 10 times the expected threshold. This will enable picking up the spot in the field. Then, while observing the spot, adjust the light value until the threshold is reached. The nautical-mile-candle rating can now be read directly from the scale.

3-5.27 ELECTRONIC TEST EQUIPMENT

The replacement of components or units in the electronic portion of infrared equipment may necessitate the resetting of one or more adjustments. Adjustments are always necessary after the replacement of variable components or of fixed components that influence the operation of a variable. Even when exact replacements are used, differences must be accommodated. The differences lie within the manufacturer's tolerance, but in critical circuits they sometimes assume considerable importance. The adjustments made necessary by replacements and repairs may, in some instances, require a complete realignment procedure. After all replacements and repairs have been made, the complete equipment should be checked for proper operation, with follow-through on all adjustments as necessary. Many infrared equipments contain a method of calibrating the sensitivity of the amplifying circuits. This check serves as a performance standard for equipment operation, and also as a test to determine whether or not the infrared detector unit is at fault when the equipment fails. This is accomplished by feeding a calibrated signal into the amplifier immediately following the infrared detector unit. The known voltage introduced into the amplifier unit gives an indication of the over-all amplifier response. Electronic-type instruments may be tested on an infrared optical tester. To test a receiver, adjust the infrared source for a circle of 1-inch diameter on the photocathode; then multiply the dial threshold reading by the receiver conversion factor to obtain the true threshold reading. These readings should not be more than 1.5 nmc. Resolution tests may be conducted by projecting the resolution pattern on the photocathode. Set the focus voltage for equal horizontal and vertical resolution in the central areas; the resolution should be at least 8 lines per millimeter in the center of a 0.3-inch-diameter circle from the cathode center, and 2.2 lines per millimeter in the periphery (0.7-inch-diameter circle). The term "lines per millimeter" means pairs of equal width of both black and white. The number and size of the spots and blemishes within the 0.3-inch-diameter and 0.7-inch-diameter circles should conform to the data provided in Table 3-12. Electronic instruments must be handled with care, both before and during testing. Observe the following precautions carefully: do not permit light to enter the objective or the eyepiece while taking threshold readings, or from 10 minutes prior to taking threshold readings. Do not take threshold readings in rapid succession with the same instrument because the instrument, retaining a previous glow, will give a false indication.

Table 3-12. Infrared Receiver Resolution Data

MAXIMUM NUMBER OF SPOTS-SPOT SIZE		
CENTER (0.3-INCH DIAMETER)	TOTAL (0.7-INCH DIAMETER	(IN MILLI-METERS)
0	0	Over 0.35
0	8	0.25 to 0.35
5	16	0.08 to 0.25
5	20	

3-5.28 STANDARD INFRARED DETECTOR TEST EQUIPMENT AND PROCEDURES

The test procedures employed with infrared detectors may be divided into two groups; detector efficiency tests, which require three separate series of measurements with radiation sources, and tests which yield the root power spectrum of the noise. The ratio of the detector efficiency to the noise can be used to indicate the quality of the detector. This ratio is called the "detectivity of the detector." To measure the detector efficiency, the signal radiation shall be normally incident on the detector, and the amount of signal radiation will be confined to the range in which the output signal is proportional to the incident power. The calibrated signal generator required for infrared detector tests must produce a voltage sine wave of accurately known rms amplitude in the proper frequency range. Normally, the range will be adjustable from 1 to 100,000 Hz, with an approximate amplitude of 1 volt rms. This signal will be passed through a calibrated attenuator for a predetermined amplitude reduction. A tunable filter, with a center frequency which can be operated over the required range and with a stable gain (i.e., independent of filter tuning) is required for detector efficiency measurements. A multirange voltmeter, to measure both the signal and noise voltages, is required. The voltmeter should read true root-mean-square amplitude of any arbitrary waveform. Three sources of radiation are required. A black-body source, which must be frequency-chopped and stable in temperature, should produce an accurately known spectral irradiance uniformly over the responsive surface of the detector area. A monochromatic source of radiation consists not only of radiation, but of a monochromator with a chopper placed between it and the source. The chopper may be for a single fixed frequency, the same as was used for the black-body source. The irradiance of the monochromatic source should also be uniform over the responsive surface of the detector. A variable-frequency source must also be available, with a stable source of radiation and a variable frequency chopper. The source of radiation should be a black-body source, and a filter should be available. The detector-impedance and pulse-time-constant measurements may be performed with normal test equipment, such as that used with other electronic or communication circuits.

3-5.29 AN/SAR-7 INFRARED VIEWING SET

The AN/SAR-7 is an instrument designed to convert infrared radiation into visible light by means of an electronic image converter tube in the receiver. An objective lens assembly in front of the image tube focuses the infrared radiation on the photosensitive surface of the tube. An eyepiece lens assembly behind the tube is used to view the visible light emitted from the fluorescent screen in the tube. The infrared receiver is used to observe distant sources of infrared radiation for purposes of reconnaissance or for blinker signaling. The receiver is frequently used in conjunction with a signaling searchlight that is fitted with an infrared filter assembly. The searchlight serves as a source of infrared radiation for both reconnaissance and signaling purposes. Infrared Viewing Set AN/SAR-7 is used for signaling as well as for observation and reconnaissance in the dark by means of invisible in frared radiation. The viewing set converts the infrared radiation to visible light that is observed on the fluorescent screen of the image converter tube through the eyepiece assembly. For signaling, the viewing set is focused on a distance signaling searchlight or blinker light that is infrared filtered. For observation and reconnaissance, the area to be observed is illuminated by an infrared-filtered searchlight and the reflected infrared radiation is converted to visible light by the viewing set.

3-5.29.1 Operating Conditions

The viewing set operates most effectively on clear, dark nights. It can be used, at reduced ranges, on moonlit nights and under conditions of smoke, haze, or fog. Rain and snow impede the operation of the viewing set to the same extent that they impede normal vision. The viewing set can also be used to receive infrared signals in daylight.

3-5.29.2 Operator's Vision

The operator's eyes should be dark-adapted before attempting to use the viewing set at night. Remain in total darknewss of 30 minutes, or wear dark-adaptation goggles for 20 minutes and then remain in total darkness for 10 minutes, before starting operation.

3-5.29.3 Precautions

Observe the following precautions when preparing the viewing set for use:

1. Never point the objective end of the receiver toward the sun or toward any other source of intense visible or infrared light. An intense light will damage the image converter tube.
2. When using the 8.6 power eyepiece lens assembly, keep the eye as close as possible to the eyepiece. This prevents light reflections from the fluorescent screen from illuminating the face. When using the 4-power or 8.6 power eyepiece lens assembly, do not allow the eyepiece to be turned toward the area being observed. The glow of the fluorescent screen may be visible to enemy observers.
3. If the viewing set is being used in conjunction with an infrared-filtered searchlight, thoroughly scan the area for enemy infrared illumination before turning on the searchlight.
4. Always handle the viewing set with the same care as with any other precision optical or electronic device.

3-5.29.4 Preparation Procedure

To prepare the viewing set for operation, proceed as follows:

1. Remove the infrared receiver from the case.
2. The receiver is stowed with the 8.6 power eyepiece assembly in place. If desired, unscrew the 8.6 power assembly and replace it with the 4-power assembly. Use the 4-power assembly for signaling and preliminary scanning, reserving the higher powered 8.6 power assembly for situations requiring greater magnification and sensitivity. For prolonged viewing, because it has long eye relief and can be used with both eyes simultaneously, the 4-power assembly is less fatiguing to the eyes.) To attach either eyepiece to the viewer, screw it on until it is stopped by the stop ring. Do not use force or tools.
3. Set the rotary switch (S1) to OFF.
4. Remove caps.
5. Insert batteries.
6. Replace caps.

NOTE

If the battery has been removed, the negative terminal of the battery must be grounded for proper operation. Battery is inserted positive end first (the end with the center center post on BA-30 batteries) and the negative end out. On the rechargeable battery, the contact ring should be inserted first.

7. When the infrared set is to be used with the 12-inch signaling searchlight, attach the viewer to the searchlight by means of the clamp supplied with the equipment.

3-5.30 INFRARED SYSTEMS IN GENERAL

The Navy employs many varied and advanced types of infrared systems that affect the current state of the art. However, due to their technical complexity and current security classification they are not discussed in the EIMB. If information is required for such a specialized system, refer to the associated technical manual.

3-6 RADAR TESTING

3-6.1 GENERAL

It is possible to determine to some degree the satisfactory operation of a radio receiver by listening to its output and observing how much the gain control must be advanced to obtain sufficient volume. However, unlike most radio communication equipments, satisfactory performance of radar sets cannot be determined so readily. Tests were performed on approximately 100 radar sets that were considered by onboard personnel as being in satisfactory operating condition. The results of these tests revealed that, on the average, the maximum effective range of the radar equipments under test was only one-half the maximum range possible had the equipments been operating at peak efficiency. In fact, five radar sets were found to be operating at less than 10 percent of their possible maximum range, which means (in effect) that these radar equipments were protecting only 1 percent of their assigned tactical areas. Since such poor performance, as demonstrated by these tests, could have serious consequences, it can readily be seen that performance testing is of the utmost importance in radar work. Investigation to find the cause of this unsatisfactory situation showed that many technicians are not sufficiently familiar with the techniques and procedures necessary to test microwave-radar systems properly. In this section an effort is made to remedy this situation. It is assumed that the reader is already familiar with the basic principles of radar. By way of review, a brief discussion of the functional requirements of the range-determining components of a radar set is included in the following paragraphs.

3-6.2 RADAR SYSTEM OPERATIONAL REQUIREMENTS

The operational requirements of typical pulse radar systems are briefly described here. For purposes of study, the typical radar system can be reduced to the following functional components: synchronizer (modulator), transmitter, antenna, receiver, and indicator. Because power supplies differ so greatly in different radar sets, they will not be considered in this general discussion. While the complexity of each of the above components may vary considerably, depending on the use for which a particular radar equipment was designed (i.e., search, navigation, fire control, etc.), basically the function of each type of unit is identical in all radars. It is not the purpose of this discussion to describe any specific radar set, but rather to deal with the functional requirements of each unit necessary for the efficient overall performance of any radar facility.

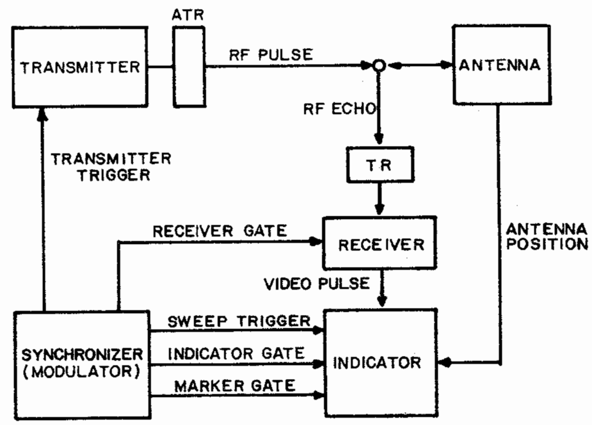


Figure 3-59. Radar Facility, Showing Timing Data Supplied by the Synchronizer

3-6.3 SYNCHRONIZER

Figure 3-59 illustrates the typical timing requirements of a radar facility which are developed and supplied by the synchronizing component. The timing circuits may all be located in one unit, or they may be distributed throughout one or more additional components, such as the transmitter, indicator, or receiver. In an externally synchronized equipment, a trigger pulse is fed to the transmitter for the purpose of timing the firing of the RF oscillator with the rest of the radar facility. In some sets, a gate pulse is also fed to the receiver in order to turn it on immediately after the transmitter fires or turn it off when the transmitter fires. When the RF performance of the transmitter and receiver is measured, some of the test equipments used must be synchronized with the radar system. To supply the necessary timing signal for these test equipments, test receptacles are usually located conveniently on these units which provide sync voltages of the proper amplitude and polarity.

3-6.4 TRANSMITTER

The transmitter consists of an RF generator and a modulator, or pulsing component. In microwave applications, the generator is usually a magnetron because of its relatively high output power. In lower-frequency applications, a ring or conventional type oscillator is used. For satisfactory operation of the transmitter, a nearly rectangular modulator pulse is required. A pulse with a flat top is desired because a magnetron tends to shift frequency if its high voltage (furnished by the modulating pulse) varies during the period of oscillation. A steep leading edge is also required, particularly in fire-control equipment, where accurate range data is necessary. In installations where minimum range data is needed, a pulse with a steep trailing edge is essential. The required width, or duration, of the pulse also depends upon the type of radar system in which it is to be used. In the case of long range air-search systems, wide pulses (2 to 200 microseconds) are utilized to maintain a high average of transmitted power. For surface search and fire-control equipments, which require high resolution, narrow pulses (0.1 to 2 microseconds) are used. Any sizeable variation from the norm in pulse width, shape, or amplitude seriously affects the range capability and accuracy of the facility. These undesired pulse variations are not necessarily apparent to a radar operator. For this reason, test methods and procedures have become incorporated in the PMS program to observe and measure transmitter operation as it affects the overall performance of a radar equipment. Figure 3-60 shows a schematic diagram of a typical radar transmitter employing a magnetron. The TR (Transmit-Receive) tube in most cases has been replaced with a limiter tube that does not require a keep-alive voltage.

3-6.5 ANTENNA

A typical antenna group consists of such components as the transmission or RF lines, with associated accessories such as rotating joints, duplexer,

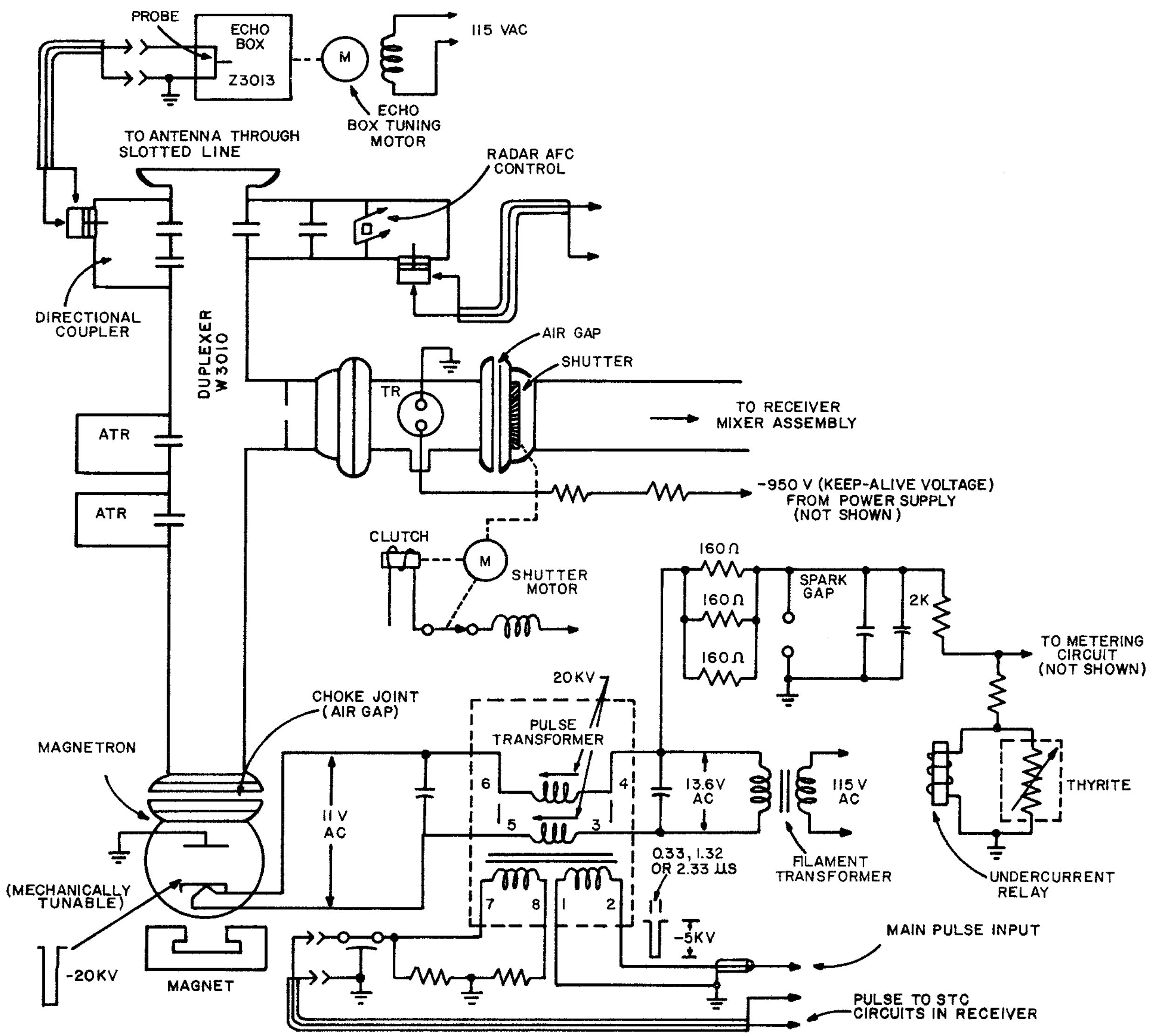


Figure 3-60. Transmitter Section, Schematic Diagram

slotted-line section and directional coupler, the antenna and the antenna reflector. Although the RF lines may be either coaxial cable or wave guides, their function is the same: that is, to deliver the maximum amount of power from the transmitter to the antenna, and to receive power from the antenna to the receiver. Many factors tend to reduce the efficiency of RF lines. Improper coupling, faulty rotating joints, dents in the line, or poor solder joints may produce an impedance mismatch between the RF lines and the antenna or transmitter resulting in a high standing-wave ratio, magnetron instability, and a reduction in equipment performance.

3-6.5.1 Duplexer

Since radar systems both transmit and receive through the same antenna, some method must be employed to block the receiver during transmission. The most common device used is the duplexer, which consists of a transmit-receive (TR) switch, and usually an anti-transmit-receive (ATR) switch. The TR switch is a gas-filled tube designed to short circuit across the spark gap each time the transmitter fires, thus preventing saturation of the receiver and damage to the crystal mixer (when used) by a strong signal.

3-6.5.2 ATR Switch

To reduce loss of the return signal in the transmitter during reception, many radar facilities incorporate an ATR switch, located an odd number of quarter wavelengths from the receiver T-junction. During transmission, both the TR and ATR gaps are fired, producing short circuits at these points. In this way transmitter power is conducted to the transmission line. During reception neither is fired, and the impedance of the ATR switch is such that the received echo is reflected into the receiver with minimum loss. In addition to this, the TR and ATR switches must be capable of quick recovery after the transmitter fires, so that echoes from nearby targets may be received. Recovery times of TR tubes range from 3 to 30 microseconds. Since the recovery time of these tubes tends to increase with use, it must be checked periodically to ensure proper operation of the radar facility. Methods of measuring recovery time will be discussed later.

3-6.5.3 Pre-set Antennas

Early radar antennas often required considerable adjustment for efficient operation, but most present-day antennas are preset at the factory, and require no further adjustments in the field. However, when making RF tests or measurements, it is important to position the antenna such that it will not be affected by strong echoes from nearby fixed targets.

3-6.5.4 Test Points

To provide efficient RF test points, two devices (directional couplers, and slotted-line sections) are used. In some equipments these devices are permanently built into the antenna group. In others they are included as part of the test equipment, and may be inserted in the RF line whenever tests are to be made.

3-6.6 RECEIVER

An efficient receiver has good sensitivity, short recovery time, and sufficient bandwidth to pass a received pulse echo without undue distortion. Figure 3-61 shows the necessary stages required to obtain these characteristics. Because of the high frequencies at which most radar receivers operate, the received signal is fed directly to a crystal mixer or RF amplifier. A silicon crystal is usually used as a mixer because of its low inherent noise level. Although the TR switch is designed to protect the receiver each time the transmitter fires, a strong signal from the transmitter may leak through. Unless additional precautions are incorporated, this undesired transmitter pulse may overdrive and block the receiver, rendering it insensitive to signals reflected from nearby targets. This blocking usually occurs in one of the resistance-coupled video stages. Several methods are used to minimize this undesirable effect, such as using a receiver gate pulse, or feeding a negative-going signal from the second detector to the first video stage.

3-6.6.1 Bandwidth

The bandwidth requirements of the receiver also depend on the type of radar system in which it is used. Fire-control radar systems require broad-band receivers for accurate range data. Search systems, on the other hand, operate satisfactorily with narrow-band receivers, since merely the azimuth and range of a target are usually all the data required.

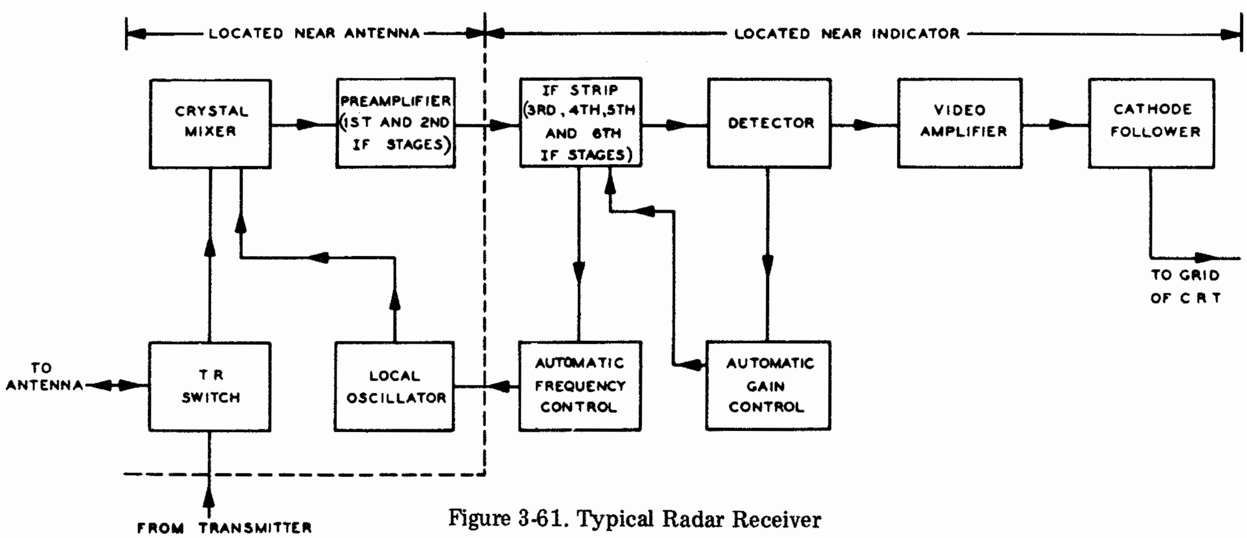


Figure 3-61. Typical Radar Receiver

3-6.7 INDICATOR

The indicator performs the important function of transforming electrical information gathered by the radar into a visual presentation on the face of one or more cathode-ray tubes which are part of the component. If the system includes an A scope or a ppi scope, visual observation of receiver performance, as indicated by echo box "ring time," may be had, as explained later in the test. It is emphasized that while the display on the ppi may indicate, to the inexperienced operator, that all the components of a radar facility are operating, it will not show how efficiently they are performing. The degreee of efficiency can be determined only be careful performance testing.

3-6.7.1 Frequency Measurement and Standards

The following subjects are pertinent to the measurement of radar frequencies: standards, coupling methods, typical test equipment, and the accuracy limitations of each are discussed. Table 3-13 shows typical bands of radar operation and their respective frequencies.

3-6.8 FREQUENCY TESTING STANDARDS

Equipments employed in frequency testing are classified as either primary or secondary standards. To summarize briefly, a primary standard provides an extremely accurate frequency source which is checked against the rotation of the earth, and is used to calibrate secondary standards. Secondary standards are extremely accurate and are not unnecessarily cumbersome or difficult to use. Consequently, for radar maintenance, accurate test equipments comparable to secondary standards are used.

3-6.9 METHODS OF COUPLING FREQUENCY STANDARDS

Frequency testing instruments are not designed to be coupled directly into the radar equipment, since the high-power transmitter pulse develops very high voltage within resonant circuits associated with the instruments, and arcing would result. Satisfactory methods of frequency coupling are are described for power sampling techniques and are discussed in later paragraphs.

3-6.10 FREQUENCY TESTING EQUIPMENTS

The two test equipments that are satisfactorily used in the measurement of microwave frequencies are: the resonant-coaxial-line frequency meter, and the resonant-cavity frequency meter. Both of these test instruments depend upon a condition of resonance to provide an accurate test indication.

3-6.11 SPECTRUM ANALYZERS AND FREQUENCY COUNTERS

Both of these equipments provide a high degree of accuracy but require additional care to prevent excessive power from being applied to the input circuit.

3-6.11.1 Spectrum Analyzer

The Spectrum Analyzer provides a visual indication of frequencies and peak power levels present at the testing point. The high impedence input of the Spectrum Analyzer causes very little loading effect, and can thus be used on RF sources having a very small output.

3-6.11.2 Frequency Counter

Frequency Counters, in general, do not have the sensitivity or high impedence of the Spectrum Analyzer. They should therefore be used to measure continuous frequencies rather than pulsed RF, due to the harmonics that are generated. Frequency counters are available to measure up to 40 gigahertz.

3-6.12 RESONANT-COAXIAL-LINE FREQUENCY METER

Coaxial-line frequency meters may be connected to operate as either transmission or reaction type indicators. When used as the transmission type, energy is fed into one coupling loop and the indicating device is connected to the other loop. When the circuit is resonant, the greatest energy transfer takes place and the indicator shows the greatest output signal. When used as the reaction type, the resonant circuit functions as an absorption device, so that at resonance the indicator shows a dip in the reading.

3-6.13 RESONANT-CAVITY FREQUENCY METER

A common type of resonant-cavity frequency meter consists essentially of a hollow metal cylinder coupled to a waveguide by means of a small hole or coupling loop and coaxial connector. The cavity within the cylinder is resonant by virtue of its dimensions. Two end plates may be thought of as the capacitance elements, and the adjoining walls as the inductance. Frequency is varied by adjusting the position of one of the end plates with a micrometer screw calibrated to indicate frequency. The degree of accuracy obtained with this type of meter is greater than that obtained with the resonant-coaxial-line frequency meter.

Table 3-13. Frequency and Letter Designations for Microwave Bands

SUBBAND	FREQUENCY IN GIGAHERTZ	SUBBAND	FREQUENCY IN GIGAHERTZ
	P Band		*X* Band-*Continued*
	0.225	*l*	9.00
	0.390	*s*	9.60
		z	10.00
		f	10.25
	L Band	*k*	10.90
	0.390		
p	0.465		*K* Band
c	0.510		
l	0.725		10.90
y	0.780	*p*	12.25
t	0.900	*s*	13.25
s	0.950	*e*	14.25
x	1.150	*c*	15.35
k	1.350	*u*†	17.25
f	1.450	*t*	20.50
z	1.550	*q*†	24.50
		r	26.50
		m	28.50
	S Band	*n*	30.70
		l	33.00
	1.55	*a*	36.00
e	1.65		
f	1.85		
t	2.00		*Q* Band
c	2.40		
q	2.60		36.0
y	2.70	*a*	38.0
g	2.90	*b*	40.0
s	3.10	*c*	42.0
a	3.40	*d*	44.0
w	3.70	*e*	46.0
h	3.90		
*z**	4.20		
d	5.20		*V* Band
			46.0
	X Band	*a*	48.0
		b	50.0
	5.20	*c*	52.0
a	5.50	*d*	54.0
q	5.75	*e*	56.0
*y**	6.20		
d	6.25		
b	6.90		*W* Band
r	7.00		
c	8.50		56.0
			100.0

*C Band includes S_z through X_y (3.90-6.20 gigahertz).
†K_1 Band includes K_u through K_q (15.35-24.50 gigahertz).

3-6.14 FACTORS AFFECTING MEASUREMENT ACCURACY

The accuracy of a microwave frequency measurement is expressed in terms of maximum error in megahertz, and may be either absolute or relative. Absolute accuracy states how much error with respect to a standard (usually WWV) is involved in a single frequency measurement, whereas relative accuracy indicated how much error is involved in the difference in frequency (or increment) between two microwave signals. For example, assume a measurement accuracy of ± 4 MHz absolute and ± 1 MHz relative; if the frequency of a certain transmitter measures 9300 MHz and the local oscillator frequency is 9330 MHz, the following conclusions can be reached:

1. The transmitter frequency is somewhere between 9304 MHz and 9296 MHz (or 9300 MHz ± 4 MHz absolute); or
2. The local-oscillator frequency is somewhere between 9334 MHz and 9326 MHz (or 9330 MHz ± 4 MHz absolute); or
3. The local-oscillator frequency is 29 MHz to 31 MHz above the transmitter frequency (9330 MHz minus 9300 MHz or 30±1 MHz relative).

Thus it can be seen that the difference between two measured values is much more accurate than the measured values themselves.

3-6.14.1 Accuracy

In receiver frequency testing, an absolute accuracy of ±4 MHz is not considered good enough. In order to obtain required accuracy, manufacturers of frequency testing equipments carefully calibrate by hand that part of the test equipment which concerns frequency testing, with the result that these equipments have an absolute accuracy better than ±1 MHz in the X-band.

3-6.14.2 Meter Tuning

Most of the early frequency meters were tuned by means of a micrometer screw, and the readings were converted into frequency with the aid of a calibration chart. Some of the newer-type meters make use of a dial, geared to the screw, which indicates frequency directly. Thus, dial readings are greatly simplified, but the gear mechanism associated with the dial introduces a certain amount of backlash which affects the accuracy of the indication. The backlash effect may be minimized by always approaching the final dial setting from the same direction. This direction should be the same as that used during factory calibration, and should be specified in the instructional literature accompanying the test equipment.

3-6.14.3 Atmospheric Effects

Atmospheric conditions of temperature, relative humidity, and atmospheric pressure exert an appreciable effect upon the accuracy of a frequency meter. This effect is minimized by constructing the resonant sections of materials that minimize or compensate for changes in temperature, and by hermetically sealing the units against moisture, with the result that sufficient accuracy is obtained for most applications. Where extreme accuracy is necessary, the correction charts for varying atmospheric conditions (supplied with the equipment) must be used. Atmospheric pressure variations under normal conditions, are not great enough to require compensation; however, the effect of reduced pressure at high altitudes can be appreciable. It is emphasized that the conditions discussed above exist inside the frequency meter, rather than outside it.

3-6.15 FREQUENCY MEASUREMENT

The measurement of frequencies employed in radar operation falls into two general categories: transmitter frequency, and receiver frequency.

3-6.15.1 Transmitter Frequency

Radar transmitter may be either fixed frequency or tunable. If the frequency of a fixed-frequency transmitter is measured and found to be outside the operating band, the magnetron or the defective component of the magnetron assembly must be replaced. Tunable transmitters may be adjusted throughout the operating band; this is a desirable feature when several radars are in use in a limited area, since the radar may be tuned to a particular frequency to prevent or avoid interference. In addition, when jamming signals are present, a change in radar frequency may be effective in eliminating or reducing their effect. Since the operating bands are fairly wide, however, transmitter frequency tests do not require extreme accuracy.

3-6.15.2 Receiver Frequency

Testing the radar receiver frequency consists of measuring the frequency at which the receiver operates most efficiently, or of measuring the local-oscillator frequency. For radar reception, a knowledge of the receiver frequency is not important so long as the receiver is carefully tuned to the transmitter frequency. In receivers using acf, the local oscillator must, of course, be operated either above or below the signal frequency in accordance with the design specifications. Here again, a knowledge of the exact frequency is not very important, but receiver bandwidth is an important factor and should be checked.

3-6.16 REACTION-TYPE INDICATION METHOD

This method of frequency measurement is shown in Figure 3-62. The meter absorbs power from the crystal detector at resonance; thus, a reaction type indication is obtained. Figure 3-63 shows the appearance of the RF enevelope as the frequency-meter tuning is varied. Resonance is obtained when the center of the pulse reaches its lowest point, indicating maximum reaction. If desired, a microammeter may be used instead of the oscilloscope. In that case, the frequency meter is adjusted for a dip in the current reading.

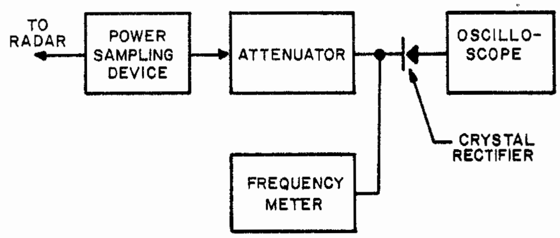


Figure 3-62. Frequency Measurement, Reaction-Type

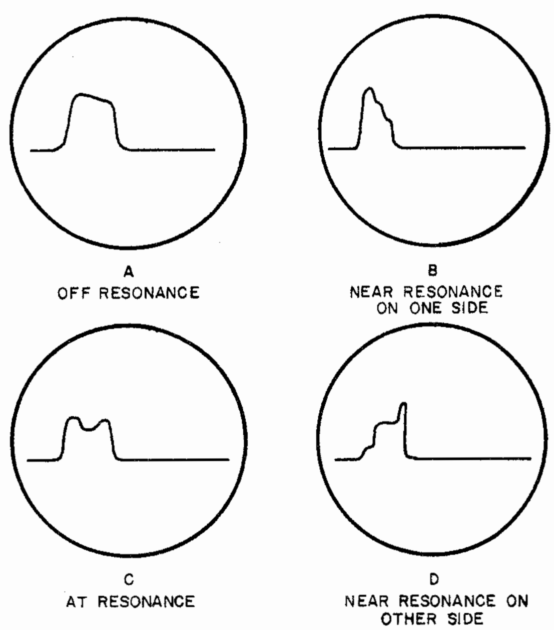


Figure 3-63. Change of Waveform Observed During Frequency Measurement

3-6.17 TRANSMISSION-TYPE INDICATION METHOD

Figure 3-64 shows a widely used test method for frequency measurement, using the transmission-type indication. The procedure for this method is as follows:

1. The equipment is connected as shown in Figure 3-64 and the power sample is coupled into the frequency meter.
2. The measurement starts with maximum attenuation and the frequency meter is then tuned through the frequency range.
3. If no indication is observed, the attenuation is reduced about 10 dB and the frequency control is again tuned through the frequency range. This process is repeated until a reading is observed.
4. The frequency dial is set for maximum reading with sufficient attenuation to keep the reading below full-scale value.
5. Ordinarily it is necessary to convert the dial reading to frequency.

3-6.18 COMBINATION POWER AND FREQUENCY TESTING

In modern power-testing equipment, a frequency meter is often included as an integral part of such equipment. The frequency meter is usually connected as shown in Figure 3-65. For power testing, the frequency meter must be tuned off-resonance so as not to affect the accuracy of the power measurement. The test procedure is very simple and is usually performed directly after a power measurement. Frequency measurement is performed as follows:

1. A 1-mW reading is established as will be discussed in the power-testing procedure.
2. The frequency meter is tuned for a minimum meter reading. (The meter must be tuned slowly or the resonance point may be passed before the thermistor can respond.)
3. Ordinarily it is necessary to convert the dial reading to a frequency.

3-6.19 LOCAL-OSCILLATOR FREQUENCY MEASUREMENT

The local-oscillator frequency can be measured by feeding the output of the local oscillator (directly, if possible) to the frequency meter and then making the test previously described. If desired, the local oscillator may be set to some predetermined frequency by setting the frequency meter above or below (as specified in instructional or maintenance literature) the frequency to be received by an amount equal to the intermediate frequency. The local

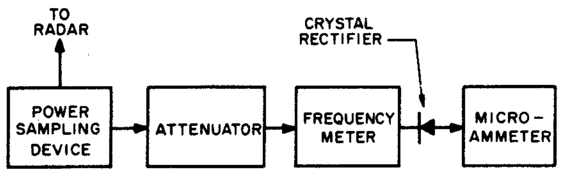


Figure 3-64. Frequency Measurement, Transmission-Type Indication

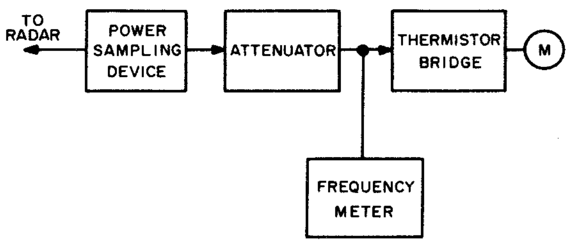


Figure 3-65. Combination Power and Frequency Measurement

oscillator is then tuned for the required indication. This method is especially useful for a radar system employing afc, which requires that the local-oscillator frequency be set on a certain side of the signal frequency.

3-6.20 POWER MEASUREMENTS

When testing radar systems, it is often necessary to measure the power output of the radar transmitter, or to determine the output level of a signal generator so that the test equipment can be used to make accurate measurements on the radar receiver. It is important, therefore, that the technician have a thorough knowledge of the principles involved in testing power output.

3-6.20.1 Units

Modern testing methods require that the absolute power in watts be the unit of measurement. Power can be measured in terms of relative values, but for most purposes such measurements are considered unsatisfactory. For example, a crystal rectifier and dc meter may be used to indicate power in units of meter deflection instead of watts. If periodic tests were made on a certain type of equipment, with the same crystal and meter and using the same procedure, any change in power would probably be discovered. However, this method has several faults. First, there is a danger that the initial reading, which must serve as a reference for comparison with later readings, may be taken at a time when the equipment is not operating properly. Second, manufacturers' specifications are rated in watts, rather than in relative values. Third, the results of this method cannot be compared with the results obtained by another person or by means of other test methods. Fourth, a different crystal-and-meter combination is very likely to give a different reading, so that if either the meter or crystal were damaged, the test procedures might need to be started all over again. It is clear, therefore, that accurate, calibrated test standards must be used if maximum performance and maintenance efficiency are desired.

3-6.21 POWER TESTING DATA

The material under this heading provides the basic information necessary to understand microwave power measurements and the techniques involved. The subjects covered are: pulse power and average power, the decibel and its use, power sampling methods, attenuators, and a brief description of the thermistor.

3-6.22 PEAK POWER AND AVERAGE POWER

Power measurements are classified as either peak power or average power. Although the actual transmitter output occurs at peak level, most modern test methods measure the heating value of the RF energy to obtain the average value. It is correct to use either value for reference so long as one or the other is consistently used. Frequently it is necessary to convert from peak power to average power, or vice versa; therefore, the relationship between the two must be understood. Figure 3-66 shows the comparison between peak power and average power when a square pulse is used. The average value, which represents the actual heating value of the pulses, is located at a point

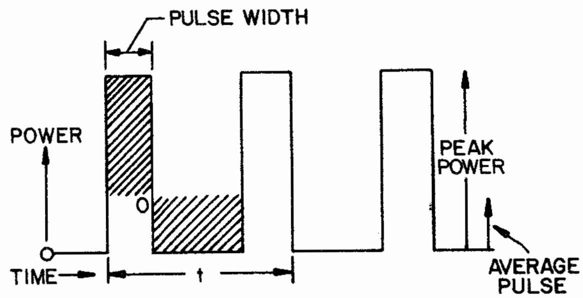


Figure 3-66. Transmitting Pulses, Showing Peak and Average Power

somewhere between zero and peak. The level of the average value is defined as that level where the pulse area above average equals the area below average between pulses. If the pulses are evened off in such a way as to fill in the space between pulses, the level obtained is the average value, as shown in Figure 3-66, where the shaded area of the pulse is used to fill in the space between pulses. In the same figure, the area of the pulse is equal to pulse width multiplied by peak power, and the area of the average value is equal to the average power multiplied by the period (T). Since the two values are equal, it is permissible to express the equation as follows:

$$\text{Pulse width x power} = \text{Average power x T.}$$

Transposing terms in the equation produces:

$$\frac{\text{Average power}}{\text{Peak Power}} = \frac{\text{Pulse width}}{\text{T}}$$

and since t=1/prf then:

$$\frac{\text{Average power}}{\text{Peak power}} = \text{Pulse width x prf}$$

The ratio of average power to peak power is called the "duty cycle," and represents each second the transmitter is on. Duty cycle is simply a numerical value, and may be used to describe power, voltage, or current as long as the terms are consistent. For example, if a certain radar has a pulse width of 1/2 microsecond and a pulse recurrence frequency of 2000 pulses per second, the duty cycle is 1/2 (10^{-6}) x 2000 or 0.001 (0.1%). If the peak power is 200 kw, the average power is 200 kw x 0.0001, or 200 watts. If the peak current is 10 amp, then the average current is 0.01 amp.

3-6.23 THE DECIBEL AND ITS USE

The decibel is part of a larger unit called the "bel." As originally used, the bel represented a power ratio of 10 to 1 between the strength of two sounds. To gain a better understanding of the bel, consider three sounds of unequal power intensity. If the power intensity of the second sound is 10 times the power intensity of the first, its power level is said to be 1 bel above that of the second. However, since the third sound is 100 times as intense as the first, its level is 2 bels above that of the first. Thus, a power ratio of 1000 to 1 by 3 bels; a power ratio of 10,000 to 1, by 4 bels, etc. It is readily seen, therefore, that the concept of bels represents a logarithmic relationship, since the logarithm of 100 to the base 10 equals 2(corresponding to 2 bels); the logarithm of 100 equals 3(corresponding to 3 bels); etc. The exact relationship is given by the formula:

$$\text{Bels} = \log \frac{P2}{P1}$$

where $\frac{P2}{P1}$ represents the power ratio.

3-6.23.1 Power Ratios

This logarithmic characteristic of the bel makes it a very convenient means for expressing power ratios. For example, assume that we desire to find the attenuation ratio of an RF attenuator which is to be used to measure transmitter power output. On test, it is found that 60,000 watts of RF input to the attenuator produces an output of 6 milliwatts. To find the attenuation ratio, use the equation:

$$\text{Attenuation ratio} = \frac{P2}{P1}$$

$$= \frac{60{,}000}{0.006}$$

$$= 10{,}000{,}000$$

This ratio can be expressed much more conveniently in terms of bels.

$$\text{Bels} = \log \frac{P2}{P1} = \log \frac{60{,}000}{0.006} = \log 10{,}000{,}000 = 7 \text{ bels}$$

In this case, the attenuation ratio is 7 bels. In other words, P2 is said to be 7 bels up with respect to P1. In all instances where P2 is numerically greater than P1, as in the above example, the final result is expressed in a positive quantity. When P2 is smaller than P1, the numerical result is the same, but it is expressed as a negative quantity. If, for example, P2 is 0.006 watt and P1 is 600,000 watts, then:

$$\text{Bels} = \log\frac{P2}{P1} = \log\frac{0.006}{60{,}000} = \log 0.0000001 = -7 \text{ bels}$$

In this case, P2 is said to be 7 bels down with respect to P1.

3-6.23.2 Decibel

Since the bel is a rather large unit, its use may sometimes prove inconvenient. Usually, therefore, a smaller unit, the decibel, is used. Ten decibels equals 1 bel. A ten-to-one ratio, which can be represented by 1 bel, can also be represented by 10 decibels (10 dB), a 100-to-1 ratio (2 bels) can be represented by 20 dB, a 1000-to-1 ratio (3 bels) by 30 dB, etc. The previous formula for bels may now be written to give a result in decibels merely by multiplying by 10. Thus, the formula becomes:

$$\text{Decibels (dB)} = 10\log\frac{P2}{P1}$$

It should be clearly understood that the term "decibel" does not in itself indicate power, but is rather a ratio or comparison between two power values. In radar testing, however, it is often desirable to express performance measurements in decibels. This can be done by using a fixed power level as a reference. The original standard reference level was 6 milliwatts (0.006 watt), but to simplify calculations a 1-milliwatt standard has been adopted, and will be used hereafter in the part of this manual dealing with radar testing. (Note: A few equipments use one watt as a standard.)

3-6.24 REFERENCE LEVEL (dBm)

When 1 mW is used as a reference level, the ratio is expressed in dBms. The abbreviation "dBm" indicates decibels relative to a 1-milliwatt standard. Thus, a pulsed radar transmitter having an average power output of 100 watts is said to have an average power output of 50 dBm. The conversion from power to dBm is made as follows:

$$\begin{aligned}\text{Average power (dBm)} &= 10\log\frac{P2}{P1}\\ &= 10\log\frac{100}{.001}\\ &= 10\log 100{,}000\\ &= 50 \text{ dBm}\end{aligned}$$

Conversions from power to dBm can be made more readily be means of graphs shown in Figures 3-67 and 3-68. If, as in the above example, the average power output is 100 watts, reference to the graph in Figure 3-67 shows that the line representing 100 watts intersects the curve at point A, indicating that 100 watts is equivalent to 50 dBm.

3-6.24.1 Voltage/Current Ratios

Voltage and current ratios may also be expressed in terms of decibels, provided the resistance remains constant. For equal resistances, the formulas are:

$$\text{dB} = 20\log\frac{E2}{E1}$$

$$\text{dB} = 20\log\frac{I2}{I1}$$

The difference in the multiplying factor in these formulas (20 rather than 10, as in the case of power ratios) arises from the fact that power is proportional to voltage or current squared, and when a number is squared, the logarithm of that number is doubled. For power ratios, the dB value is 10 times the logarithm of the ratio. For voltage or current ratios, the dB value is 20 times the logarithm of the ratio. As was stated previously, power measurements are peak power or average power. When testing the overall performance of a radar system, it is necessary to know the peak power output of the transmitter. However, modern test equipments can measure only the average power output, which, depending upon the specific test equipment, may be given either in watts or in dBm. Therefore, the relationship between pulse power and average pulse power for a rectangular pulse must be determined. It is most convenient to express this relationship in terms of dB.

3-6.24.2 Peak Power

The peak power output of a transmitter (in decibels relative to 1 mW) can be found from the relationship:

Peak power (dBm) = Average power (dBm) + duty cycle (dB)

3-6.24.3 Average Power

For a given radar transmitter, the two latter quantities are easily determined. The average power output is measured by means of test equipment. If the value obtained from the test equipment is expressed in watts, it must be converted to dBm. The

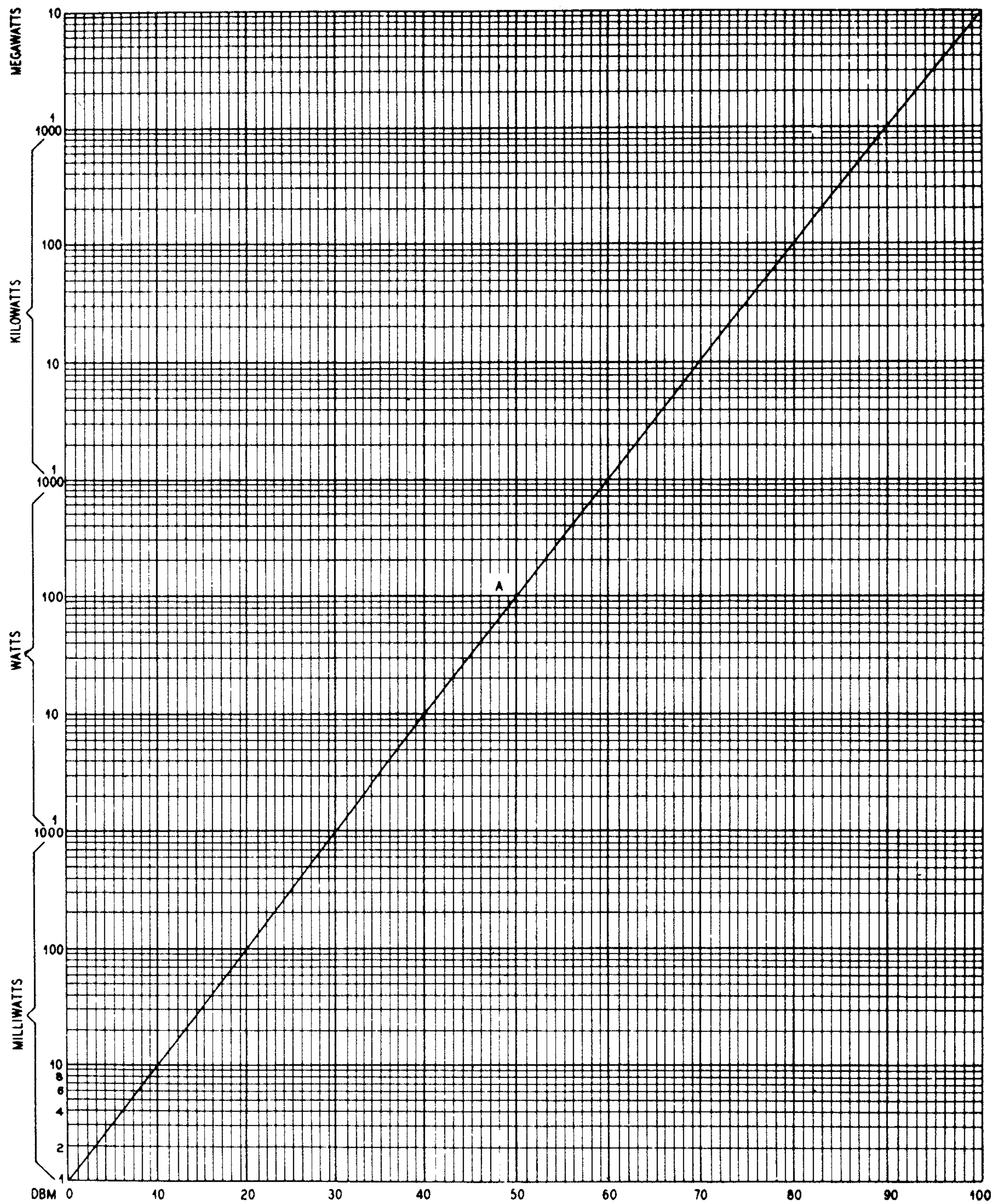


Figure 3-67. Power to dBm Conversion Chart, 1 Milliwatt to 10 Megawatts

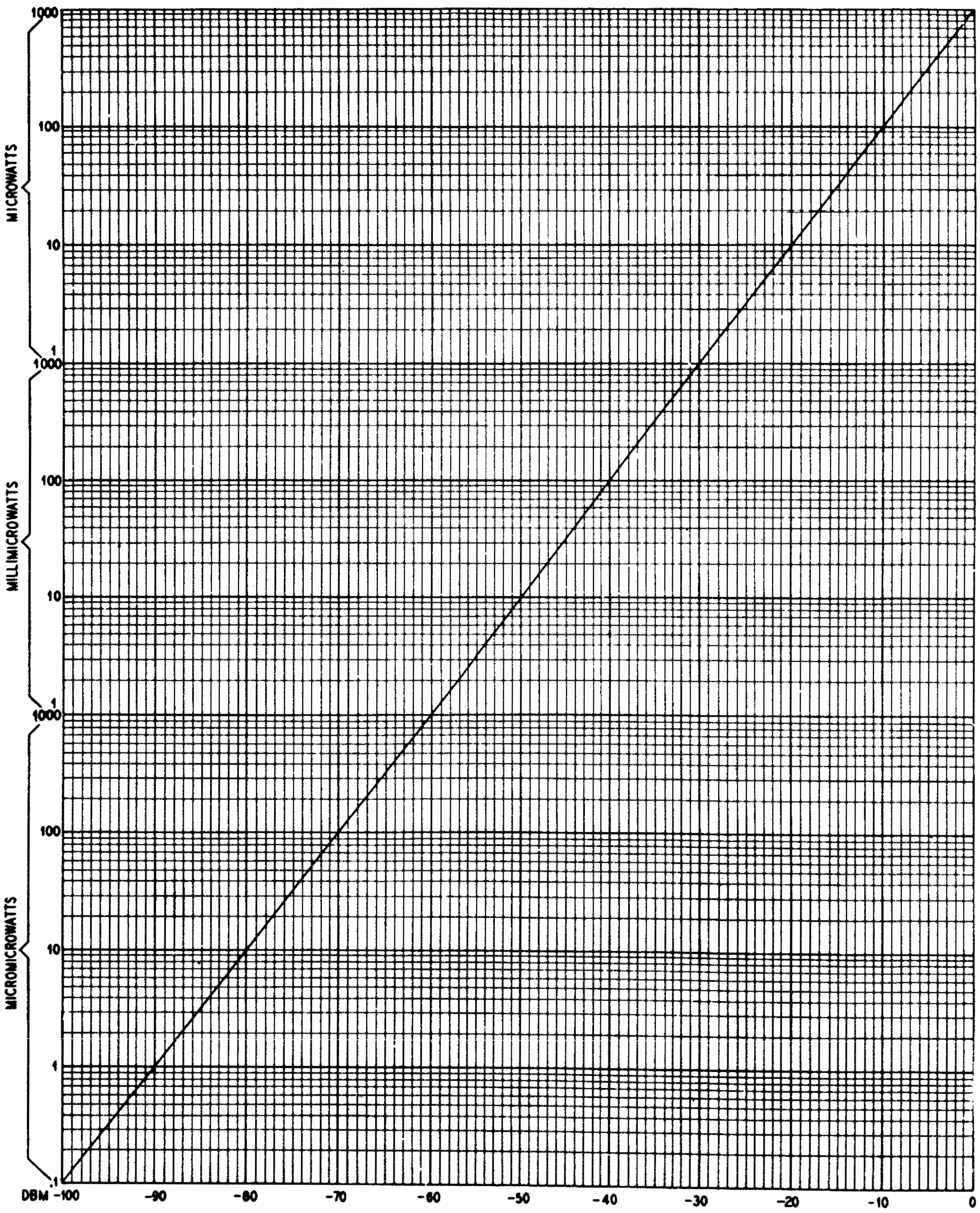


Figure 3-68. Power to dBm Conversion Chart 1 Milliwatt to 0.1 Micromicrowatt

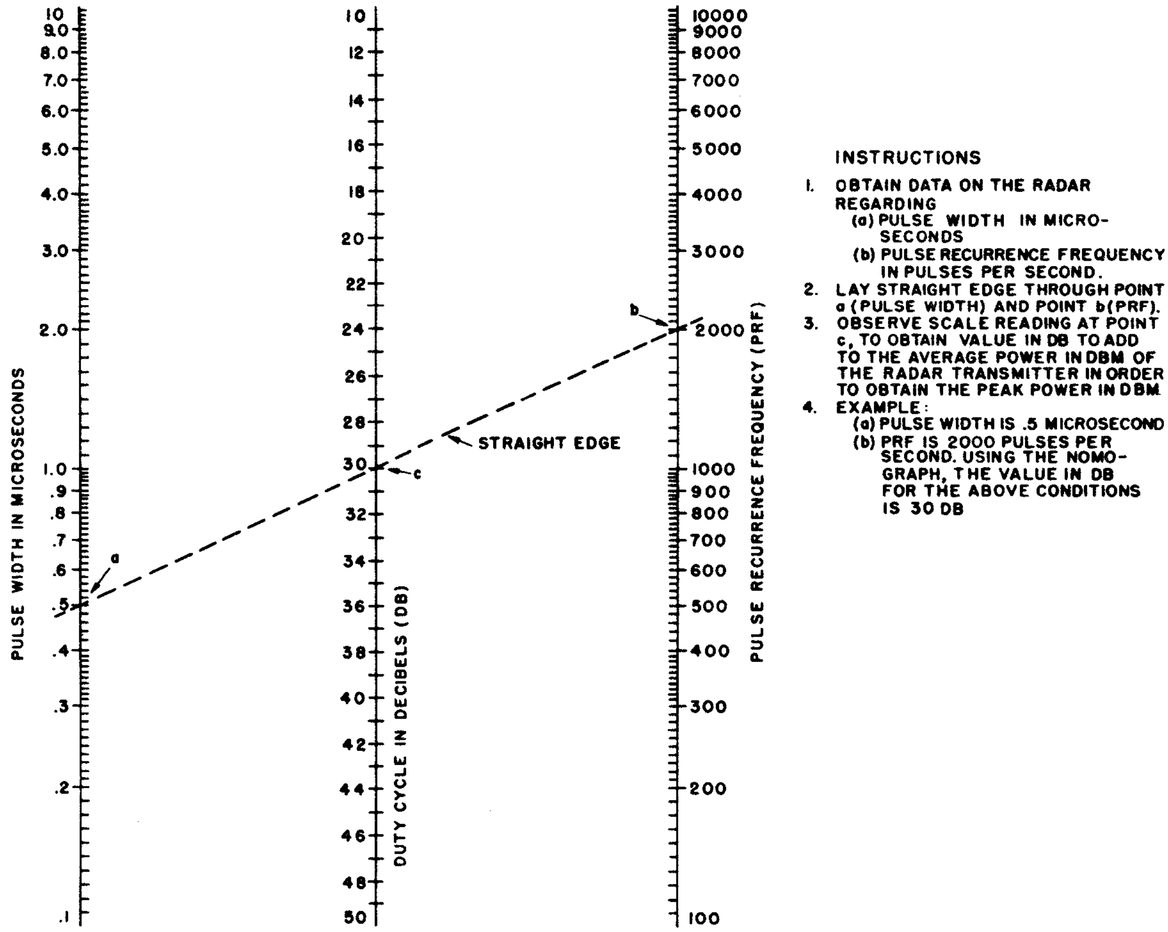


Figure 3-69. Average to Peak (Duty Cycle) Power Conversion Chart

conversion can be made by means of the graphs shown in Figures 3-67 and 3-68. Some test equipments are calibrated directly in dBm, and no conversion is required. The duty-cycle figure, which depends on the duration of the transmitted pulse and on the pulse repetition rate, can be found directly from the nomograph shown in Figure 3-69.

3-6.25 POWER SAMPLING TECHNIQUES

The testing of radar power always requires some method of removing or inserting the power to be measured. There are three principal devices used to accomplish this. They are the test antenna, the RF probe, and the directional coupler.

3-6.26 TEST ANTENNA

The test or pickup antenna consists of a directional antenna array which is broadly tuned to the radar band to be used. This antenna is placed in the radiation field of the radar antenna, and picks up a certain percentage of the radiated signal. The test antenna may be made portable by mounting it on a tripod frame, or it may be fixed by means of a bracket installed as a part of the radar system. It is common practice to locate the pickup antenna at least one diameter of the radar reflector away from the radar antenna, as shown in Figure 3-70A, and to orient the two antennas for maximum pickup. With this procedure, the space attenuation is approximately 30 dB. The exact loss either will be given for the particular installation or must be measured. Any subsequent testing should be done with exactly the same antenna spacing. Another placement method is to clamp the pickup antenna to the edge of the radar reflector in such a manner that the pickup is directed toward the radar antenna feed array. This method is shown in Figure 3-70B. With the pickup in this position, antenna leakage power is used in the test, rather than direct radiation. This procedure has the advantage of allowing operation of the test equipment at various radar antenna positions, and the radar antenna does not require careful orientation. The use of a pickup antenna has the important advantage of testing the entire radar system (including the radome), if the pickup antenna is placed outside the radome. This enables the testing to show operating efficiency, with all controllable factors included.

3-6.26.1 Disadvantages

Four primary disadvantages are associated with the pickup antenna method of sampling power. First, the placement of the antenna is critical; second, antennas are sensitive to frequency changes; third, it is difficult to make tests during radar scanning; fourth, nearby objects can modify the signal picked up by the antenna. Nearby objects, or propagation from other sources, can cause reflections and result in large errors in signal pickup. The presence of these reflections can be detected in the following manner. While observing the signal picked up by the antenna, carefully move the pickup antenna closer to the radar antenna. A smooth increase in signal strength should be noted and, if the pickup antenna is moved away, a smooth decrease in signal strength should be noted. Any sudden or erratic variations or minimum points indicate that nearby objects are influencing the pickup, and another pickup position must be chosen. The test antenna method is used primarily by shipyards and antenna research facilities.

3-6.27 RF PROBE

The RF probe consists of a small capacitive probe inserted into the electrostatic field in the RF transmission line. The greater the penetration of the probe, the greater the power pickup. The penetration of most RF probes is sufficient to provide 20 dB or more attenuation between the main line and the probe output. The probe is fitted with a coaxial connector to facilitate connection to test equipment. In older radar facilities, the RF probe was used extensively, but is now considered obsolete since the development of the directional coupler. The RF probe allows normal radar operation during test, but has some disadvantages. First, reflections from nearby objects in the RF line have a great effect on the attenuation figure; second, probe penetration is very critical; third, the probe is quite sensitive to frequency; and fourth, the attenuation figure depends upon the load connected to the probe.

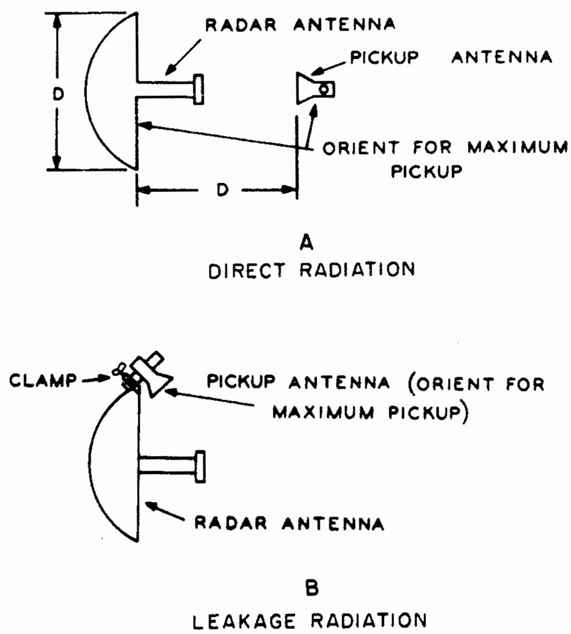


Figure 3-70. Placement of Pickup Antenna

3-6.28 DIRECTIONAL COUPLERS

The directional coupler, as the name implies, couples, or samples, energy only from a wave traveling in one particular direction in the waveguide. By the proper use of one or more directional couplers, reflected signal power can be prevented from affecting the accuracy of power measurements. Figure 3-71 shows a common type of directional coupler, which consists of a short section of waveguide coupled to the mainline waveguide by means of two small holes, and containing a matched load in one end and a coaxial transition in the other end. The degree of coupling between the mainline waveguide and the auxiliary is determined by the size of the two holes.

3-6.28.1 Waveguide Action.

The action of this waveguide is explained by the diagrams in Figures 3-72 and 3-73.

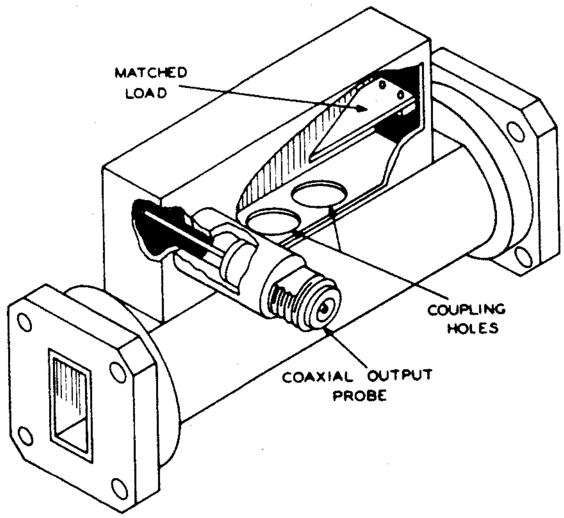


Figure 3-71. Directional Coupler, Cutaway View

In Figure 3-72, power is shown flowing from left to right, and two small samples are coupled out at points C and D. Since the two paths, (represented by C-D-F and C-E-F), to the coaxial probe are the same length, the two samples arrive at point F in phase and are picked up by the coaxial probe. With regard to the paths to the matched load, however, path C-D-F-E is one-half wavelength longer than path C-E because the two holes are one-quarter wavelength apart. Therefore, the two samples arrive at point E 180 degrees out of phase, producing cancellation, and the load receives no power. Figure 3-73 shows the same coupler with power flowing in the reverse direction. Again, samples are removed at points C and D. The two paths (D-F-E and D-C-E) are the same length. The two samples therefore arrive at point E in phase, and are absorbed by the load. However, path D-C-E-F is a half wavelength longer than path D-F, and the resulting 180-degree phase-shift causes cancellation at point E. The result is that the coaxial probe receives power only from a wave traveling from left to right in the main line, thus any reflections causing power to flow from right to left have no effect upon the coupled signal. In practice, the attenuation between the coaxial output and the main line for power flowing from left to right is usually adjusted to be 20 dB or over and is called the nominal attenuation, or simply the attenuation, or the coupling factor. The ability to reject power in the reverse direction is called the directivity attenuation, or simply the directivity, and is usually 20 dB or greater. If a certain coupler has a nominal attenuation of 20 dB and a directivity of 20 dB, the forward attenuation of 20 dB and the reverse attenuation is 40 dB. If the main line carries a 50-kW pulse the forward output is 500 watts peak power and the reverse output is 5 watts peak power. Five watts, as compared with 500 watts, is too small to have any great effect.

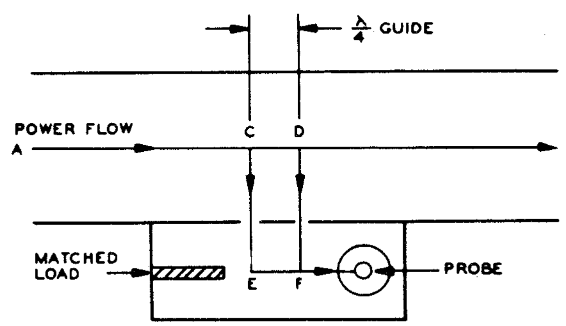


Figure 3-72. Directional Coupler, Direct Power Flow

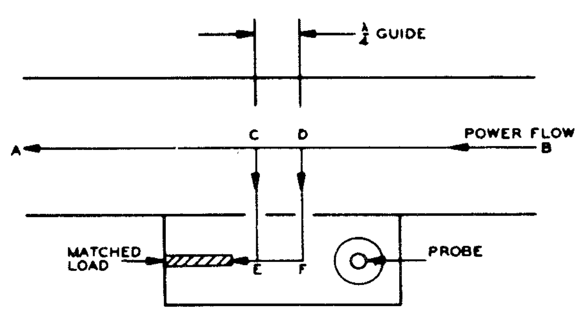


Figure 3-73. Directional Coupler, Reversed Power Flow

3-6.29 BROAD-BAND COUPLER

Forward, or nominal attenuation does not vary rapidly with frequency, but directivity does. The rate of variation can be reduced by the use of designs, whereby the directional coupler can be operated over a broad band of frequencies. One type of broad-band coupler is the 3-hole coupler. If two directional couplers one-quarter wavelength apart are used, a broader bandwidth is obtained. Since the holes are one-quarter wavelength apart, two couplers have one hole common to both. The three-hole coupler therefore employs the action of two directional couplers, and the center hole serves as a common coupling to the two end holes. Another type of broad-band directional coupler is shown in Figure 3-74. In this unit, the coupling holes are one-quarter wave-length apart and elongated. In addition, the two holes are in opposite halves of the main waveguide, which has the effect of causing a 180-degree phase shift between the coupled signals. This phase shift reverses the direction of coupling, so that when power enters at point A, the two signals arrive in phase at the coaxial output; and when power enters at point B, the two signals arrive in phase at the load and are therefore absorbed. The result is that the coupler in Figure 3-74 operates in reverse manner to that shown in Figure 3-72. In this case, the directivity is relatively independent of frequency, but the coupling factor varies rapidly with frequency.

3-6.30 SINGLE-HOLE COUPLER

A third type of directional coupler, shown in Figure 3-75, uses a single hole as the coupling element. This is called the "Bethehole coupler." Through a single hole, waves are excited in the auxiliary guide because of the electric field and the magnetic field in the main guide. Due to the phase relations involved in the coupling process, the waves generated by the two types of coupling cancel in the forward direction, but reinforce in the reverse direction. Therefore, in Figure 3-75, power entering a point A is coupled to the coaxial output, while power entering a point B is absorbed in the dummy load. If the two wave guides were parallel, the magnetic component would be coupled to a greater degree than the electrostatic component, and the directivity would be poor. By placing the auxiliary waveguide at the proper angle, the amplitude of the magnetically-excited wave is made equal to that of the electrostatically-excited wave without changing the latter), and good directivity is obtained. The angle required depends upon the frequency of operation.

3-6.30.1 Directional Couplers

Directional couplers serve as stable, accurate, and relatively broad band coupling devices, which can be inserted into a transmission line so as to sample either incident or reflected power. In most cases a directional coupler is made a part of the radar system and is connected so as to sample the transmitted RF signal. Thus, any undesired reflection from nearby objects, a mismatch between the line and antenna, or line discontinuities are virtually eliminated as a source of error in power measurements. Directional couplers are also made for use with coaxial transmission lines, and operate in a manner very similar to the two-hole coupler.

3-6.31 BIDIRECTIONAL COUPLER

In many cases it is desirable to measure the power reflected from the antenna as well as the direct power from the transmitter. A bidirectional coupler provides a convenient method to measure direct and reflected power. As shown in Figure 3-76, it consists of a straight section of waveguide, with an enclosed section attached to each side along its narrow dimension. Each enclosed section contains an RF pick-up probe at one end and an impedance termination at the other end. The impedance termination (in this case)

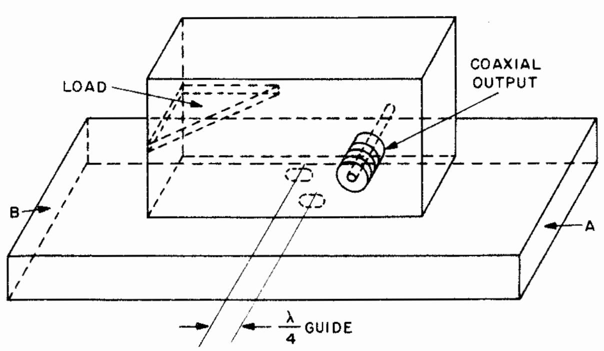


Figure 3-74. Reverse Directional Coupler

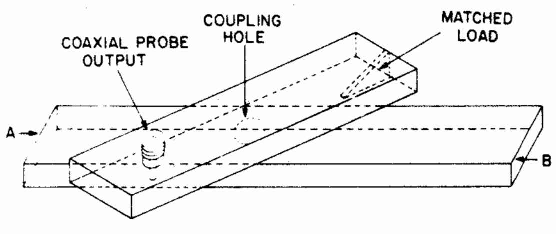


Figure 3-75. Single-Hole Directional Coupler

is in the form of a resistance card. The sections are supplied with energy from within the main waveguide through three openings spaced one-quarter wavelength apart. The RF probe farthest away from the transmitter is used to measure direct power, and the one nearest the transmitter is used to measure reflected power. Energy from the transmitter going toward the antenna enters the enclosed sections through the three openings on each side. Because the openings in each section are spaced one-quarter wavelength apart, the energy travels a quarter wavelength between each of the three openings. The energy coupled into the enclosed section is attenuated by the coupling medium, a predetermined amount below that in the waveguide. As shown in Figure 3-76, the center opening in each of the enclosed sections is larger than each of the holes to either side, thus allowing twice as much energy to enter through the opening. The energy entering the enclosed section farthest away from the transmitter [section (A) of Figure 3-76] is considered first. Part of the transmitted energy enters this enclosed section, and the rest of the energy goes to the antenna. Because of the location and dimensions of the openings in Section A, this energy enters the enclosed section, combines in phase, and is measured by the direct power probe. A power meter connected to the probe gives a direct indication of the transmitted power in the waveguide. The transmitted energy which entered the other enclosed section [section (B) of Figure 3-76] is zero because of the phase displacement of the three openings at that enclosed section. The energy passing the first opening is 180 degrees out of phase with the energy at the third opening. Since the center opening is sufficiently large to supply twice the magnitude of energy as that supplied by either of the two openings, the energy is cancelled. The end of the enclosed section is terminated as described earlier.

3-6.31.1 Reflected Energy

To measure reflected energy caused by either standing waves or energy received from targets,

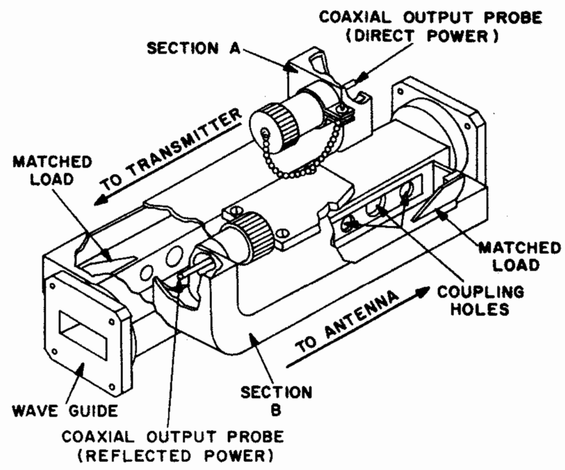


Figure 3-76. Bidirectional Coupler

section (B) acts in exactly the same manner as section (A) when making direct power measurements. Since the direction of energy flow is reversed, the energy now appears at the reflected power probe in section (B). This reflected energy, upon entering section (A), is cancelled out in the same manner that section (B) cancelled the transmitted power when performing direct power measurements. Typical nominal attenuation of a bidirectional coupler is 40 dB for both the direct and reflected outputs. This attenuation must be considered when making power measurements with a bidirectional coupler. The use of the bidirectional coupler for making standing-wave measurements is discussed later in this section.

3-6.32 ATTENUATORS

Attenuators are classified as "dissipative" or "non-dissipative". The cutoff waveguide section is a good example of a non-dissipative attenuator, in that the attenuator merely rejects signals instead of converting them into heat. Dissipative coaxial attenuators are usually short coaxial sections which use resistive material for a center conductor. One such attenuator uses a glass rod; upon which a thin deposit of metal has been sprayed, for a center conductor. Aquadag is sometimes used in place of the metal film.

3-6.32.1 Pads

Short, fixed attenuator sections, called "pads" come in a large variety of forms and loss values.

For example, the CN-42/UP is a 10-dB 50-ohm coaxial attenuator and the CN-43/UP is a 16-dB, 50-ohm coaxial attenuator. The coaxial attenuator can be made variable by constructing the resistive section in two telescoping sections, so that the length, and therefore the attenuation, can be varied.

3-6.32.2 Strips

Dissipative waveguide attenuators consist of strips of resistive material placed inside the waveguide, parallel to the electrostatic field. Where the exact value of attenuation need not be known, the strips are made of bakelite or fiber, with an aquadag coating on one side, as shown in part (A) of Figure 3-77. Calibrated attenuators are usually of the metalized glass variety, an example of which is shown in part (B) of Figure 3-77. As the moveable resistive element approaches the center of the waveguide, the power loss is greater. The resistive element is driven by a dial-and-cam arrangement that is calibrated in dB; the cam surface can be shaped to give any desired spacing of the calibration marks. The ends of the resistive element are tapered to produce as little reflection as possible over a wide band of frequencies.

3-6.32.3 Cascades

Most waveguide attenuators do not have sufficient range to cover all values required in normal use. To overcome this limitation, it is common practice to use two attenuators in cascade, such that the total attenuation is the sum of the individual readings. In some cases, both attenuators are continuously variable. In other cases, only one attenuator is continuously variable and the other is adjusted in graduated steps. One type of attenuator unit has one attenuator with an attenuation range of 7 to 45 dB and the other with a fixed attenuation of either 0 or 35 dB. Thus the combination provides two ranges of attenuation, from 7 to 45 dB and 42 to 80 dB. The over-all range is said to be 7 to 80 dB in two overlapping steps.

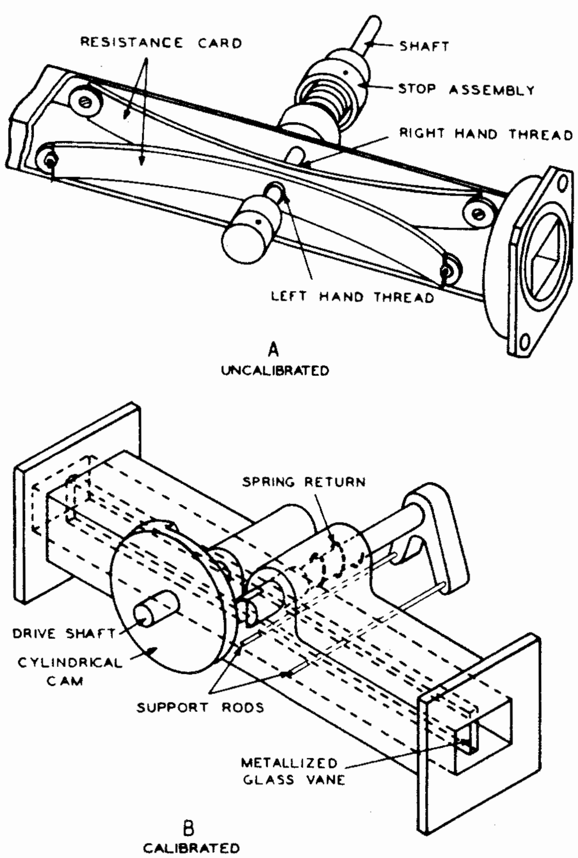


Figure 3-77. Waveguide Attenuators, Showing Construction

3-6.32.4 Power Overloads

Since dissipative attenuators are easily damaged by application of too much RF power, care must be taken to keep the applied power below the maximum rating of the equipment in use. Power overloads cause the resistive element to blister and peel away from the supporting section. This condition, in turn, causes the attenuation value to change, and produces excessive power reflection. Once an attenuator is damaged, it should be discarded. Damage can be detected by inspecting the surface of the attenuator material; it should be smooth and of an even color. The use of a directional coupler practically eliminates the possibility of power burnout, since a nominal loss of 20 dB normally reduces the radar output to a level below the maximum safe value. The following equation provides measurement of the power into the attenuator:

$$Pi = (Po) + (Pc)$$

where:

Pi = Power being applied to attenuator (in dB)

Po = Transmitter average output power (in dB)

Pc = Collective losses; algebraic addition of directional coupler, cable loss, connectors

Pa = Power handling capability of attenuator.

and:

If burnout is to be avoided, the following should be maintained:

$$Pi \leq Pa$$

A common practice is to keep Pi at 3dB below Pa.

A typical example can be found in PMS checks of receiver minimum range when the transmitter is required to be keyed. This test allows the transmitted RF energy to be coupled to the signal generator through a directional coupler and connecting cable. An additional attenuator should be used, as explained below. The input attenuator, as most signal generators, is rated for one half watt (+27dBm) of average power. If a radar having two hundred watts of average power (+53dBm) is coupled to the signal generator through a -20 dB directional coupler and a -3dB loss connecting cable that has been previously calibrated for loss at the frequency being used, the power being applied to the input of the signal generator can be calculated as follows:

$$Pi = (Po) + (Pc)$$

$$Pi = (+53dBm) + [(-20dB) + (-3dB)]$$

Therefore:

$$Pi = +30dBm$$

3-6.32.5 Power/dBm Conversion

Refer to Figure 3-67 for conversion of dBm to power.

$$Pi = +30\ dBm = 1000mw = 1\ watt$$

Therefore excessive power is being applied to the signal generator input circuit. Use of an additional -3dB attenuator is mandatory in this case, and a -6 or -10 dB attenuator is recommended.

3-6.33 ATTENUATION CHECKS

The directional coupler is a type of coupling unit for which the coupling loss is accurately predetermined. Each attenuator is accompanied by a tag or stamped nameplate which states the coupling loss and, in some cases, the mid-band directivity.

3-6.34 CALIBRATION STANDARDS

If reliable attenuation figures are to be attained, some type of calibration standard must be employed. The calibration standard generally used is the cutoff calibrated attenuator. This type has the advantage of possessing linear dial markings; that is, the degree of dial rotation required to change the attenuation from 20 to 23 dB is the same as for a change from 50 to 53 dB. This linear characteristic makes the attenuator very useful for calibration purposes where the exact attenuation is not important but the change in attenuation must be known. Under the conditions where resistive coaxial or waveguide attenuators are used, the technician must necessarily rely upon calibration figures supplied in the maintenance literature accompanying the attenuators. Correct results are obtained as long as the unit is not subjected to mishandling or overload. Once a calibrated resistive attenuator is damaged, either mechanically or electrically, it should be discarded, because the calibration is no longer reliable, and the SWR may be excessive. In the waveguide attenuator employing a metalized glass vane as the resistive element, the glass vane is easily cracked or broken by mechanical shock, and must be handled with care.

3-6.35 CALIBRATION ACCURACY

The calibration procedures described below are usually performed when a particular device is suspected of being in error or is not marked. The accuracy of such measurements is dependent upon the accuracy of the calibrated attenuators and directional couplers used. Errors can result from operational errors caused primarily by carelessness. The precautions listed below should always be observed.

1. Allow all associated equipment to reach operating temperature. A 1-hour warm-up is preferable; however, a half hour can be considered a minimum time.
2. Subsequent testing should be done under as nearly identical conditions as possible, so as to minimize corrections otherwise made necessary by variations in temperature, pressure, relative humidity, and other operating conditions.
3. If conditions do change, apply corrections as specified in instructions or maintenance literature.
4. Make sure that all cable connections and fittings are tight.
5. Repeat those measurements considered critical several times and strike an average of the readings obtained.

3-6.36 CABLE-ATTENUATION CALIBRATION

During a transmitter power test, it is a simple matter to calibrate the attenuation of connecting cables. The method outlined below requires the use of an additional connecting cable whose attenuation figure need not be known, together with a coupling adapter. The procedure for the test is as follows:

1. Use a radar with a supplied directional coupler, if possible. If a pickup antenna is used, be careful not to disturb its position while the following steps are performed.
2. Connect thermistor bridge and calibrated-attenuator power meter to the directional coupler by means of a connecting cable. The attenuation figure for this cable need not be known.
3. Set calibrated attenuator to give 1-mW reference power level on power meter. Note the reading.
4. Using a coupling adapter, connect cable under test in series with cable used in step 2 above.
5. Check all connections for tightness.
6. Decrease reading on calibrated attenuator to produce the reference level on the meter. Note the new reading.
7. The difference in the readings obtained in steps 3 and 6 is the total attenuation of the coupling adapter and the cable under test. Since the loss of the coupling adapter is usually less than 0.1 dB, it may be ignored. Therefore, the cable attenuation is the difference in the readings obtained in steps 3 and 6 above.

3-6.37 ATTENUATOR CALIBRATION

The method of calibrating cable attenuation just described can also be used to calibrate attenuators of both the fixed and variable types. The use of this method requires that the attenuator under calibration have cable-type fittings, or that low-loss adapters be available. Since a cutoff-type attenuator is very nonlinear at low attenuation levels, a stop is provided to prevent operation in the nonlinear region. Therefore, when the attenuator dial reads zero, there is still a fixed loss, called "attenuator zero loss." Attenuators are usually furnished with calibrated charts that give zero loss versus frequency. When an attenuator is used, the dial readings must be increased by the value indicated on the chart for the particular frequency in use. Zero loss can be calibrated by turning the dial to zero and measuring the attenuation in the same manner as in cable attenuation. Each time a different frequency is used, the zero loss should be rechecked. Zero loss can be checked, by the test setup shown in Figure 3-78, without using a second calibrated atteunator, by performing the steps given below:

1. To calibrate a cable which is to be used as reference, use the test method described previously. The loss in this cable should be greater than the attenuator zero loss. If necessary, lossy cable, such as RG-222/U, may be used.
2. Use either an FM or pulsed-type signal generator to provide the necessary signals.
3. With the transmitter not operating, synchronize the signal generator with the same pulse used to trigger the oscilloscope.
4. If possible, use a directional coupler to feed the signals into the radar system. An adapter may be used, but these are not very reliable.

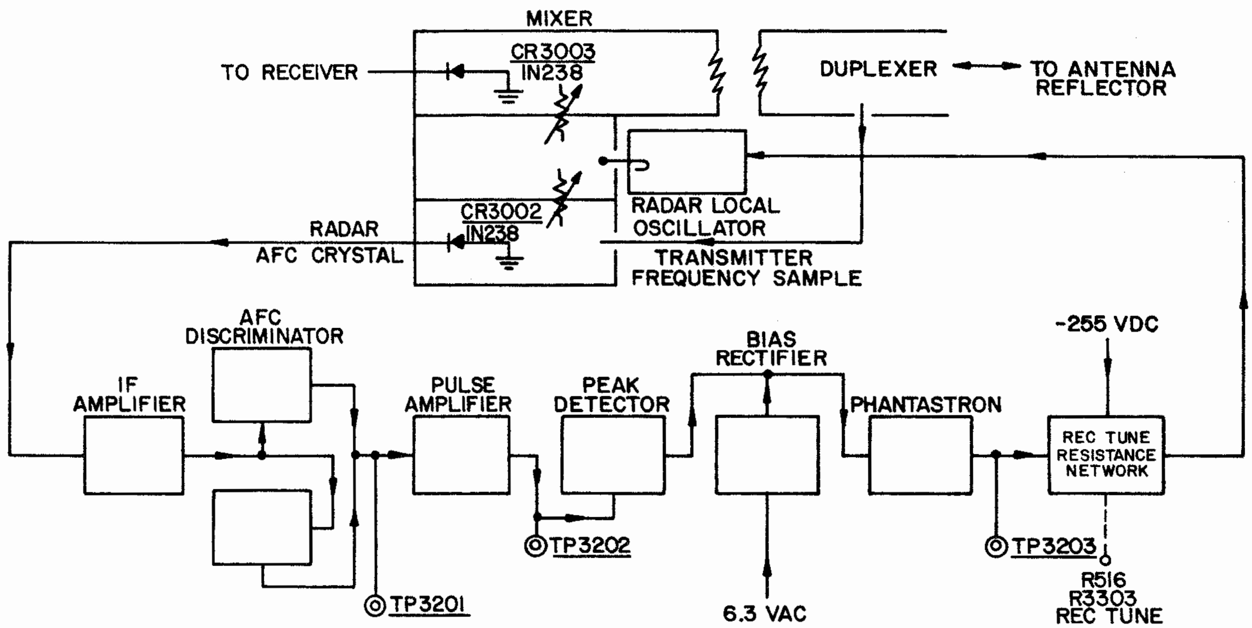


Figure 3-78. Radar AFC. Block Diagram

5. Connect test setup as shown by the broken line in Figure 3-78.

6. Tune radar receiver and signal generator to the desired frequency.

7. Adjust the receiver gain to produce a pulse just a little over one-half-inch high on the oscilloscope. Reduce signal-generator output until pulse is one-half-inch high to prevent the possibility of limiting in the receiver.

8. Remove calibrating cable and restore connections as shown by the solid lines in Figure 3-78.

9. Adjust attenuator for a half-inch pulse amplitude on the oscilloscope.

10. The attenuator zero loss is the difference between the cable loss and the attenuator dial reading.

11. Repeat the above steps, using at least four different frequencies in the operating band, and draw a graph showing attenuation versus frequency. It can be stated generally that attenuation increases with frequency.

3-6.38 RECEIVER PERFORMANCE TESTING

The performance of a radar receiver is determined by a number of factors, most of which are involved and established in the design engineering of the equipment. In the test to follow, only those factors concerned with maintenance will be considered. The most important factors, which will be discussed in detail, are: receiver sensitivity, which includes noise figure determination and minimum-discernible-signal measurement; TR recovery time; receiver recovery time; and receiver bandwidth. Many radar installations include circuits serving a special function. Four of these special circuits commonly encountered are: 1) moving target indication (MTI); 2) instantaneous automatic gain control (IAGC); 3) sensitivity time control (STC); and 4) fast time constant (FTC). These circuits may be found in combination or singly, depending upon the purpose of the radar. In the test methods and procedures about to be described, the special functions should be disabled. If an automatic-frequency-control (AFC) circuit is included in the radar equipment, it may be permitted to operate during receiver tests. A good check on AFC is to make the tests specified for manual tuning, then switch to AFC. If the AFC circuit is normal, the signal indications should not change.

3-6.38.1 Testing Receiver Sensitivity

Inefficient range performance of a radar set can result from troubles in the radar receiver. Loss of receiver sensitivity has the same effect on range as a decrease of transmitter power. For example, a 6-dB loss of receiver sensitivity shortens the effective range of a radar just as much as a 6-dB decrease in transmitter power. Such a drop in transmitter power is very evident in meter indications and, therefore, is easy to detect. On the other hand, a loss in receiver sensitivity, which can easily result from a slight misadjustment in the receiver, is very difficult to detect unless accurate measurements are made. The sensitivity of the receiver determines the ability of the radar set to pick up weak signals. Greater sensitivity, then, indicates that the receiver can pick up weaker signals. Sensitivity of a radar receiver is measured by determining the power level of the minimum discernible signal (MDS). MDS is defined as the weakest signal that produces a visible receiver (scope) output, and its value is determined by the receiver output noise level, which tends to obscure weak signals. It follows, therefore, that an MDS measurement is dependent upon the receiver noise level, and that measuring either one of these quantities will provide an indication of receiver sensitivity. The input to the I-F amplifier does not go through a typical RF amplifier as in HF receivers. Radars in the UHF and above range couple the received signal into a mixer to reduce noise normally generated in the RF head. Figure 3-79 shows a typical mixer and AFC circuits.

3-6.39 NOISE ANALYSIS

In any conductor there is a certain amount of random electron motion resulting from thermal agitation. This motion produces a voltage within the conductor that varies in a random manner. Since this voltage is a pure noise voltage, it produces signals that contain frequencies randomly distributed throughout the RF spectrum. The signals that occur in the portion of the RF spectrum covered by a given receiver are picked up and appear at the receiver as noise. The input power in watts, represented by

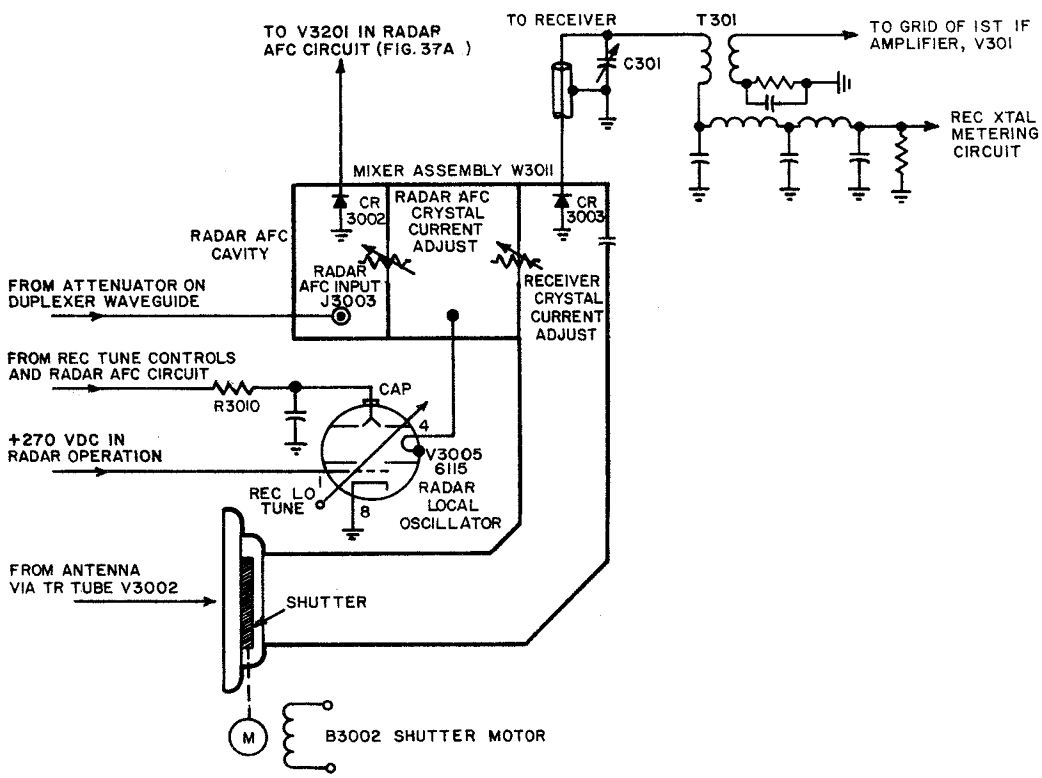


Figure 3-79. Typical Radar Mixer and AFC Circuits

this form of noise is given by the formulas:

Noise power = KT△F

Open circuit noise voltage series with resistor = 4KT△FR

where:

K = Boltzmann's constant (1.37×10^{-23} watt-seconds (joules) per degree Kelv.)

T = Temperature in degrees absolute (Kelvin scale)

= (C°+273)

△F = Range of frequencies involved (bandwidth) in cycles per second

R = Resistance

3-6.39.1 Thermal Agitation

The formula given above shows that thermal agitation noise is determined by bandwidth and temperature. The constant merely serves to convert the noise units into units of power. A decrease of temperature causes less random electron motion, and, at absolute zero, all motion and noise theoretically cease. Since noise covers all frequencies, it is apparent that a greater bandwidth encompasses a greater range of signals, and means more noise power. In a theoretically perfect receiver, this noise could be considered as a voltage across the antenna terminals, and the power represented could be calculated on the basis of temperature and bandwidth. In practice, however, the actual noise developed in a receiver is greater than the calculated value because of the generation of other types of noise within the receiver circuits. For example, a carbon-type resistor, which is comprised of fine particles of carbon, generates a noise signal when current flows through the resistor, because of small changes in the contact area of the particles. Various resistors have widely varying noise levels, and those that are used in the input circuits of a radar receiver must be chosen so as to have as low a noise level as possible. Electron tubes also generate noise signals, because of random variations in electron emission from the cathode, random variations in the current division between the plate and screen grid, etc. Since electron tubes produce noise in proportion to the number of electrodes employed, it follows that triode tubes are generally used where noise limitation is an important consideration.

3-6.40 NOISE FIGURE DETERMINATION

The term "noise figure" (NF), as applied to a radar receiver, indicates the amount of noise that is to be expected. Thus, NF is defined as the ratio of measured noise to calculated noise, and may be expressed as a power ratio or in dB. Therefore, NF can be said to be the input signal-to-noise ratio in comparison to the output signal-to-noise ratio. Since the input ratio is larger than the output ratio, NF is always greater than 1. The input should be at a temperature of 290 degrees Kelvin. In the microwave range of operation, virtually all of the noise originates within the receiver. Atmospheric and manmade noise or static is normally too small to be considered. The three main sources of noise in a radar receiver are: the crystal mixer; the I-F preamplifier (usually the first two I-F stages); and the local oscillator.

3-6.40.1 Crystal Characteristics

Conversion loss and noise figures are functions of the rectified crystal current and are determined by the local oscillator signal level. Noise figure increases and conversion loss decreases as the rectified crystal current increases. These crystal characteristics have been correlated to the front-to-back resistance ratio of the crystal and the back crystal current at one volt. Therefore, a measurement of this ratio and of the back current become indications of the crystal condition. In most cases, the most expedient and successful check of a mixer crystal is to replace it with a new or known good one. It is of interest to note that the output noise of a reflex klystron is much greater than normal when the tube is tuned off the center of a mode. Early radar receivers had noise figures in excess of 20 dB while modern receivers have noise figures of only 0.5 to 18 dB. In general, lower receiver frequencies result in lower noise figures. The noise figure of a radar receiver can be determined by the use of either a noise generator or a CW signal generator.

3-6.41 TEST METHOD USING NOISE GENERATOR

A noise generator produces a random noise signal which covers a frequency range in excess of the radar bandwidth. One such instrument uses a temperature-limited diode, operated at saturation, as the noise-signal source. When a diode is operated under these conditions, the noise produced is proportional to the DC current, the generator frequency range must be adequate, and the diode cannot be used for over 1000 MHz. Therefore, the DC input power can easily be converted to obtain the true noise power. The procedure for determining the noise figure of a radar receiver, using a noise generator, is given in the following steps:

1. Connect a milliameter (0-1 ma) in series with the diode load of the second detector of the receiver.

2. Terminate the receiver input in an impedance equal to the normal source impedance. Adjust receiver gain to produce a 0.5-ma reading. This reading results from noise alone. Remove receiver input from termination and attach generator to the receiver input. Decrease the generator noise excess to zero.

3. Connect noise generator to receiver input.

4. Adjust output of noise generator until meter reads 0.707 ma (1.4 x 0.5). At this point make certain that a further increase in noise causes a corresponding increase in meter reading. If this does not occur, then the receiver is limiting and the readings will not be accurate. In this case, use a lower value of current in step 2; for example, start with 0.3 ma and increase this value to 0.42 ma (0.3 x 1.4) in step 4 (detector is assumed to be linear).

5. The noise generator power output is now equal to the receiver noise power. Note the dial reading. A chart is usually furnished with the instrument for converting the dial reading to power.

6. Calculate the noise figure by using the following formula:

$$NF(dB) = 10 \log \frac{P\ measured}{P\ calculated}$$

where "P measured" is the amount of noise being fed to the receive by the noise generator. This figure must be in micromicrowatts. "P calculated" is the figure arrived at by using the following formula:

$$P_c = 4KT\Delta F I^2_{dc} R_s$$

K = Boltzmann's constant [(1.37 x 10^{-23} watt-seconds (joules per degree Kelvin)]

T = Ambient temperature of the equipment under test measured in degrees Kelvin (°C +273)

ΔF = Receiver bandwidth (in cycles per second)

I_{dc} = DC plate current of the noise diode

R_s = Source resistance of the antenna

As an example, to calculate the noise figure of a receiver having a bandwidth of 4 megacycles (4x 10^6), and ambient temperature of the equipment in degrees centigrade at 20(20+273):

$$\begin{aligned} NP &= 4KT\Delta F \\ &= 4 \times 1.37 \times 10^{-23} \times (20+273) \times (4 \times 20^6) \\ &= 6422.56 \times 10^{-17} \\ &= 0.06423 \text{ picowatts} \end{aligned}$$

The measured noise power being fed to the receiver under test is 1.018 picowatts. Using the formula below, the noise figure of the receiver is arrived at as follows:

$$NF(dB) = 10 \log \frac{P\ measured}{P\ calculated}$$

$$\begin{aligned} NF(dB) &= 10 \log \frac{1.018}{0.0642} \\ &= 10 \log 15.85 \\ &= 10 \times 1.2 \\ &= 12 \text{ dB} \end{aligned}$$

The noise figure for the receiver under test is calculated as 12 dB, this being a normal noise figure for this radar receiver. A gaseous discharge tube may be used as a noise source. The noise figure (NF) is doubled when the noise generated is turned on and is given by:

$$NF_{(c/B)} = 15.4 - \text{Attenuation (dB)}$$

Figure 3-80 illustrates this procedure when the noise figure is greater than 15.4 or when a calibrated output power meter is being used.

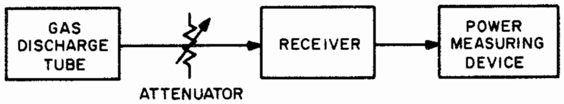


Figure 3-80. Gas Discharge Tube Noise Source

3-6.42 TEST METHOD USING CW SIGNAL GENERATOR

The CW method of measuring the noise figure uses a calibrated signal generator in the same manner as the noise generator. This method is not as accurate as the noise-generator method because the detector characteristics of the receiver under test may affect the ratio of signal power to noise power.

3-6.43 MINIMUM DISCERNIBLE SIGNAL MEASUREMENT

The measurement of a minimum discernible signal (MDS) consists of measuring the power of a pulse whose level is just sufficient to produce a visible receiver (scope) output. It follows that if a radar receiver has the specified MDS level, the noise figure should be correct also. Therefore, measurement of the MDS is a satisfactory substitute for a noise-figure determination, and is less complicated. Correct pulse length must be used, and when readings are taken periodically for comparison purposes, the identical pulse length must be used each time.

3-6.44 RF LEAKAGE DETERMINATION

In the measurement of MDS, a very high degree of attenuation (approximately 98 dB for the average radar) and a very low power level (about one pico) are involved. Because of these factors, very little RF leakage from the signal generator can be tolerated, or the amount of leakage signal picked up by the receiver will be appreciable, compared to the signal fed through the attenuator. Since leakage signals are independent of attenuator setting, very inaccurate MDS readings can be obtained when leakage is present. If the leakage signal reaches the receiver in phase with the signal through the attenuator, the MDS reading will be low, and thus will indicate that the receiver sensitivity is much better than it actually is. In such a case there is a good possibility that a defective receiver may appear to be normal. On the other hand, if the leakage signal reaches the receiver out of phase with the signal through the attenuator, the MDS reading will be high, and thus will indicate that the receiver sensitivity is worse than it actually is. In the construction of a signal generator, special attention is given to the problem of minimizing RF leakage. The RF oscillator is carefully shielded, and then it and the attenuator assembly are enclosed in a second shield which serves as the case of the instrument. In addition, all connecting cables and couplings are provided with shields and close-fitting connectors. In spite of these precautions, however, a small amount of RF leakage exists, even in the most modern equipment. The presence of leakage makes it imperative to locate all equipment associated with MDS tests outside the radar-antenna radiation field. In addition, the equipment should never be operated outside its case or with loose cable connections. Also, on early types of signal generators where a door is provided on the front panel for access to the oscillator adjustments, the door must be kept closed during measurements. These precautions must be observed; otherwise erroneous measurements will be obtained.

3-6.45 LEAKAGE DETECTION METHOD

The presence of RF leakage is detected by the following method. Determine the MDS level, then move the test set to another position and determine the MDS level at that point. Observe whether the first MDS reading differs from the second; if it does, RF leakage is present in one of the two positions. When leakage is found to be present, locate the test set as far from the radar antenna and receiver as practicable. Find that position where movement of the test set does not affect the MDS. In general, if rotation of the test set does not change the MDS level, the RF leakage can be considered negligible.

3-6.46 MDS MEASUREMENT USING PULSED-SIGNAL GENERATOR

A pulse generator block diagram is shown in Figure 3-81. To provide internal sync, a small portion of the output is applied to a crystal detector, and the resulting rectified output signal is made available as a trigger to the pulser circuit. Thus, the radar pulse can produce sync automatically, and, if desired, external sync may be used. The RF oscillator may operate in either CW or pulsed condition, and may be turned off when desired. With the oscillator off, the equipment may be used to measure transmitter power and frequency. The purpose of the uncalibrated variable attenuator, sometimes called the "power set," is to drop the output of the RF oscillator to the standard 1-mw level used in the test set. The thermistor bridge monitors the power entering the calibrated attenuator. Calibration of this attenuator is such that the dial reading includes the zero loss.

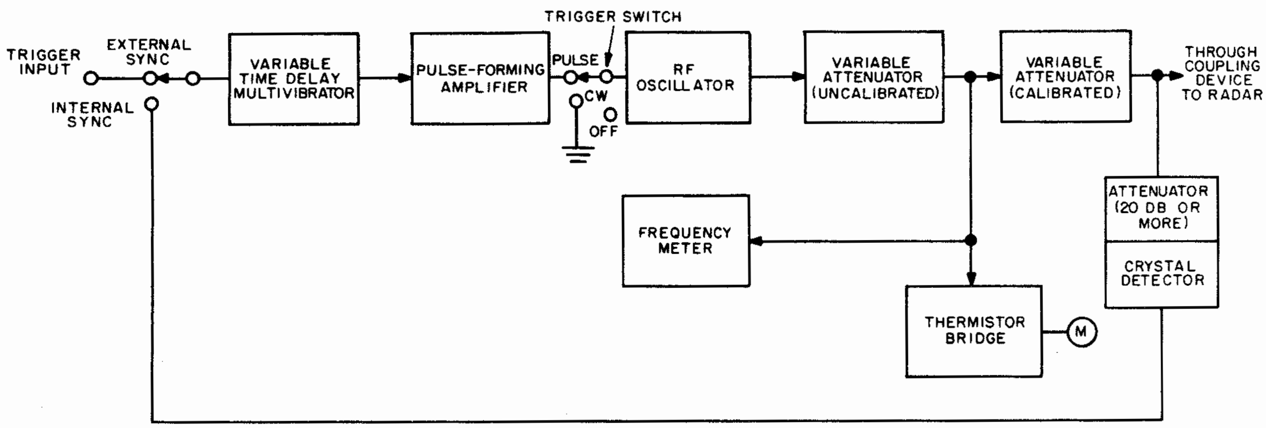


Figure 3-81. MDS Measurements Using Pulsed RF Signal Generator

3-6.46.1 MDS Measurement

An MDS measurement, using a pulsed signal generator, is given in the following steps:

1. If internal sync is not to be used, connect radar trigger pulse to trigger input.
2. Connect attenuator output to coupling device. A directional coupler is usually employed, but a waveguide terminator may be used.
3. Set function switch to the off position and zero the thermistor bridge.
4. Turn function switch to CW and adjust uncalibrated attenuator for midscale reading on thermistor bridge.
5. Using the frequency meter, adjust the RF oscillator to the radar frequency. The freqeuncy meter should be kept detuned, except during frequency checks.
6. Set the uncalibrated attenuator accurately for a l-mw bridge reading.
7. Set function switch to the pulse position.
8. Adjust the calibrated attenuator for the MDS indication previously described in the early test method. The variable time delay must be set so that the artificial echo does not occur at the same range as a radar target.
9. Find the total attenuation. The value obtained is the receiver MDS in terms of -dBm. Total attenuation = coupling loss + cable loss + attenuator reading (all in dB). If desired the -dBm value may be converted into terms of power by utilizing the chart shown in Figure 3-67, but the dBm reading conveys more information to the technician.

3-6.47 MDS MEASUREMENT USING FM SIGNAL GENERATOR

Use of pulsed-signal generator for the measurement of MDS has one major disadvantage: the accuracy of the results depends upon how accurately the signal generator is tuned to the radar frequency. This difficulty has been overcome in the design of an FM signal generator incorporated in a test set, as shown in Figure 3-82. The RF section of this signal generator is very similar to that of the pulsed generator shown in Figure 3-81. Since the RF oscillator is a reflex klystron, the oscillator frequency may be varied by means of a sawtooth voltage fed to the repeller plate of the tube. This voltage is developed by a sawtooth generator activated by a trigger pulse. The sawtooth voltage rise is nearly linear and lasts for about 50 microseconds. This voltage is fed to a circuit shown in basic form in Figure 3-83, which contains two controls. The signal-width control determines the amplitude of the sawtooth fed to the klystron, and the phase control determines the level of the sawtooth by supplying a variable negative DC voltage to fix the operating point of the klystron repeller. When the signal-width control is in the maximum clockwise position, no sawtooth voltage is applied to the tube's repeller and the signal generator supplies a CW output. Hence with the signal-width control in this position, the klystron mode pattern can be explored by manually varying the repeller voltage with the phase control. In operation, the FM generator creates an artificial echo pulse by rapidly sweeping an RF signal through the receiver pass-band at a given time after the transmitter fires. This action is shown in Figure 3-84. The transmitter pulse starts the sawtooth sweep, which

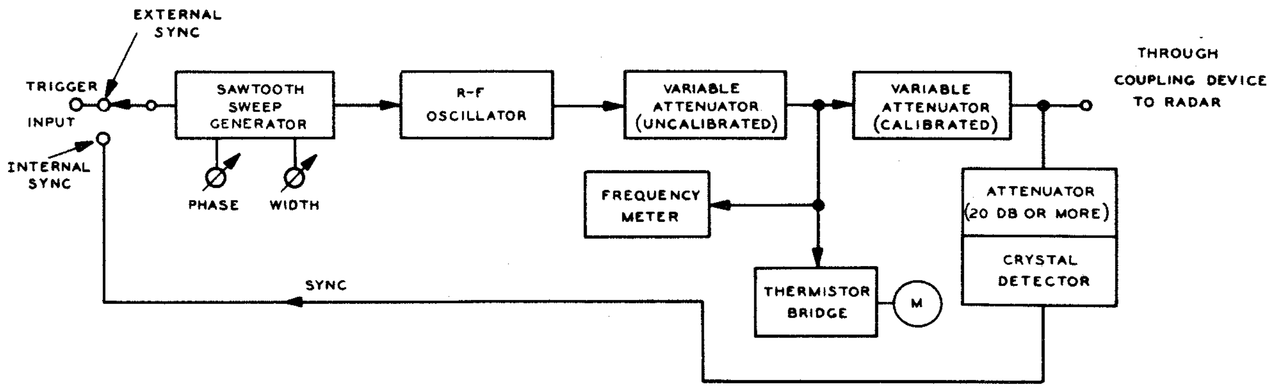


Figure 3-82. MDS Measurement Using FM Signal Generator

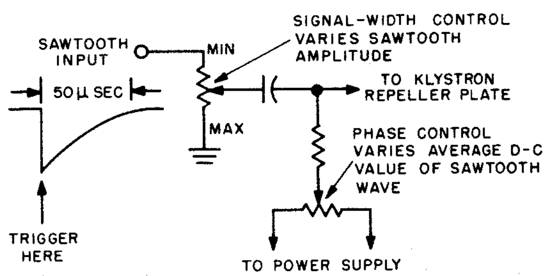


Figure 3-83. Signal-Width and Phase-Control Circuit

varies the signal frequency. As the generator frequency is swept through the receiver passband, a plot of frequency versus output is obtained, with the result that a pulse, very similar to an echo pulse, is reproduced on the radar "A" scope. The width of the pulse obtained depends upon the receiver bandwidth and the rate at which the frequency is swept. For example, if the receiver bandwidth is 4 MHz and the klystron frequency is swept at a rate of 4 MHz per microsecond, the signal will be within the frequency range of the receiver pass-band for only one microsecond; hence, the pulse seen on the "A" scope will be one microsecond wide. The effect of decreasing the amplitude of the sawtooth voltage is shown in Figure 3-84. A decrease in amplitude allows the frequency to be swept at a lower rate and and thus results in a wider pulse on the "A" scope. By varying the signal-width control, the desired pulse width can be obtained. When measuring MDS, the pulse width is usually made equal to the transmitter pulse width; however, this is not a critical factor.

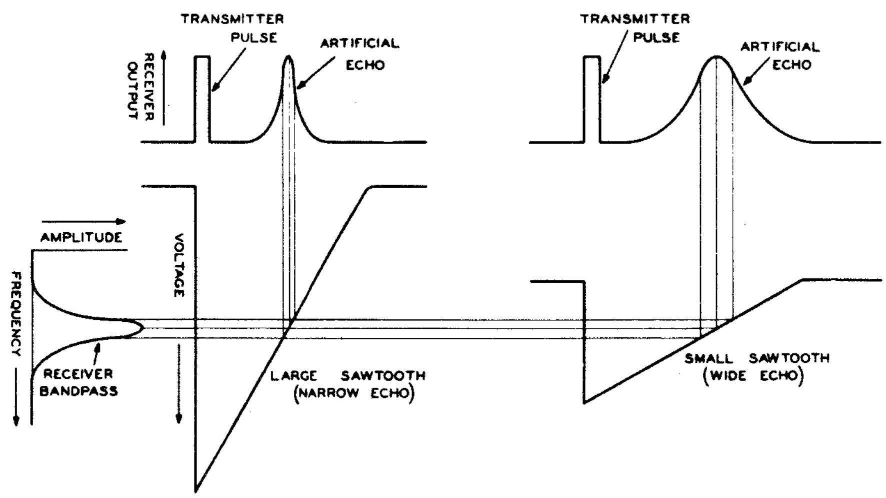


Figure 3-84. Effect of Sawtooth Amplitude on Presentation of Artifical Echo

3-6.47.1 Range Control

The phase control is really a range control, although it is seldom labeled as such. The higher the setting, the greater the echo range. Because each klystron mode produces an echo indication, increasing the setting of the phase voltage sufficiently with the signal-width control makes it possible to obtain a series of echoes. One echo is thus received for each mode, with the one for the most negative mode having the shortest range. However, for normal operation, only one mode is used.

3-6.47.2 MDS Measurements

When an FM signal generator is used, the procedure for measuring MDS is given in the steps below:

1. Connect radar trigger-pulse output to trigger input jack on FM signal generator. (Omit this step if internal sync is to be used.)
2. Connect RF input through coupling device to radar. (A directional coupler is preferred.)
3. Set signal-width control to maximum.
4. Adjust phase control for maximum klystron output as indicated on thermistor bridge.
5. Tune klystron (cavity) to approximate radar frequency by adjusting the frequency meter to the frequency of the radar transmitter and tuning the klystron for a dip in the thermistor-bridge meter reading. Since the klystron mode is fairly broad, extreme accuracy in tuning the klystron is not necessary. For example, the width of the flat portion of an X-band klystron is about 10MHz; therefore, the tuning accuracy required is ± 5 MHz.
6. If necessary, adjust the phase control again for maximum output. If a sizable adjustment is required, repeat step 5.
7. Adjust uncalibrated attenuator for a 1-mw indication.
8. Set receiver gain control for a 1 cm noise level on the oscilloscope.
9. Reduce the signal width control setting until the pulse is seen on the oscilloscope. This will need to be done at a low setting on the calibrated-attenuator control.
10. If necessary, adjust phase control to position echo pulse in a target free area.
11. Adjust signal width control for desired echo-pulse width.
12. Set calibrated attenuator for the MDS previously described. Rock the phase control during this step to distinguish the echo more easily.
13. Find the total attenuation in dB. The value obtained is the MDS in dB below 1 mW (-dBm). Total attenuation = coupling loss + cable loss + attenuator reading (all in dB).

3-6.48 CW (DOPPLER) - TYPE RECEIVER CONSIDERATIONS

Since the maximum Doppler frequency (difference frequency) is dependent upon the maximum relative velocity between target and receiver, the bandwidth of the receiver need only be wide enouth to pass the entire expected band. Ordinarily, the range of the Doppler frequency is from about 1 to 10 kilohertz. The receiver operates with a filter gate (selective filter) generally following the video amplifiers. This effectively reduces the bandwidth and allows only the range of Doppler effected frequencies to pass, thus increasing the signal-to-noise ratio. In this type radar, the same antenna is generally used simultaneously for transmitting and receiving so that the receiver must be isolated from the transmitter. The transmitter energy is supplied to the waveguide thruplexer plane-polarized such that the electric vector (plane of polarization) is parallel to the vertical wall of the thruplexer. The receiving crystal detector is sensitive only to energy reflected in the horizontal plane.) Receiver tests of this type equipment therefore require that energy being fed into the receiver be properly polarized. The output of the receiver is limited before being detected so that the dc output will be directly proportional to the frequency deviation. The dc voltage can then be sent to a counting-type indicating device.

3-6.48.1 Receiver Gain

This equipment is rarely used for detecting distant targets; therefore it is seldom necessary to amplify exceedingly weak signals. However, since the gain of the receiver must be sufficient to cause limiting of all signals received for accurate operation, the sensitivity of the receiver should be checked periodically. The sensitivity of this receiver can be checked with an FM signal generator set for an appropriate deviation modulation. When possible, the signal is fed through the antenna so that the sensitivity of the entire radar set can be checked. The output indication is best obtained from the input of the limiter. If an FM signal generator is not available, a standard generator may be used, but the receiver sensitivity should be checked at several points throughout the Doppler range. Procedures for checking Doppler receiver sensitivity are the same as for communications receivers.

3-6.48.2 CW Radar

CW radar equipment usually operates in the microwave region and most of the noise is inherent in the receiver. Therefore, a much better method of determing the performance of the receiver is to take a noise figure measurement, since noise is the limiting factor in determining maximum sensitivity. Besides the considerations and procedures regarding noise-figure measurements that are given under RECEIVER

PERFORMANCE TESTING discussed previously, the bandwidth of a CW radar receiver should be taken to include the filter gate (selective filter), since a much higher figure can be obtained from the resulting narrowing of the bandwidth. The output measurement is obtained by measuring the input signal of the first limiter tube.

3-6.49 TESTING RECEIVER BANDWIDTH

Receiver bandwidth is defined as the frequency spread between the half-power points on the receiver response curve. Receiver bandwidth is specified for each radar, although wide variations are often tolerated. If either the bandwidth or the shape of the receiver response curve is not correct, a considerable change in the value of circuit components is required to alter the response materially. The receiver response should be checked after an extensive repair to any IF amplifier. Figure 3-85 shows a typical response curve of a radar receiver. The half-power points are shown as 3 dB below maximum (mid-frequency) response. Since the curve is plotted in terms of voltage, these points are also represented by the 70.7 percent voltage points (1/2 = 0.707 as shown in the figure.)

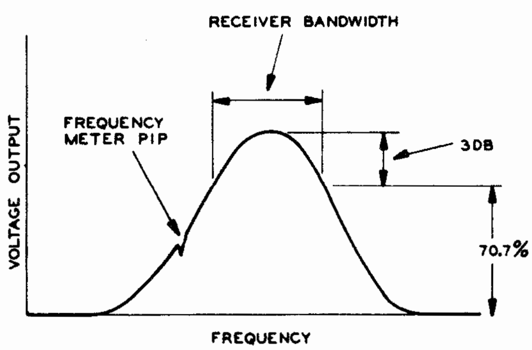


Figure 3-85. Receiver Response Curve

3-6.50 INTEGRATED RECEIVER METHOD

The bandwidth test procedure given is used when the receiver is operating as an integral part of the radar facility, and can very easily be reformed after checking the MDS. When the radar receiver is to be tested as an individual component, another test procedure is used which will be described immediately following this method. The integrated receiver method, which is considered superior to other methods, makes use of the test setup for measuring MDS, using a sweep signal generator as previously described. The procedure is given in the following steps:

1. With the equipment connected in the same manner as for an MDS measurement, turn the signal-width control to obtain a response curve about 1-cm wide.
2. Reduce receiver gain such that the noise amplitude is just barely visible on the scope.
3. Adjust calibrated attenuator to produce a pulse amplitude below receiver saturation level.
4. Tune frequency meter until response curve shows an absorption pip at one of the half-power points. Read the frequency, then repeat for the other half-power point. The difference between these two frequencies is the receiver bandwidth.

When the foregoing procedure is used, the half-power points may be located very easily as outlined in the following steps:

1. Note the attenuator dial reading following step 3 given below.
2. Increase the attenuator reading 3 dB and mark the level at the top of the response curve.
3. Return the attenuator to the previous setting.
4. The half-power points are at the level marked in step 2.

3-6.51 PREFERRED RECEIVER METHOD

This method is employed when the radar receiver is to be tested as an individual component rather than an integral part of the radar equipment. Figure 3-86 shows the test setup for checking a receiver which is detached from the radar facility. A sweep generator produces a variable-frequency signal that is fed into the receiver IF input.

NOTE

> The sweep width and the center frequency of an FM signal generator is adjustable to cover any standard radar intermediate frequency.

The receiver video output is fed to the vertical-deflection circuit of an oscilloscope. In addition, a sync voltage is supplied by the sweep generator to maintain horizontal motion of the electron beam in synchronism with the frequency sweep. The oscilloscope, therefore, indicates frequency horizontally and receiver output vertically. A second signal generator, called the marker oscillator and which is internal to the sweep generator, produces an accurately calibrated CW signal which is mixed with the sweep generator output. When the varying sweep passes the marker-oscillator frequency, a beat signal results producing a marker pip on the response curve as shown in Figure 3-87. The marker-oscillator dial indicates the fequency at which the pip occurs.

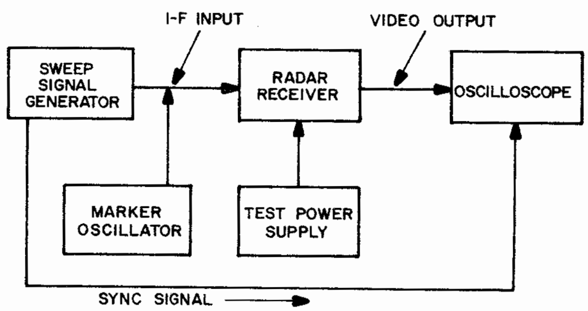


Figure 3-86. Test Setup for Checking Receiver Response

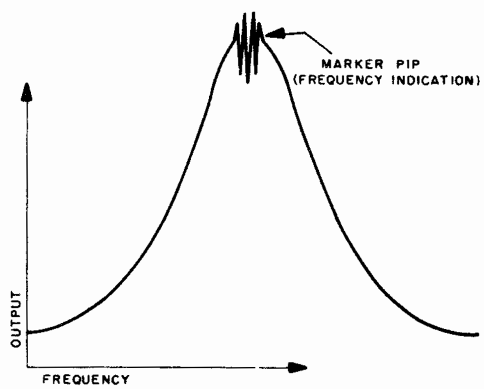


Figure 3-87. Response Curve, Showing Marker Pip at Mid-Frequency Point

3-6.51.1 Marker Pips

To check receiver bandwidth using the test arrangement discussed above, the marker pip is positioned until it rests at the 70.7-percent point on the curve and the frequency dial is read. The frequency at the other half-power point is determined in the same manner. The spread between these two points, expressed in frequency, is the measured bandwidth. In most cases, the radar receiver does not need to be removed and cable connections can be made to the input and output of the receiver. The principal of bandwidth testing can be applied to any part of a radar that has a bandpass or band-stop system. By using the sweep generator to supply a broad source of RF energy and an oscilloscope to monitor the system response, adjustments can be made or repairs facilitated. When RF signals are being passed through non-amplifying circuits, such as passive preselectors or filters, a spectrum analyzer should be used.

3-6.52 TESTING TR RECOVERY TIME

The time required to permit TR recovery is determined by the time it takes the TR switch to deionize after each transmitter pulse. It is usually defined as the time required for the receiver to return to within 6 dB of normal sensitivity after the end of the transmitter pulse. However, some manufacturers use the time required for the sensitivity to return to within 3dB of normal or to full sensitivity. The TR recovery time is one of the factors that limits the minimum range of a radar because the radar receiver is unable to receive signals until the TR switch deionizes. The recovery time may vary from about 0.3 to 20 microseconds, depending upon the particular radar set.

3-6.52.1 TR Function

The primary function of the TR section is to protect the crystal detector from the powerful transmitter pulse. Even the best TR switches allow some power to leak through, but, when the switch is functioning properly, the leakage power is so small that it does not damage the crystal. It has been found, however, that the useful life of a TR tube is limited because the amount of leakage and the recovery time increase with use. To ensure efficient performance, some technicians make it a policy to replace the tube after a given number of hours. A better practice is to measure the TR recovery time at frequent intervals, as called for in preventive maintenance procedures, and make a graph or chart, which will immediately disclose any change in performance. Figure 3-88 shows in an approximate manner how recovery time is correlated with leakage power. Note that the end of the useful life of the TR tube is indicated by an increase in recovery time increases before the leakage power becomes excessive. In practice, the TR tube is re-placed when any sharp increases in recovery time become apparent. Ambient temperature also has an effect on recovery time. The colder a TR tube, the greater its recovery time. For example, tube type 721A recovers in about 7 microseconds at 28 degrees C; however, at -186 degrees C, the recovery time is about 100 microseconds, and at -20 dgreees C, the recovery time is about 14 microseconds. When tests are conducted under widely varying temperature conditions, this effect must be considered.

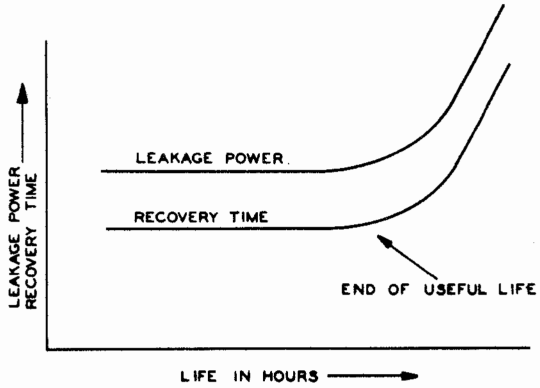


Figure 3-88. Graphic Comparison of TR Recovery Time and Leakage Power

3-6.53 PULSE OR FM SIGNAL GENERATOR TEST METHOD

TR recovery time can be tested by means of a setup that uses either an FM or a pulsed signal generator. When an FM or pulsed-signal generator is used, the TR recovery time test is conducted by performing the following steps:

1. The same test setup used for the measurement of MDS is utilized for this test (Figure 3-81).
2. Set receiver gain to indicate about 1 cm noise on the oscilloscope.
3. Adjust calibrated attenuator to give a pulse amplitude about halfway between the noise level and saturation. Note the attenuator reading.
4. Reduce attenuator setting by 6 dB.
5. Rotate phase control (time delay) to position the pulse closer to the transmitter pulse. Continue rotation of control until the pulse amplitude drops to the level established in step 3.
6. Read the range at which the pulse is now located. The value obtained is the "6 dB recovery time" or the recovery time to within 6 dB of normal receiver sensitivity. Recovery time may be indicated in either microseconds, miles, or yards as long as subsequent readings are in the same units.

If desired, the test procedure given above can be modified so that the "full-sensitivity TR recovery time" is measured. Steps 5 and 6 are then modified as follows:

5. Omit
6. Using the phase control, move pulse toward the transmitter pulse until the pulse amplitude just starts to decrease. The use of this modified procedure will result in a longer recovery time, but when the results of a series of measurements are plotted, the curve obtained will be similar to the one shown in Figure 3-88. Consistency is the most important factor; therefore, the type of recovery measurement used should always be noted in the maintenance records.

3-6.54 CW SIGNAL GENERATOR METHOD

It is also possible to test TR recovery time by using a CW signal generator. The test is conducted by performing the steps listed below:

1. Connect signal generator to coupling device.
2. Adjust radar-receiver gain to indicate about one cm of noise on "A" scope.
3. Tune signal generator to the frequency of the radar receiver. Proper tuning will be evidenced by a rise in the "A" scope trace.
4. Adjust output of signal generator (calibrated attenuator) to a point just below receiver saturation. The indication should now appear similar to that shown in Figure 3-89.
5. Measure the range between the transmitter pulse and the point on the "A" scope where the noise amplitude is one-half of the maximum noise level.

The procedure described above gives the TR recovery time to within 6 dB of normal. If the full sensitivity TR recovery time is desired, the time is measured to the point where the noise just barely reaches full value. In many cases, nearby targets will interfere with the testing of recovery time. When this occurs, position the radar antenna so as to point it at free space. In case this still does not eliminate the interference, the use of an absorption screen or a dummy RF load is recommended.

3-6.55 CURRENT AND VOLTAGE CHECKS

One method used in testing a TR tube is to measure the keep-alive current. This current maintains the TR tube partially ionized to make the firing more reliable, and thus helps protect the crystal. The current is usually about 100 microamperes, and falls off as the end of the TR tube life approaches. Another method is to measure the keep-alive voltage between the plate and ground of the TR tube when the voltage is known to be good, and to record this voltage for use as reference for future checks. However, these checks are not as reliable as a recovery-time test.

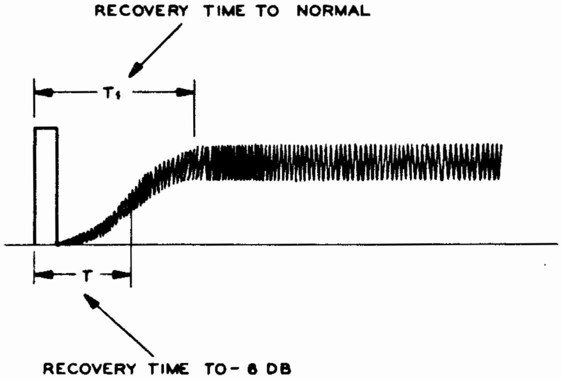


Figure 3-89. TR Recovery Test Indication Using A CW Signal

3-6.56 TESTING RECEIVER RECOVERY TIME

Radar-receiver recovery time is defined as the time required for the receiver sensitivity to return to normal after a saturation echo is received. This time is determined in the original radar design and is of very short duration. Receiver recovery is not discussed in terms of minimum range since TR recovery is much longer. The receiver recovers from a transmitter pulse long before the TR tube recovers. Figure 3-90 illustrates the effect of receiver recovery. Note that immediately following the echo pulse the noise is at reduced amplitude, and that the recovery time is the period of reduced noise level. No absolute measurement of receiver recovery is necessary. A noticeable time interval, however, usually indicates trouble. From that point, the AGS and STC circuits should be checked for proper operation if they cannot be disabled during testing.

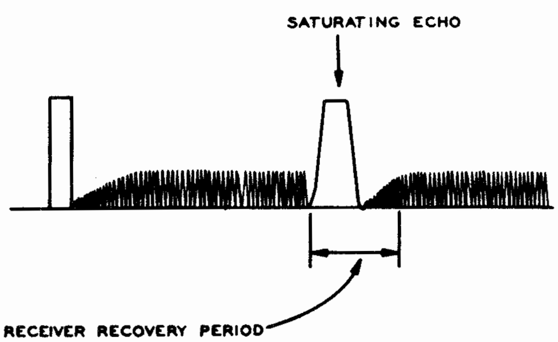


Figure 3-90. Receiver Recovery Time

3-6.57 TRANSMITTER PERFORMANCE TESTING

The performance of a radar transmitter is determined by several factors, most of which are evolved and established in the design engineering of the equipment. In the text to follow, only those factors involving maintenance will be considered. The most important factors which will be discussed are the magnetic field of the magnetron magnet, pulse repetition rate measurements, pulse duration measurements, and the measurement of the modulator pulse.

3-6.58 MAGNETRON MAGNETIC FIELD

The magnetron is a high-power transmitting tube used almost universally in modern microwave radar. As a source of high-power microwaves, the multicavity magnetron represents a very great advance over both conventional space-charge and velocity-modulated or medium power klystron-type tubes. Magnetrons produce pulse power on the order of hundreds of kilowatts at frequencies as high as 24,000 MHz. These tubes are basically self-excited oscillators whose purpose is to convert DC input power into RF output power. Magnetrons are generally constructed so that they are inserted between the pole pieces of a permanent or electromagnet.

3-6.59 PRECAUTIONS

Certain precautions should always be taken when handling magnetron magnets. The field strength of these magnets is greatly reduced if they are jarred or hit even lightly. The magnetic field is very strong and if magnetic tools are used when working close to the magnet, the strong field may pull them sharply against the magnet.

3-6.59.1 Nonmagnetic Tools

A test was made to determine the effect of allowing tools to strike a magnet, and it was found that only one touch with a steel screwdriver reduced the main magnetic field by 50 gausses. Since the magnetron used with this particular magnet was designed to operate properly within limits of 50 gausses above or below its rated 2500 gausses, two or three such light taps on the magnet would seriously affect its performance. This difficulty can be avoided by the use of nonmagnetic tools. A nonmagnetic screwdriver is essential. If magnetic tools must be used, special precautions must be taken to prevent them from jumping toward the magnet. A nonmagnetic cover such as cloth or tape, wrapped around the pole pieces to a depth of three sixteenths inch, reduces the effect of touching a screwdriver to the magnet to about one-tenth of what it would be without the cover. In general, iron, nickel, and other magnetic objects should not be brought near the magnet. When drilling or filing in the vicinity of the magnetron assembly, cover it completely so as to prevent any metal filing from becoming attached to the magnet. During storage, care should be taken to prevent the interaction of the fields of two or more magnets. A safe rule is to allow not less than six inches between them. In addition, always store a magnet with the keeper in position.

3-6.60 MAGNETRON MATERIAL STORAGE

All material having a strong magnetic field should be stowed away from cables carrying heavy current. Such cables, in themselves, produce a strong magnetic field and can induce or remove magnetic properities in nearby ferrous metals.

3-6.61 PULSE REPETITION RATE MEASUREMENTS

Pulse repetition rate (PRF) (also termed pulse recurrence frequency) is an important factor in radar performance. The pulse rate is a direct function of average transmitter power, and also serves to set the upper limit or maximum range of a radar equipment. Short range radars (e.g., fire-control or ground controlled approach radars) have a high PRF, while long-range (search) radars have much slower rates, since the required "rest" time is considerably longer for the greater ranges. Pulse recurrence frequency of 200 to 400 pulses per second are typical for radar equipments operating at 250 or 300 mile ranges. Fire control and ground approach radars have pulse repetition rates that are as high as several thousand pulses per second. Regardless of the type of pulse radar, the PRF is in the low audio frequency range.

3-6.61.1 Frequency Counter

Two methods for measuring PRF employ a frequency counter and an oscilloscope. The frequency counter can be used to measure either the PRF in terms of frequency or as time between the pulses. The accuracy of this method is far superior to any other method.

3-6.61.2 Oscilloscope

The oscilloscope method requires two main triggers to be displayed. The time separation must be interpolated from the display screen. Calculations are then performed if the PRF in frequency, rather than time, is to be ascertained. Small changes in PRF may go undetected due to the small display or to large pulse-repetition time. Procedures for determing the pulse recurrence frequency are as follows:

1. Connect the trigger or a sample of the transmitter output to the vertical input of the oscilloscope.
2. Adjust the sweep frequency until at least two pulses appear on the time trace.
3. Using the sweep calibration, determine the time elapsed between either the leading or trailing edges of two successive pulses.
4. Using the following formula, calculate the pulse repetition rate.

$$\text{PRF} = \frac{1}{\text{Pulse width + rest time}}$$

Since one complete transmitting time cycle equals the sum of the pulse width and the rest time, the number of cycles in one second equals the pulse recurrence frequency.

3-6.62 PULSE WIDTH MEASUREMENTS

Pulse width can be measured by means of a frequency counter, using the "Time Internal" function. The frequency counter's timing circuits are started on the leading edge of the pulse and then stopped on the trailing edage. The time is displayed on the readout. The oscilloscope method of measurement requires one pulse to be displayed and then expanded either by delayed intensity function or simple timebase expansion. An interpolation from the display is made from the 70.7 percent points of the leading and trailing edges, as shown in Figure 3-91.

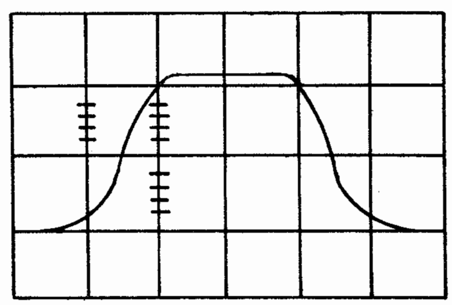


Figure 3-91. Pulse Width Measurement

3-6.63 MODULATOR PULSE MEASUREMENT

Since modulator pulses in the order of thousands of volts are common in radar equipment, several different pieces of test apparatus have been developed so that these high-voltage pulses may be viewed on an oscilloscope. These pieces of apparatus all perform the basic function of dividing the modulator pulse so that a small portion of the pulse voltage, which is proportional to both the magnitude and shape of the pulse, can be applied to the oscilloscope at a voltage low enough to be within the operational range of the oscilloscope.

3-6.64 RESISTIVE LOAD

The "resistive" load is a form of dummy load capable of providing a termination for making overall performance tests on radar modulators. The termination is comprised of high-voltage resistors in the form of a voltage divider of known ratio for the purpose of measuring and viewing the output pulse of the modulator with an oscilloscope.

3-6.64.1 Dummy Load

A dummy load typical of 50-ohm termination consists essentially of two resistive elements made up of one 49-ohm resistive element and a 1-ohm element connected in series, thus providing a 50-to-1 ratio voltage divider. A tap or coaxial connector, with a nomenclature such as MEASURE PULSE, is connected across the 1-ohm resistor. The electrical connection from the divider to the modulator is made with the high-voltage pulse cable of the radar set. The modulator is operated in accordance with the standard instructions of the radar set. Before any measurements are made, the modulator must be operative long enough to heat the load thoroughly (often 10 minutes). This is important, because the values of the resistances are generally unstable, decreasing with temperature. The total resistance and divider ratio is determined with the resistors thoroughly heated. To measure the modulator pulse, connect an oscilloscope, capable of displaying video pulses, to the MEASURE PULSE terminal video jack. Observe the shape and measure the pulse voltage. The voltage is determined by means of the oscilloscope and a calibrating voltage. Measure the pulse by taking the voltage indicated by the oscilloscope and mulitply it by the ratio factor of the resistance load (50 in the case illustrated).

3-6.65 VOLTAGE DIVIDER

The voltage divider (in this application) is a complete portable test equipment used to measure, in conjunction with an oscilloscope, video pulses from about 200 volts to 20,000 volts (or higher in high-impedance circuits). The voltage divider differs from the resistive load basically in that the pulse voltage division is accomplished by capacitors, rather than by resistors.

3-6.66 POWER MEASUREMENT

Power measurement within the scope of the radar technician can be accomplished with either a thermistor-mount type power meter or by a spectrum analyzer. The TS-177 meter is a common thermistor-mount type which measures average power. This meter can be used additionally to measure the attenuation of cables and for checking the performance of attenuators suspected of being defective. Spectrum analyzers measure peak power and should be used when distortion in power output is suspected. Additional attenuators will be necessary, and their value can be calculated from the data in paragraph 3-6.32 of this section.

3-6.67 OVERALL SYSTEM TECHNIQUES

The judging of the range capability and data accuracy of a radar by visual observation alone is inaccurate and valueless. Numerous field tests made on radar sets have verified this statement. In fact, these tests disclosed that may field radars were performing at no more than one-half their maximum capability, although the maintenance and operating personnel considered the operation to be normal. Since any performance less than optimum reduces the tactical area protected by these radars, the measurement of performance is of the utmost importance, especially in time of war. Investigation of the above situation disclosed that many technicians are not completely familiar with the latest techniques of radar system testing. In the following text, the overall system techniques of radar testing which were not previously covered are discussed. The topics included are: 1) Timing-Circuit Calibration; 2) Standing-Wave-Ratio Measurement; 3) Spectrum Analysis; and 4) Overall System Performance.

3-6.68 TIMING-CIRCUIT TESTING

Although pulsed radar was developed primarily for detecting and range finding of various objects, certain specialized requirements have come into later applications. Radar equipments are classified as to the particular use for which they are designed. Any radar equipment performs one or more of the following functions: 1) Search—location of targets with respect to the position of the radar of the ship (this may include IFF applications); 2) Navigation—location of the radar with respect to targets or beacon stations; 3) Ground Control—control of aircraft and direction of air traffic (blind landings and fighter direction); 4) Fire Control—aiming of guns controlled by radar information; 5) Interception—directing fighters toward enemy and, if necessary, enabling blind firing. All of these functions (except search) require a high degree of range accuracy. Therefore, in all radars except those used for search, it is most important that the timing circuits be accurately calibrated.

3-6.69 RADAR TRIGGERING

The overall effectiveness of the triggering system of a radar set depends in large measure upon each trigger's being correctly delayed from the main or pretrigger.

3-6.69.1 Trigger Loop

A trigger loop is discussed in the following text, and its effect upon overall performance of the system is described. In such a loop, one or more triggers may be dependent upon the generation of a previous trigger. Figure 3-92 shows such a trigger loop, and triggers A through F will be discussed.

3-6.69.2 Pulse "A"

This is the main trigger generated from an oscillator or noise generator. If a noise generator is used, a "jutter" PRF is made which is more difficult to jam with synchronous jamming. This "A" trigger determines the PRF of the radar set.

3-6.69.3 Pulse "B"

This is a pre-trigger for the radar receiver and for the final amplifier. It turns the receiver "off" and turns "on" the final amplifier's secondary control circuits. Trigger B is the source for generating pulse "C", and provides a blowing trigger for ECM equipment.

3-6.69.4 Pulse "C"

This trigger turns on the primary control circuits for the final amplifiers and allows the gate between C and D to be generated. In turn, this gates "on" the intermediate amplifiers. It is important to note at this point that if the pretrigger was not generated until "C" time, there would be added interference to the radar receiver, and with the secondary control circuits not turned on soon enough, a clean pulse will not be generated because its leading edge will have been distorted.

3-6.69.5 Gate C - D

This is the actual "on" time of the radar transmitter. An excessive gate time will cause an increase in the average power and may thus exceed a manufacturer's specification or may overload the power supply.

3-6.69.6 Pulse "D"

Trigger D determines the trailing edge of the transmitter pulse and activates the receiver.

3.6.69.7 Pulse "E"

This is a resetting pulse used to regenerate the "A" trigger and to generate "F", which is the zero range trigger.

3.6.69.8 Pulse "F"

This is the zero range trigger, and is employed to synchronize the ship's repeaters. Pulse "F" is variable in relation to "E" trigger. Figure 3-93 shows the synchronizing relationship between the various pulse generators. Care must be taken when adjusting the foregoing time relationships, and should be second in consideration only to that of power supply voltage levels.

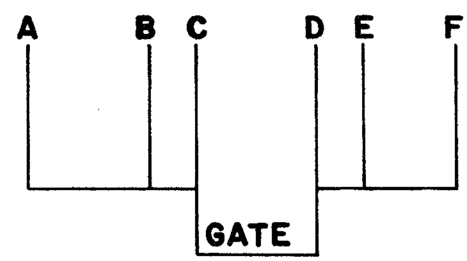


Figure 3-92. Trigger Pulses

3-6.70 RANGE DATA ACCURACY

Since the range of a target is determined by the time interval between the transmitted pulse and the echo return to that point, all time delays introduced by the radar components will add to the range indication. Thus, the range indicated will be in error by an amount equivalent to the time delays which have been introduced.

3-6.70.1 Time Delays

In every radar installation, time delays occur within the equipment between the time the radar trigger is initiated and the time the echo pulse arrives at the indicator. The causes of these time delays are as follows: 1) the modulator output pulse occurs a short time after the trigger input pulse; 2) the RF output from the transmitter takes time to increase to the proper amplitude after application of the modulator pulse; 3) time is required for the transmitted RF energy to travel to the antenna, and for the reflected RF energy to travel from the antenna, and for the reflected RF energy to travel from the antenna to the receiver; 4) time is required for the RF pulse to travel through the receiver; 5) time measurement errors are caused by incorrect scale factors in the ranging circuits; e.g., the oscillator frequency which determines the range mark position may not be correct, or the linear sawtooth slope which generates timing delay may slope too much or

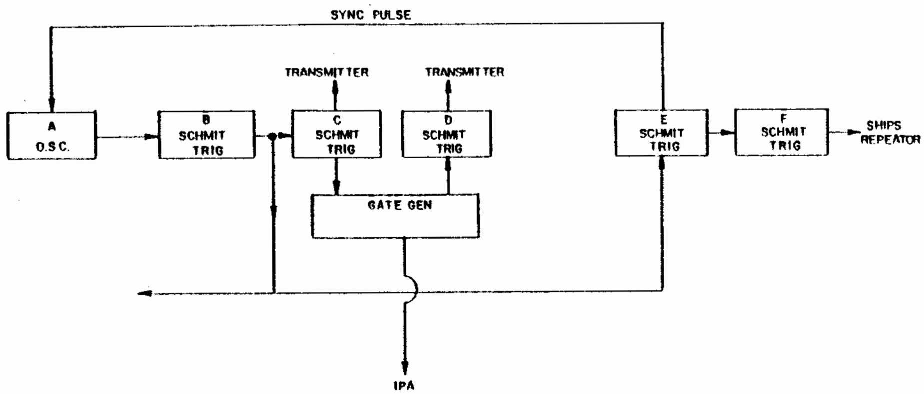


Figure 3-93. Pulse Generator Relationships

too little. Therefore, this is called a slope or rate error. Nonlinearity of the ranging circuits is caused by nonlinearity of the "linear" sawtooth waveform, and there is also a small delay in starting the range circuit. This latter, however, is a negative or compensating delay, as it subtracts from the other delays.

3-6.71 ZERO RANGE ERROR

The difference between real range and indicated range is termed "zero range error." A certain amount of range error may be due to the target distance being measured on one radar set and then measured on another set which may be separated from the first by up to 100 yards, therefore, an error of 100 yards exists that cannot be compensated for. An even greater problem exists if the further radar is being used for gunfire control in which the gun mount may be another 20 yards away and therefore a total of 120 yards of error. During shipyard availability periods, the delay (and therefore the range error) inherent in the transmission line from the transmitter to the antenna, plus equipment reaction time, will delay the transmitted wave with respect to the zero range trigger. These delay periods are carefully measured and rested so that corrections can be applied to the zero range trigger to give an accurate range reading on the PPI.

3-6.72 FIXED-TARGET METHOD

The fixed-target method of zero-error determination is the most reliable method in common use. This method involves the use of a fixed target at some accurately known distance. A natural target may be used, but a portable reflector gives more reliable results. The target range indicated by the radar is carefully read and compared with the known range. The range indicated by the radar should be greater than the known distance; the difference is the zero error.

3-6.73 DOUBLE-ECHO METHOD

The double-echo method of range correction is normally associated with fire-control radar equipments, but any radar set that is capable of receiving two echoes from the same target during one sweep on the range scope is capable of determining range accuracy by the double-echo method. An example of how the double-echo method is used is given below:

1. The radar antenna is trained on a target at a range of between 1000 and 3000 yards. A transmitted radar pulse strikes the target, is reflected, and returns to the point of transmission. A small portion of this returned radar pulse is accepted by the radar antenna and presented as target "A" (Figure 3-94) on the indicator of the radar equipment. The remainder of the returned pulse is re-radiated and directed back toward the target for the second time. There the pulse is again re-radiated back to the point where the radar antenna accepts the pulse and presents it as target "B" on the scope.

2. The time elapsed between the reception of target pulse "A" and target pulse "B" indicates the true range of the target from the radar set, provided, of course, that both pulses were received

during the time of one sweep of the range scope. By placing the ranging device (step, notch, or marker) at target "B" and noting range "d" and then moving the ranging device to target "A," and noting range "c," the correct range may be computed. Subtract the range recorded for target "A" from the range recorded for target "B" ("d" minus "c"). This resultant figure should equal the range recorded for target "A." If it does not, then with the ranging device set at target "A", the range correction control is adjusted, or the range counter mechanically disengaged and "slipped" until the counters indicate the correct range. This procedure of noting the ranges and altering the counters is repeated as many times as necessary to obtain accurate ranging. One note of caution is observed - the measurement of ranges "c" and "d" is made as quickly as practicable, to prevent the possibility of the target range varying by any appreciable amount during the time interval between two range measurements.

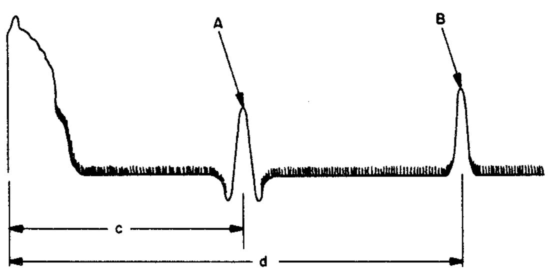


Figure 3-94. Double-Echo Range Scope Presentation

3-6.74 STANDING-WAVE-RATIO MEASUREMENT

In a radar installation a low SWR is maintained principally for the following reasons: (1) reflections occurring in the RF line cause magnetron pulling, and can result in faulty pulsing (this effect is more pronounced when the line is long, as compared to a wavelength); (2) arc-over may occur in the RF line at maximum-voltage points; (3) mechanical breakdown in the line may sometimes occur, because of the development of hot spots and (4) reflection of power causes excessive heating of the transmitting device, the SWR should be less than 1.5 to 1, which represents a reflection of less than 5 percent of the incident power. In the maintenance of radar, SWR measurements are useful in two ways. First, defective RF line components can be located by checking the SWR of each component or by substitution. Second, radars having RF tuning adjustments can be adjusted with the aid of SWR test equipment.

3-6.75 SLOTTED-LINE METHOD

The slotted-line method of measuring SWR can be used with the aid of an RF probe and a slotted line. The slotted line is a coaxial or waveguide section of transmission line, with a longitudinal slot cut into its outer conductor that permits insertion of the RF probe. The slot is constructed at least a wavelength long, but is not wide enough to cause appreciable loss by radiation. To explore the voltage field existing in the line, the probe is placed in the electrostatic field through the slot and the moved back and forth. The probe feeds an RF detector, and the rectified output operates a meter which indicates the SWR. Many radars have slotted-line test sections that are integral parts of the RF section. In such cases, a removable protective plate usually covers the slot. For those facilities not having built-in slotted sections, a series of slotted-line sections to fit any radar set have been devised. In some cases the section is inserted into the radar set by means of coupling sections, or a test setup is developed whereby the radar units may be tested. The SWR is measured by performing the steps given below:

1. If the radar facility does not have a line, insert a slotted section into the radar transmission line (with the adapters provided), as close to the magnetron as possible.
2. Connect RF probe to amplifier, and adjust probe for a penetration of a few thousandths of an inch.
3. Operate radar transmitter and probe to a maximum point.
4. Set probe penetration to provide a scale meter reading of 1.
5. Move probe to a minimum point, and read SWR on meter.
6. If SWR is too high (1.5 to 1 or higher) and the radar has tuning stubs, adjust the stubs for an SWR of as near 1:1 as possible. The latter adjustment is made with the antenna pointing at free space.

3-6.76 DIRECTIONAL COUPLER METHOD

The directional coupler method of determining SWR is frequently used in the field. To determine the SWR of an assembly, the coupler is inserted into the transmission line and the incident power is measured. The coupler is then reversed, and the reflected power is measured. The SWR can then be calculated. This method is not very accurate if the coupler directivity is low. For example, if a given coupler has a nominal loss of 20 dB and a directivity of 20 dB, the SWR obtained as a result of its use is 1.6 to 1; however, the actual SWR may be any value from 1.4:1 to 1.8:1. It is seen that the greater the directivity, the greater the accuracy. The above method has been largely superseded by the use of the bidirectional coupler, which greatly simplifies the application. This coupler, which consists of two directional couplers mounted on the same line but coupling in opposite directions, has been developed to a point where the accuracy of the method is quite good. The bidirectional coupler does not require reversing, is easier to use than the slotted line, is more rapid in operation, and can be made a permanent part of a pressurized RF assembly. In addition, the coupler can be used in connection with power and frequency testing and spectrum analysis, which is discussed in detail following standing-wave-ratio measurement.

3-6.77 BIDIRECTIONAL COUPLER METHOD

In many equipments, a bidirectional coupler is included in the transmission line. The operation of this coupler for use in waveguides was discussed previously under DIRECTIONAL COUPLERS. Only the method for using the coupler for measurement of SWR is given here. Bidirectional couplers are also designed for use in coaxial transmission lines. The value of the SWR is given by the ratio of the voltage at a maximum point to that at a minimum point; that is

$$SWR = \frac{E_{max}}{E_{min}}$$

Since E_{max} is the sum of the incident voltage E_1 and the reflected voltage E_r and E_{min} is the difference of these two voltages, the equation above can be rewritten as:

$$SWR = \frac{E_1 + E_r}{E_1 - E_r}$$

Making use of the relationship $P\frac{F^2}{Z}$, the equation becomes:

$$SWR = \frac{\sqrt{Z_L}\sqrt{P_1}\sqrt{Z_T}\sqrt{P_r}}{\sqrt{Z_L}\sqrt{P_1}\sqrt{Z_T}\sqrt{P_r}}$$

where:

P_i = incident power from the transmitter.

P_r = reflected power from the transmission line termination

Z_L = the load impedance of the terminated transmission line as seen from the coupler

Z_T = the transmitter output impedance as seen from the coupler

In most cases, $Z_L = Z_T$ (the condition for maximum power transfer). For this condition, the equation is simplified to:

$$SWR = \frac{\sqrt{P_1 + P_r}}{\sqrt{P_1 - P_r}}$$

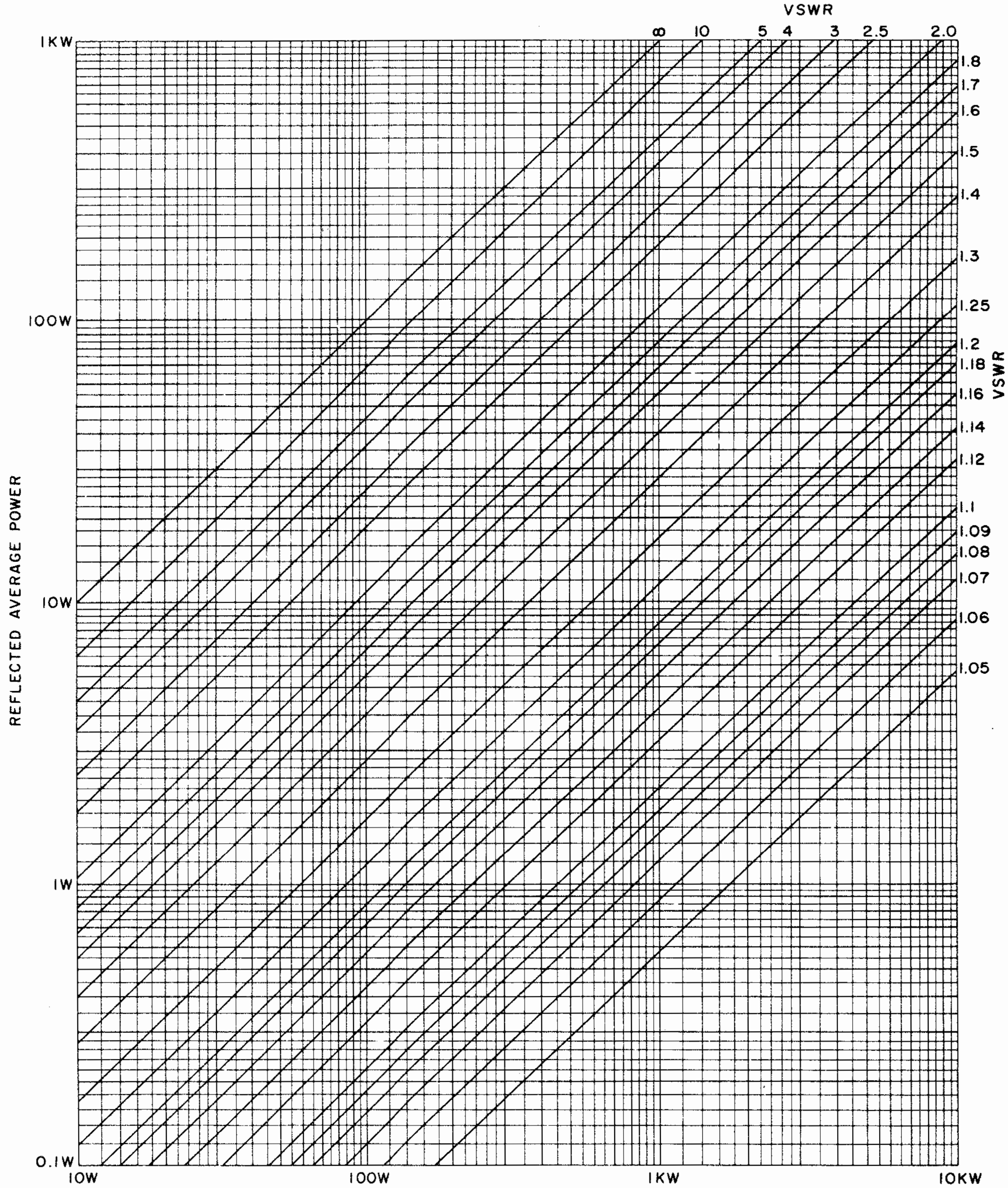
Figure 3-95 is a plot of this equation for common values of incident and reflected power. An important item in using this method for finding SWR is that the power meter used with the bidirectional coupler will not, in general, directly provide actual values of incident and reflected power. The total attenuation in the coupling to the meter must be taken into account when calculating incident and reflected power. These values are then used to determine SWR using the figure.

3-6.78 CAUSES OF STANDING WAVES

Any discontinuous change along an RF line, such as might be introduced by a change in dimensions, or a change in geometry introduced by a sharp bend or dent, or an obstacle in the line, produces reflections. Some of the most common causes of an excessively high standing-wave ratio are: (1) dirt or moisture in the RF line; (2) dented or bent line; (3) burrs or poorly soldered joints; (4) defective coupling joints; (5) defective rotating joint; and (6) mismatched antenna.

3-6.79 LOCATING DISCONTINUITIES

If an increase is noted in the standing-wave ratio, check the RF transmission lines for the common causes listed above, as well as for any other damage which may result from battle, storms or normal wear. Check the antenna also, since any bending of the reflector or dipoles changes its impedance and results in an increased SWR. Many RF transmission-line faults are visible and easily located; however, in some cases the trouble may be of such a nature that the



$$VSWR = \frac{\sqrt{P_I} + \sqrt{P_R}}{\sqrt{P_I} - \sqrt{P_R}} \qquad P_R = \frac{(VSWR-1)^2}{(VSWR+1)^2} P_I$$

Figure 3-95. Graph of Relation Between Incident Power, Reflected Power, and SWR defective part can be found only by making special tests, such as a time-domain reflectometer, or terminating the transmission line by sections to locate the defective section.

3-6.80 DUMMY RF LOAD METHOD

A dummy RF load is a resistive section of transmission line which absorbs RF power without causing appreciable reflection. A typical waveguide, filled with a mixture of sand and aquadag, which serves as a resistance to absorb power. To minimize reflections the front surface of the resistive element is constructed so as to present an oblique surface to the incident RF power. The exterior of the dummy load is fitted with cooling fins, and is painted a dull black for greater heat transfer. A typical X-band dummy load gives an SWR of less than 1.05 to 1, and absorbs 150 watts of power (average). If necessary, the power rating of dummy RF loads is increased by the use of forced air or water cooling. Coaxial dummy RF loads are similar in operation to the waveguide-type loads. The resistive mixture forms a tapered contact between the inner and outer conductors. Two distinct advantages are gained by the use of dummy RF loads in radar maintenance: 1) when military security prohibits RF radiation, maintenance can still be carried out with the aid of the loads; and 2) the load can be used to absorb power without reflections, regardless of surroundings. It may also be used for measurement of power output.

3-6.80.1 Dummy RF Load

When the SWR of an RF transmission line becomes excessive, the causes of the standing waves may be located with the aid of a dummy RF load by performing the steps given below:

1. Remove antenna feed, and substitute dummy RF load. If the substitution corrects the condition, the trouble may possibly be in the antenna, or it may be the result of reflections from nearby objects.

2. If the SWR is still too high, change the antenna scanning position and recheck the SWR. If the SWR changes, a defective rotating joint is sometimes indicated. In some cases, the SWR may not change, even though a rotating joint is defective. Therefore, this test will not eliminate the possibility of a joint, but will locate faults caused by its rotation.

3. If the rotating joints are not defective, remove the section of the line next to the antenna feed section, and replace the dummy load at the open end of the RF assembly. Recheck the SWR. If the section removed is defective, the SWR will improve. Continue the process of removing sections until the offending section is located. The SWR should be checked again after the trouble is corrected and the RF components are reassembled.

3-6.80.2 Time-Domain Reflectometer

Equipment using rigid coax as a transmission line may use a Time-Domain Reflectometer to measure discontinuities along the transmission line, as well as measurements on any other type of cable. A display of length versus impedance is provided, as shown in Figure 3-96. Each connector or joint can be located through the use of the length unit and a multiplier.

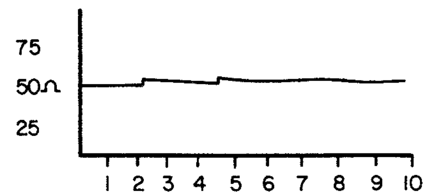


Figure 3-96. Time-Domain Measurement of Line Length vs. Impedance

3-6.81 SPECTRUM ANALYSIS

It is possible, by means of a spectrum analyzer, to observe a selected portion of the electromagnetic spectrum on the screen of a cathode-ray tube. The display consists of vertical pulses distributed along the horizontal axis; the position of each pulse indicates the frequency of a particular signal, while the relative height of each pulse indicates the relative strength of the signal. In other words, the display viewed on the cathode-ray-tube screen is, in effect, a graph of energy plotted against frequency. By analyzing the spectrum of a radar transmitter, a great deal of information is obtained regarding the condition of the radar. The spectrum analyzer can show the presence or absence of frequency modulation, and can also indicate the presence or absence of amplitude modulation in the signal. By means of a frequency meter, which is normally an integral part of the spectrum analyzer, it is possible to determine the bandwidth necessary to transmit each condition; only a limited number of harmonics are involved in the usual square wave. However, in actual practice a good square wave may contain frequencies up to the 100th harmonic.

3-6.82 TRANSMITTER SPECTRAL DISPLAY

When a transmitter is modulated by short rectangular pulses occuring at the PRF (pulse repetition frequency) of the radar, two distinct modulating components are present. One component consists of the PRF and its associated harmonics; the other consists of the fundamental and odd-harmonic frequencies that comprise the rectangular pulse, as was previously discussed. Figure 3-97 shows an ideal display of that part of the spectrum covered when an RF carrier is pulse-modulated. The vertical lines in the figure represent the modulation frequencies produced by the PRF and its associated harmonics, while the lobes represent the modulation frequencies produced by the pulse frequency and its associated harmonics. The vertical lines are separated by a frequency equal to the PRF. The amplitude of the main lobe falls off on either side of the carrier until it is zero at the points corresponding to the second harmonic of the fundamental pulse frequency. The first side-lobe is produced by the third harmonic of the pulse frequency; the second zero point, by the fourth harmonic. In the ideal spectral display, each frequency above the carrier has a counterpart frequency equally spaced below the carrier, so that the pattern is symmetrical about the carrier. The amplitude of the side lobes is considered important, because in the ideal spectral display, the first side-lobe represents 4.5 percent of the carrier amplitude, and the second side-lobe represents 1.6 percent of the carrier amplitude. The main lobe carries the major portion of the transmitted energy.

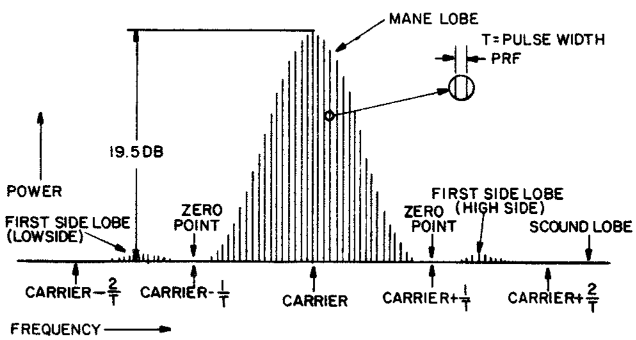


Figure 3-97. Ideal Spectral Display of a Modulated RF Carrier

3-6.83 TRANSMITTER OUTPUT VERSUS RECEIVER RESPONSE

The importance of the transmitter output characteristics as compared with the receiver response becomes readily apparent by inspection of Figure 3-98, which shows an optimum receiver response curve superimposed upon an ideal pulse spectral display. The receiver bandwidth is broad enough to include all the energy between the first zero points. Note that the receiver also responds to the first side-lobes, but at reduced level. Any RF energy that exists outside the limits of the receiver response is, of course, lost, and the effect is the same as if the transmitter power were reduced. Since practically all of the transmitted energy is within the limits of the receiver response, as shown in Figure 3-98, further broadening of the receiver bandwidth results in very little increase in energy pickup. It is apparent, however, that a decrease in bandwidth causes a definite reduction in energy pickup.

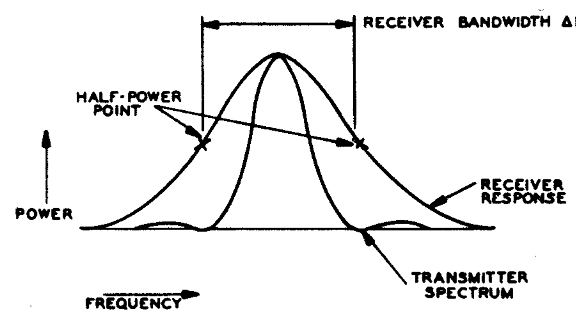


Figure 3-98. Transmitter Spectral Display Compared with Receiver Response Curve

3-6.83.1 Side Lobes

The spectrum side-lobes contribute very little in terms of pulse amplitude, but they contribute to the steepness of the edges on the output pulses, as shown in Figure 3-99. From this cursory examination, it can be seen that an ideal receiver has sufficient bandwidth to include a great many side lobes, in order to reproduce the transmitted pulse with a high degree of accuracy. However, if the above condition is obtained, the increased bandwidth allows the receiver to respond to more than the normal amount of noise and limits its sensitivity. A reduction of bandwidth, within limits, does not lessen the pulse amplitude, but reduces noise response. Too great a reduction of bandwidth results in decreased pulse amplitude, as shown in Figure 3-99, because of the loss of some energy in the main lobe. Optimum bandwidth results in the greatest receiver sensitivity, but causes a slight distortion of pulse shape. Since accurate pulse shape is important in precision ranging and tracking operations,

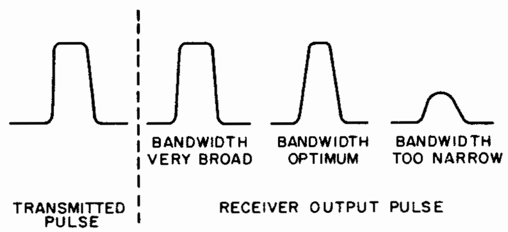


Figure 3-99. Effect of Receiver Bandwidth upon Pulse Shape

certain radar systems designed for this type of service have a receiver bandwidth that is broader than optimum, to provide a sharp leading edge on the pulse.

3-6.84 MODULATION DISTORTION

In a properly functioning transmitter, the period of the transmission interval is as specified, the oscillations during the transmission interval are of constant frequency and amplitude, and the time required for oscillation to start and stop is approximately zero. Any deviation from these conditions produces distortion that is visible on the spectral display in the form of either pulse amplitude distortion or frequency modulation (FM), or both. Figure 3-100 shows the spectral dispaly when FM is present. The zero points are lost, thus indicating the presence of new frequencies in the spectral display. This has the effect of placing more of the transmitted power in the sidebands, therefore results in the loss of energy outside the receiver passband. Part (B) of the same figure shows the spectral display when the amount of frequency modulation is excessive. In part (B), the magnetron is operating at two distinct frequencies (double moding), and the receiver, if tuned throughout its range, would thus have two tuning points. When the above condition prevails, more than half of the transmitted power is wasted. The presence of pulse amplitude distortion in the transmitted output has the effect of producing dissymmetry in the display, as shown in Figure 3-101. The zero points are still clearly defined, but the lobes on one side of the carrier are much larger than normal. In general, distortion resulting from frequency modulation is far more undesirable than distortion from amplitude modulation. Figure 3-102 shows a combination of both types of distortion, which results in a very poor quality spectral display.

3-6.84.1 Display Troubles

The troubles which give rise to a poor transmitter spectral display are sometimes difficult to locate. Briefly, at this point it can be stated that trouble may arise from the following causes: (1) defective magnetron; (2) defective magnet; (3) mismatch in RF section (pulling); (4) improper pulse shape or amplitude (pushing); and (5) reflections from nearby objects (pulling). Two methods of obtaining a graph or display of the spectrum are described in the following paragraphs. The first method requires that a graph of power versus frequency be plotted. The method is relatively slow, and demands considerable experience on the part of the technician. The second method is simplified by the use of a spectrum analyzer, which provides, as was mentioned earlier, a display on the screen of a cathode-ray tube corresponding to the graph plotted in the first method. The circuit of a typical spectrum analyzer is discussed briefly. Some spectral displays are examined and interpreted, and a method for frequency measurement is described.

3-6.85 FREQUENCY METER METHOD

The use of a frequency meter is a rather simple method of obtaining readings to plot a spectral graph. A high-Q, transmission-type, resonant-cavity meter, such as is found in most echo boxes is utilized, together with a rectifier-meter indicator. The test arrangement is shown in Figure 3-64. Readings are

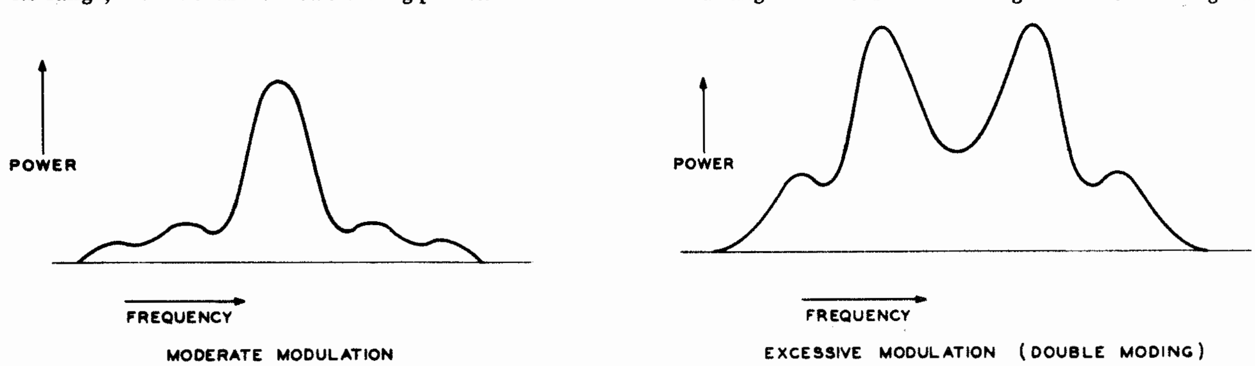


Figure 3-100. Transmitter Spectral Displays, Showing Distortion Resulting from Frequency Modulation

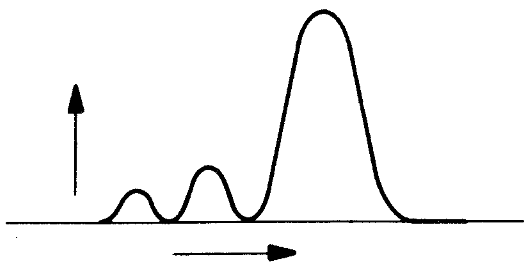


Figure 3-101. Transmitter Spectral Display, Showing Distortion Resulting from Amplitude Modulation

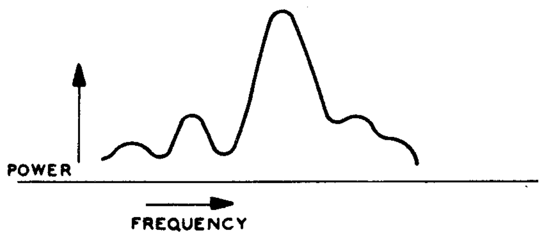


Figure 3-102. Transmitter Spectral Display, Showing Distortion Resulting from Combined Frequency and Amplitude Distortion

taken at frequent intervals throughout the frequency range of the transmitter, and a graph is made to indicate meter readings vertically and frequency-meter indications horizontally. If the graph is very carefully plotted, a rough outline of the spectrum is obtained. A good idea of the spectrum is gained by noting how the meter reading varies as the frequency meter is tuned through resonance. All spectral readings must be obtained by rotating the frequency-meter dial in one direction only. If the dial is rocked into position, backlash in the dial drive mechanism will cause an appreciable error. The usual procedure is to approach each reading from the low-frequency side.

3-6.86 SPECTRUM ANALYZER METHOD

The spectrum analyzer, which is a form of panoramic receiver, provides a simplified method of analyzing spectral phenomena. A small portion of the transmitter output is coupled into the signal input circuit of the spectruum analyzer. Care must be taken to keep the input low enough to prevent burnout of the attenuator. A directional coupler provides an ideal coupling system, but a pickup antenna may also be used.

3-6.87 SPECTRUM ANALYZER CIRCUIT ANALYSIS

In the spectrum analyzer, a narrow-band receiver is electrically tuned through a range of frequencies. The output, in terms of power, is displayed vertically on an oscilloscope whose horizontal sweep is synchronized with the frequency sweep of the receiver. A block diagram of a typical spectrum analyzer is shown in Figure 3-103. The receiver employed is a superheterodyne type. The input, which usually consists of a coaxial-line termination, a broad-band attenuator, and a crystal mixer, is untuned, and therefore responds equally well to all signals within the operating band. The local oscillator is usually a reflex-klystron type. The I-F amplifier is a high-gain, narrow-band (50 kHz or less) amplifier, usually operated above 20 MHz. In some cases, double, or even triple, superheterodyne action is used to obtain the narrow bandwidth required. The I-F section is followed by a detector and amplifier which feed the vertical plates of a cathode-ray tube. The sweep generator produces a variable-frequency sawtooth voltage which sweeps the local oscillator repeller and, therefore, the receiver frequency and the horizontal deflection plates simultaneously. A reaction-type frequency meter is included which is designed to absorb local-oscillator power at resonance, thereby indicating the local-oscillator frequency.

3-6.87.1 Modes

On the front panel of the analyzer is a function switch, usually labeled MIXER-SPECTRUM. In the SPECTRUM position, the indicator displays the output of the receiver. In the MIXER position, the indicator displays the crystal-mixer current, which is a function of the reflex-klystron local-oscillator output. Figure 3-104 shows a typical reflex-klystron chart. Note that the tube will oscillate only at certain voltages and, as the voltage is varied, the power output varies. Each separate voltage range of oscillation is called a MODE. The modes are relatively flat on top, and each succeeding mode encountered becomes stronger as the repeller is made more negative. Within any given mode, the frequency is proportional to the negative voltage on the repeller. A frequency range of 60 MHz is common in X-band tubes. However, the frequency at the top of each mode is determined by the size of the resonant cavity in the tube; therefore, all of the modes have the same center frequency. The sweep generator produces a sawtooth voltage which is adjustable in both amplitude and average voltage value. The sawtooth amplitude control, usually called the spectrum width control, has sufficient range to cover at least one

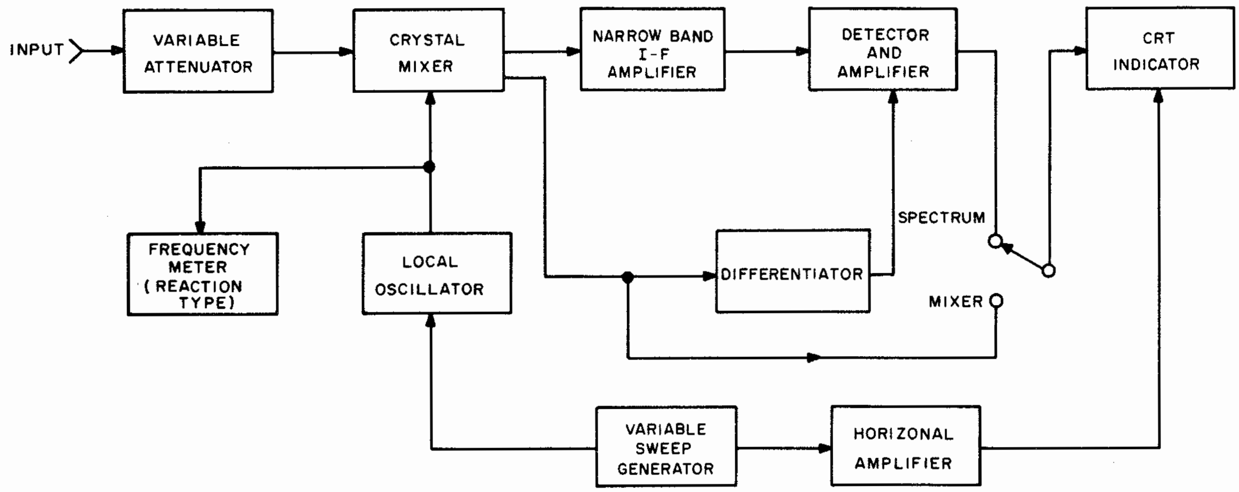


Figure 3-103. Typical Spectrum Analyzer, Block Diagram

mode, and quite often, two. The average voltage control of the sawtooth, usually called the spectrum center control, allows the choice of any desired klystron mode, or the use of any range within a particular mode. In normal use, only a limited section of one mode is employed.

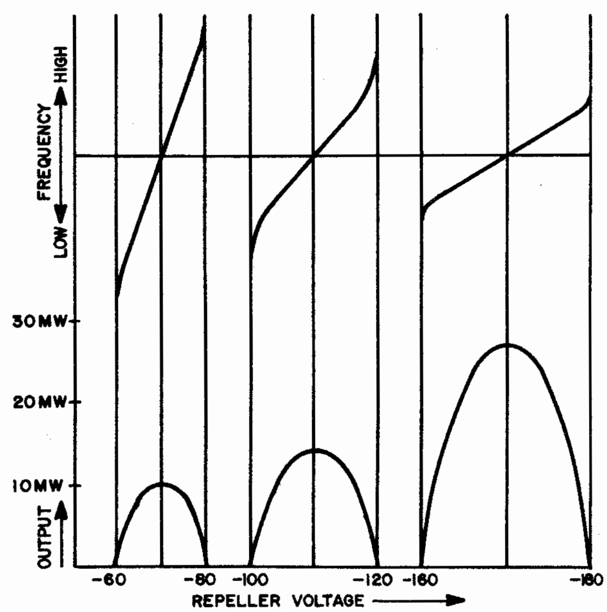


Figure 3-104. Typical Reflex Klystron Chart

3-6.87.2 Klystron Testing

The spectrum analyzer can be used as a klystron tube tester. When the function switch is in the MIXER position, the presentation is similar to that shown in Figure 3-105, which illustrates one complete klystron mode and part of another. The pip shown in the center of each mode is the frequency indication introduced by the reaction-type frequency meter. The mixer function of the analyzer allows the condition of the local oscillator to be checked; if desired, the oscillator frequency can be set to any specified value. The klystron to be tested is substituted for the local oscillator in the spectrum analyzer, and the mode pattern observed. The amplitude of the mode indicates power relative to that of the regular oscillator. The tuning range is examined and any irregularities noted. Each mode should present a smooth regular curve. If desired, the tube under test is pretuned to the approximate frequency before insertion into the radar in order to simplify radar tune-up.

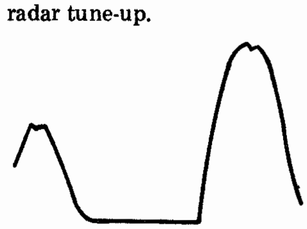


Figure 3-105 Klystron Modes as Presented on Spectrum Analyzer

3-6.88 TRANSMITTER SPECTRAL DISPLAY ANALYSIS

As the spectrum analyzer frequency is swept, the spectral display appears upon the cathode-ray tube indicator in the form of a series of vertical pulses. These pulses are not to be confused with the vertical lines shown in Figure 3-97, which are separated by a frequency equal to the PRF. If the spectrum analyzer has a bandwidth of 50 kHz, a large number of PRF lines are included in each pulse because the analyzer samples a 50 kHz segment of the spectrum each time the transmitter fires. Thus, each pulse in the display represents the energy contained in a 50 kHz band at the freqeuncy of the analyzer at that instant. If the radar PRF is 200 pulses per second and the analyzer sweep rate is 10 Hz, the display consists of 20 pulses across the screen of the cathode-ray tube. Figure 3-106 shows a typical magnetron spectral display. These pulses indicate only the general outline of the display, and are much too coarse to reveal the internal structure. Figure 3-107 shows the same conditions with the spectrum width control advanced to produce a greater spread.

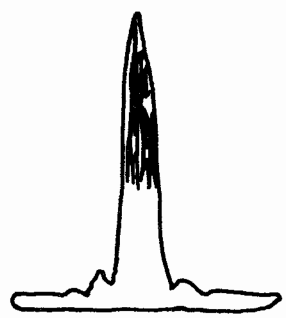


Figure 3-106. Typical Magnetron Spectral Display

3-6.89 FREQUENCY MEASUREMENT

The measurement of frequency is greatly facilitated by the use of a differentiator. (Refer to the block diagram shown in Figure 3-103.) A portion of the crystal-mixer current is applied to a differentiator, and the differentiated waveform is applied to the amplifier section of the spectrum analyzer. Figure 3-108 shows the result of differentiating and amplifying the mixer signal. Part (A) of the figure shows the display with the function switch in the MIXER position; and part (B) is the display with the switch in the SPECTRUM position. Note that the frequency-meter pip now appears as an "S" curve, and that the mode ends are marked by pips. This signal is combined with the spectral display, and appears superimposed on the base line of the pattern, as shown in Figure 3-109. The exact frequency is taken at the center of the "S" curve, where it crosses the base line. The pips marking the mode end limits should never be seen on the display, since no spectral indication may be obtained outside the mode limits.

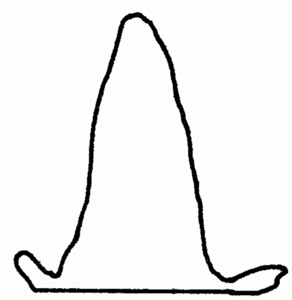


Figure 3-107. Typical Magnetron Spectral Display, Analyzer Width Control Advanced

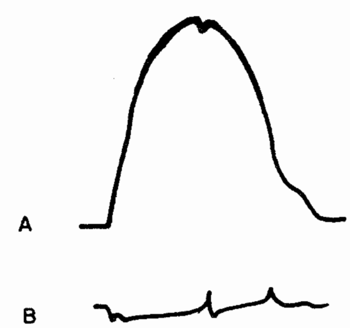


Figure 3-108. Effect of Differentiator upon Mixer Output

3-6.90 INSTALLATION TESTING

The spectrum analyzer can also be used for some installation tests. In this case, both the transmitter and local-oscillator signals can be conveniently sampled by means of a small pickup antenna placed near the base of the local-oscillator socket. In this position, the pickup antenna is the RF leakage field, and the intensities of the two signals are approximately equal, because of the proximity of the pickup antenna to the weaker source. Because the RF section of the analyzer is untuned, image signals are also received.

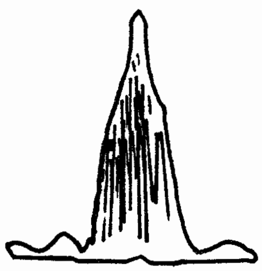


Figure 3-109. Typical Spectral Display, Showing Frequency-Meter Pip

Thus, the signal picked up appears at two points on the analyzer tuning scale. In practice, however, an image is just as useful as the real frequency, and is often used in measurements even though the frequency scale is reversed. Since the analyzer frequency meter is designed to indicate its local-oscillator frequency rather than the input-signal frequency, the most accurate frequency-measurement method is to measure the analyzer local oscillator when the oscillator is tuned above the input signal, and then measure it when the oscillator is tuned below the signal. The signal frequency is then halfway between the two readings. The signal frequency meter also is tuned for maximum absorption of the input signal to obtain a direct indication of the input-signal frequency. Resonance is indicated by a slight reduction in the amplitude of the signal; however, this is difficult to observe. Figure 3-110 shows an overall spectral representation of a transmitter and local oscillator of a particular radar installation. In this figure, it is assumed tha the intermediate frequency of the spectrum analyzer is 25 MHz, the radar transmitter freqeuncy is 9375 MHz, and the radar local-oscillator frequency is 9405 MHz. This produces a radar intermediate frequency of 30 MHz. If the spectrum analyzer was capable of showing the entire range of frequencies, the transmitter display would be recorded at 9350 MHz and 9400 MHz, and the local-oscillator display, which appears as a single frequency, would be recorded at 9380 MHz and 9430 MHz. (Note that the frequencies shown represent the local-oscillator frequency of the spectrum analyzer, and not the signal frequency.) Most spectrum analyzers, however, cannot show the entire range of frequencies given in Figure 3-110. A typical X-band analyzer, for example, is able to present a continuous range of only 50 to 60 MHz. To examine various portions of the entire range, the analyzer turning dial is turned to vary the range of frequencies being covered. In this way,the entire band (8500 MHz to 9600 MHz) May be covered by one instrument.

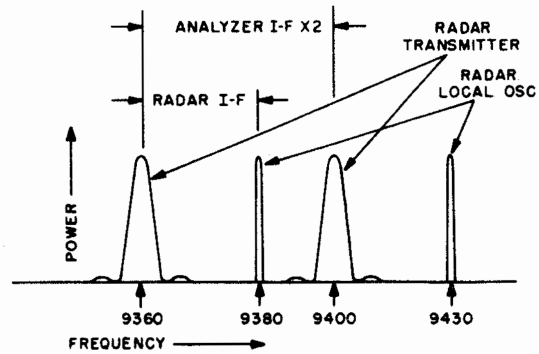


Figure 3-110. Overall Spectral Representation of Transmitter and Local-Oscillator Output

3-6.90.1 AFC Checks

The spectral range can be made broad enough to display both the radar transmitter and local-oscillator frequencies simultaneously. Because it has this feature, the spectrum analyzer is recommended for use in tuning a radar local oscillator to a specified frequency. It is also recommended for use in checking AFC action. The procedure is also recommended for use in checking AFC action. The procedure for an AFC check is as follows: set the antenna scanning unit in motion and note the pulling action on the magnetron. Any lateral motion in the display position indicates a change in frequency. With the AFC in operation, any shift in the radar-transmitter frequency should be accompanied by a corresponding shift in the local-oscillator frequency. Therefore, the distance between the local-oscillator and transmitter patterns should remain constant. Excessive pulling is usually evidenced by a distortion of the shape of the transmitter display. A marginal check on AFC operation to detect incipient trouble is made by turning on the AFC and using a wavemeter to measure the normal oscillator frequency. Manually detune the receiver until the AFC circuit fails to hold the oscillator frequency when it just jumps out of tune and the normal frequency is an indication of how well the circuits are functioning. The greater the difference, the better the circuits are functioning.

3-6.91 OVERALL RADAR PERFORMANCE

Radar performance testing involves a series of measurements which are primarily intended to indicate the ability of the radar to detect targets. The combined results of the tests then indicate the overall radar performance. Two distinct and separate factors are involved in the consideration of radar performance: 1) minimum-range performance; and 2) maximum-range performance. Both of these factors are discussed in detail in the text to follow.

3-6.92 MINIMUM-RANGE PERFORMANCE

Certain radar facilities are designed to detect and find the range of nearby targets. Examples of these are radars used for fire control, aircraft interception, and ground-controlled-approach applications. The TR recovery time and transmitter pulse width are important factors in determining the effective minimum range. If a target has range of 200 yards, the echo is returned to the radar in about 1¼ microseconds after the occurrence of the transmitter pulse. If the receiver is to respond to this echo, the TR switch must recover sufficiently during this short interval to allow passage of energy to the receiver. Since a nearby target returns a strong signal, the recovery need not be complete, because the receiver responds to a strong signal, even at reduced sensitivity. Minimum-range performance is also influenced by the transmitter pulse width. Long-range search radar facilities may use a pulse width of 2 microseconds or more,which represents a free-space range of over 320 yards. Radars designed for close-range work have pulse widths as short as one-quarter-microsecond, which represents only a little over 40 yards of free-space range. Furthermore, high-power radars may require the use of a pre-TR tube to prevent transfer of harmonic energy through the TR; as a result, the recovery period may correspond to 2000 or more yards of range.

3-6.93 MAXIMUM-RANGE PERFORMANCE

The factors which determine the maximum range of a radar are rather diverse; however, for the purpose of the discussion to follow, they may be divided into two general categories. The first category, which is not controllable by the maintenance activity, consists of target reflection and wave propagation factors. The second category is made up of performance factors, which are controllable to some degree by the maintenance activity.

3-6.93.1 Target Reflection

Target reflection is a direct function of target size. Four principal factors enter into the determination of target sync or the amount of energy reflected by a radar target: 1) the material of which the target is constructed; 2) the surface area presented to the radar; 3) the configuration of the surface presented to the radar; and 4) the operating frequency of the radar. In general, it can be said that the amount of energy reflected from a target is a very complex consideration, and that the reflected-signal energy cannot be predicted with any degree of accuracy. Therefore, when information on the reflected-signal strength is required, it is best found by direct measurement. For different target configurations, the reflected-signal strength varies considerably. An approximate target area for some aircraft and vessels is presented in Table 3-14.

Table 3-14. Radar Target Area of Aircraft

OBJECT	TARGET AREA IN SQ METERS
Small aircraft	16
Large aircraft	74
Cruiser	14
Surface submarine	0.09
Large freighter	15

The maximum range which may be expected from a radar may be calculated using the following equation:

$$R_{max} = K \frac{R_k G_o{}^2 \lambda^2 f_p \ 1/3}{NF} A_E$$

where:

R_{max} = Nautical miles

P_T = Transmitter power (peal) in watts

G_o = One-way antenna gain in dB over that of isotropic radiator

λ = Wave length in centimeters

t_p = Pulse duration in μ sec

f_p = Pulse repetition rate per sec

A_L = Target size in square meters

N_F = Receiver noise figure in c/B

K = 5.5, which is a constant for converting equation to nautical miles

All of the above factors can be obtained from the radar maintenance manual except for A_Σ can be approximated from the preceding table, and G_o can be computed by the following equation, assuming an illumination factor of 0.6:

$$G_o = 20 \log \frac{D}{\lambda} + 10 \log f + 9.938$$

D = Diameter of antenna

f = Illumination factor

λ = Wavelength (same unit as diameter)

3-6.94 RADAR SYSTEM PERFORMANCE FACTORS

Maximum range performance also depends upon the condition of the radar. Radar condition is the only factor which may be controlled to some degree by maintenance, and therefore is the most important factor. Radar equipment performance is dependent upon such items as transmitter power, frequency, and spectrum, receiver bandwidth and sensitivity (MDS), TR recovery time, and AFC operation.

3-6.95 RADAR SYSTEM SENSITIVITY

System sensitivity is the ratio of the transmitted power admitted by the receiver passband to the MDS power. A precise determination of system sensitivity, therefore, involves a check of both the transmitter spectrum and the receiver passband. It follows that if half the transmitted energy is outside of the receiver passband, the power is effectively cut in half. System sensitivity is proportional to the fourth power of the maximum range, as shown in the following expression:

$$\text{Range (maximum)} = \sqrt[4]{P_t/P_{mds}}$$

where:

P_t = transmitter peak power encompassed by the receiver passband

P_{mds} = is the minimum discernable signal power.

Note that in the above expression the fourth root is taken rather than the square root. This is explained by the fact that the inverse square law, which is used to determine the strength of a transmitted signal over a given distance, is applied twice: once for the forward path, and once for the echo return. When transmitted power and MDS power are measured in terms of watts, the sensitivity of the radar system is calculated in terms of a power ratio by simple division. However, these two power figures are usually measured in dBm, and the system sensitivity is more conveniently calculated in dB as follows: System sensitivity (dB) = P_t (dBm) - P_{mds} (dBm). (Note that the MDS figure is a negative dBm quantity and, as such, must be subtracted algebraically from the transmitted power.) The sensitivity of a radar system is normally specified in the technical manual for the equipment. Any loss of sensitivity results in a corresponding decrease in the maximum range. Figure 3-111 shows graphically how a decrease in system sensitivity, given in dB below optimum performance, causes a decrease in the maximum range, given in percentage loss. The shaded area shows the limits of best and worst propagation condition. The lower curve is based on the fourth-root range equation above and represents ideal propagation conditions which are typical of the best air search conditions. The upper curve is based on the 16th root of the ratio P_t/P_{mds} and represents the worst propagation conditions, typical for ground radars operating under substandard atmospheric conditions. The dotted curve in the figure is based upon the eighth root of the ratio P_t/P_{mds} which centers it between the extremes and represents approximate sea-search conditions. To see the effect of loss of system sensitivity upon maximum range, consider a hypothetical radar which has operated under all three propagation conditions shown in the figure. Assume only that under the fourth-root condition the maximum range (with no system sensitivity loss) is 100 miles. The maximum ranges for the three propagation conditions, with optimum system performance and with the performance 20dB down, are then given in Table 3-15. It can be seen from this table that the greatest percentage loss in range because of system sensitivity loss (performance 20 dB down) occurs when the propagation conditions are optimum.

3-6.96 ANTENNA AND TARGET ALTITUDES

Most radars do not have stabilized platforms that provide a constant horizon reference.

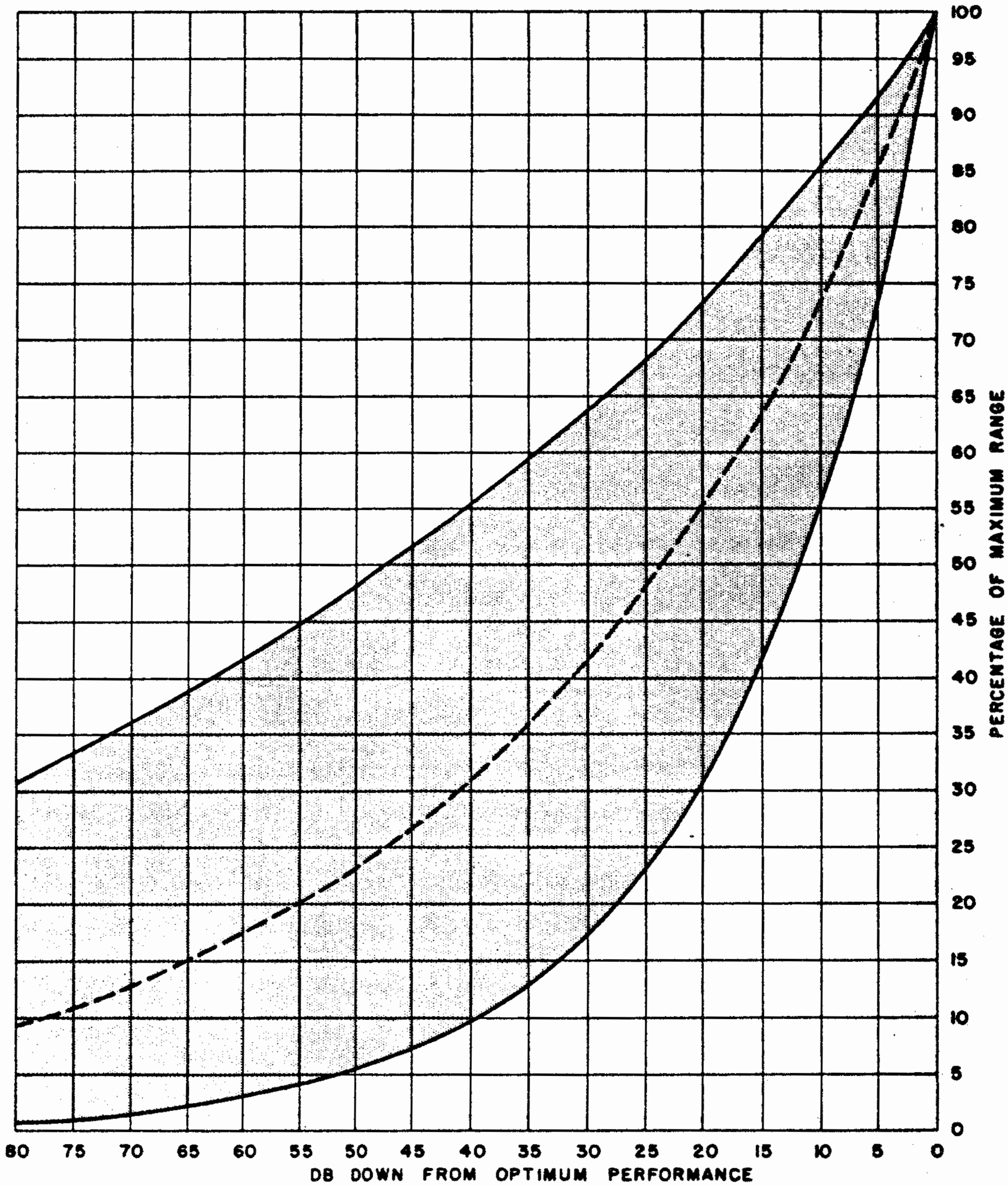


Figure 3-111. Radar Performance Versus Maximum Range

The radiated beam will therefore be aimed well below and above the horizon at times. This will cause "fading" of some valid targets and can produce false targets due to delayed reflections. Blind areas and "fade zones" are common troubles in the lower UHF radars. Blind areas caused by the radar line of sight can be calculated with the aid of Figure 3-112 for surface targets. Air target compensations can be calculated from Figure 3-113. The angles given are referenced to the antenna. It can be seen that a radar antenna with a vertical beam width of 10 degrees could detect an air target at 100,000 feet and present it until it was 90 miles distant, when the indication would fade out. Air-search radars employ on the order of 45° vertical beam widths for detection of air targets that may be re-entering the atmosphere; tracking them as they are passing overhead; or for detection of high altitude, long range targets

3-6.97 PROPAGATION FACTORS

Atmospheric conditions play a very important part in radar performance. Some of the more common factors are: 1) "duct" formation; 2) temperature inversion and atmospheric refraction; 3) rain echoes and scattering; and 4) atmospheric absorption.

Table 3-15. Estimated Range for Different Propagation Conditions

MAXIMUM RANGE		
PROPAGATION CONDITION	OPTIMUM PERFORMANCE	PERFORMANCE 20 dB DOWN
4m	100 miles	31 miles
8 ✓	10 miles	5.6 miles
16 ✓	3.2 miles	2.4 miles

3-6.98 DUCT FORMATION

Duct formation occurs when there is a sharp discontinuity in the atmospheric conditions close to the ground. The discontinuity reflects a transmitted signal in about the same manner as a metallic surface, and thus directs the wave back to earth, where reflection again occurs. Therefore the space between the earth and the discontinuity, in effect, acts as a waveguide. As a result, an abnormally long radar range may be observed.

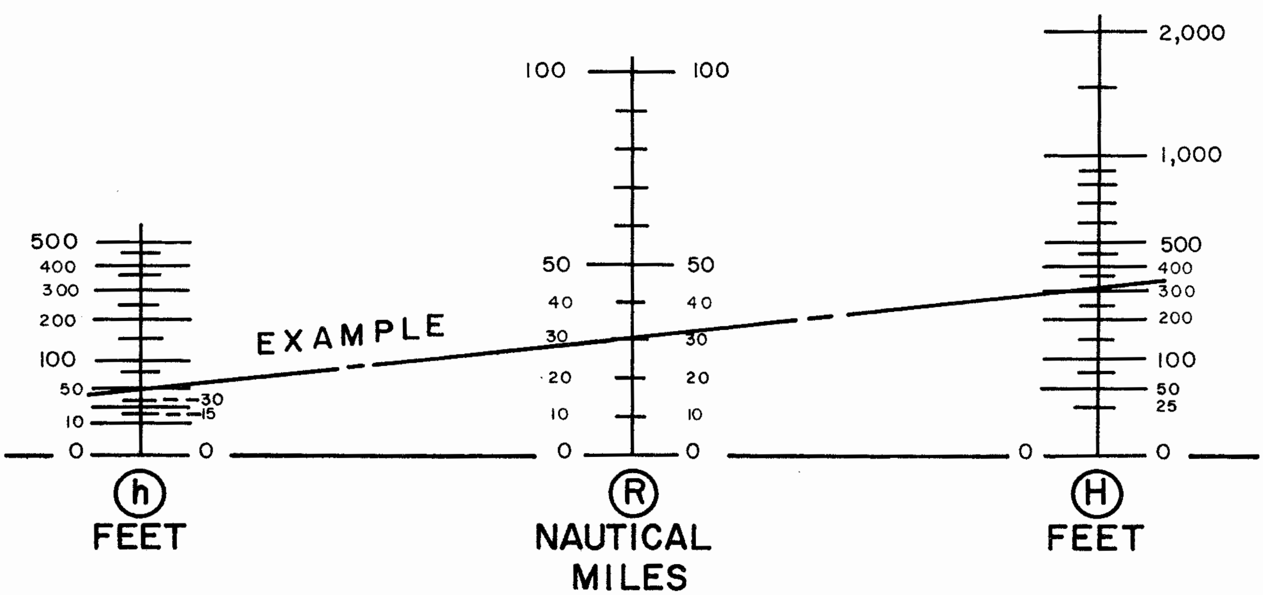


Figure 3-112. Surface Target Nomograph

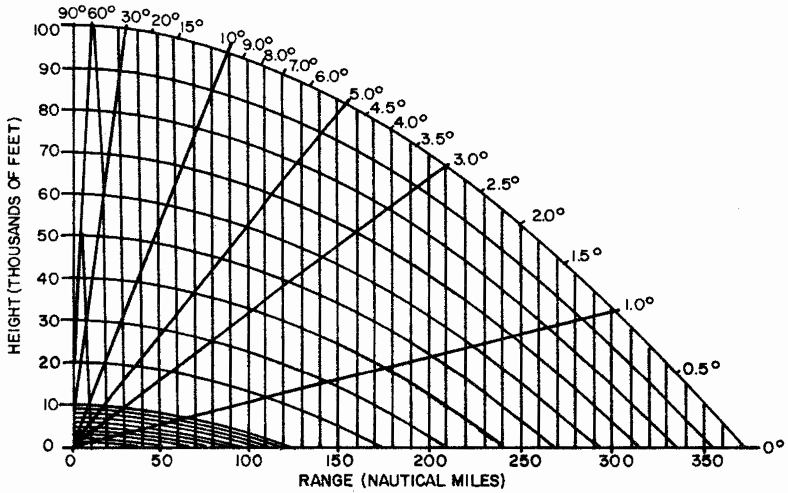


Figure 3-113. Air Target Compensation

3-6.99 ATMOSPHERIC REFRACTION

Atmospheric refraction is a phenomenon in which radar waves are bent in the earth's atmosphere. Under normal conditions the atmosphere is more dense at the surface of the earth, becoming less dense as altitude increases. As a result, electromagnetic energy travels more slowly at lower altitudes, and is effectively bent downward. The radar horizon is therefore extended about 15 percent beyond the calculated horizon under normal conditions. This phenomenon may be further augmented by a condition known as temperature inversion, which is caused by a warm air mass surmounting a colder air mass. The increased temperature at higher altitudes further decreases normal atmospheric density, as compared to the surface density, and the radar horizon is greatly extended. Temperature inversion is very common where warm air masses from land move over the cool air directly over a large body of water. The opposite condition can also prevail if the gradation of density is reversed, in that a colder air mass surmounts a warm air mass; refraction will then cause the radar wave to bend upward and thus greatly reduce the radar horizon.

3-6.100 RAIN ECHOES AND SCATTERING

Moisture in the atmosphere may cuase microwave signals to be either scattered or reflected, depending upon the size of the droplets. If the droplets are rather large, as in a heavy rain cloud, reflection occurs and causes an echo. This effect is very noticeable at the higher microwave frequencies. Smaller droplets may cause scattering rather than reflection, causing the range to be greatly reduced.

3-6.101 ATMOSPHERIC ABSORPTION

Atmospheric gases have the property to absorb certain microwave frequencies. Each gas has its own absorption spectrum. Of the gases studied thus far, each is unique in regard to the absorption of frequencies. For example, water vapor absorbs strongly above 10,000 MHz, showing a peak at approximately 23,000 MHz. Oxygen absorbs very strongly at about 60,000 MHz, and ammonia gas at about 24,000 MHz. The fact that absorption characteristics of various gases differ markedly has made it possible to analyze gases by means of their absorption spectra. The absorption effect is very undesirable in radar operations, because it results in reduced range at the frequencies of maximum absorption. Fortunately, this effect is not pronounced in the X band and at lower frequencies, but it does make the K band very unreliable.

3-6.102 RESONANCE CHAMBER (ECHO BOX)

An echo box, or resonance chamber, consists basically of a resonant cavity, the dimensions of which are determined by the frequency band in which operation takes place. The resonant cavity is tuned by a plunger, which can be adjusted back and forth in the cavity. This plunger is mechanically connected to a calibrated tuning dial. Connection to the radar set is made by a pickup dipole or a coaxial horn placed in the antenna radiation field, or by means of a cable connected to a directional coupler in the transmission line of the radar. An output power meter circuit made up of a microammeter, a crystal, a filter capacitor, and an attenuator (to prevent overloading) are usually included as a part of the echo box test equipment. The output meter indicates the relative power output of the radar transmitter. Refer to the block diagram of an echo box, Figure 3-114.

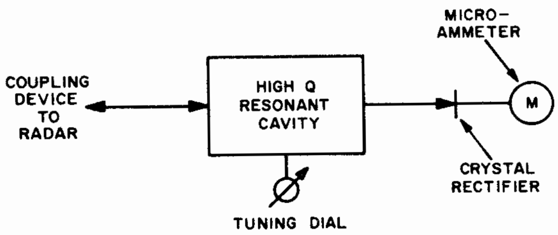


Figure 3-114. Typical Echo Box

3-6.102.1 Echo Box Operation

Any tuned circuit can be shock-excited by a sudden application of energy. When the excitation is removed, the tuned circuit continues to oscillate (ring) for a length of time. The greater the Q of the resonant circuit, the longer the ringtime. In use, the echo box picks up RF energy from a transmitter pulse. When the cavity is tuned to the frequency of the pulse, the RF energy picked up causes oscillations to build up in the resonant cavity. These oscillations continue after the radar pulse is cut off; however, the amplitude of each succeeding oscillation decreases, because of internal losses and output meter dissipation, and because of internal losses and output-meter dissipation, and because some of the energy is coupled back to the radar set. The energy coupled back is detected by the receiver, and appears as a pattern on the radar indicator (Figure 3-115). Ringtime is measured in terms of either yards or microseconds between the start of the transmitter pulse and the point where the ringing signal reaches the noise level of the radar receiver. The value of ringtime is influenced by the following factors: (1) receiver sensitivity; (2) peak transmitter power; (3) coupling loss width; and (4) the Q of the echo box. It should be noted that the first two factors provide a check of system sensitivity. This check, however, is not reliable unless the other three factors are either known or kept constant.

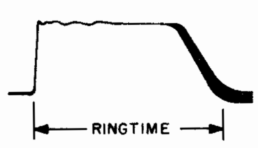


Figure 3-115. Ringtime Indication on a Scope

3-6.103 COROLLARY DATA

Because of its simplicity and compactness, the echo box is a very valuable test equipment for periodic system testing. However, an echo box presents only relative information. The echo-box installation must first be calibrated with standard test equipment before the information has any practical value. In fact, the echo box must be recalibrated at regular intervals, to ensure that the information gained is reliable.

3-6.103.1 Echo Box Installation

Correct installation of an echo box is very important. If the radar has a directional coupler, the echo box is located at some convenient point, and an RF cable is used to connect the echo to the coupler. It is important that the same cable and echo box be used for all subsequent testing. If the radar has no directional coupler, a pickup antenna is permanently installed at a point where reflections are at a minimum, and the echo box is connected to the pickup antenna with an RF cable. Again, it is important that the same pickup antenna, cable, and echo box be used for any subsequent measurements.

3-6.103.2 Multiresonant Boxes

The multiresonant type of echo box, which is used to some extent in the field, is made up of a cavity of irregular shape and of a size corresponding to several wavelengths. Because of its construction, the multiresonant echo box effectively functions as many cavities of different sizes with overlapping response curves. Consequently, it is resonant over a broad band of frequencies and does not require tuning. In most cases, a pickup antenna is built into one end of the box. The multiresonant echo box has an extremely high value of Q. However, because frequency and size are interdependent, only those frequencies of the X band or higher permit construction of reasonably sized boxes.

3-6.104 Calibration

It is possible to calibrate the echo box so that ringtime may be correlated with system sensitivity; future ringtime readings can then be converted into sensitivity readings. The conversion is easily made, because the change in ringtime per dB change in sensitivity is specified for an individual echo box. A common figure encountered in the field is about 100 yards per dB. Thus, if a radar has lost 1000 yards of ringtime, the sensitivity has decreased about 10dB. When the ringtime is found to be low, the meter reading is noted and then compared to the calibrated reading. Since the meter measures relative transmitter output, a low reading indicates trouble in the transmitter. A normal meter reading, coupled with a low ringtime indication, however, points to trouble in the radar receiver. To calibrate an echo-box installation, proceed as follows:

1. Orient both the radar and pickup antennas for maximum pickup. Unless reflections are found to be present, this step represents the final adjustment of the pickup antenna.

2. Record all settings of radar controls that affect PRF and pulse width.

3. Adjust radar receiver gain for about 1 cm of noise as shown on an oscilloscope connected to the receiver output.

4. Tune echo box for greatest ringtime indication on the oscilloscope.

5. Adjust coupling in echo box to give a standard meter reading of 75 percent of full scale.

6. Carefully read ringtime; at the same time note the echo box temperature.

7. Measure system sensitivity, using standard test equipment and procedure. Ringtime is then correlated with system sensitivity so that future ringtime readings can be converted to sensitivity readings.

Step 1 above can be deleted in the case of equipments using duplexes with connections for echo boxes. Large nearby objects must be taken into consideration so as not to increase ringtime or affect overall performance of the radar set. During later tests of sensitivity, using the echo box, the temperature-correction factor is included in the technical manual. As the echo box temperature decreases, the Q increases and ringtime will increase.

3-6.105 TR RECOVERY TIME CHECK

The TR recovery time is checked by the use of an echo box, as follows:

1. Note the slope of the response between receiver saturation and noise level.

2. Detune echo box until this slope just starts to change.

3. Read ringtime. This reading is the TR recovery time.

3-6.106 SPECTRUM ANALYSIS

The spectrum of a transmitter is analyzed by the use of an echo box, as follows:

1. Detune echo box by rotating tuning dial in one direction until meter indicates zero.

2. Rotate tuning dial in the opposite direction, and record the meter readings at various dial positions. Be sure to record all maximum and minimum dial readings.

3. Plot a graph of meter readings (vertical) and dial readings. The graph represents the transmitter spectrum.

3-6.106.1 Precautions

The following precautions should be observed when using an echo box:

1. The same echo box, cables, and pickup device should be used each time the tests are performed.

2. Make certain that the same radar test conditions are established each time the echo box is used. Record all control settings.

3. Measure ringtime very carefully. It is recommended practice to take several readings and average the results. Use a precision range marker if possible.

4. Keep accurate records, as specified in the technical manuals for the radar.

5. The echo box employs a crystal diode to detect the signal for meter. The echo box should therefore be either disconnected or detuned when not in use.

3-6.107 SYSTEM TROUBLE SHOOTING

A radar may show a gradual decline in performance and eventually reach a point where corrective maintenance is required. On the other hand, the radar may suddenly develop a fault. The suddenly-developed fault is immediately obvious. However, the purpose of periodic testing and recording of performance is to anticipate possible troubles. Periodic testing shows any trend as it develops, and in many cases minor corrective action at this time prevents a future major breakdown. One of the most difficult jobs confronting the technician is locating the specific cause of a certain trouble. In localizing the cause of a loss of performance, the first step is to determine whether the trouble is located in a particular unit, such as the modulator-transmitter or the receiver. This may be done by the use of the echo box and the information listed in Figure 3-116, or by the use of the various test equipments and procedures previously described in this section on Radar Testing. From there the specific circuits or source of power for that unit are tested. In the following text, both receiver and transmitter are analyzed in detail (the power-supply troubles, being rather straightforward, are not covered).

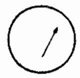
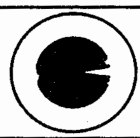
EFFECT	APPEARANCE ON		PROBABLE CAUSE
	RADAR INDICATOR	ECHO BOX METER	
RINGTIME AND TEST SET OUTPUT SATISFACTORY			RADAR PERFORMANCE SATISFACTORY
RINGTIME LOW, OUTPUT READING SATISFACTORY			RECEIVER TROUBLE: DETUNED MIXER OR LOCAL OSCILLATOR, BAD CRYSTAL, EXCESSIVE I-F NOISE, ADJUSTMENT OF PROBES IN MIXER CAVITY DETUNED T-R BOX.
RINGTIME LOW, TEST SET OUTPUT VERY LOW.			LOW POWER OUTPUT. CHECK SPECTRUM.
RINGTIME LOW, TEST SET METER READING LOW.			TROUBLE PROBABLY IN TRANSMITTER AND RECEIVER AND/OR TROUBLE IN TRANSMISSION LINE.
RINGTIME ERRATIC, TEST SET METER READING STEADY.			TEST SET DETUNED. BAD PULSING, DOUBLE MODING TRANSMITTER, OR LOCAL OSCILLATOR POWER SUPPLY TROUBLE. CHECK SPECTRUM.
RINGTIME ERRATIC, TEST SET OUTPUT READING ERRATIC.			FAULTY TRANSMISSION LINE OR CONNECTION – CONDITION WORSE WHEN LINE IS RAPPED
END OF RINGTIME SLOPES GRADUALLY, PERHAPS EVEN EXCESSIVE RINGING. GRASS APPEARS COURSE. TEST SET OUTPUT READING STEADY AND SATISFACTORY.			OSCILLATING I-F AND/OR NARROW BAND RECEIVER
PRONOUNCED DIP IN RING-TIME AT END OF PULSE.			FAULTY TR OR DUE TO RECEIVER GATING ACTION
RINGTIME VERY SLIGHTLY LOW, POOR OR BAD SPECTRUM.		 POOR SPECTRUM	TRANSMITTER TROUBLE.
BLANK SPACES OR ROUGH PATTERN ON PPI RINGTIME INDICATOR, TEST SET OUTPUT READING VARIES AS RADAR ANTENNA IS ROTATED.			FREQUENCY PULLING OF TRANSMITTER DUE TO BAD ROTATING JOINT OR TO REFLECTING OBJECT NEAR RADAR ANTENNA.

Figure 3-116. Echo Box Indication of Radar Trouble

3-6.108 LOW RECEIVER SENSITIVITY

Low receiver sensitivity is evidenced by a high MDS test figure. The reasons for this condition may be either excessive noise generation or excessive signal loss preceding the I-F amplifier section. As long as the noise present in the receiver ouput is excessive, the I-F amplifier cannot contribute to the sensitivity. When defective or improperly adjusted, the following items may cause low receiver sensitivity: crystal mixer, local oscillator, I-F amplifier (first two I-F stages), and the TR and ATR tubes.

3-6.109 AFC OPERATIONAL DIFFICULTIES

Because proper operation of the AFC circuit is primarily dependent upon correct coupling between the AFC crystal mixer, the local oscillator, and the output of the magnetron, this portion of the circuit should be checked first. A variable coupling is usually provided to adjust the amount of local-oscillator signal injection to the AFC crystal mixer. Care should be taken to prevent overloading of the AFC crystal mixer, and reference should be made to the applicable technical manual as to the correct crystal current for proper operation. One milliampere of crystal current is usually the allowable limit. If the degree of coupling is found to be correct, the trouble may lie in a defective local-oscillator tube. The local oscillator must operate smoothly over the desired pull-in frequency range if normal AFC operation is to take place. In addition, it is essential to make sure that the local oscillator is operating in the correct mode on the proper side of the transmitter frequency. A magnetron with an improper frequency spectrum may also cause the AFC circuit to seem faulty.

3-6.110 POOR MINIMUM-RANGE PERFORMANCE

Minimum-range performance is controlled by the recovery time of the TR tube (and, if used, the pre-TR tube). Excessively long recovery time, of course indicates the end of the useful life of a TR tube.

3-6.111 INCORRECT OPERATING FREQUENCY

Incorrect operating frequency usually breaks down into two possible causes: 1) the magnetron may be defective; or 2) pulling may exist because of some fault in the RF assemblies or from strong reflections from a nearby object. When a new magnetron is inserted to correct off-frequency operation, it is not necessarily true that the original magnetron is defective. Individual constructional differences of magnetrons may vary, causing one to be pulled more easily by external conditions than another of the same type number. Irresponsible replacement of apparently defective magnetrons may result in the rejection of good tubes. It is first necessary to check for the presence of pulling, to determine whether the magnetron actually is at fault. This check is made by measuring the SWR of the RF assemblies, with the slotted line placed as close to the magnetron as possible, or by feeding the magnetron output into a dummy RF load and rechecking the frequency. When off-frequency operation occurs with a low SWR, the indication is that the magnetron should be replaced, unless, of course, it is of the tunable type.

3-6.112 POOR SPECTRUM

As was previously discussed, spectrum analysis is of considerable importance in the maintenance of a radar facility. The reason for a poor magnetron spectral display or graph can be magnetron pulling or pushing, a defective magnet, or a defective magnetron.

3-6.113 MAGNETRON PULLING

The test for magnetron pulling is made by means of SWR measurements or by the use of a dummy antenna, as mentioned above. Magnetron pulling may cause frequency shift, but this may go unnoticed if the frequency is still within the operating band.

3-6.114 MAGNETRON PUSHING

A poor spectral display or graph is often evidence of magnetron pushing, and this fault is the result of improper modulator operation. When the output pulse is of improper shape or amplitude, especially at lower power levels, excessive AM or FM may be present. The test applied to the modulator is made with the aid of a synchroscope and voltage divider. The voltage divider serves to reduce the modulator pulse to a usable amplitude. This amplitude is observed and multiplied by the appropriate factors. The pulse shape is observed and compared with available waveform charts. Under certain conditions, the magnetron causes improper loading of the modulator, and thus introduces pulse distortion. The use of a dummy load for the modulator eliminates this condition. The modulator dummy load is a resistive impedance equal to the firing impedance of the magnetron; in most cases a voltage divider is built into the test equipment to facilitate the measurement of pulse amplitude. This load replaces the magnetron during pulse measurements, and therefore helps to isolate trouble definitely to the modulator.

3-6.115 DEFECTIVE MAGNETRON

A poor spectral display or graph may indicate a defective magnetron. A weak magnet may cause the magnetron input to exceed rated values; if so, continued operation results in a damaged unit. Missing lines in the spectral display are the result of magnetron arcing and, if excessive, may completely destroy the shape of the spectrum. Many magnetrons display moderate arcing until seasoning is completed. This should therefore be allowed a sufficient breaking-in period before the spectrum is analyzed. As mentioned previously, the end of the useful life of a magnetron is characterized by an increase in arcing and general instability. When the output power is low, it usually inicates a weak magnetron or a low modulator output. This uncertain condition may be resolved by testing the modulator output pulse; normal pulse indicates that the trouble is in the magnetron.

3-6.116 BEAM WIDTH DETERMINATION

The upward trend in frequency for radar and microwave communications uses has been largely the result of tremendous antenna gain which can be realized by using moderate size paraboloidal reflectors as a part of the antenna system. The reason for this is that, for a given size paraboloid, an increase in frequency produces a decrease in radiated beamwidth. The equation for approximating beamwidth at the half-power points is:

$$\theta = 70 = \frac{\lambda}{D}$$

where:

θ = beamwidth in degrees

λ = wavelength — in the same units

D = width of paraboloid

3-6.117 RADAR MTI TESTING

All surface-to-air search radar equipment employ moving-target indication (MTI) receiving equipment in addition to the normal receiver. The MTI receiver output consists of a video signal in which (at least theoretically) all returns from fixed targets have been eliminated, leaving only moving targets to be seen on the MTI display. The ground clutter that masks targets on the normal display is therefore removed on the MTI display, enabling the operator to "see" moving targets which would otherwise go undetected. Figure 3-117 shows the video section of a typical MTI Receiver.

3-6.118 CANCELLATION RATIO

The cancellation ratio is a measure of the effectiveness of cancellation of fixed targets in an MTI receiver. It is obtained by measuring the amplitude of the uncancelled residue of the same target with the cancellation circuits in operation, and then determining the difference in amplitude. The amplitudes are measured at the output of the MTI circuits, using an RF signal generator and a synchroscope.

3-6.118.1 Measurement

In making the amplitude measurements, the output from a pulsed signal source is coupled into the duplexer or, in some equipments, into a built-in

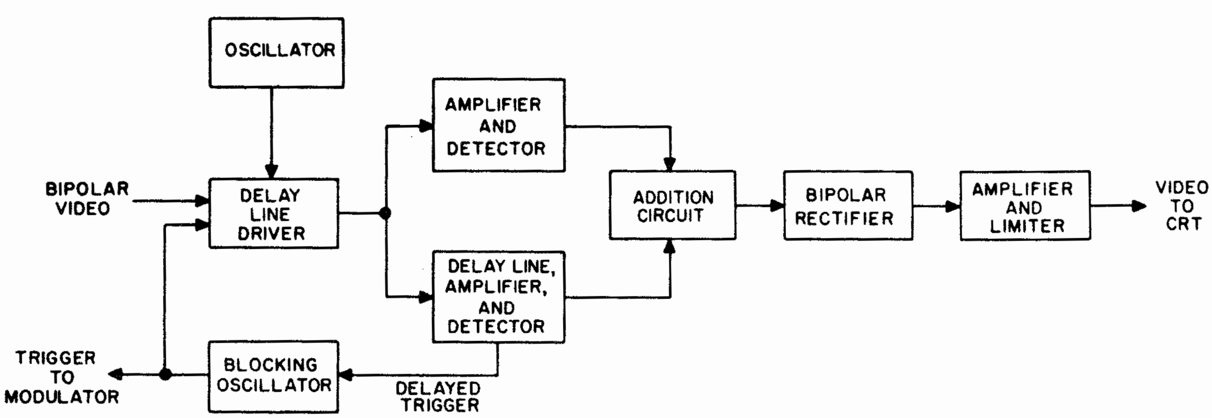


Figure 3-117. Video Section of a Typical MTI Receiver

input connector. Antenna rotation is stopped in a direction which allows display of a suitable fixed target. The MTI is then disabled by changing the PRF or by causing the coho oscillator to become non-coherent (thus effectively disabling the cancellation circuits). The signal generator pulse amplitude is then adjusted by means of the calibrated attenuator until it is equal to the amplitude of the uncancelled target return, and the antennuator dial reading noted. Upon restoration of the MTI to normal operation, the amplitude of the fixed target should be greatly reduced. The calibrated attenuator on the signal generator is adjusted until the signal generator output pulse is equal in amplitude to the uncancelled residue of the fixed target, and the attenuator dial reading is again noted. The cancellation ratio is the difference of the two values of attenuation, and is usually given in dB. The actual value depends upon the type of equipment, but is usually on the order of 20 to 30 dB.

3-6.119 SUB-CLUTTER VISIBILITY

The sub-clutter visibility is a measure of ability to detect a small moving target in ground clutter. The test consists of determining the difference in amplitude between a cancelled fixed target and a moving target at the same range. Ideally, any target visible in the random noise should be visible over a fixed target, but because of the fact that a fixed target leaves some uncancelled residue, this condition is not generally attained.

3-6.119.1 Measurement

In performing this test, the cancellation circuits are disabled as in the previous test, and the antenna is rotated until it is centered upon a suitable fixed target. The receiver gain is then lowered until the target return is not limited, and an RF signal is introduced from the signal generator. (Since the signal generator oscillator is not phase-locked to the transmitted RF energy, the output pulse appears as a moving target.) The signal generator pulse delay is adjusted until the pulse appears next to the fixed target in range, and the calibrated attenuator is varied until the signal generator pulse is equal in amplitude to the fixed target. The dial reading is then noted. The MTI equipment is restored to normal operation, and the signal generator pulse delay is adjusted until the moving target (signal generator pulse) coincides in range with the fixed target. The calibrated attenuator is then varied until the moving target is just visible above the fixed target. This is accomplished by increasing the attenuation until the target is invisible and then slowly decreasing the attenuation until the moving target is just visible. The attenuation value is noted. The sub-clutter visibility is the difference of the two attenuator dial readings, and is usually expressed in dB. The actual value depends upon the type of equipment, but is generally on the order of 15 to 20 dB for older equipments and may be as high as 35 to 40 dB on the most modern types of equipment.

3-6.120 MTI TROUBLES

Moving-target-indicator (MTI) circuits are susceptible to the troubles commonly encountered in other electronic circuits. Since the delay line and the coherent oscillator are common only to the MTI circuit, these will be discussed briefly.

3-6.121 EXCESSIVE DELAY LINE ATTENUATION

Excessive delay line attenuation may be caused by dirty mercury. The mercury should be drained off and then replaced with clean mercury. For detailed instructions, refer to the applicable technical manual. When a delay line is refilled with mercury, or if it has been subjected to severe vibration, excessive attenuation may result. Allowing the delay line to rest for approximately 24 hours results in the mercury settling and the condition is corrected.

3-6.122 TOTAL FAILURE OF DELAY LINE

Total failure of the delay line is usually caused by either broken crystals or by a shorted coaxial cable connector in one of the two transducers. A broken crystal is manifested by leakage of mercury at the transducer tank. If no leakage is present, it may be assumed that the trouble is in the coaxial connectors. If a crystal must be replaced, refer to the applicable technical manual for detailed instructions.

3-6.123 COHERENT OSCILLATOR TUNING

The coherent oscillator has been designed to be exceptionally stable. The usual frequency tolerance of the MTI circuits requires not more than a 10-hertz deviation in frequency during the interval between transmitting pulses. Because of this extremely small tolerance, the tuning of the coherent oscillator should be checked carefully and (if necessary) readjusted. Since different types of MTI equipments differ so greatly, no attempt will be made to describe any special procedure. For complete detailed instructions, refer to the applicable technical manual.

3-6.124 POWER GRID TUBES

These tubes are widely used in equipment that may range up to several gigahertz in frequency. High average power can be sustained, with ruggedness required to cope with combat situations. Considerations must also be given to long life and the operational limitations of such tubes. With respect to filament, a 5% overvoltage will decrease the life of a tube by as much as 40-50%. The tube should be operated in the "knee" region, which is a function of filament voltage versus output power. The tube should be operated on the lower edge of this "knee". Figure 3-118 displays such a graph. Screen voltage should be kept high and the grid bias adjusted for best spectral output. Grid drive should be moderate to compensate for transient "back-heating" time. This is defined as heating due to secondary bombardment of the tulo elements due to the electrons traveling from the cathode to the plate. This travel is actually reversed temporarily while transitting the space between the tube elements.

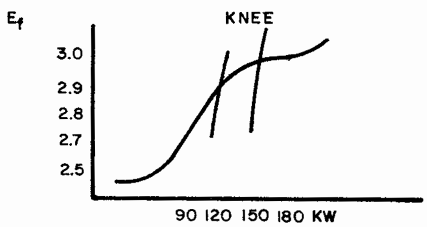


Figure 3-118. Filament Voltage vs Output Power

3-6.125 COOLING

Both air or water cooling should be maintained at top performance to prevent arc over of elements or the establishment of current paths due to contaminated cooling lines or air passages. Systems using distilled water should be provided with storage tanks that vent outward only. This prevents the water from aerating and absorbing oxygen which could cause equipment corrosion and compromise water purity. All systems have "targets" for minimizing electrolysis, and these should be checked periodically. The resistance of water used for radar cooling should be maintained above 1×10^6 ohms. A megger should be used to check hoses and insulators for current leakages, with readings of over 40 megohms for all involved. Reduced power should be used as much as possible except when jamming or small targets are encountered. A 3dB reduction in output power can increase tube life by as much as 40%.

3-6.126 MODULATOR PULSE

The modulator pulse produced from a PFN (Pulse Forming Network) is affected only by changes in load; voltage being applied to charge the PFN; triggering the thyrotron; and the PFN itself. The output pulse is a square wave that can be observed on an oscilloscope through a test point provided, or acquired through a pickup probe that can be placed near the output cable. Figure 3-119 is a typical block diagram of a modulator incorporating a PFN. Note that the charging choke (L102) and the pulse transformer (L103) are designed to have sharp leading and trailing edges for the mode in which the modulator operates; therefore, both the electrical and physical properties of each are critical. The shunt diode (V110) prevents the PFN from reverse charging. Test point TP101 shows the input pulse to the PFN. Both DC level and pulse amplitude should be noted for use in troubleshooting.

3-6.127 DRIVER AMPLIFIERS

Systems incorporating tuned amplifier stages involve special considerations. Where possible, a sweep generator should be used in aligning circuits which must operate over a particular band. Tube interelectrode capacitance and transistor junction capacitance are critical when used in VHF/UHF equipment. The effects can be noted in Figure 3-120 which shows a circuit (A) tuned to the low end of the harmonics, and the same circuit (B) when it is tuned to the high end of the harmonics that generate a square wave. Part (C) represents a correctly-tuned square wave amplifier. The curving up or down determines whether high or low frequency attenuation is involved. The lower rounded leading/trailing edges, as shown in Figure 3-121A, indicate a lack of the harmonics necessary to generate the square wave. This can be due to low bias, insufficient bandwidth of the amplifying stages, or to lack of drive to the amplifiers. Increased rounded edges, as shown in Figure 3-121B indicate excessive harmonics necessary to generate the square wave. A recheck of the sweep generator deviation should be included to ensure adequate output and bandwidth.

3-6.128 KLYSTRON AMPLIFIERS

Klystron amplifiers are used in all UHF and above radars, either as local oscillators or as amplifying devices. Gains of +105 dB with relatively low noise figures of +4dB are available. The impingement of electrons on the elements and cavities of the klystron creates most of the noise generated. The most frequent cause of circuit noise, however, is the power supply ripple or intermodulation which has caused premature removal of a good amplifier/oscillator. Improper tuning of the input-output circuit may cause unwanted harmonics which add significantly to the noise figure

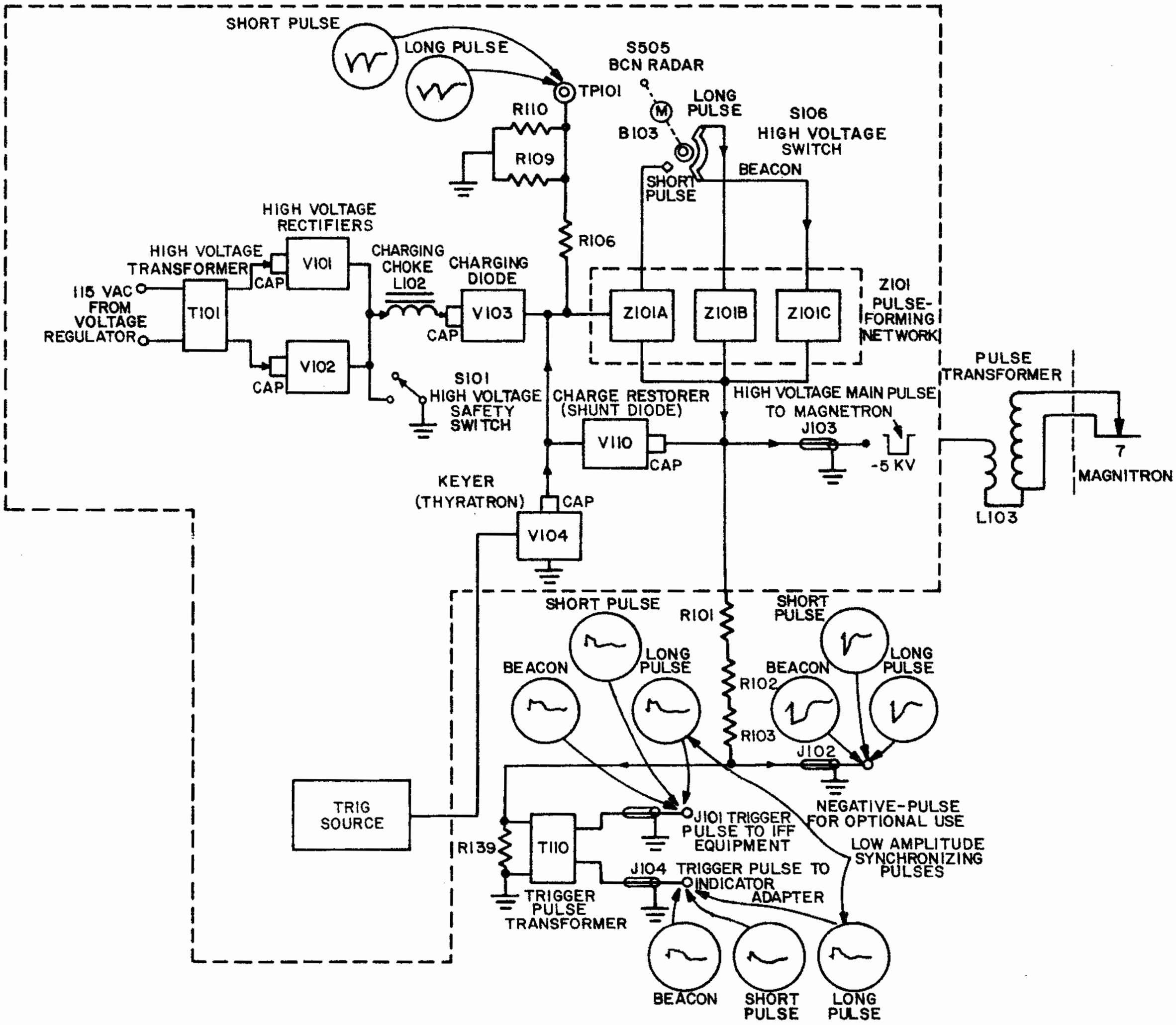


Figure 3-119. Typical Block Diagram of a Modulator Incorporating a PFN

of a klystron. Klystron amplifiers should not be used to amplify amplitude-modulated signals if the RF output is driven higher than 80% of the saturated level. Pulsed and CA methods are similar to standard triode operation, with the exception that most multicavity klystrons that are to be used over a band must have each cavity tuned for each change in frequency. Frequency changes should therefore be kept to a minimum. Beam current should also be kept to a minimum. The PMS covers the performance requirements of the concerned equipment. Notes of frequency, beam current, collector current, input power and output power should be maintained. Figure 3-122 is an electrical representation of a 3-cavity klystron amplifier. A general description of its operation is as follows. The electron beam is generated in the electron gun section, which also will focus the beam. In the RF section, the input RF is coupled to the beam via the tunable cavity. Since the output is RF-coupled, beam current will exist without input or output signals being present: the first cavity ("buncher") is

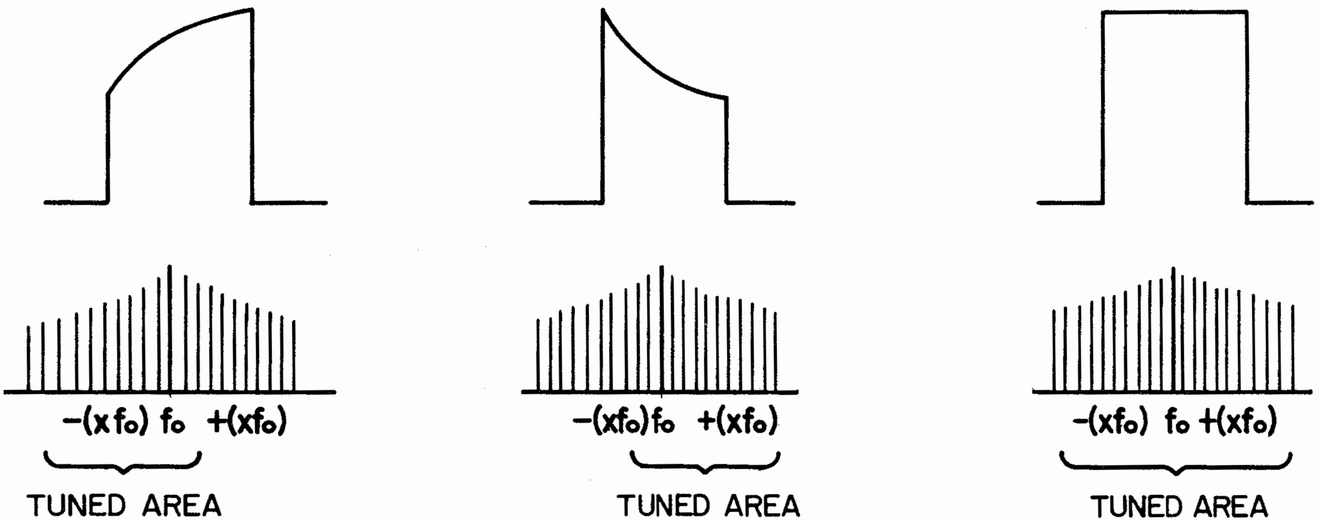


Figure 3-120. Square Wave Generation

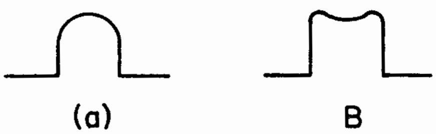


Figure 3-121. Effects of Harmonics on Wave Shape

cavities greatly affects the amplification factor. Spacing and size affect the frequency of operation. The collector serves to return propelled electrons to the power supply. Figure 3-123 shows power connections for a typical klystron amplifier. Either an electromagnet, a permanent magnet, or both may be employed. The efficiency of this type of amplifier is on the order of 30 to 60 percent. Figure 3-122 earlier showed an electrical representation. Figure 3-124 shows another representation with tuning diaphragms in each cavity and other applicable elements. The input cavity is a high Q tank circuit and the Q of each individual cavity decreases to the output cavity which is the lowest Q circuit. Looking at Figure 3-124 it can be seen that electrons emitted from the cathode are formed into bunches as they pass through the tube. The electromagnet surrounding the tube focuses this electron beam and directs it through the cavities and drift tubes to the collector.

3-7 IFF SYSTEMS

The IFF systems operate on the challenge/response principle. Five modes are employed to transmit coded challenge interrogator signals to unidentified radar targets. In response, the systems receive coded recognition replies from "friendly" targets equipped with identification (Transponder) components. The IFF system has at times been referred to as the "Identification: Friend or Foe" system. The IFF system also enhances the tracking of friendly targets (air and surface) which may otherwise be obscured by radar clutter. Physically, the Interrogator and Transponder interface with the radar system and its associated displays. Functionally, the system interfaces with Army and Air Force AIMS systems and with the FAA Air Traffic Control Radar Beacon System (ATCRBS). The system also interfaces functionally with its predecessor, the MK X (SIF) IFF System. Most equipments are micro-miniaturized solid state and employ integrated circuits mounted on PC Card Modules for about 85% of the circuitry. Redundancy of the interrogation function on major combatant vessels is currently provided by using interrogator sets on a one-per-radar basis. The decoding subfunction is redundant on a one-per-display basis. Two IFF systems currently in use are the MARK X and MARK XII. The MARK X, which is being phased out, has three modes of challenge and reply operations. The MARK XII (AIMS) has five conditions of interrogate/reply. Mode "C" is designed for altitude reporting and Mode "4" is employed for computer-encrypted identification.

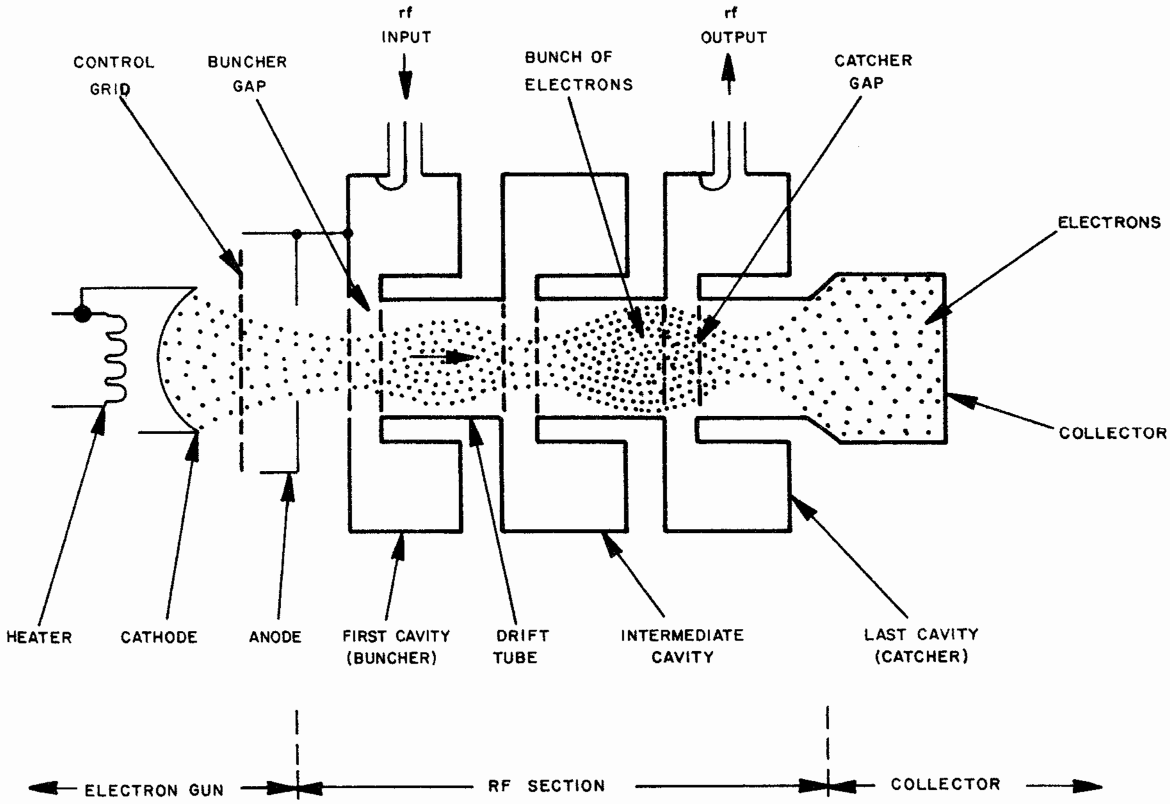


Figure 3-122. Klystron Electrical Representation

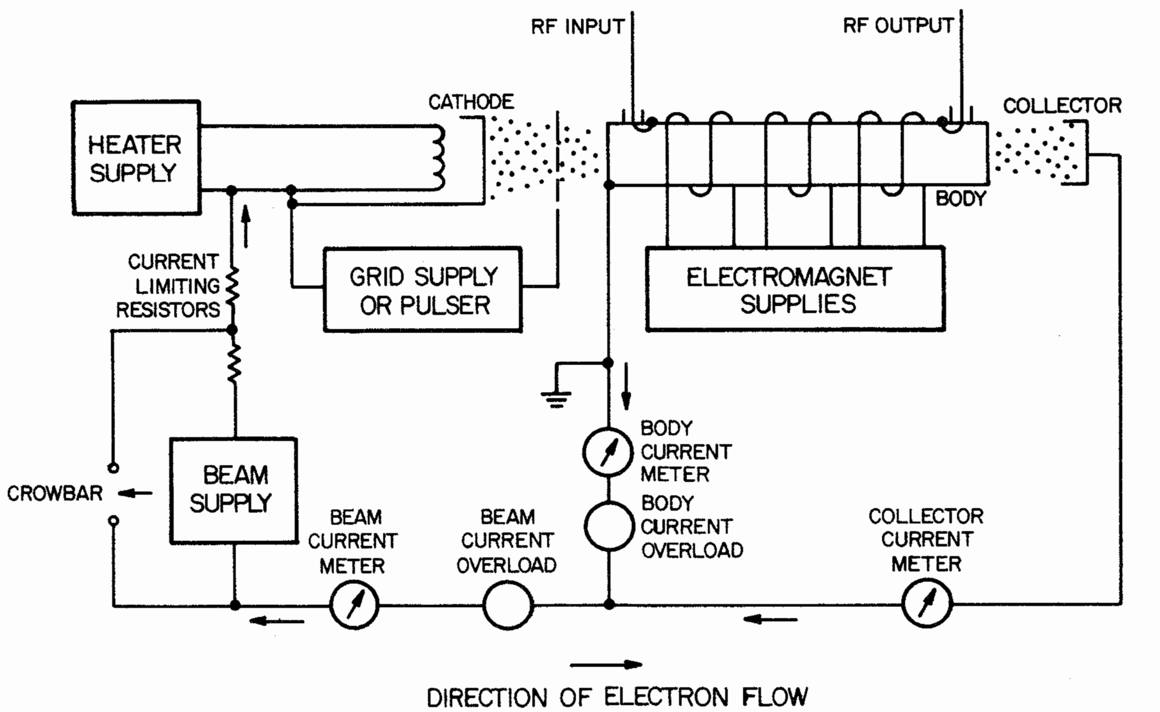


Figure 3-123. Klystron Power Supply Connections

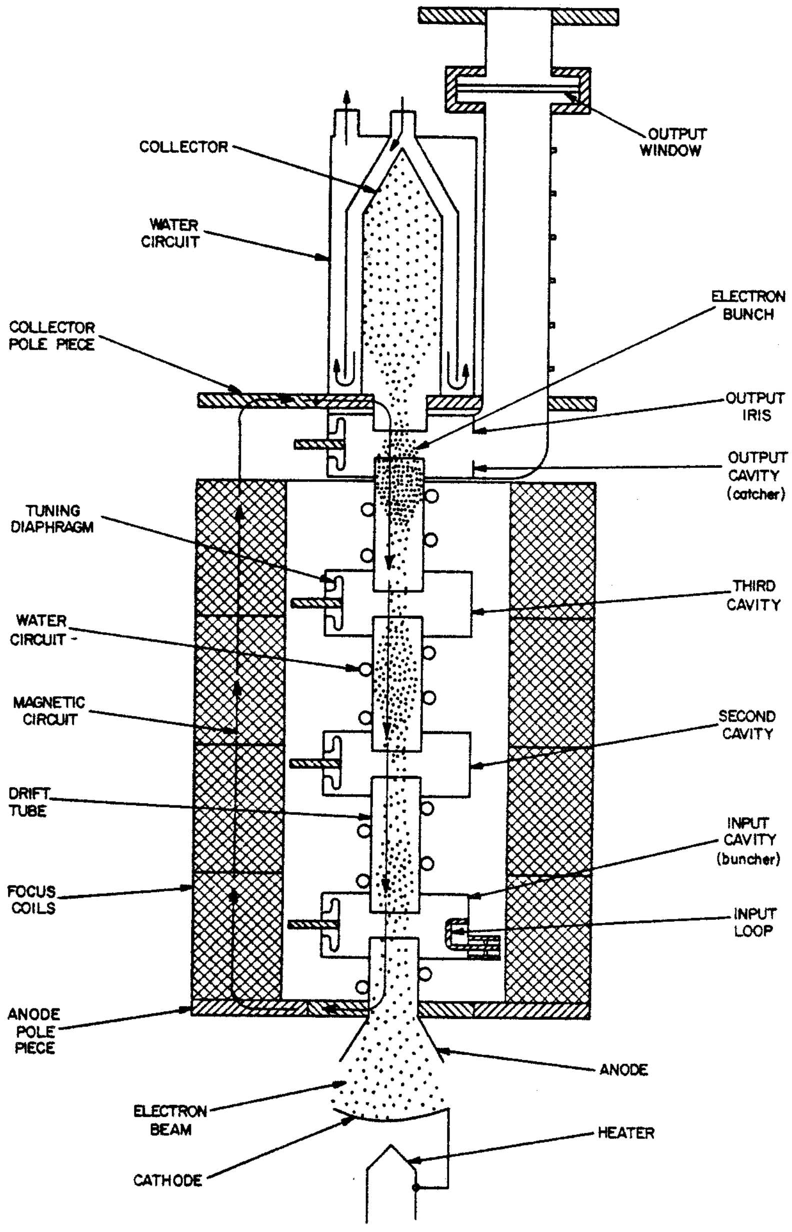


Figure 3-124. Klystron Sectional View

3-7.1 "AIMS" SYSTEM COMPONENTS

The "AIMS" system is composed of an Interrogator (Challenge) subsystem and a Transponder (Reply) subsystem. The Interrogator subsystem is normally used in conjunction with a radar set to which its operation is synchronized. The Interrogator subsystem permits the radar operator to interrogate (challenge) other platforms and to interpret this data as specific identification of friendly radar targets. The Transponder subsystem accepts a challenge from other platforms and provides coded replies that permit identification of its own platform. An Interrogation Side Lobe Suppression (ISLS) capability inhibits the Transponders from replying to interrogations from other than the main lobe of the interrogating antenna.

3-7.2 INTERROGATOR SYSTEM

The AIMS interrogation system is comprised of an Interrogator (challenge unit) and a Transponder (coded response unit). The Interrogator is customarily used in conjunction with a radar set with which its transmission and reception are synchronized. Radar contacts of interest may thus be interrogated, and the proper coded response indicates the interrogated contact (ship or plane) is friendly. The Transponder also accepts challenges from external transmitters (those from other ships or planes) and emits the proper coded response that identifies it to the challenger. Responses not incorporating the proper identification code are interpreted as coming from hostile units. A special feature, the Interrogation Side Lobe Suppression (ISLS) inhibits the Transponders from replying to challenges from other than the main lobe of the interrogating antenna.

3-7.2.1 Interfacing

The Interrogator and Transponder interface physically with the radar system involved and with its associated PPI displays. Functionally, the IFF system interfaces with Army and Air Force AIMS systems, and with the FAA's Air Traffic Control Radar Beacon System (ATCRBS). The system also interfaces functionally with its predecessor, the MK X (SIF) IFF system. Micro-miniaturized solid state modules and integrated circuitry mounted on printed circuit cards are employed in approximately 85 percent of IFF equipments. Redundancy of the interrogation function is ensured by using the Interrogators on a one-per-radar set basis. The decoding subfunction is redundant on a one-per-PPI basis.

3-7.3 AN/UPA-59A(V) SYSTEM COMPONENTS

Table 3-16 lists current Interrogator System components for the AN/UPA-59A(V). Antennas used in conjunction with the AN/UPA-59A(V) are the AS-2188/UP and the OK-217/UPX.

Table 3-16. Interrogator System Components

COMPONENT	FUNCTION
BZ-173A/UPA-59A(V)	Alarm Monitor
ID-1844/UPA-59A(V)	Indicator, Intratarget Data
KY-761(P)/UPA-59A(V)	Decoder, Video

3-7.4 AN/UPX-23 INTERROGATOR SET

This equipment has been designed to meet AIMS shipboard requirements. It consists of a transmitter and receiver capable of interrogating ships or planes equipped with MK X IFF/SIF and MK XII IFF radar identification sets (Transponders), receiving their RF response, and then processing such response into video signals to be applied to decoders and indicators for identity verification. A subsequent version of the UPX-23 (the AN/UPX-27) is electrically and mechanically interchangeable with the UPX-23.

3-7.4.1 AN/UPX-23 Subsystems

Table 3-17 lists the subsystems and respective functions of the AN/UPX-23 Interrogator Set.

3-7.5 ANTENNAS

The antenna configurations are tailored to suit specific Interrogator subsystems, as described in Table 3-18.

3-7.6 TRANSPONDER COMPONENTS

Table 3-19 lists the Transponder Set components and describes their respective functions.

3-7.7 TEST EQUIPMENT

Test equipment associated with the AIMS shipboard Interrogator/Transponder system is listed and described in Table 3-20.

3-7.7.1 Test Conditions

The test conditions for Transponder tests must fall within a specified set of limits before a "GO" status, as indicated by TS-1843A/APX, can be achieved. If any one or more of the listed characteristics fail to fall within the specified test limits, the test set will indicate a "NO GO" condition exists and must be corrected. In the "Test" function, a lighting of the "GO" lamp occurs on the Transponder Set Control. In the "Monitor" function, a "GO" status is indicated by a two-second illumination of the "GO" lamp, which then extinguishes. The process then repeats so long as the proper input is applied to the test set. Conversely, a "NO GO" condition is indicated by failure of the lamp to illuminate.

Table 3-17. AN/UPX-23 Interrogator Set Subsystems

SUBSYSTEM	FUNCTION
C-8430/UPX	Control monitor; employs switches and indicator lights to set up and monitor system operation
KIK-18/TSEC	Crypto Code Key; sets up individual code for KIR-1A/TSEC on one-code-key-per AIMS installation basis
KIR-1A/TSEC	Crypto computer; provides coding for the AN/UPX-23 Interrogator and crypto decoding of Transponder replies
MX-7647 (XN-1)/UPX or MK8110/UPX	Interference Blanker; preproduction unit wave used on some early ships
MX-8758/UPX	Currently used Interference Blanker (single channel "defruiter"); eliminates random nonsynchronous signals which appear on the PPI as momentary flashes of light or clutter, colloquially known as "fruit"
SG-841/UPX	Pulse Generator; samples trigger of associated radar and provides all necessary timing triggers for the shipboard AIMS operation
SG-993/UPX	This version of the pulse generator is the renamed SG-841/UPX after Mode B modification has been accomplished

Table 3-18. Antenna Configurations

TYPE	DESCRIPTION
AT-1688/SPS-48(V)	Directional "hog trough" array mounted "piggy-back" on the antenna of some AN/SPN-43 radar sets
AS-177A/UPX	Omni-directional antenna; used with Transponder Sets APX-64 and APX-72. Generally installed on a ship's yardarm. In some installations used as an ISLS antenna. Also used with AIMS Transponders as a transmitting antenna, and with the ASSECTIS for system tests. Classed as a consumable item because no piece-part support exists in the Navy Supply System, but can possibly be repaired if care is exercised in the process
AS-2168/U or AS-2787/UPX	These are lightweight "hog trough" ten-foot span arrays mounted "piggy-back" on search radar antennas or, when pedestal-mounted, are slaved to a given antenna. Capable of generating sum and difference patterns for ISLS operation
AS-2189/U	Five-foot span antenna having ISLS capability. Gain is 3 dB less than and its horizontal beam width is twice as great as those of the AS-2188/U antenna. Can either be mounted "piggy-back" or slaved on a separate pedestal as the radar installation may dictate. If a separate pedestal is required, the AN/UPA-57 Antenna Pedestal should be utilized

Table 3-19. Transponder Set Components

COMPONENT	DESCRIPTION
AN/APK-64(V) or AN/APX-72	When properly challenged, automatically transmits a coded reply. Primarily Aircraft Transponders adapted to shipboard use; can reply to a single mode, a combination of modes, or to all five modes, depending on control setting
CY-2332/APX-64(V)	Electronic Frequency Converter; changes 60 Hz input to the 400 Hz required by the AN/APX-64(V) Transponder Set
PP-6099 ( )/APX-72	Power Supply; converts 115VAC to the 28VDC required for the AN/APX-72 Transponder Set
CY-6816/APX-72	Case, Control; adapts the C-6280/APX-72 Transponder Set Control (initially designed for Aircraft) for shipboard installation
See Note	NOTE The following are common to the AN/APX-64(V) and AN/APX-72 Transponder Sets
C-6280/APX-72	Controls operation of the RT-728/APX-64 or the RT-859/APX-72 Receiver-Transmitter and the KIT-1A/TSEC Crypto Computer. Mounted in CY-6816/APX-72 enclosure. (The CY-6816 enclosure is not used for an AN/APX-64(V) installation.)
KIK-18/TSEC	Crypto Code Key; used for setting the individual codes for KIK-1A/TSEC and KIT-1A-TSEC Crypto Computers. Only one code key per AIMS installation
KIT-1A/TSEC	Crypto Computer; decodes interrogations and produces appropriate coded response
MT-3951A/U	Mounting; used in installation of KIR-1A/TSEC and KIT-1A/TSEC Crypto Computers
TD-937( )/SPX	Electronic Gate (Suppression Mixer); acts to suppress the Transponder when one of the own ship's interrogators is radiating challenges from its own ship.)

Table 3-20. Interrogator/Transponder Test Equipment (Continued)

TEST SET	DESCRIPTION
AN/UPM-98A or AN/UPM-98B	Basically an AN/UPM-98 Test Set redesigned to meet AIMS requirements. Contains most of the features of the AN/UPM-137 except not of modular construction; utilizes vacuum tubes; will not interleave different pulse trains
AN/UPM-136	Transponder-only capability; for use on small surface vessels and submarines; similar to that portion of the AN/UPM-137 that applies to Transponders
AN/UPM-137	Universal IFF Test Set capable of calibrating and testing AIMS Interrogators, Transponders, Video Coders, Decoders, and Radar Video Distribution and Display Systems. Consists of four modules: 1. RF Module; 2. SIS Module; 3. Oscilloscope Module; and 4. Interrogator Module. Will ultimately replace the AN/UPM-98 for shipboard use.
TS-1843A/APX	In-line Test Set that provides "GO - NO GO" indication. Evaluates (via PMS checks) the following preset conditions of the Transponder: 1. Receiver Sensitivity 2. Receiver Frequency 3. Decoding 4. Reply Frequency 5. Reply code bracket-pulse spacing 6. Reply peak-pulse power 7. Antenna VSWR 8. Reply rate (percent)

3-7.8 SYSTEM OVERVIEW

Figure 3-125 shows a typical MK XII IFF System. As stated previously, such a system can operate independently without the aid of an associated radar set to supply the triggering action. The principal function of the electronic gate (TD-937A), as explained in Table 3-19, is to inhibit the Transponder from responding to interrogations radiated by its parent vessel.

3-7.8.1 Operational Checks

In the absence of specific test equipment, operational checks on IFF equipment can be performed sucessfully by using the following: 1) an RF Signal Generator of the correct range; 2) a specialized pulse generator for modulating the RF Signal Generator; and 3) a differential bridge, or an oscilloscope. Power measurement can be made either by the bridge or by applying an internal dummy load and attenuation to the oscilloscope. A suitably calibrated conversion chart is then employed in determining the output power of the IFF equipment under test. In this way, operational checks and a limited degree of system troubleshooting can be achieved.

3-7.8.2 Back-to-Back Testing

While it is unlikely that simultaneous failure of the Interrogator and the Transponder will occur, back-to-back testing can be performed provided that continuity of the Electronic Gate Generator (TD-937A) Cable is interrupted for the test. Such an arrangement permits the necessary coded pulses to be present through the Interrogator and the Decoder. With Transponder output connected directly to Interrogator output, however, suitable attenuation must be employed. When radiating through antennas for test purposes, the 77 to 76 codes are not to be used. These particular codes are for emergency use only. However, since the PMS requires checking transponder operation for Code 7777, the Transponder must be returned to its assigned code once the PMS is completed. Under no conditions is the PMS to be conducted with a Transponder connected to an antenna.

3-7.8.3 Antenna Testing

The IFF systems are normally supplied with an omni-directional test antenna that is identical

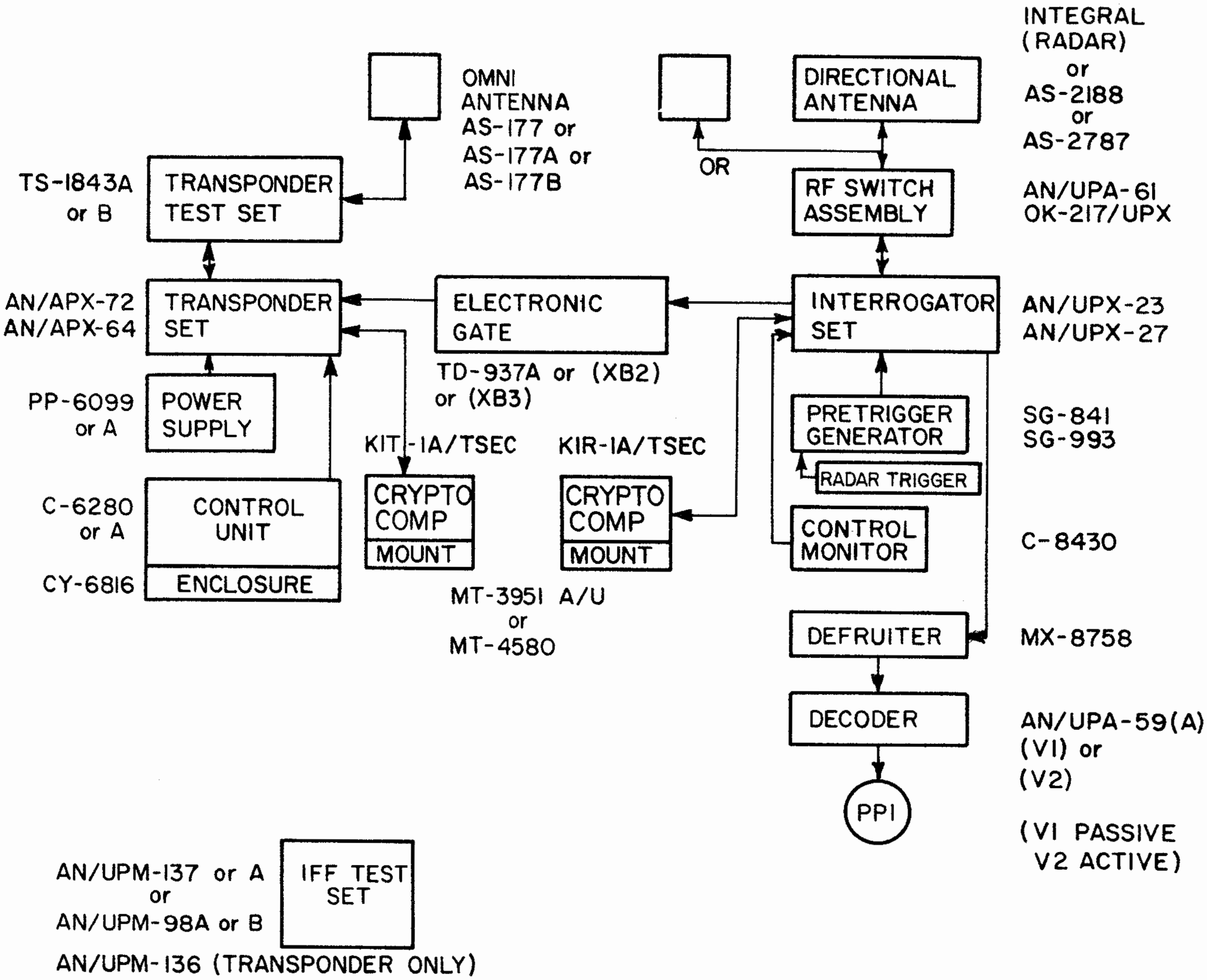


Figure 3-125. MK XII IFF System, Simplified Block Diagram

in characteristics to the Transponder's normally-used antenna. If a normally-used antenna becomes damaged or inoperative, temporary use of the test antenna will at least keep the system operative. The test antenna can also be used in determining if the normally-used antenna is defective. The test antenna, in this instance, is connected to an external test set and a signal whose frequency corresponds to that of the IFF receiver is then injected. The receiver's output is connected to an oscilloscope, and a notation is made of the indicated video output. The external test set (or signal generator) is then connected to the input side of the suspect antenna under test. The output power of the IFF is then adjusted until it attains the same level of video output (refer to Figure 3-126). This value is noted, and the difference between the two power output settings then represents the attenuation due to the suspect antenna, cables, connectors, etc. In other words, the derived attenuation is a combination of the IFF set's antenna system, the distance between the test antenna and the suspect antenna, and the test antenna system. An IFF antenna system can be checked for opens or shorts by using a Time Domain Reflectometer (TDR) as described in Section 5 of this manual. A multimeter can serve as a rough check on antenna electrical characteristics.

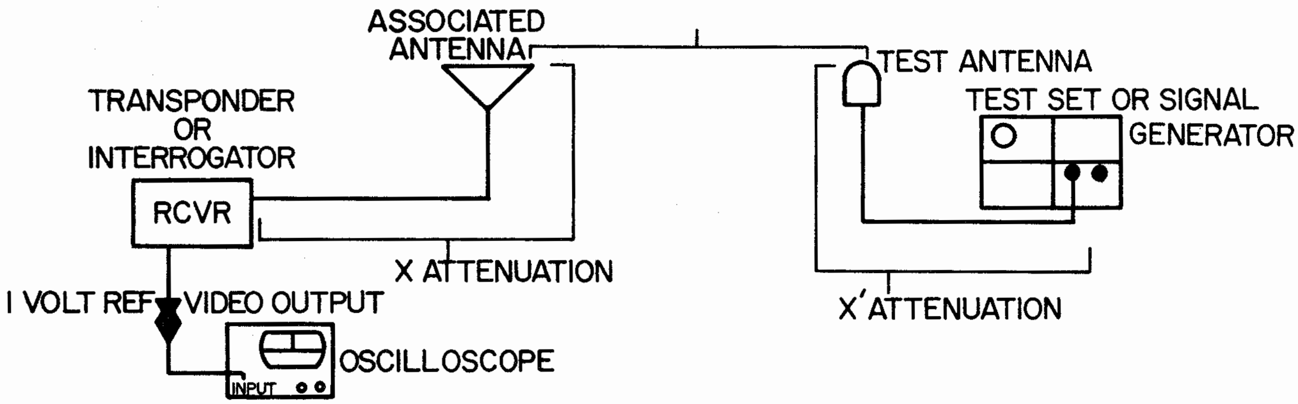


Figure 3-126. Antenna System Attenuated Measurement

3-8 NAVIGATIONAL AIDS

3-8.1 GENERAL

Electronic navigation implies navigation by means of electronic equipment and by radio wave emissions received from an outside source. The radio time signal was the first such aid to come into navigational use. It made precise time available to the navigator for use in connection with celestial navigation. Subsequently, he was able, on request, to obtain radio bearings from a limited number of shore stations. Usually only one such bearing was available to him at a time. This system was replaced by the shipborne radio direction finder (RDF), which permitted the navigator to obtain a bearing on any radio station within range of his receiver which was transmitting signals. In many coastal areas he was able to obtain several such bearings, which enabled him to determine his position with considerable accuracy. Extensive research has been carried out in the field of electronic navigation over the past several decades. The development of long-range airplanes established a need for suitable electronic navigational systems. Subsequently the need arose for systems suitable for the Fleet Ballistic Missile submarines, and more recently, systems have been required for the navigation and guidance of space vehicles. In each instance these needs were twofold: systems for maintaining dead reckoning position (inertial systems) and systems for position fixing (navigation aids). Under the first heading extremely sensitive and accurate, but expensive, gyros and accelerometers were designed. For position fixing, highly accurate instruments for determining the time of travel of radio signals have been produced, as well as for determining the measurement of altitude angles of celestial bodies by automatic electro-optical and radiometric tracking. In addition, equipment for measuring the Doppler shift of very precisely timed radio signals transmitted by man-made satellites has yielded excellent results. The Doppler shift (named for the Austrian scientist who reported the effect in 1842) is the apparent change in frequency of radiated energy when the distance between the source and the receiver is changing. In piloting, excellent position fixing is being achieved by both radar and sonar, as currently instrumented. Bathymetric navigation, or navigation by means of continuous soundings of the ocean bottom analyzed by a computer, holds great promise. However, it requires data in the form of very precise bathymetric charting of the operating area. The ideal navigation system has yet to be developed. Such a system should be worldwide, self-contained, passive, completely reliable, and highly accurate. Currently, the most promising systems, although they do not meet all the above requirements are Omega and Satellite Navigation. Omega uses radio signals from land-based transmitting systems, while the Satellite Navigation system depends on signals from a satellite travelling in a precisely determined orbit.

3-8.1.1 Atmospheric Effect

Electromagnetic energy, as transmitted from the antenna, radiates outward in all directions. A portion of this energy travels parallel to the earth's surface, while the remainder travels upwards as well as outwards, until it strikes the ionosphere and is reflected back to earth. This latter process may be repetitive, as illustrated in Figure 3-127. (Sky wave "A" in this figure is called a "one hop" wave; sky wave "B" is a "two hop" wave.) That portion of the energy which follows the surface of the earth is called the "ground wave"; the portions that are reflected back are termed "sky waves." In the employment of low frequencies, ground waves become very important, and the conductivity of the earth's crust becomes a major factor in signal attenuation (the decrease in amplitude of a wave or current with increasing distance from the source of transmission) by absorption, and by its effects on propagation velocity. Because of this conductivity factor, the electromagnetic field to some extent penetrates the earth's surface. The lower limit of the wave becomes slightly impeded by its penetration into this medium of increased conductivity, while the upper portion of the wave is not so affected. This results in the line of force leaning away from the signal source, causing the movement of the electromagnetic wave to follow the curvature of the earth's surface. It is pertinent that the lines of force of the electric field are perpendicular to the lines of force of the magnetic field, and the direction of motion of the electromagnetic wave is perpendicular to both. (Figure 3-128). It is this tendency to follow the earth's curvature that makes possible the transmission of ground waves over great distances. Combined with this curvature of the motion of the electromagnetic wave is the energy dissipation through absorption in the penetration of the earth's surface. This latter effect necessitates the use of high power to achieve long distance transmission of the ground wave. The variation in the characteristics of the surface of land areas complicates any prediction of its effects on ground wave transmission. The conductivity of the ocean surface, however, is quite constant, and propagation velocity over ocean areas can be predicted with considerable accuracy. Only the low frequency radio transmissions curve sufficiently to follow the earth's surface over great distances. Electromagnetic fields at higher frequencies do not penetrate as deeply into the surface and therefore encounter less impedance of velocity from the ground. They are slightly curved but not enough to provide ground wave signals at great distances from the transmitting antenna.

3-8.1.2 Interference

If two or more radio waves arrive simultaneously at the same point in space, interference results. The combination of such waves is in accordance with the principle of superposition of fields. Each field may be represented by a vector, indicating spatial direction and intensity, as shown in Figure 3-129. The resultant field direction and intensity of either the electric fields or of the magnetic fields may then be determined by following the rules for vector addition.

3-8.1.3 Ionization

The daylight portion of the earth's atmosphere is subjected to bombardment by intense ultra-violet rays of the sun. At extremely high altitudes in the atmosphere, the gas atoms are comparatively sparse. Electrons are excited by the powerful ultra-violet electromagnetic forces which reverse polarity approximately 10^{17} times per second. This violent oscillation causes the electrons to separate from the positive ions with which they were combined. These freed electrons would eventually find their way to other electron-deficient atoms, but this is prevented by the continuing forces of the ultra-violet rays while in direct sunlight. Thus, the freed electrons form ionized layers which reach their maximum intensity when the sun is at its highest. Ionization has a tendency to form

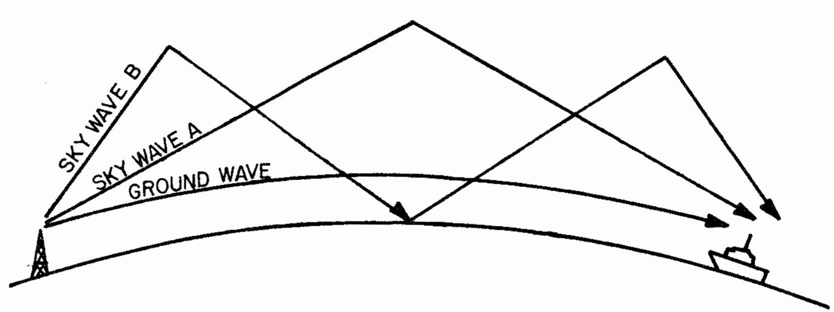


Figure 3-127. Electromagnetic Wave Propagation

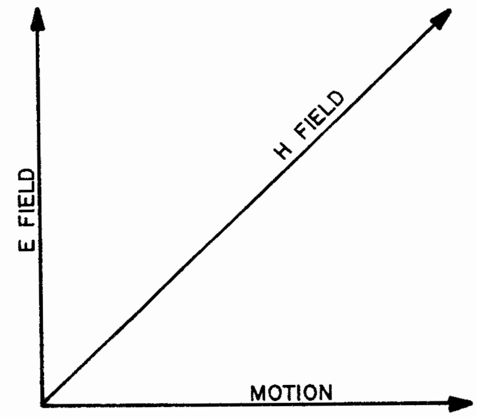


Figure 3-128. E & H Field Vectors

layers in the atmosphere and these layers change, disappear, combine, and separate as they are affected by the local time of day, the time of the year, and the phase of the eleven-year sun-spot cycles. The layers are also affected by apparent random changes occuring from moment to moment.

3-8.1.4 Ducting

Many irregularities occur, especially at the higher frequencies, in the propagation of electromagnetic waves. A phenomenon called radio refractive ducting occurs over much of the radio frequency spectrum but particularly on frequencies of the VHF and UHF bands. This phenomenon seems to occur more frequently over oceans than over land. It is generally associated with a temperature inversion at a very low altitude, perhaps 200 or 300 feet, and a sharp decrease in moisture content of the warm air. Very long ranges have been reported when low-power UHF transmitters were employed in experimenting with this phenomenon. Ducting can be responsible for limiting, as well as extending, the range of radio transmissions.

3-8.1.5 Hyperbolic Navigational Systems

Hyperbolic navigation systems are based on the theory that the known velocity of travel of electromagnetic waves through space is constant, within acceptable limits. The capability of measuring this difference in time of the arrival of signals from two separate sources makes possible the determination of position. A major advantage of the hyperbolic navigation systems is that position line data may be computed in advance of its use, and plotted or printed on charts, at convenient units of time difference value, eliminating the necessity for the navigator to make such computations. A disadvantage lies in the deterioration of accuracy inherent in spherical hyperbolic system geometry. The hyperbola is the locus of the points at which synchronized signals from two transmitters comprising a system will arrive at a constant time difference. This time difference is expressed in microseconds, or millionths of a second. The receivers employed in these systems therefore provide a readout in microseconds of time difference. In Figure 3-130, signals from stations A and B transmitted

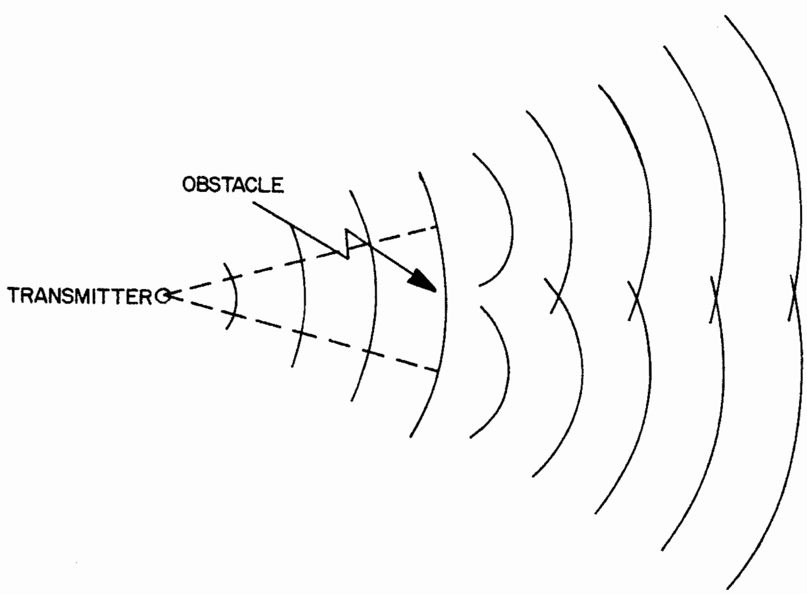


Figure 3-129. Wave Scattering from Interference

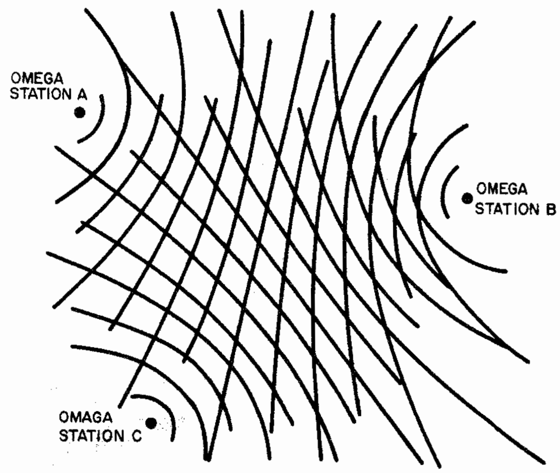


Figure 3-130. Omega Transmission Intercepts

simultaneously would have no time difference but would arrive simultaneously at any point along the center line, as all points along this line are equidistant from points A and B. All other lines, represented by hyperbolas, would represent points of equal time difference. In actual hyperbolic navigation systems, the transmission for the slave station (A) is delayed, rather than being simultaneous with the master (B). The figure represents hyperbolic lines on a plane surface. The appearance of the lines on a navigation chart representing a portion of the spherical surface of the earth will vary somewhat with the chart projection used. The computation of hyperbolas for a given pair of stations is tabulated for the navigator, giving the coordinates at which each hyperbola intersects a whole meridian or parallel, whichever is applicable. Charts on which the hyperbolas are printed are commonly employed. At least one receiver is now available which includes a digitial computer and coordinate converter, thus providing a readout in geographical coordinates. Two types of time-difference measurement are employed in hyperbolic systems. In one, the matching of electromagnetic wave envelopes of pulses transmitted from the two stations is measured in time difference, resulting in a rough measurement which locates the receiver on a hyperbolic line with known geographical coordinates. In the other system, matching the electromagnetic wave phase provides a fine measurement within an area or lane defined by two time-difference hyperbolas. This is in addition to the rough measurement obtained by time difference. In phase matching systems, extreme accuracy is possible under favorable conditions. A precision of about 0.05 microseconds can be obtained, which is equivalent to about 50 feet in position. This precision is, of course, degraded by system geometry as distance from the base-line increases. To establish position, a ship or aircraft employs two or more pairs of stations to acquire two or more intersecting position lines. As with other methods of determining position by means of intersecting lines of position, precision in positioning varies with the angle of intersection of the lines. Where lines cross at right angles, the area of most probable position is circular, with its center at the intersection of the lines. Where the lines intersect at an acute angle, the area of most probable position is elliptical. Thus, the minor axis of this ellipse will be equal to the diameter of the circle formed when the intersection is at 90°. The major axis will be greater because the ratio of its length to the diameter of the circle varies with the cosine of the angle of the intersection. Short-range hyperbolic systems designed for survey and oceanographic use include Decca Survey and Raydist, among other specialized equipment.

3-8.1.6 Rho-Theta Navigation

Rho-Theta navigation, or (more specifically) range-direction navigation utilizes a combination of circular, or ranging systems for distance measurements, with azimuthal, or directional measuring systems (Figure 3-131). The Omnirange (VOR) system, in general use for aviation throughout the U. S., provides bearing information. A large number of the stations are equipped with distance-measuring equipment (DME) to provide a complete rho-theta system. The military version is known as TACAN. Although these systems are sufficiently accurate for general navigation purposes, they are limited to a line-of-sight range. In the design of all electronic position systems, two major factors are carefully taken into consideration to obtain precision: repeatability, and predictability. The first factor, repeatability, is the ability of a system to repeat a position indication. In other words, if the position of a point on the surface of the earth is given in coordinates of the system at one time, the question is asked, "How closely may we return to that exact position at some later time?" The second factor, that of predictability, is one of knowing, given the atmospheric conditions, the propagation characteristics of the signal. Predictability is influenced primarily by refraction in the atmospheric medium, and by the conductivity of the surface.

3-8.2 LORAN "A"

As of December 1977, use of LORAN "A" by the U. S. Navy has been discontinued as a navigational aid.

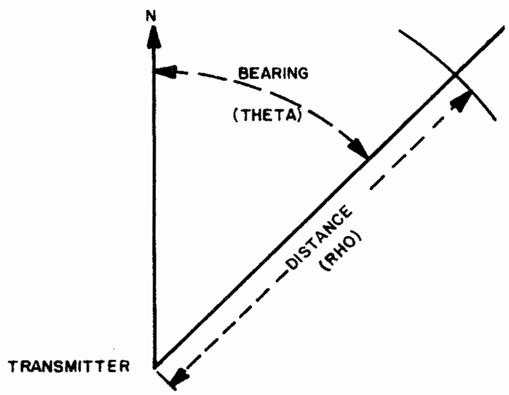


Figure 3-131. Rho-Theta Diagram

3-8.3 LORAN "C"

LORAN-C is a pulsed, hyperbolic long-range navigational aid system. It operates in the internationally-allocated frequency spectrum of 90 to 110 kHz, which is the long-range navigation frequency spectrum established by International Radio Regulatory Considerations of the Geneva (1959) Radio Regulations. LORAN-C is centered on a carrier frequency of 100 kHz. The need for an accurate long-range navigation system was recognized during World War II, and the first operational chain was established along the East Coast of the United States in 1957. Since then, LORAN-C coverage has been greatly expanded.

3-8.3.1 Characteristics

LORAN-C is a pulsed, hyperbolic system of radio navigation available to ships and aircraft by day or night, in all weather conditions, over land and sea. Each station radiates a multipulse transmission in the form of eight pulses spaced 1,000 microseconds apart. A Master Station transmits a ninth pulse for station identification. LORAN-C can supply position information with a high degree of accuracy at great distances. In LORAN-C, the time difference reading is obtained by comparing the arrival times of signals from two stations. In addition, the system employs a phase comparison technique. An approximate position is obtained by using the difference in arrival time of the pulsed signals. This is refined by a comparison of the phase of the signal within the pulse. The phase comparison is accomplished automatically within the receiver and does not involve a separate operation on the part of the operator. Due to its low-frequency (100 kHz) base-line distance (500-700 miles) LORAN-C is able to provide reasonably accurate information up to 1,200 miles by means of ground waves, and over 3,000 miles with sky waves.

3-8.3.2 Station Network

In a LORAN-C network, three or more stations transmit pulses which are radiated in all directions. One of these stations is designated as the "master" station, which sends out the master signal, and the remainder are "slave" stations. In the present LORAN systems the signals are not sent simultaneously and the slave station signal is delayed by a controlled amount. Therefore the master station pulse is always received first and the time-differences increase from a minimum at the slave station to a maximum at the master station. When the master pulse is received by the "slave" stations the transmitters at these stations are actuated and, after the appropriate delay, they in turn transmit similar groups of pulses, accurately synchronized with the received signals. The constant time-difference between the reception aboard ship of the master and slave pulses establishes the LORAN LOP (Lines Of Position). The time-difference remains constant along a hyperbolic line and a series of lines of constant time-difference are computed for each pair of stations and the data made available in the form of charts and tables. When the navigational position of a vessel is desired, the time-difference of a pair of stations is determined from the LORAN receiver. By consulting the charts and/or tables, interpolating where necessary, the LOPS can be plotted corresponding to the measured values. When a pulse is transmitted the amplitude starts at zero, rises to a maximum and recedes back to zero. This pulse shape can be varied. In LORAN-C there is a fast build-up amplitude to the peak, and the leading edge of the pulse is used for timing signals. The purpose of this sampling point being on the leading edge of the pulse is to differentiate between sky and ground waves. The ground wave path is always shorter than the reflected path from the ionosphere, and will be received first. The sky wave delay is so short, between 25 and 55 microseconds, that it is necessary to use the leading edge of the pulse to assure that the ground wave is received before being contaminated by the effects of the sky wave. The ability to use ground waves without contamination from sky wave permits use of visual techniques in time-difference measurements, and permits the use of long base lines with high accuracy synchronization between master and slave stations. Within each of the multi-pulse groups from the master and slave stations, the phase of the RF carrier is changed with respect to the pulse envelope in a systematic manner from pulse-to-pulse. The phase of each pulse in an eight- or nine-pulse nine-pulse group is changed in accordance with a prescribed code so that it is either in phase (+) or 180° out of phase (-) with a stable 100 kHz reference signal. The phase code used at a master station is different from the phase code used at a slave, but all slave stations use the same code. The use of phase-coded pulses by the system provides a measure of protection against interference from outside sources, and also reduces contamination of the ground wave of pulses transmitted subsequent to sky waves from preceding pulses, i.e., the sky wave of the first pulse arriving at the same time as the ground wave of the second pulse. Contamination by preceding sky waves without phase coding would nullify the effect of sampling only the ground wave, thereby degrading the inherent accuracy of the system. The use of phase coding also provides the receiver with necessary logical information for automatic search for the master and slave signals. Automatic search can be utilized for convenience or when the signal-to-noise ratio of the received signals precludes visual identification. Use of multipulses for LORAN-C makes possible the sharing of the same RF channel by all stations in the system. Identification of particular groups of stations must be provided by some means other than channel selection. Accordingly, provision has been made in LORAN-C equipment for 48 different pulse recurrence intervals or rates. The 48 rates are divided into six basic rates, each subdivided into eight specific rates. The Basic Pulse Recurrence Rates are shown in Table 3-21. The combination of numbers and letters in the designation of a LORAN-C line indicates the basic pulse recurrence rate, the specific pulse recurrence rate, station type designator of a particular station pair, and the time-difference in microseconds found on charts, tables, and indicators. Specific Pulse Recurrence Rates assigned for identification (following H, L, S, SH, SL, or SS): 0, 1, 2, 3, 4, 5, 6, 7. Station Type Designators (not station letter designators) and Transmission Sequence: M-Master, X-Slave, Y-Slave. For example, the complete legend SO-X-13300 denotes the following: 1) basic pulse recurrence rate, 20 pulses per second; 2) specific pulse recurrence rate, 0; 3) station type designator, slave X; and 4) time-difference reading, 13,300 ms. A LORAN-C network is composed of one master station and two or more slave stations. The transmitting stations may be arranged in triads (Figure 3-132), stars (Figure 3-133), or squares to provide optimum geometric accuracy for position fixing in the desired coverage area. Stations are so located that signals from two or more pairs of stations may be received in the coverage area. The accuracy of this system depends upon the transmitting stations keeping their signals properly timed or synchronized. Since each slave station transmits a series of eight pulses and the master station nine pulses, a built-in master station identification and warning system exists. In case of loss of synchronization, the ninth pulse from the master

Table 3-21. Basic Pulse Recurrence Rates

SPECIFIC PRR	BASIC PRR SS	SL	SH	S	L	H
0	100,000	80,000	60,000	50,000	40,000	30,000
1	99,900	79,000	59,900	49,900	39,900	29,900
2	99,800	79,800	59,800	49,800	39,800	29,800
3	99,700	79,700	59,700	49,700	39,700	29,700
4	99,600	79,600	59,600	49,600	39,600	29,600
5	99,500	79,500	59,500	49,500	39,500	29,500
6	99,400	79,400	59,400	49,400	39,400	29,400
7	99,300	79,300	59,300	49,300	39,300	29,300

Legend:

H ———— 33 1/3 pulses per second
L ———— 25 pulses per second
S ———— 20 pulses per second
SH ———— 16 2/3 pulses per second
SL ———— 12 1/2 pulses per second
SS ———— 10 pulses per second

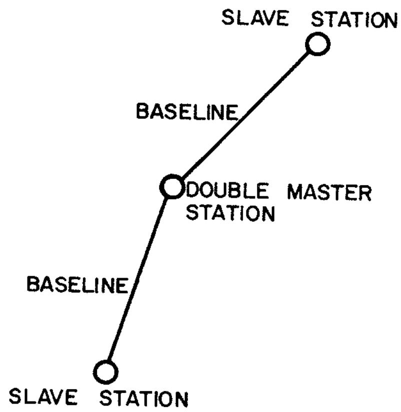


Figure 3-132. LORAN-C Network, Triad Arrangement

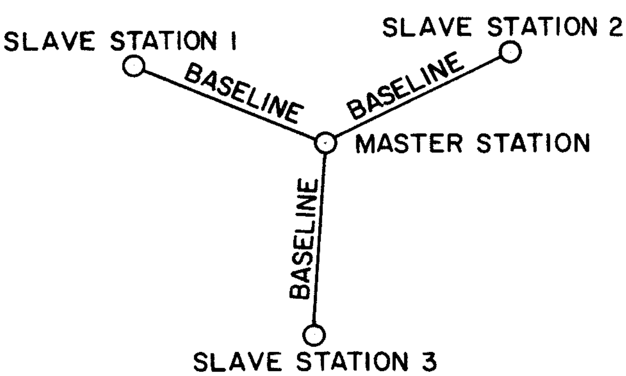


Figure 3-133. LORAN-C Station Network, Star Arrangement

station blinks back and forth thus warning the operator of a receiver that the signal is not valid. The inherent accuracy capability of this system makes it extremely useful for additional purposes, besides precise electronic navigation. Some of these uses are as follows. It can serve as a long-range time distribution system with an accuracy in the order of one microsecond. It permits microsecond-order relative time standardization between widely separated receiving locations. It is useful for electromagentic wave propagation studies.

3-8.3.3 Reception Ranges

Ground wave coverage may be considered to be a function of signal propagation strength, and the strength of signal to noise ratio. During periods of favorable atmospheric conditions the ground wave range is approximately 2,000 miles. However, during periods of high static noise and interference, this range may decrease to less than 1,000 miles. As a general rule, at peak pulse powers of 300 KW, the ground wave range for reliable signals may be taken as 1,200 miles. It is this favorable power output plus the low frequency of the system that allows reception by submerged vessels. First-hop sky waves extend out to about 2,300 miles, and second-hop sky wave signals have been picked up and used at a range of 3,400 miles from the transmitting stations. The accuracy of LORAN-C is dependent upon atmospheric conditions, noise and interference. Consequently ground wave accuracy is on the order of 0.1 percent of the distance traveled, and sky wave accuracy is between 3 and 5 nautical miles. It is the phase comparison and the longer baseline between stations which give LORAN-C its greatly increased coverage and accuracy compared with the earlier hyperbolic navigational systems.

3-8.3.4 Receivers

LORAN-C receivers are specifically designed for this particular system, and must be capable of accepting a carrier frequency of 100 kHz, with a band width of approximately 25 kHz. They are fitted with manual controls for selecting the proper recurrence rate; the remainder of the operations, including synchronization, may be accomplished manually or by automatic means. Phase measurement is automatically performed and once the receiver has been synchronized on a master and slave signal, the arrival time-difference is directly and continuously displayed as the receiver moves in the service area of the stations. The time-difference is then translated into geographic coordinates by plotting on charts, by the use of tables or by a computer.

3-8.3.4.1 Receiver Operation

When using a receiver specifically designed for LORAN-C, it is not necessary to advance or retard LORAN-C lines of position because all readings are taken simultaneously. A single observation provides readings which establish lines of position for all of the pairs within the particular network being used. Normally, these lines will not coincide with the lines printed on the charts, therefore interpolation will be required. When maximum accuracy is required, the appropriate LORAN-C tables should be used. From these tables the latitude and longitude of two points are obtained. These points are computed close enough so that a straight line drawn between them gives an acceptable line of position. Sky wave corrections are printed at numerous grid intersections on the chart. Corrections are given for both daytime and nighttime operations and also for matching ground waves with sky waves during both of these periods. The corrections are applied directly to the reading of the LORAN-C receiver after necessary interpolation.

3-8.3.5 Time Measurement

The extremely high order of stability in the LORAN-C system has generated considerable interest in its use of time-measurement. The National Bureau of Standards and the U. S. Naval Observatory have indicated that emissions from the LORAN-system provide the capability for synchronizing and setting clocks to an accuracy of better than one microsecond in the areas covered by ground waves. Such timing information is approximately 1,000 times better than service available by most other means.

3-8.3.6 Position Accuracy

To obtain high position accuracy over long transmission paths, receivers must be designed specifically for LORAN-C system use. With the proper equipment, the user is capable of obtaining the maximum amount of information available for the transmitted LORAN-C signals. The procedure for obtaining this information follows. The receiver is designed to accept a carrier frequency of 100 kHz with a bandwidth commensurate with requirements for pulse type reception (approximately 25 kHz). The receiver operator selects the pulse recurrence rate of the LORAN-C chain to be observed. This selection aligns timing within the receiver to the timing of the signals to be observed. When the received signals are of sufficient amplitude to be observed on an oscilloscope swept at the rate of transmission, the received signals appear stationary. Signals on rates other than those selected at the receiver, drift through the stationary signals. After selecting the proper recurrence rate, the operator then synchronizes the receiver with the master signals. Synchronization can be accomplished in either of two ways on most LORAN-C receivers. In the first method, the operator utilizes the ninth pulse of the master station for visual identification of the time sequence of transmission. In the second method, the automatic search feature of the receiver is used. Automatic search is made possible through the use of the phase-coded pulses and logic circuits in the receiver. In either of the two methods, the sampling gates are aligned in time with the receiver master pulses allowing the receiver to commence automatic tracking of these signals. Synchronization of the receiver on slave signals is then accomplished by visual or automatic means. When the receiver has synchronized on a master and a slave signal, the arrival time-difference is directly and continuously displayed as the receiver moves in the service area. These time-difference readings are then translated into geographic coordinates by the use of computers, charts, or tables. Automatic alarms have been incorporated in most receivers to inform the operator when the receiver is tracking on a combined ground wave-sky wave signal. Alarms also inform the operator when the receiver has lost a particular signal either through improper sampling or when a transmitting station is off-air. LORAN-C charts are prepared and sold by the U. S. Naval Oceanographic Office, Washington, D. C. The charts are standard projections and show the chain rate identification, isogonic lines, and the first-hop sky wave corrections for day and night. The ionosphere heights for day and night corrections are assumed to be 73 and 91 kilometers, respectively. The hyperbolic lines of position, the main feature of the charts, are spaced approximately 10 to 50 microseconds apart depending on the local geometric accuracy potential of the area.

3-8.4 OMEGA SYSTEM

Omega transmitting stations operate in the internationally-allocated Very Low Frequency (VLF) navigational band between 10 and 14 kHz. This very low transmitting frequency enables Omega to provide adequate navigation signals at much longer ranges than other ground-based navigation systems. The Omega system has the potential of providing a position-fixing accuracy of 2 to 4 nautical miles (95% of the time) for most ships and aircraft. The Omega Navigation System involves the use of land-based transmitting stations and special Omega receivers. Charts and propagation tables are required to determine ship position. When a ship is within range of three or more Omega stations, the Omega receiver, once set, will continually display a series of lane values or numbers on its panel indicator. These numbers correspond only roughly to the Line of Position (LOP) on the Omega plotting chart. This display, along with the proper correction data obtained from the appropriate Propagation Correction tables, is plotted on an Omega plotting chart to establish the ship's actual position. In most airborne Omega systems the entire procedure is completely automatic as well as pilot operable. Omega is now available over a significant portion of the earth's surface with seven transmitting stations at North Dakota, Norway, Hawaii, Japan, La Reunion, Liberia, and Argentina (See Table 3-22). The installation of an eighth station in Australia, to be completed in 1980, will complete the network of transmitting stations. In the interim a temporary station in Trinidad has been operating in the G segment of the Omega format.

3-8.4.1 Theory

Omega, like other radio navigation systems, depends on the table transmission of radio signals. As can be seen from Figure 3-134, signals radiate out from each transmitting station in circular "waves". By knowing the time it takes for a signal to travel to a

Table 3-22. Omega Transmitting Stations

STATION LETTER DESIGNATION	LOCATION	LATITUDE	LONGITUDE
A	Aldra, Norway	66°25'N	13°08'E
B	Monrovia, Liberia	6°18'N	10°40'W
C	Haiku, Hawaii	21°24'N	157°50'W
D	La Moure, North Dakota	46°21'N	98°20'W
E	La Reunion	20°58'S	55°17'E
F	Golfo Nuevo, Argentina	43°03'S	65°11'W
G	Australia	38°29'S*	146°56'E*
G_1	Trinidad (temporary)	10°42'N	61°38'W
H	Tsushima, Japan	34°37'N	127°27'E

*Approximate

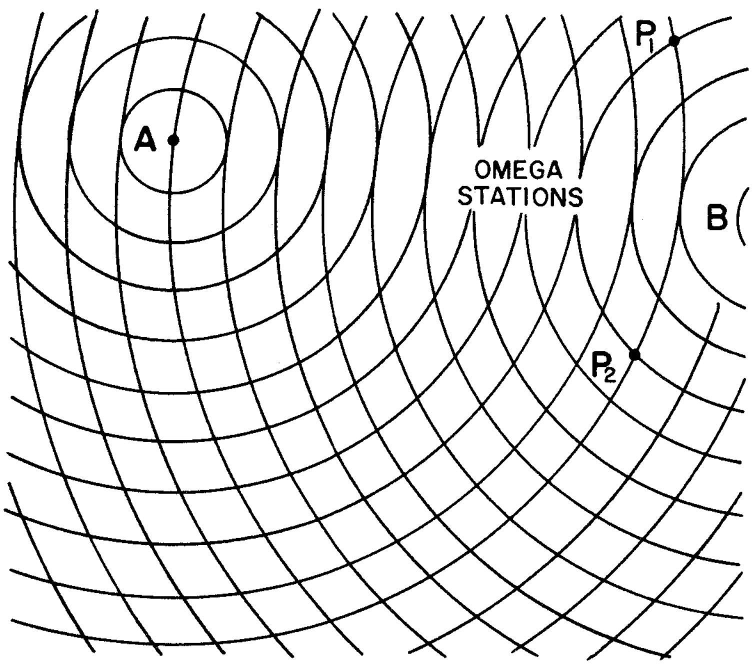


Figure 3-134. Omega Wave Radiation

certain point and also knowing the speed at which a signal is traveling, it is possible to calculate the distance to that point. The travel times of signals can be indirectly measured by an Omega receiver, and radio propagation research has provided us with the velocity or speed required for the distance calculation. The distance from any Omega station defines a circular Line of Position (LOP) on which the receiver is located. As can be seen from Figure 3-134 by using two stations, two circular LOPs place the receiver at two possible locations, P_1 or P_2. Since in Omega the baseline between stations A and B is extremely long (greater than 10 megameters - 5,000 n. mi.), this ambiguous situation usually reduces to first knowing the ship's general position. More importantly, other stations received can provide additional circular lines of position to further reduce this uncertainty. The use of a circular LOP method, however, requires that the receiver be equipped with a very accurate and properly set "clock" or precision oscillator. Therefore, its use is limited to the most sophisticated shipboard and airborne receiving systems. The most common use of radio navigation systems is in what is called the "hyperbolic mode". In this situation, the difference in signal travel times or difference in distance serves as the basic principle. From Figure 3-135, it can be seen that on the baseline half-way between the stations is a line (XY). In this case it is called "perpendicular bisector," and contains all the possible locations that are equidistant from the transmitting stations. On either side of the perpendicular bisector, however, there are hyperbolic lines of position that locate all points where the difference in distance is constant. For example, at P_3 and P_4 the difference in distance to each transmitting station is the same and they both fall on the same hyperbolic line of position. It takes two transmitting stations to form one hyperbolic LOP, thus at least three Omega stations are required to produce two LOPS. Two LOPs are always required to provide the intersection which is the navigational fix. (See Figure 3-136). To explain how hyperbolic and circular radio navigation systems operate, the relationships of time and distance were used in the pre- ceding dis cussion. In practice, however, Omega receivers do not directly measure time or time differences but measure phase or phase differences. This is a characteristic of systems that transmit a continuous wave (CW) signal rather than a series of pulses. Referring to Figure 3-137, assume that two transmitting stations, A and B, are located at some point near the baseline. This is also the baseline between stations A and B shown in Figure 3-135. Now assume that signals are simultaneously transmitted from stations A and B. At certain points, the sinusoidal wave transmitted from each station will pass through 0°, 180°, 270°, and 360° over the distance of one wavelength.

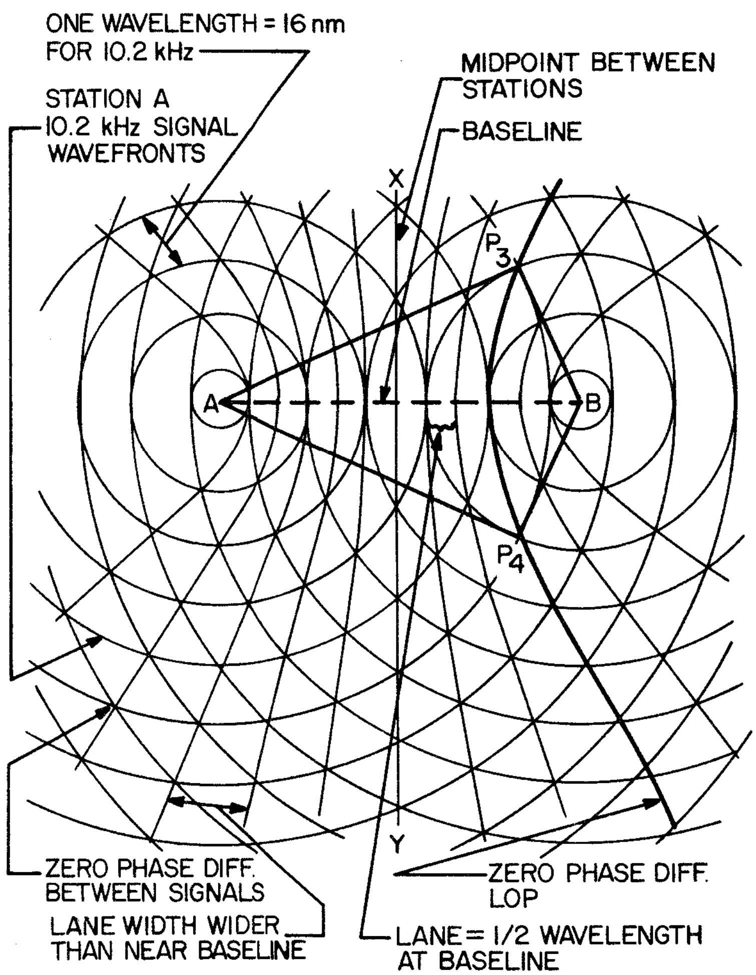


Figure 3-135. Omega Wave Interlacing

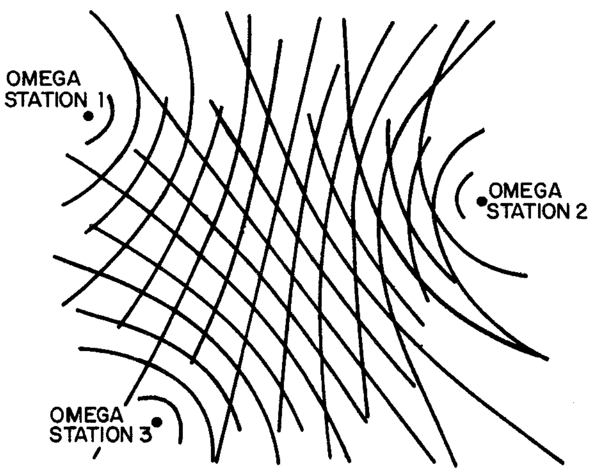


Figure 3-136. Lines of Position

The standard Omega receiver, however, measures the difference in phase between the signals transmitted simultaneously from stations A and B. From Figure 3-137, it can be seen that when the phase of signal B (-135°) at a point on the line Q-Q' is subtracted from the phase of signal A (+135°) the algebraic sum is 270°. Thus, traversing only the distance of 1/2 wavelength the receiver will detect a phase difference change of 360°. The important thing to understand here is that the receiver repeats the process again as it travels along the baseline, since it cannot measure changes in phase greater than 360° and resets at 0° This characteristic of the system will again be discussed in the section dealing with lane identification. Although receivers could be suitably calibrated and designed with counters that display phase differences in degrees, it is more practical to work with numbers to the base 10. The distance between two zero phase difference contours, which is 1/2 the wavelength on the baseline in a hyperbolic system (see Figure 3-137), is referred to as a "lane." For convenience, this distance is divided into one hundred parts, called "centilanes." The term "centicycle" is used interchangeably with centilane. Since typical receivers have a resolution of about one centilane, 100 hyperbolic LOPs are possible in a lane that a receiver could detect. The discussion thus far has been limited as to what takes place near the baseline. As the ship moves from the baseline, the hyperbolas begin to diverge, as shown in Figure 3-138. The actual distance on the earth's surface which is represented by a lane becomes greater. As a result, any initial error in an Omega measurement

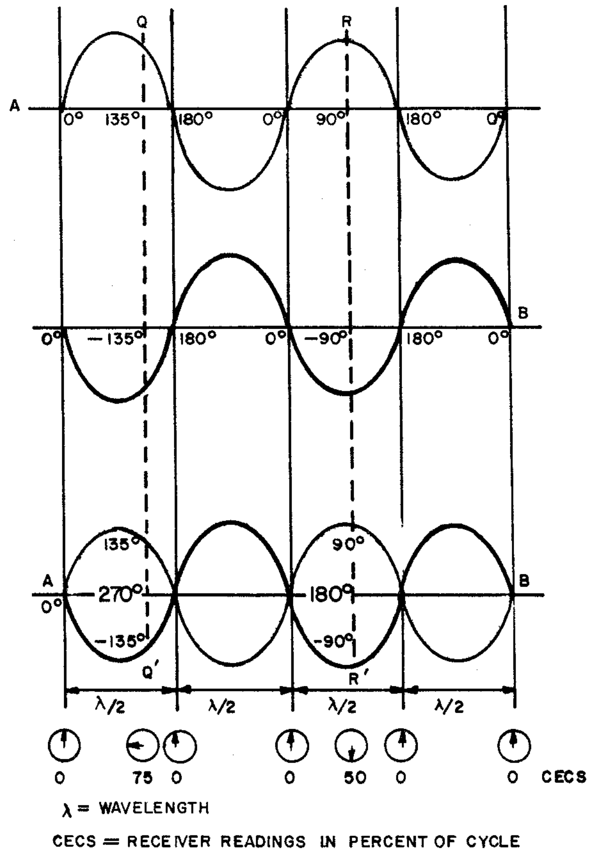


Figure 3-137. Phase Relationships

is magnified. This is true of all hyperbolic radio navigation systems, but constitutes a lesser problem in Omega since the stations are very far apart. The Omega user must therefore be at considerable distance from the stations before the effects of poor geometry become significant. Figure 3-139 illustrates this point. Hyperbolic lines-of-position drawn on Omega charts are based on a theoretical set of propagation conditions. The navigator cannot plot his receiver readings on an Omega chart without first correcting them for his local propagation conditions. At this point it is appropriate to examine the propagational aspects of Omega signals. The data collected at monitoring sites has shown that the transmitted Omega signal is very stable and predictable over long distances during most portions of the day. Omega signals are propagated through what is known as the "waveguide" which exists between the earth's surface and the lower D region of the ionosphere. Within the waveguide, Omega signals undergo changes in propagation velocity due to contraction of the waveguide during daytime when the

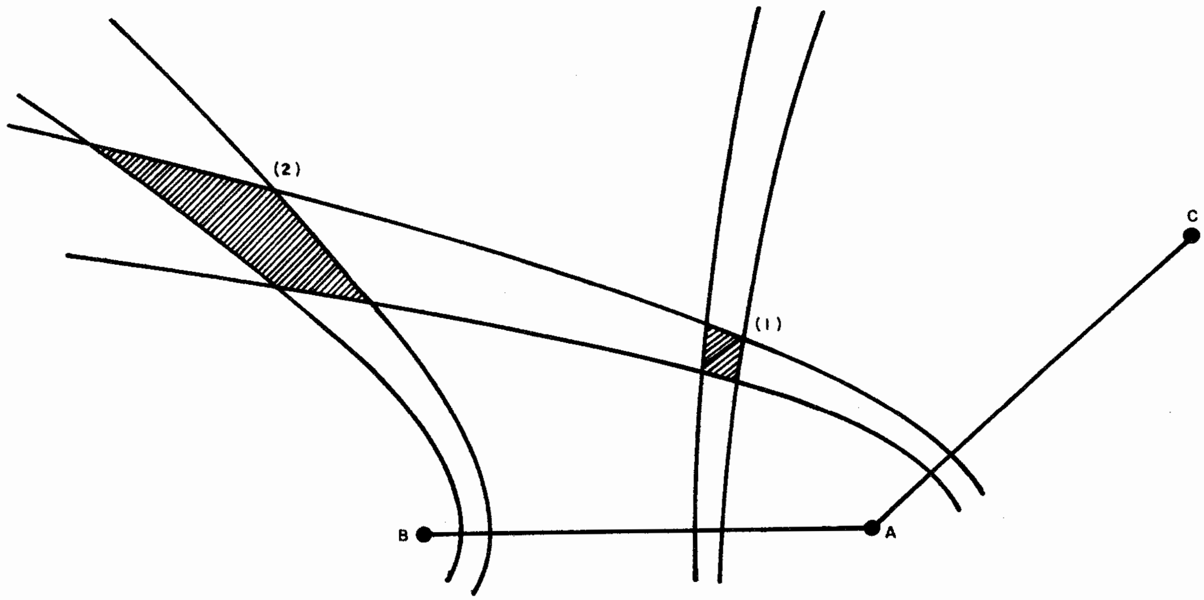


Figure 3-138. Uncertainty of Ship's Position

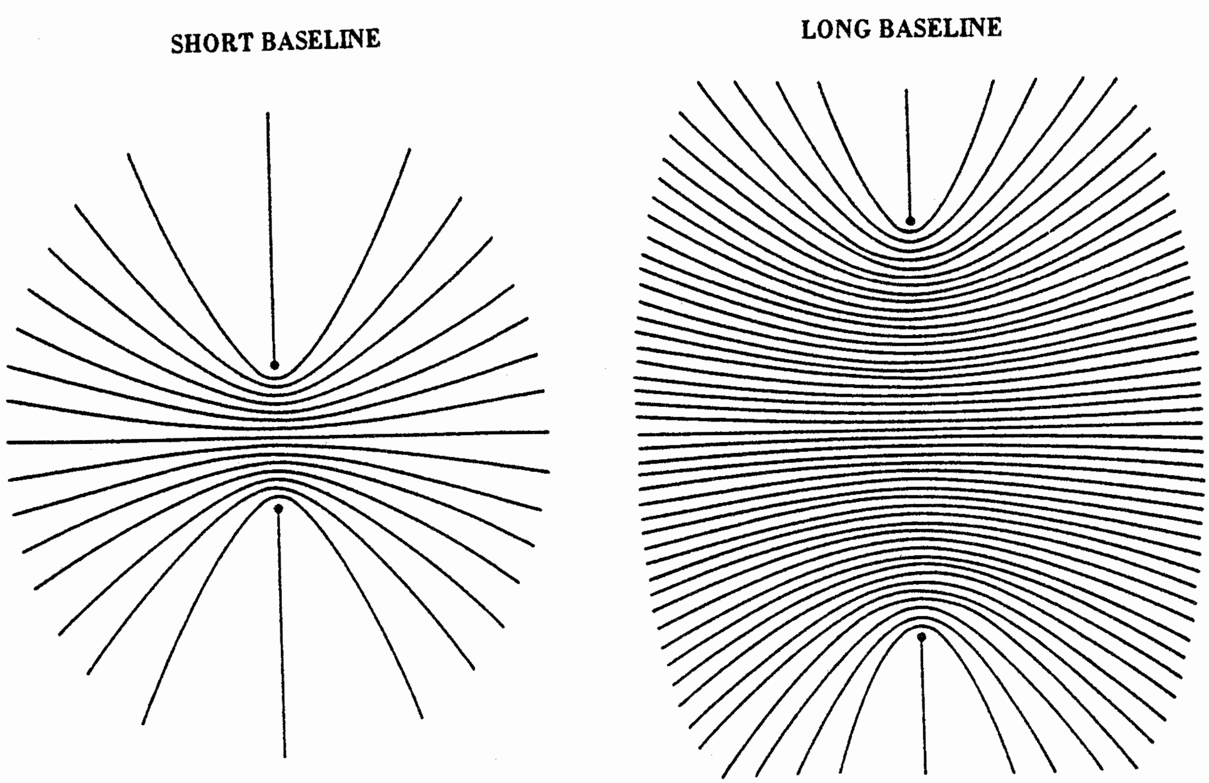


Figure 3-139. Divergence of Lines-Of-Position (LOPS)

ionosphere lowers to within about 70 km of the earth's surface, and expansion of the waveguide at nighttime when it moves to 90 km above, see Figure 3-140. It is possible to predict to within about ±five centilanes (centicycles) what the actual signal arrival time will be over daytime or illuminated paths. At nighttime this prediction accuracy decreases by a factor of two. On transitional paths, or ones that are partially illuminated by sunlight, predictions are extremely hard to make since the phase changes so rapidly. During a transitional period, it is usually recommended that the navigator not use signals so affected. Other factors affecting the phase of Omega signals besides the height of the ionosphere are the orientation of the earth's magnetic field and the earth's surface conductivity. All of these factors are included in the Propagation Correction tables (PPCs) used by the navigator to correct his receiver readings before plotting on the navigational chart.

3-8.4.2 Operating Characteristics

The operating characteristics of the system can be categorized as follows, signal format and control; the requirement for lane identification typical of phase comparison systems; and the process of handling errors attributable to signal propagation.

3-8.4.2.1 Signal Format

All stations now transmit three basic navigational frequencies (10.2 kHz, 11 1/3 kHz, 13.6 kHz) more or less omni-directionally. To prevent interstation signal interference, transmissions from each station are time-sequenced as illustrated in Figure 3-141. This pattern is arranged so that during each transmission interval (approximately 1 second), only three stations are radiating, each at a different frequency. The duration of each transmission varies from 0.9 to 1.2 seconds, depending on the station's assigned location within the signal pattern. With eight stations in the implemented system and a silent interval of 0.2-second between each transmission, the entire cycle of the signal pattern repeats every 10 seconds. Besides the three basic navigational frequencies, other frequencies can be added to the Omega signal format. Original plans were made to transmit two unique frequencies at each station for the purpose of inter-station time-synchronization, but this requirement has been removed through use of highly stable cesium

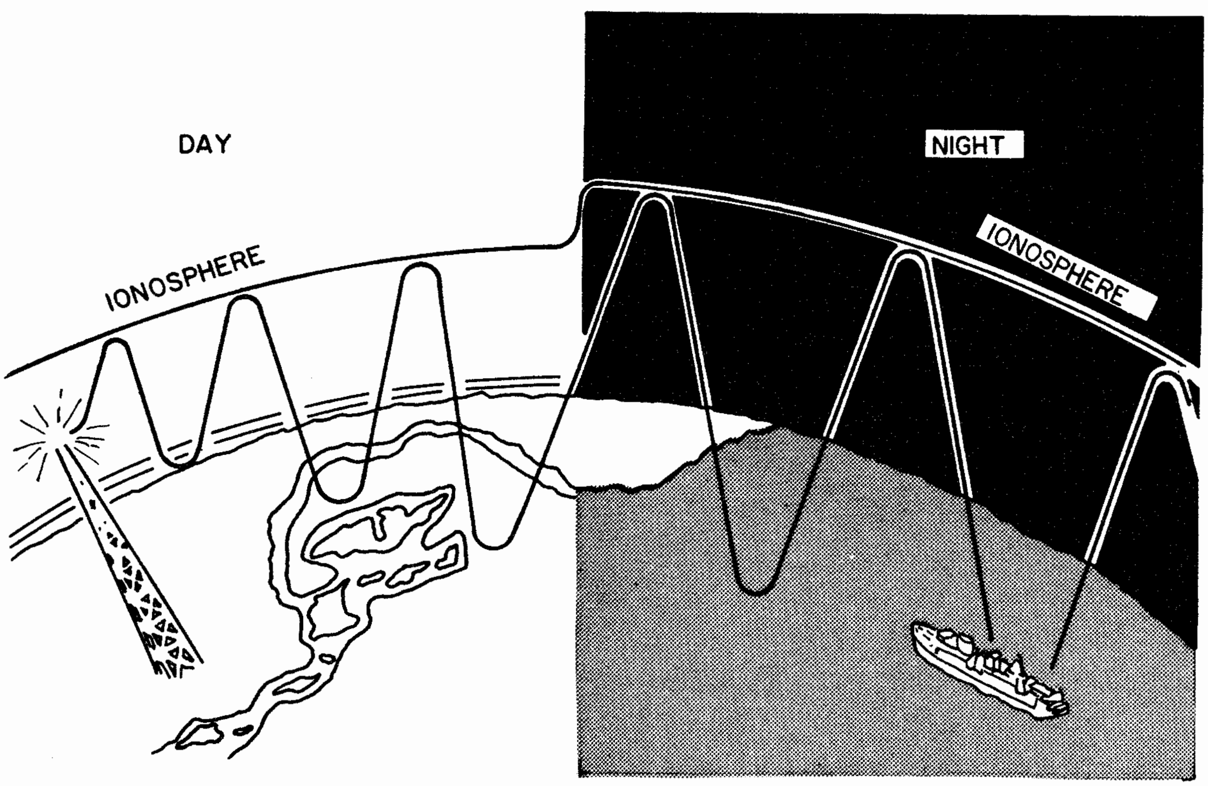


Figure 3-140. Transitional Paths of Omega Transmission

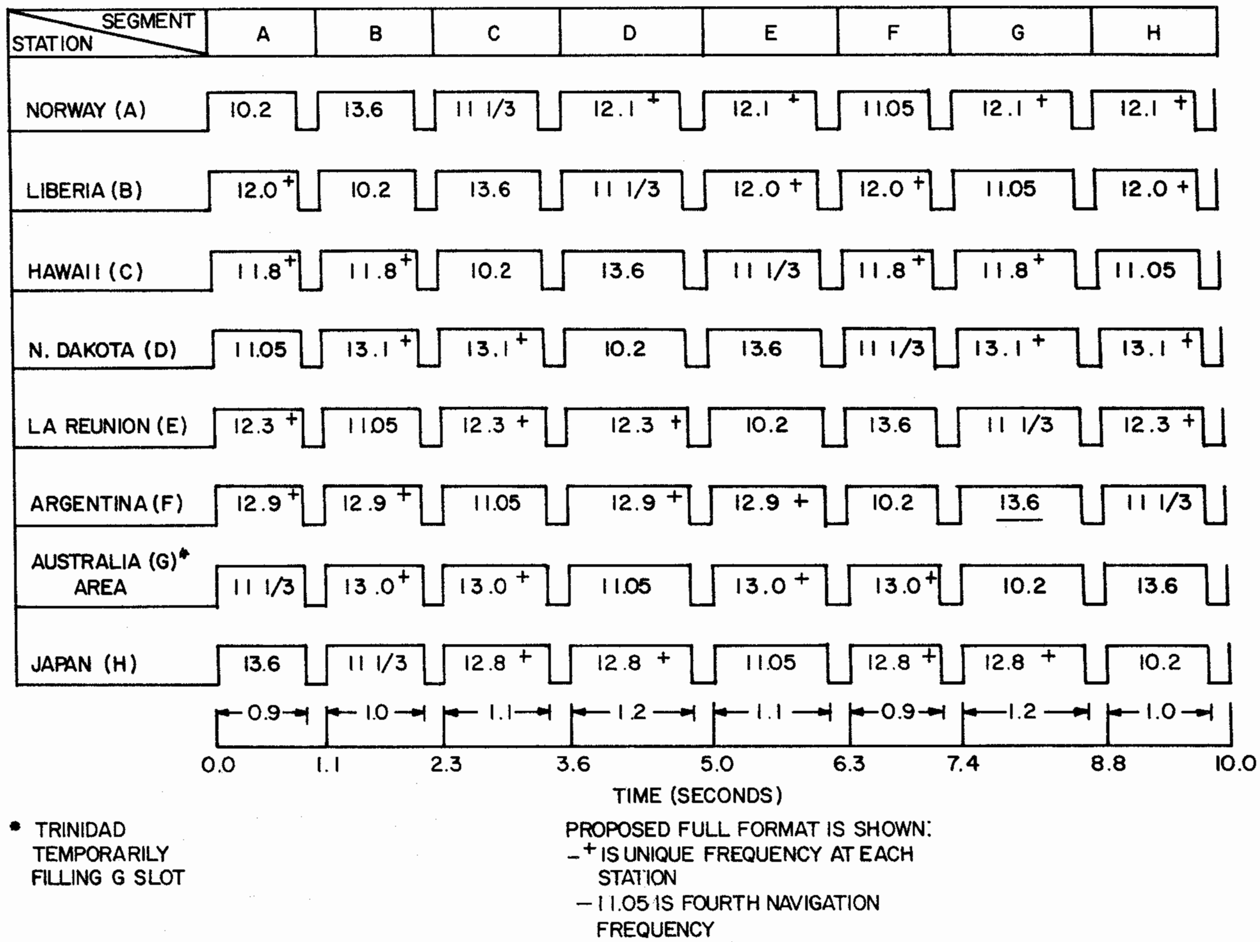


Figure 3-141. Omega Signal Transmission Format

frequency standards. Present plans call for the incorporation of a fourth navigation frequency (11.050 kHz) transmission which will allow for a lane resolution capability as great as 288 nautical miles. In addition, a unique frequency transmission for each station can be added which will aid in time-dissemination by providing a beat frequency and a high duty-cycle at that frequency.

3-8.4.2.2 Synchronization Control

The Omega signal format is designed such that each station within the network can be identified by its transmission on a particular frequency at a prescribed time. In addition, the synchronization of all transmissions is tightly controlled and the phase relationships between all signals are maintained to within a few centihertz. With this high degree of phase stability in the transmissions, the accuracy of the navigational fix is then primarily limited to the receiver and to the accuracy of the navigator's propagation correction tables. All Omega transmitting stations are synchronized by means of very stable cesium beam frequency standards. These standards or clocks are referenced to the atomic time-scale, which differs from Coordinated Universal Time (UTC) more commonly in use. Thus, the Omega epoch or time-reference is seconds ahead of UTC since the yearly adjustments for earth motion are not made to bring Omega Epoch into agreement with UTC.

3-8.4.2.3 Lane Identification

As was pointed out in the theory section, receiver readings are observed and corrected to determine the ship's location within a particular lane.

Each phase difference reading of zero indicates that a lane has been crossed. Although a receiver indicates the boundary of a lane, it cannot assign a specific lane number without external reference with regard to a Dead Reckoning (DR) position or other source of geographic coordinates. Most mariners will use the Omega plotting chart, which contains a geographic grid and numbered hyperbolic lanes, to determine proper lane values. Unless some estimate of position is available, a navigator faces a problem of lane ambiguity. Most receivers keep track of lanes through an automatic counter which advances whenever a zero phase difference reading is detected. In addition to a lane counter, chart recorders which graphically display lane crossings are particularly helpful in identifying a possible loss in lane count. Loss of lane count can occur due to such circumstances as power interruption, receiver malfunction, propagation disturbances and unusually high local noise conditions. To assist the navigator in establishing or reestablishing lane count, frequencies in addition to 10.2 kHz have been added to the Omega signal format (see Figure 3-142). In this technique, two received signals such as 10.2 kHz and 13.6 kHz are differenced within the receiver through the principle of heterodyning, or outside the receiver through mathematical differencing. This difference between the original frequencies (13.6 kHz and 10.2 kHz) is a resultant frequency of 3.4 kHz. This produces a wider lane, just as if 3.4 kHz were transmitted and measured directly. As can be seen from Figure 3-142 in a 3.4 kHz lane the distance between zero phase differences is 24 nautical miles. When 11.3 kHz and 10.2 kHz are differenced, a 1.1 kHz signal results whereby the lane width increases to 72 nautical miles, and so forth. For most marine navigation, this lane identification aid is not required since a DR position can usually be obtained that meets the 8-nautical-mile requirement for 10.2 kHz lane identification. For rapidly moving vehicles, such as aircraft, lane resolution capabilities of 72 nautical miles are generally required. Identification and retention of the proper 10.2 kHz lane is probably the most important procedure to observe in the use of Omega. This includes proper initialization of the receiver; selection of the most favorable LOPs during the voyage based on signal strength, geometry, charts, navigational notices and experience; proper resetting

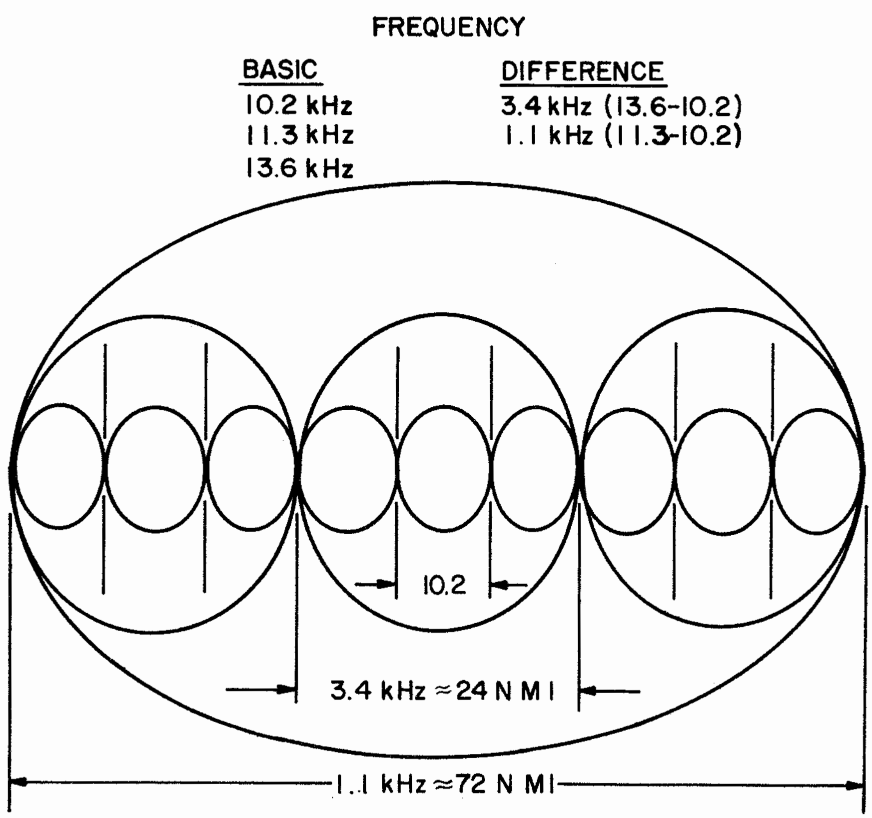


Figure 3-142. Omega Phase Differences

of the receiver in the event of a receiver power failure or temporary loss of transmitted signal; and in recognition of inaccurate Omega readings due to lane slips, propagation anomalies, local precipitation static and related factors which can affect position-fixing capability. Most marine navigators accomplish this by maintaining an accurate DR plot and depend on the higher lane resolution frequencies as a backup. In aircraft systems, the function of "laning" is accomplished by a computer, due to the rapid computations needed by a high-speed vehicle.

3-8.4.2.4 Propagation Characteristics

The propagation characteristics that permit the use of Omega at a great range also introduce certain limitations. Two areas that require special attention are normal time variations and model interference. Since Omega signals are propagated within the waveguide formed by the earth and ionosphere, changes in propagation parameters such as velocity may be expected as a result of changes in the ionosphere or ground. The most obvious navigationally undesirable variation is a daily or diurnal phase change. Normal changes in illumination of the ionosphere by the sun throughout the day may cause an uncorrected phase measurement to vary as much as one complete cycle. Since these variations are highly repeatable, prediction and correction are possible. Unpredictable short-term variations may also occur. Ninety-five percent of the time these are minor variations related to random propagational variations which will not degrade normal navigational accuracy. Occasionally, however, large disturbances can occur as a result of solar emission of X-ray or particle bursts. The emission of X-rays from the sun occasionally causes a short-term disruption of Omega signals, referred to as a Sudden Phase Anomaly (SPA). The duration of an SPA is generally not greater than one hour, but an LOP may experience a shift of several miles. The SPAs occur on an average frequency of about 7-10 per month. They usually affect signals from only a few stations at a time, since X-rays from the sun tend to enter a limited illuminated portion of earth's surface. The release of a large quantity of protons from the sun, although an infrequent occurence, produces what is known as a Polar Cap Disturbance (PCD). The effect of a PCD may be to shift an LOP 6-8 miles for a period of several days. This disturbance generally lasts for several days and varies in magnitude during the period. The PCDs affect only those transmissions involving arctic propagation paths. Because of its possible long duration and large LOP shift, PCD notices are broadcast as navigational warning messages. Modal interference is a special form of signal interference wherein the various waveguide modes of signal propagation interfere with each other, causing irregularities to appear in the phase pattern. Ideally, one mode would be completely dominant at all times and the resultant phase grid would be regular. In practice, competing modes do not completely disappear. Three situations are recognizable. If the competing mode is very small, then the dominant mode will establish a nearly regular phase pattern, as is intended. This is usually what happens during the day. A second possibility is that the competing mode may be almost equal to the dominant mode. The potentially serious case is that in which modal dominance can change. This may occur, for example, if one mode is dominant during the day and a second mode is dominant at night. Clearly, somewhere during sunset and sunrise (the transitional period) the two modes must be equal. Depending upon phasing of the modes at equality, abnormal transitions may occur in which cycles are "slipped" or lost. Positional errors of a full wave-length are possible under such conditions, and use of a station so affected should be avoided. If this is not possible, particular attention must be given to proper lane identification. Propagation Corrections (PPCs) must be applied to each Omega receiver reading in order to compensate for ionospherically induced signal variations and thereby improve position fixing accuracy. Omega propagation correction tables for each transmitting station (A through H) contain necessary data for correcting Omega receiver readouts affected by prevailing propagation conditions relative to the nominal conditions on which all charts and tables are based. A brief introduction, which also describes the arrangement and application of the corrections together with illustrative examples, precedes the tabular data within each PPC table.

3-8.4.3 System Usability

A comprehensive program is underway to collect and analyze technical data on the transmitted signal and to perform special tests of the system on a required basis around the world in order to validate actual coverage and accuracy and to define limitations when and where they may exist. Until this effort is completed, the Omega system is considered usable for navigation with the understanding that users may experience various levels of coverage and accuracy, depending upon such factors as type and sensitivity of receivers utilized, time of day, geographical location, and level of operator training and knowledge.

3-8.4.4 Omega Notices and Navigational Warnings

As with other navigational aids, information is disseminated concerning station off-the-air periods, VLF propagation disturbances, and other pertinent data affecting usage of the system. Status information on Omega stations can be obtained on the telephone by calling (202) 245-0298. Navigational warnings are broadcast to marine users via the HYDROLANT/HYDROPAC message system (Consult Defense Mapping Agency Hydrographic Center (DMAHC) Pub. 117A and 117B), and to aviation users via the Federal Aviation Administration Notice to Airmen (NOTAM) system. Published notices affecting charts and tables appear in Notice to Mariners. Hardcopy of the Omega Navigation System Operation Detail (ONSOD) telephone recording can be obtained by writing to ONSOD and authorizing a collect TELEX. Omega status messages are also broadcast by various time-service stations. The National Bureau of Standards Station, WWV, Boulder, Colorado, transmits a 40-second message at 16 minutes past the hour, and Station WWVH, Kauai, Hawaii broadcasts the same message at 47 minutes past the hour. Norway Radio Station Rogaland, broadcasts notices in international morse code on HF four times daily. Other stations in the global time service and maritime information network are expected to be added in the future. In addition to the regularly issued notices on station off-air periods, it is important to bear in mind that major planned maintenance may take place in the months listed for each station:

March	- Argentina
April	- Liberia
May	- Hawaii
June	- La Reunion
July	- Norway
February	- Australia
September	- North Dakota
October	- Japan

The actual off-air times information is disseminated as noted above sufficiently in advance, and may vary from a few days to several weeks depending on the required maintenance or repairs.

3-8.4.5 Ancillary Uses and the Future of Omega

The future of Omega shows great promise due to the unusual versatility that has been built into the system. With several transmitting frequencies available, increasing station redundancy for any given geographic area, and new advances in electronic technology, many essentially new and independent measurements can now be performed using Omega navigation signals. These measurements give rise to special techniques such as Differential Omega, precise time transfer, frequency standard calibration, and weather balloon tracking.

3-8.4.5.1 Time and Frequency

Omega transmissions can be used as sources of precise time due to the inherent reliability of VLF propagation and almost continuous signal availability. The format provides various carrier frequencies which may be used either singly or in various combinations for timing.

3-8.4.5.2 Differential Omega

Many large commercial and U.S. Navy ships are now equipped with Omega receivers providing enroute navigation capabilities of two to four nautical miles (95% of the time). This accuracy is adequate at sea, but not for most coastal regions. Omega accuracy can be improved to within 1/4 nautical miles or better in coastal areas by a technique known as Differential Omega. A radio link is used to relay the differential corrections from a local monitor station to Omega users in the vicinity. Differential Omega can be employed to increase the repeatable accuracy of the system. Since propagational errors vary in the same pattern for a local area, a fixed monitor station in the area measures all diurnal changes from its computed or standard coordinates. Upon receiving differential corrections (e.g., via radio data link), users in the same area (within a 200-nautical mile radius) obtain fixed accuracies that vary from 0.25 natucial miles at distances of 50 miles to 0.5 nautical miles at 200 miles distance. Figure 3-143 illustrates the differential Omega concept.

3-8.4.6 Charts and Publications

All U.S. Omega plotting charts and publications are issued by the Defense Mapping Agency Hydrographic Center (DMAHC), Washington, D.C. Both Omega propagation correction tables and hyperbolic lattice tables carry the Publication Series number 224 and are so indexed in DMAHC Publ. No. 1-N-A, Catalog of Nautical Charts and Publications. Nautical charts overprinted with the Omega lattice are indexed in the Regional catalogs (Pubs. 1N-1 through 1N-9). The tabular data has been subdivided and arranged such that the Omega Propagation Correction Table series will be printed, distributed, and made available for each individual station, enabling the navigator to acquire only the stations and areas desired. The publication number, together with the suffix and station letter, will fully identify the table for requisitioning purposes. A standard phase velocity is used for each

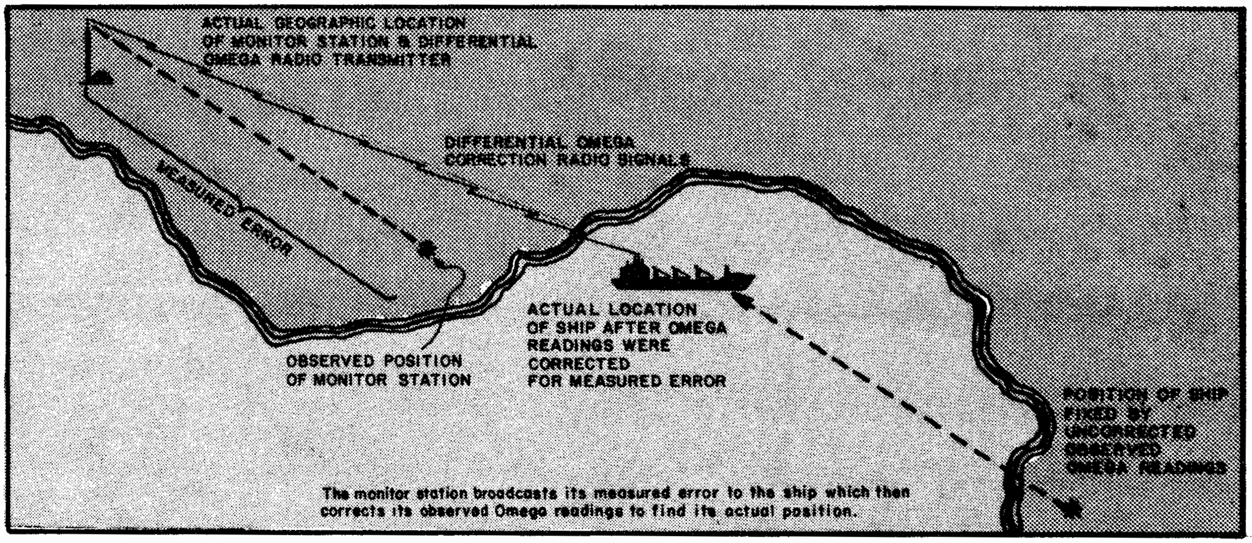


Figure 3-143. Differential Omega Concept

chart (propagation velocity changes are accounted for in PPC tables), and only the basic 10.2 kHz frequency LOPs are printed. LOPs for other frequencies can be derived from this basic frequency. Additionally, lattice tables are being produced for the navigator, who can then construct his own LOPs on any plotting sheet or chart desired. Instructions for using the lattice tables are provided in the introduction to the tables.

3-8.4.7 System Testing

Due to the varied Omega system configuration, the technician should consult the equipment technical manual, along with the related PMS and maintenance standards, when performing maintenance or repair on Omega components.

3-8.5 SATELLITE NAVIGATION

The Navy Navigational Satellite System (NAVSAT) provides an accurate, all-weather, worldwide navigational capability for naval surface vessels, aircarft, and submarines. The measurement of radio signals transmitted by NAVSAT is based on the Doppler shift phenomenon - the apparent change in frequency of the radio waves received when the distance between the source of radiation (in this case the satellite) and the receiving station, is increasing or decreasing because of the motion of either or both. The amount of shift in either case is proportional to the velocity of approach or recession. The frequency is shifted upwards as the satellite approaches the receiving station and shifted downward as the satellite passes and recedes. The amount of this shift depends on the exact location of the receiving station with respect to the path of the satellite. Accordingly, if the satellite positions (orbits) are known, it is possible by a very exact measure of the Doppler shift in frequency to calculate the location of the receiver on earth. The Doppler shift is also affected by the earth's rotation, but this effect is allowed for and corrected by the computer in providing the fix. The accuracy obtained by using this Doppler shift technique is possible because the quantities measured, frequency and time, can readily be determined to an accuracy of one part in a billion. The NAVSAT system (Figure 3-144) consists of one or more satellites, ground tracking stations, a computing center, an injection station, Naval Observatory time signals, and the shipboard receiver and computer. Each satellite is placed in a nominally circular polar orbit at an altitude of about 600 nautical miles, orbiting the earth in approximately 105 minutes. Only one satellite is used at any given time to determine position. The satellite stores data which is updated from a ground station approximately every twelve hours and it broadcasts the following data every two minutes: fixed and variable parameters described in its own orbit; a time reference. Two frequencies (150 mHz and 400 mHz) are employed because the ionosphere, which is a dispersion medium bends and also stretches radio waves, causing the satellite to seem closer than it actually is. Each frequency is somewhat differently affected, precise allowance can be made for the

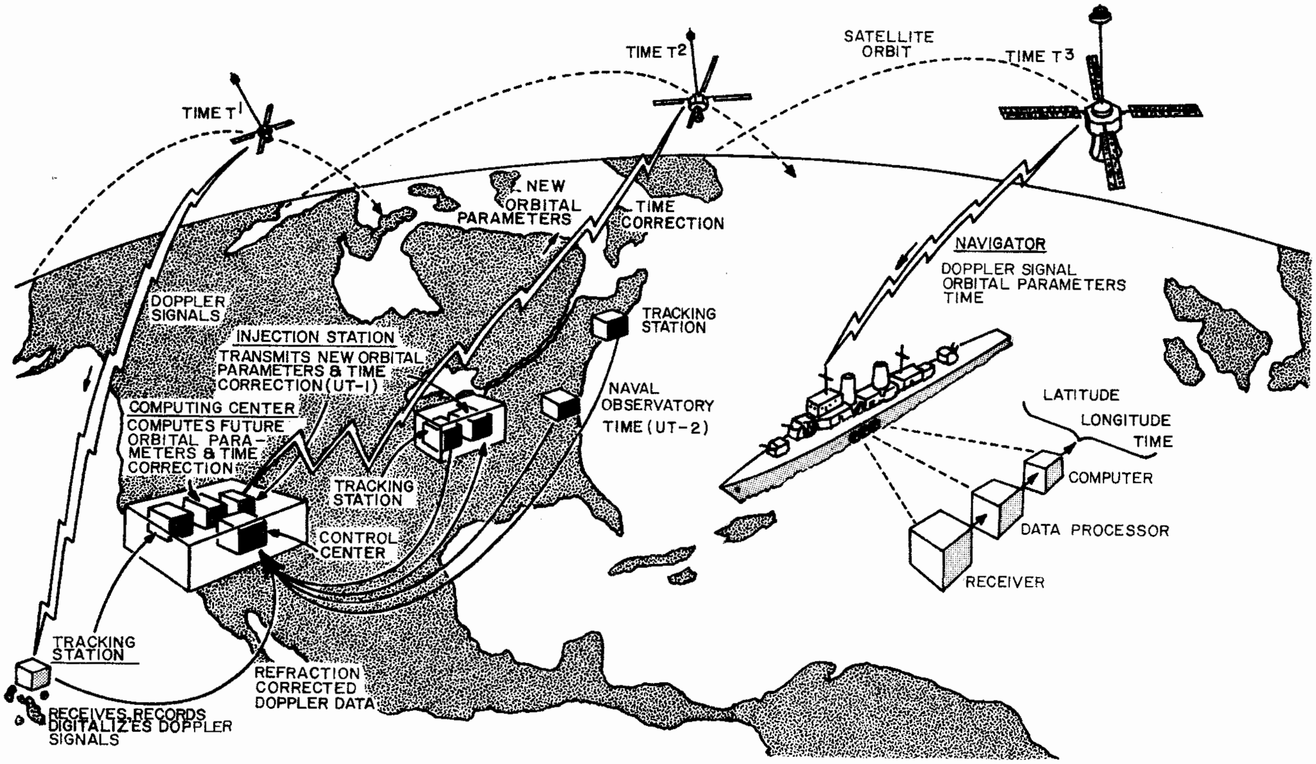


Figure 3-144. Navy Navigational Satellite System

ionosphere's effect on the waves. The parameters describe the satellite orbit as a function of time and are correct only for the 2-minute time interval for which they are transmitted by the satellite, and for those intervals immediately preceding and following that period. In each 2-minute transmitting period, data on eight periods is given: four before, and four after the time of the message. The time reference is synchronized with corrected GMT (UT_2) from the Naval Observatory. NAVSAT users are kept informed as to the operational status of the satellites, of the insertion of new satellites into service, and the withdrawal of satellites. This information is in the form of SPATRAK messages originated by the U. S. Naval Astronautics Group, Pt. Mugu, California.

3-8.5.1 Effects on Satellite Orbit

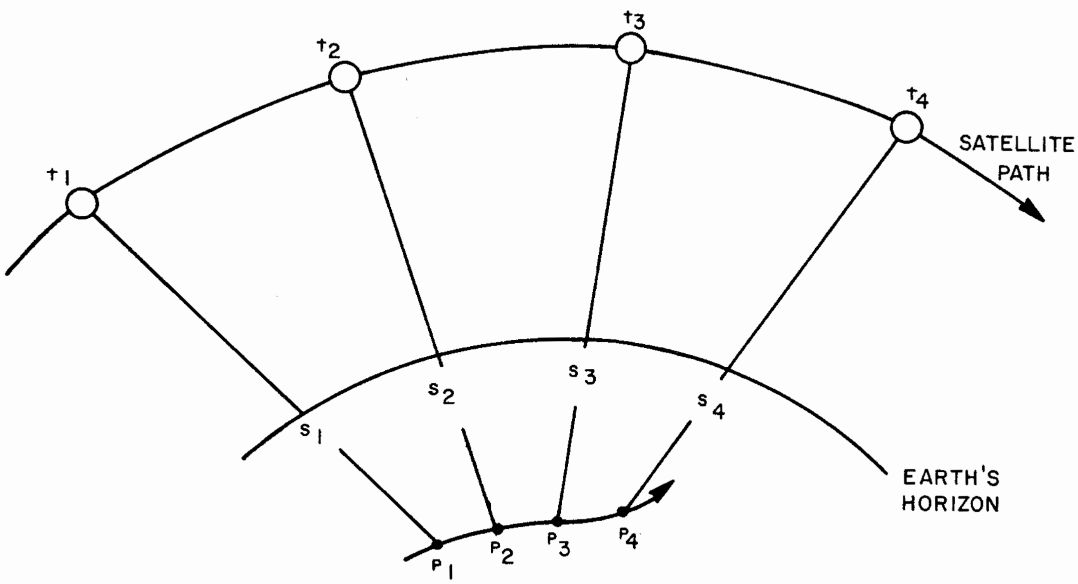
A planet in deep space follows a fixed path around its parent body in accordance with Newton's Laws of Motion. Its orbit is Keplerian, or perfectly elliptical, and its position can be predicted exactly for any given future instant of time. A NAVSAT satellite moves under the earth's gravitational attraction in accordance with the same laws, but as it operates at an altitude of about 600 miles, it is subjected to external forces that produce orbital irregularities, or perturbations. To make the system acceptable, these perturbations must be accurately predicted, so that the satellite's position can be determined for any instant of time. The most important of these forces is caused by the earth's shape. The earth is not a sphere, but is an oblate spheroid; in addition, its gravitational field is irregular. The satellite is also subject to slight atmospheric drag, as it is not operating in a complete vacuum. Other external forces which affect it are the gravitational attraction of the sun and the moon, solar photon pressure and solar wind, electrostatic and electromagnetic forces caused by the satellite's interaction with charged particles in space, and in the earth's magnetic field. Fortunately, all the forces causing perturbations are either sufficiently constant or so localized that they can be reduced to formulae which can be programmed into orbital computations. To determine the precise orbit of each satellite in the system, ground tracking stations are established at exactly determined positions in Hawaii, California, Minnesota, and Maine. These stations regularly monitor the Doppler signal as a function of time. Concurrently, the U. S. Naval Observatory monitors the satellite's time signal for comparison with corrected universal time (UT_2). The resulting information is transmitted to the computing center for processing. As the satellite is essentially moving as a planet, and as the perturbations in its orbit are determined by the computer, of all the possible paths permitted by Newton's Laws, only one can result in a particular curve of Doppler shift. Thus, at any instant of time, the position of the satellite relative to the known location of the tracking station can be determined very precisely. The computing center, having received this data, computes an orbit for the satellite that best fits the Doppler curve obtained from the tracking stations. This orbital information is extrapolated to give satellite positions for each two minutes of UT_2 for the following sixteen hours, and this data is supplied to the injection station, for transmission to the satellite about every twelve hours, for storage and retransmission on schedule. The satellite is in effect a relay station which stores and transmits the data computed at ground stations and which are inserted in its memory system.

3-8.5.2 NAVSAT Equipment Configurations

Typical NAVSAT shipboard equipment used by the navigator consists of a receiver, a computer, and a readout unit. Complete operator instructions are suppled in the operator's manual. The satellites (sometimes referred to as "birds") are completely transistorized; they are octagonal in shape, and have four windmill-like vanes, which carry solar cells. They are gravity-gradient stablized, so that the directional antennae are always pointed downwards, toward the earth. A satellite fix may be obtained when the satellite's maximum altitude, relative to the observer, is above 10° and less then 70°. As a general rule, each satellite will yield four fixes a day - two on successive orbits, and two more on successive orbits some twelve hours later. However, this sequence may be disturbed, as the satellite, while above the horizon, may pass at too great or too small an altitude relative to the observer to permit obtaining an accurate position.

3-8.5.3 Obtaining a Fix

To obtain a navigational fix, the ship's estimated position and velocity of movement must be entered in the computer. The accuracy of the estimated position is not of great importance; however, the accuracy with which the course velocity can be established is important, as will be seen in the following discussion. On ships equipped with the Ships Inertial Navigation System (SINS) (described later), the two-minute synchronization signal received from the satellite is transmitted to SINS. In some installations this signal causes the SINS to print out ship's position data coinciding with the two-minute Doppler count. In other installations the SINS general purpose computer is used to solve the NAVSAT problem rather than employing a separate computer. If inertial equipment is not available to supply automatic information on the ship's movement to the computer, the course and speed from the gyrocompass and EM log are inserted in the computer. This, of course, is a potential source of error, as the system, for high accuracy, requires an input of the ship's true velocity; that is, her speed and direction of travel relative to the surface of the earth. Unfortunately, accurate information on the existence of a current and its set and drift is rarely available to the navigator. In round numbers, the error in a NAVSAT fix will be about 0.25 miles for every knot of unknown velocity. A velocity north (or south) error causes a considerably larger error in the fix than a velocity east error. The fix determined by the ship's computing system is based on the Doppler frequency shift which occurs whenever the relative distance between a transmitter and a receiver is changing. Such a change occurs whenever a transmitting satellite passes within range of a radio receiver on earth, and consists of a combination of the motion of the satellite in its orbit, the motion of the vessel over the surface of the earth, and the rotation of the earth about its axis. Each of these motions contributes to the overall Doppler frequency shift in a characteristic manner. An increase in frequency occurs as the satellite approaches the ship, in effect compressing the waves en route. The received frequency exactly equals the transmitted frequency at the point of closest satellite approach, where for an instant of time there is not relative motion directly along the vector from the satellite to the receiver. The received frequency then decreases as the satellite recedes from the ship's position thereby expanding the radio waves between them. The shape of the curve of frequency differences and its time of reception depend both on the receiver's position on earth and the satellite's location in space. The reception of these Doppler signals and the resulting computer computations form the basis of the satellite navigation system. Figure 3-145 shows (in simplified form) the relationship of time, range, and position. In the diagram, t_1 through t_4 represent the position of the satellite in orbit at the successive transmissions which occur at two minute intervals. S_1 through S_4 represent the slant range between the satellite and the ship. P_1 through P_4 represent the position of the ship referenced to the time at which the receiver recognizes the satellite synchronization signal $t_1 + \Delta t_1$ through $t_4 + \Delta t_4$, where Δt represent the time interval for the signal to travel from the satellite to the receiver aboard ship. Figure 3-146 shows Doppler frequency variation with time. The integral Doppler measurements (Figure 3-146) are simply the count N 1-2 of the number of cycles received between



f_o	=	NOMINAL VALUE OF NAVIGATOR'S REFERENCE FREQUENCY
c	=	SPEED OF LIGHT
Δ_f	=	DIFFERENCE BETWEEN NAVIGATOR'S REFERENCE FREQUENCY AND SATELLITE TRANSMISSION FREQUENCY
T	=	2 MINUTES (I.E. $t_2 - t_1$, $t_3 - t_2$, ETC.)
λ	=	LATITUDE
ω	=	LONGITUDE
t	=	TIME OF TRANSMISSION TIMING MARK
$t + \Delta t$	=	TIME OF RECEPTION OF TIMING MARK
N	=	DOPPLER COUNT
S	=	SLANT RANGE

$$N_{12} = f_{o/c}[S_2(\lambda;\omega)-S_1(\lambda,\omega)] + \Delta f.T$$
$$N_{23} = f_{o/c}[S_3(\lambda,\omega)-S_2(\lambda,\omega)] + \Delta f.T$$
$$N_{34} = f_{o/c}[S_4(\lambda,\omega)-S_3(\lambda,\omega)] + \Delta f.T$$

Figure 3-145. Integrated Doppler Measurement

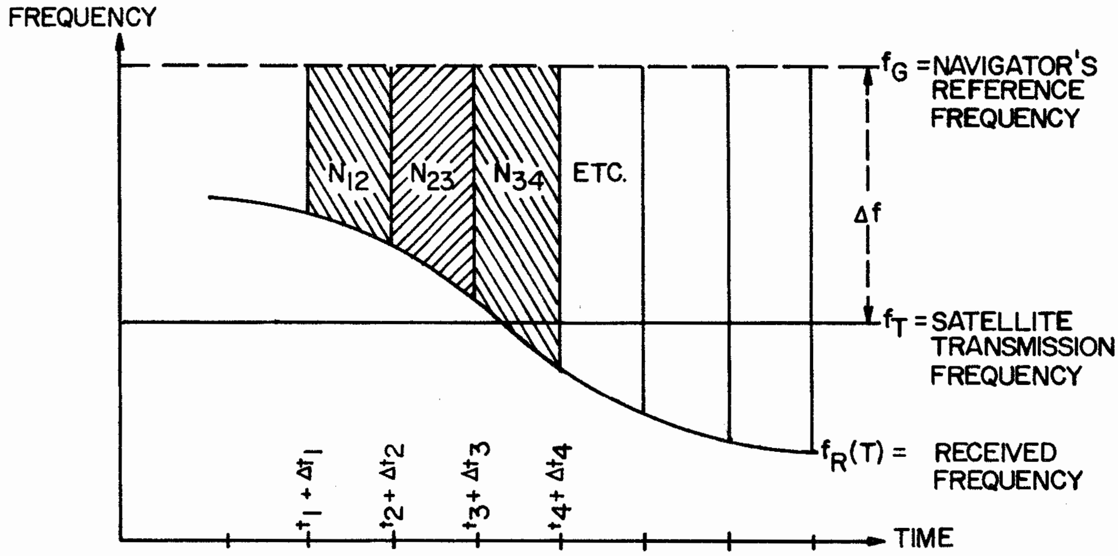


Figure 3-146. Doppler Frequency Variation with Time

$t_1 + \Delta t_1$, and $t_2 + \Delta t_2$, the count N 2-3 of the number of Doppler cycles between $t_2 + \Delta t_2$ and $t_3 + \Delta t_3$, and so on for all two-minute intervals during the satellite pass. Five or six two-minute Doppler counts are obtained during a typical satellite pass. Each Doppler count consists of a constant, plus a measured slant range difference between the receiver, and the satellite at positions defined by the navigation message. The measured range differences are truly known only if the constant but unknown frequency difference, ΔF, between the satellite's oscillator and the receiver's reference oscillator can be determined. To calculate a position fix, the Doppler counts and the satellite message are fed to a digital computer. The computer is also provided with an initial estimate of the ship's latitude and longitude and an estimate of the frequency difference, ΔF. The computer then compares calculated range differences from the known satellite positions to the estimated ship's position with those measured by the Doppler counts, and the navigation fix is obtained by searching for and finding those values of latitude, longitude, and ΔF which make the calculated range differences agree best with the measured range differences. Because the geometry is complicated, only simple, linearized equations are used, and the computations are performed repeatedly until the solution converges. No more than three or four repetitions are normally required, and a fix is obtained within a minute or less on typical small digital computers. The time signal transmitted by the satellite, which occurs at the two-minute mark, is accurate to better than 0.02 seconds. It thus may be used conveniently as an accurate chronometer check. With the NAVSAT receiver locked to the satellite signal, the two-minute signal will be heard as a "beep."

3-8.5.4 System Testing

Tests are usually performed with an interval test set that simulates in satellite signal and message. Consult PMS and the equipment's technical manual for specific test procedures for a particular NAVSAT equipment.

3-8.6 INERTIAL NAVIGATION

Inertial navigation is based on the measurement of vehicle accelerations parallel to the earth's surface. An accelerometer provides a signal which is proportional to acceleration occuring along its input-sensitive axis. Acceleration is integrated twice to obtain distance. The first integration yields velocity, and the integration of velocity gives displacement. In both bases these are relative velocities and relative displacements with respect to the initial starting velocity or starting point. The measurement of acceleration is based upon rigid physical principles as originally defined by Newton. Acceleration can be sensed with respect to inertial space by use of an accelerometer, and displacement is obtained by integration, which is inherent in the accelerometer or provided by means of external integrators. In any case, no external contact with the physical world is necessary. A requirement of an inertial navigation system is a stable system of reference with respect to which the acceleration measurements are made. The input axes of the accelerometers must be maintained tangent to the earth and aligned to the appropriate earth coordinates. To accomplish this isolation and alignment, a "platform" is needed. A platform, though a unit by itself, is necessary to maintain the proper alignment of the accelerometers regardless of rotational disturbances caused by vehicle motion. The gyroscope is utilized in sensing rotational motion, and provides signals to maintain correct alignment of the platform.

3-8.6.1 Development Criterion

Ships and airplanes have long been guided during periods of poor visibility by a system of "dead reckoning". From a known position at start, speed and direction of motion (velocity) is integrated with time to give a distance and direction (displacement) which when plotted from the starting point results in current position. The limitations of this system are the accuracies of the instruments used to measure the variables. Time can be measured very accurately, but even with the most accurate logs or air speed indicators, velocity can only be measured with respect to the medium air or water, in which the vehicle travels. Currents of wind or water can only be estimated, with the resultant velocity over the ground being largely a matter of guess work. By the application of Newton's Laws of motion, velocity may be found by the integration of acceleration with time, and acceleration may be found by measuring the force required to displace a mass. Further, this acceleration is relative to the universe and not to the medium in which the vehicle is traveling. The development of a system of dead reckoning which would be highly reliable over long distances required sensitive and accurate instruments. Such instruments have now been developed and in conjunction with the gyroscope and have resulted in the inertial navigation system. Inertial navigation introduces the problem of converting the universe- or space-oriented system to the earth coordinates which mark position on earth. Inertial navigation has provided reliable guidance of ships and planes during all types of weather conditions and has made practical the nuclear submarine which is capable of remaining submerged for long periods and to travel under the polar ice fields. The inertial navigation system is not dependent upon visual fixes, allowing navigation without approaching the surface of the water. It is independent of external electronic signals and therefore cannot be interfered with and emits no signal of any type, thus not betraying its location. This subsection will deal with the problems of orienting the system to earth coordinates, and with the instruments used to make the necessary precise measurements.

3-8.6.2 Earth Characteristics

The earth is approximately a sphere, but due to the centrifuge effect of its rotation, the diameter through the equator is approximately twenty-nine miles greater than that through the poles. The earth rotates around its axis once every twenty-four hours, which reduces to an angular velocity of fifteen degrees per hour. The earth is not homogenous; its density is greater at some places than at others. These variations must be taken into consideration in inertial navigation but for the purposes of developing a grid system are neglected.

3-8.6.3 Coordinate System

The grid system which is used for computing position on the earth is illustrated in Figure 3-147, and it consists of imaginary lines of latitude and longitude. Lines of latitude form east/west circles about the surface of the earth and are generated by the rotation of a line perpendicular to the earth's spin axis. All lines of latitude are parallel to each other and are referred to as "parallels". The equator is the greatest line of latitude and is the reference line for measuring latitude north or south. Latitude is measured in degrees, minutes, and seconds of arc. Lines of longitude form semicircles on the earth's surface which run from one pole to the other. They are formed by the intersection of the earth's surface with a plane which contains the earth's spin axis. The reference line for longitude is the Prime Meridian, which is the line of longitude that passes through Greenwhich, England. Longitude is measured in degrees, minutes, and seconds of arc east or west of the Prime Meridian to a maximum of 180 degrees. Any unique position on earth may be described by the intersection of lines of latitude and longitude.

3-8.6.4 Vehicular Navigation on Earth

The distance between two circles of latitude can be calculated by using the formula derived in physics, $S=R \times \theta$, where S is the distance, R is the radius of the earth, and θ is the difference between angles of latitude. The problem of finding the distance between lines of longitude is somewhat different. At the equator, $S=R \times \theta$, where θ is the angle of longitude, but anywhere away from the equator, the lines of longitude are closer; in fact, there is no distance between them at the poles. When calculating distances between lines of longitude, and radius of the circle of latitude on which the vehicle is traveling must be considered. Figure 3-147 shows how this new radius is calculated. This new radius must be used in the formula $S=R \times \theta$, or $S=(r \cos Lat) \times \theta$. Remember that in this formula, our difference angle of longitude (θ) is measured in radians.

3-8.6.5 Basic Requirements

Ships require an instrument in which latitude and longitude can be set before leaving port, and which will continually read out its exact latitude

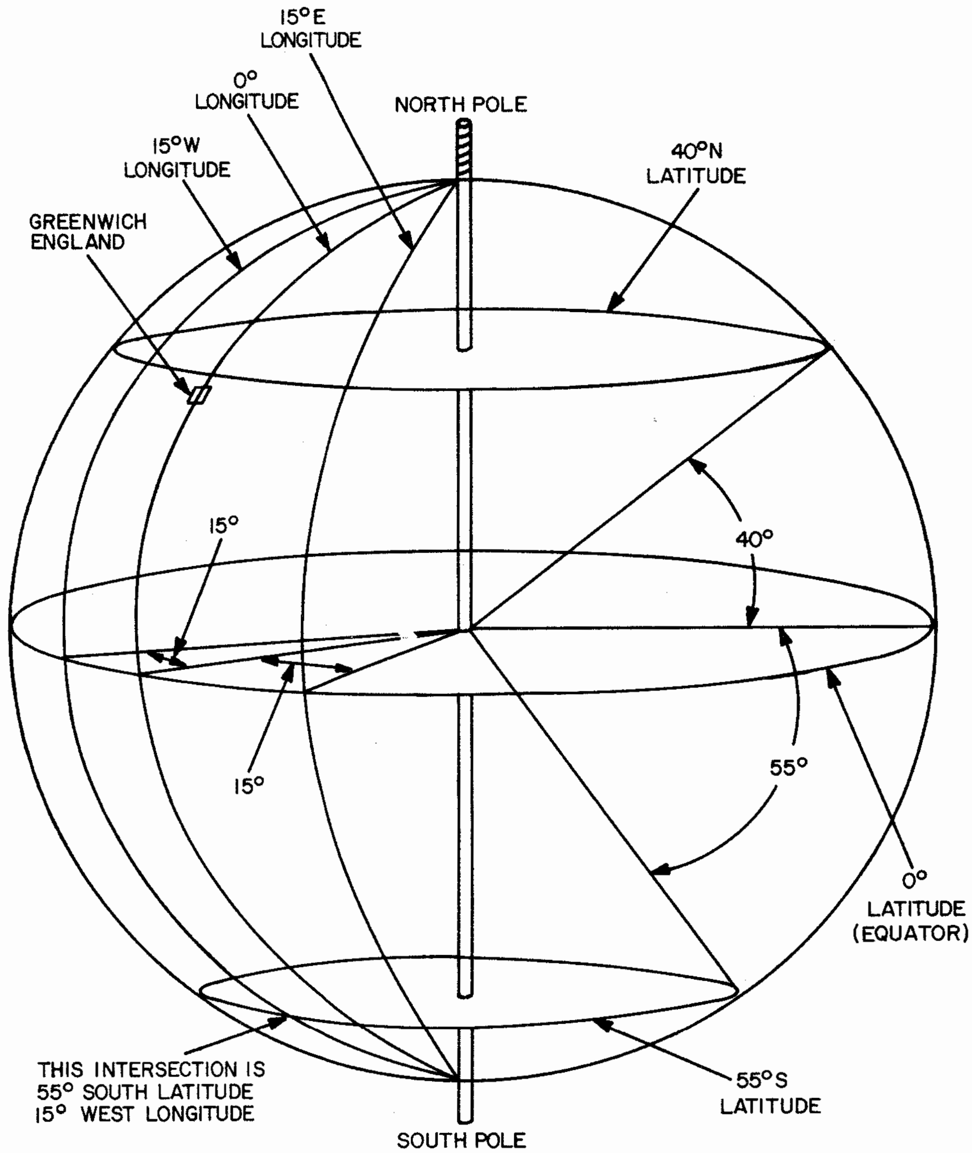


Figure 3-147. Latitude and Longitude Grid Principle

and longitude no matter where the vessel goes. This instrument must also indicate the direction of true north at all times. This could be done if there existed in the ship a platform with some mechanization so that it would always be horizontal. In other words, anything resting upon that horizontal platform would sense gravity as being straight down. A ball would therefore not roll of its own accord if set upon this platform, as illustrated in Figure 3-148. The ship's rotational movements of roll, pitch and yaw effects would thus be removed from the platform. Accelerometers could be mounted on such a platform to measure the ship's acceleration north or south, or east or west as illustrated in Figure 3-149. Now, if the ship were accelerating to the east, by timing how long it accelerated, final velocity would be known. Then, by timing how long the ship stayed at that velocity, this would determine how far the ship had gone to the east (acceleration $\rightarrow \int_{\text{time}}$ = velocity $\rightarrow \int_{\text{time}}$ = distance east). A new longitude can be computed by knowing how far east the ship has gone. If, at the same time, acceleration is measured to the north as in the above diagram, the same process would give distance north. This is easily converted to latitude.

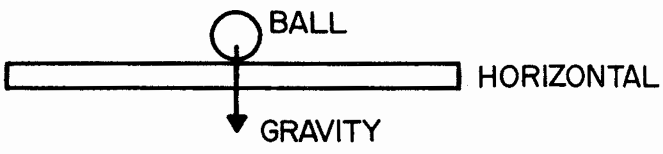


Figure 3-148. Stable Platform

3-8.6.5.1 Accelerometers

Using the familiar concept of F = ma, a mass (m) being accelerated (a) will have a certain force (F) acting on it to cause that acceleration: when the ship moves, it takes the platform with it, that is, it exerts a force on the platform. This is illustrated in Figure 3-150. With this simple mechanism, it is possible to measure the amount of force applied. The inertia of the weight makes it want to stay where it is, but the stretching spring gradually increases the force on the weight until the spring stops stretching and the weight is stationary in its box. (This is where the term "Inertial" navigation comes from). An electrical pickoff might be used to indicate how far the spring stretched, i.e., how far the weight moved from the center position. The distance the weight moved is equivalent to Force (as in any spring scale) and the mass of the weight is known. If F = ma then $a = \frac{F}{m}$, or, in other words, the distance the weight moves is directly proportional to acceleration. Since the actual acceleration of a ship varies from moment to moment, it would be necessary to read the meter very rapidly and very often. Although this would be difficult for an individual, it is very simple for a computer to take very rapid readings and do the whole computation instantly.

3-8.6.5.2 Misaligned Platform

A stable platform could be improperly aligned in two ways. It might not be exactly horizontal, as illustrated in Figure 3-151, or it might not be exactly oriented to north, as illustrated in Figure 3-152. If this were an east-west accelerometer, it would sense a component of gravity causing the weight to move to the left (down hill). This would therefore be interpreted by the computer as an acceleration to the east. The computer, not knowing the platform is tilted, would compute a new longitude. Since the ship did not actually move, this would constitue a longitude error. In Figure 3-152, the computer again assumes that the platform is properly oriented, in which case the E-W accelerometer senses only E-W motion. Actually, the E-W accelerometer is sensing some E-W motion and some N-S motion, but this enters the computer only as E-W acceleration. Likewise, the N-S acceleration, although a mixed quantity, is interpreted by the computer as being only N-S. Thus, the computer computes a direction of movement that is not quite correct, resulting in a latitude and longitude

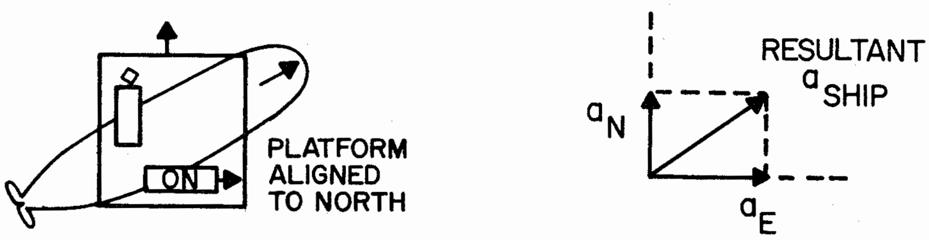


Figure 3-149. Ship's Acceleration Vector

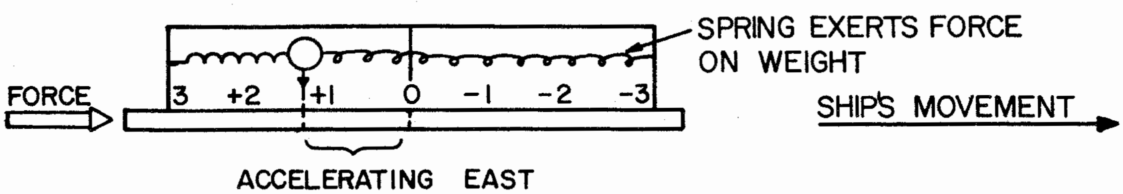


Figure 3-150. Basic Accelerometer

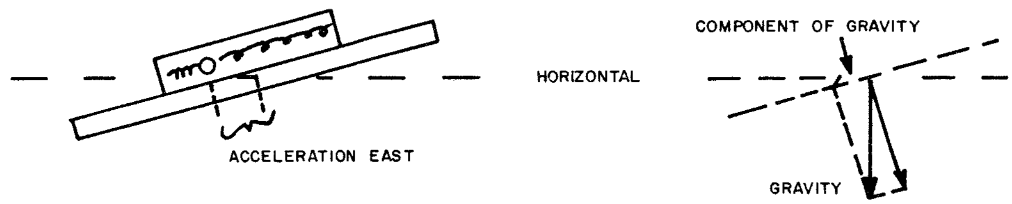


Figure 3-151. Platform Not Horizontal

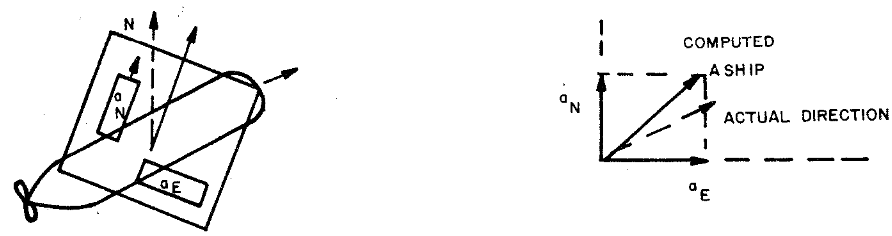


Figure 3-152. Heading Misalignment (orientation to north)

ERROR. (Note that the computer does not compute direction vectors but only the N-S component and E-W component in terms of latitude and longitude).

3-8.6.6 Basic SINS

The Ships Inertial Navigation System's (SINS) operation centers on determining and correcting the stable platforms errors by comparing the SINS latitude and longitude indications with outside "fix" information. The errors then can be determined and corrections made. Once this is done, the SINS will operate for long periods with only extremely small errors.

3-8.6.6.1 Essential Components

When a gyroscope (a device which tends to maintain its rigidity or direction in space) is mounted on a platform, the horizontal plane of the platform is established and maintained. Gimbals enable the platform free movement in any direction as illustrated in Figure 3-153. If the platform rotates about any of its three mutually perpendicular axes, the gyroscope will be forced to precess. This precession can be detected electronically. The electronic signals are sent to the "Platform Electronics" cabinet located near the platform. These electronic circuits divide the sensed motion into Roll, Pitch and Heading. After amplification, these newly-developed signals are sent to drive motors on each gimbal axis such that the platform is sent back to its original position. Since the gimbals are geared together, when the ship rolls, the platform tips; the tipping platform exerts a force on the gyroscope; the force on the gyroscope causes it to precess; the electronic pickoff measures the amount and direction of precession; the electronic circuits convert the signal to gimbal motor drive current; the gimbal motors drive the gimbal just enough to compensate for the ship roll, returning the platform to its original position. Thus, if the platform were horizontal originally, it will again be horizontal. This all occurs virtually instantaneously such that a marble sitting on the platform would not even begin to roll. Therefore, an accelerometer would not be tipped and would not sense any gravity. On a ship in port, the accelerometer would produce no output since the ship is not moving horizontally and the rolling (pitching and yawing) effect is removed.

3-8.6.6.2 Platform Alignment

Since it is desired that this electronic followup circuit act to maintain the platform in its original position, it is necessary to first align the platform to the horizontal. This operation is provided for quite simply in the SINS by using Alignment Modes of operation, which merely use gravity to indicate the horizontal. The control is accomplished through special circuitry which uses the accelerometers similar to a bubble level. Once the platform is horizontal, the SINS is switched to the Navigate Mode, which uses the followup system just described.

3-8.6.6.3 Ships Movement

With a horizontal platform, the accelerometers are free to sense any N-S E-W movement.

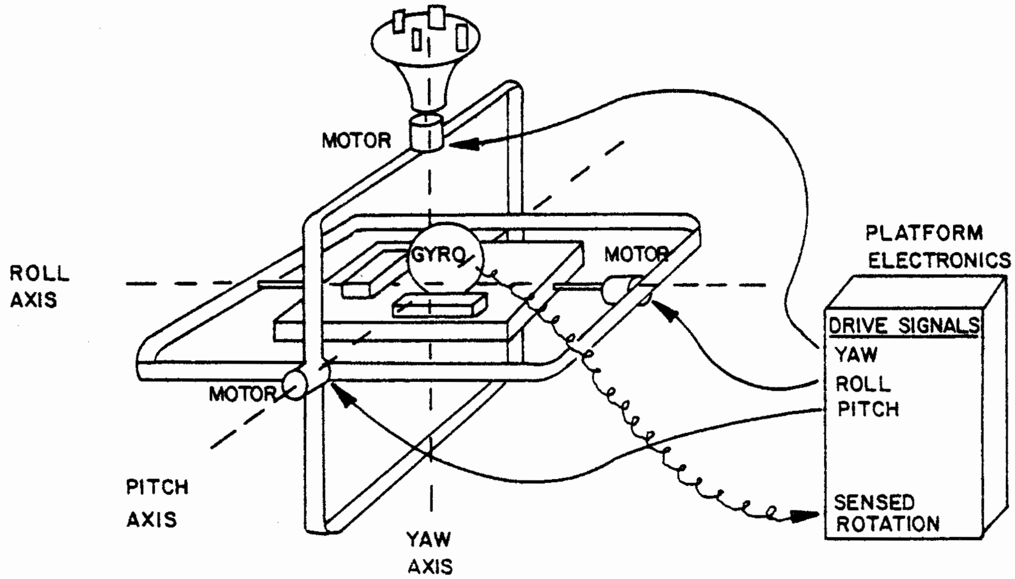


Figure 3-153. Stable Platform Principle

If the earth were flat, that would be all that is needed for an inertial navigator. Consider what occurs in Figure 3-154 when the ship moves from point A to point B to point C on the nearly spherical earth. The ship must follow the surface of the earth, and in so doing the ship is actually ROTATING, just as though it were pitching. In paragraph 3-8.6.6.1, it was explained that the gyroscope senses any rotation of the

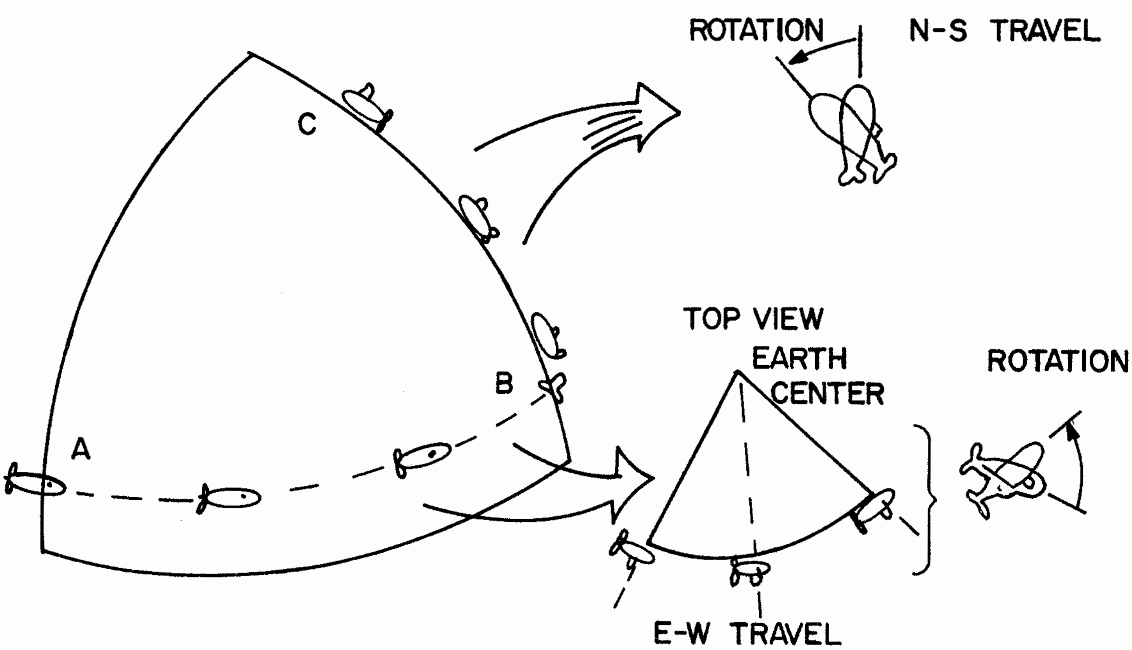


Figure 3-154. N-S E-W Movement

ship and counteracts that rotation. Therefore, if undisturbed, the stable platform would not be compensated for the rotation caused by the east to west motion in Figure 3-155 and the platform would not remain horizontal. When the ship has sailed a quarter of the way around the earth, as illustrated in Figure 3-155, the uncontrolled platform would actually become vertical. This obviously would not provide an acceptable horizontal platform for accelerometers to operate properly. However, if it is known how far the ship has traveled from the spot where the platform was initially horizontal, simple geometry can be used (knowing the radius of the earth) to calculate how much the earth's surface has curved, and therefore how much the ship has "rotated" because of the earth's curvature. The shipboard computer can add in that calculated rotation to the signals controlling the platform gimbals, and the platform will be adjusted to the new "calculated" horizontal. Accelerometers are thus used with a computer to determine velocities and distances moved. Since this is done very rapidly and frequently (every few feet) the computer calculates a new position on earth. At each such new position, the computer can solve the geometry equation and determine a new horizontal. Thus, the stable platform will maintain its horizontal base anywhere on the curved surface of the earth. The entire process of SINS navigation is represented below in Figure 3-156. Note that for technical convenience, the signals developed by the computer to compensate for the curving earth surfaces are sent to the gyroscope as torquing commands in order to delude the gyroscope into generating extra rotation signals. Thus, the followup system adjusts the platform to the new "calculated" horizontal.

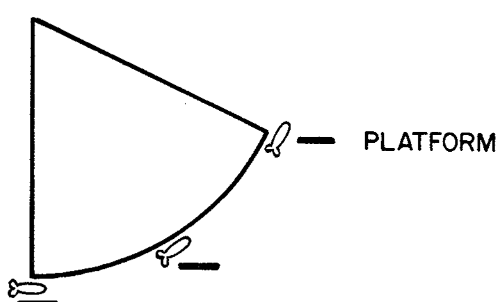


Figure 3-155. Platform Tilt as Ship Moves Around Earth

3-8.6.6.4 Heading

When the stable platform is originally aligned in the horizontal plane it must also be aligned to true north. However, once the platform is oriented to north, any change in ship heading (Yaw) is sensed as rotation by the gyroscope. The rotation of the platform is then countered by the followup system. Therefore, the platform should always maintain its initial orientation with true North.

3-8.6.6.5 Errors Developed in SINS

Any less-than-perfect SINS will develop navigational errors which are usually caused by the following misalignments:

1. Platform not EXACTLY horizontal.
2. Platform not EXACTLY aligned to true north.

As previously mentioned these two types of misalignments result in errors developing in latitude and longitude indications. In turn these errors cause HEADING errors. The result is a very delicate instrument that reads out constant latitude, longitude and heading, all of which are in error. SINS fortunately is far superior to other navigational aids because the latitude, longitude and heading errors always vary in an easily identifiable geometric pattern and therefore errors can be detected and corrected. Misalignment occurs because the operator is not always able to initially set the exact true latitude, longitude and heading into the SINS. Starting out with faulty information causes the platform to be not quite horizontal and not quite oriented to true north. However, if the error development can be observed over a period of 8 to 12 hours, the geometric pattern can be identified. The amount of error initially set in can then be calculated and the SINS corrected for that error. Properly corrected, the SINS will then operate for long periods with practically no error in latitude, longitude or heading indications.

3-8.6.6.6 Characteristic Errors

When the ship is in port, the exact position in terms of latitude and longitude is known. If this position is set into the SINS, it should continue to indicate that same position as long as the ship does not leave the berth. However, if there are intrinsic errors, the SINS will show the ship's latitude and longitude to be changing. By plotting the errors every hour, it will be seen that the latitude error varies continuously as a sine wave, as illustrated in Figure 3-157. Because latitude follows this pattern, only part of the curve is needed to draw the whole cycle, Figure 3-158. If the curve starts at zero error and after 6 hours there is (+) 20 miles error, in another

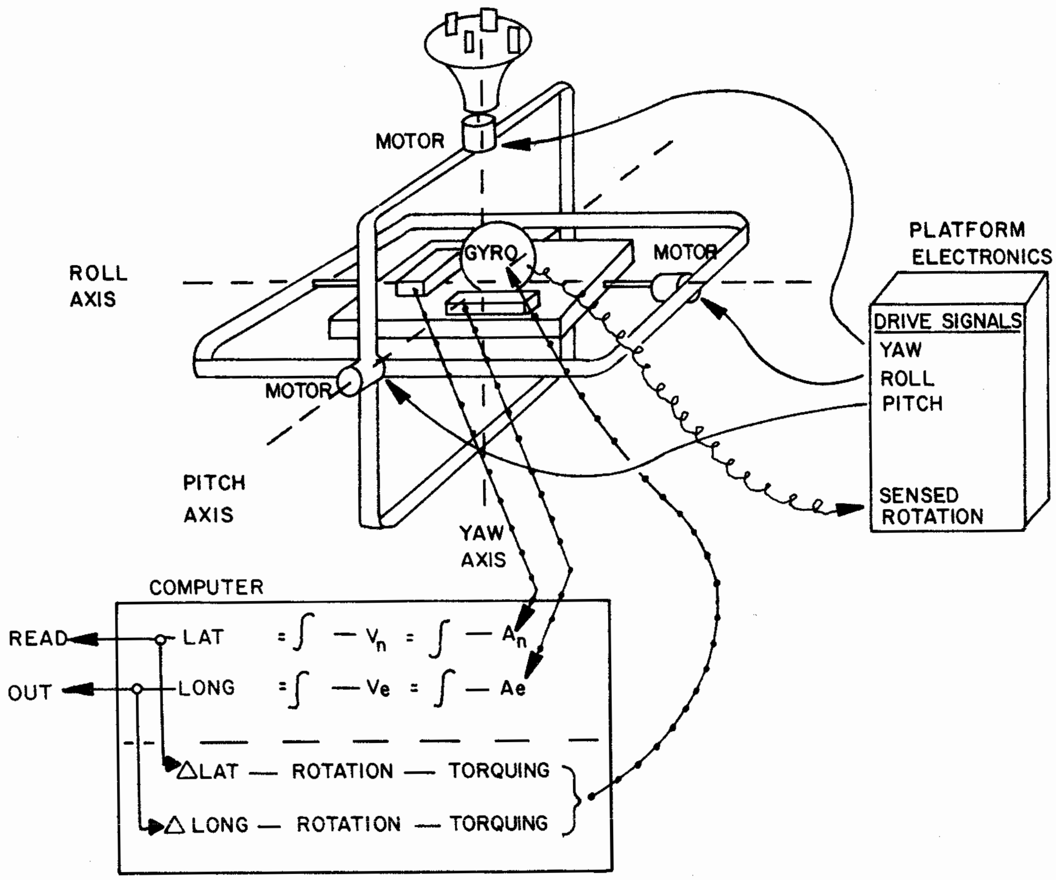


Figure 3-156. SINS Block Diagram

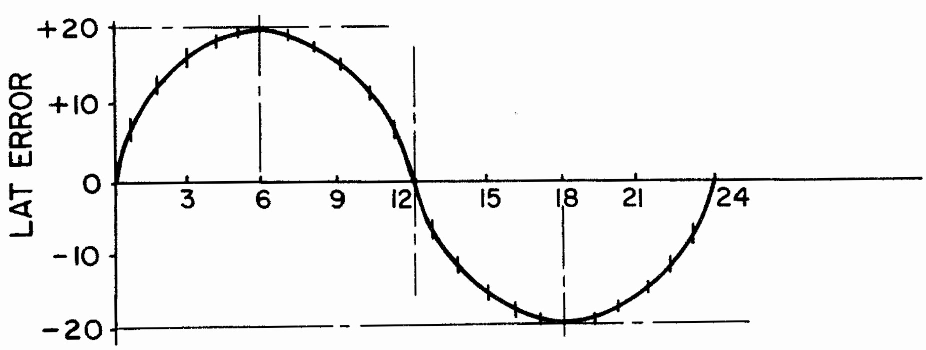


Figure 3-157. 24-Hour Latitude Error Curve

6 hours the error will be zero; and six hours after that occurrence the error will be (-) 20 miles. Thus, the latitude error follows a 24-hour sine wave. At the same time, the the operator might determine and plot the heading error every hour. It would be found that the HEADING error is directly related to latitude error but is offset by 6 hours (Figure 3-159). From these conclusions, the technician can predict SINS errors. For example: by settting the right latitude into the SINS and 6 hours later the error is (+) 10 miles, at 12

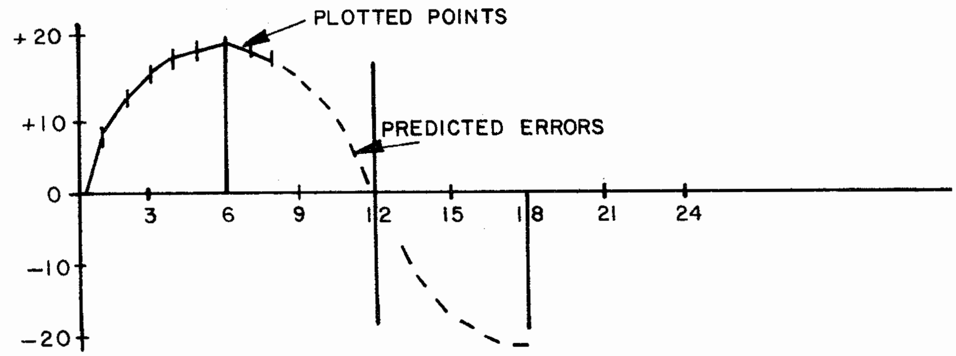


Figure 3-158. Plot of Latitude Error Curve

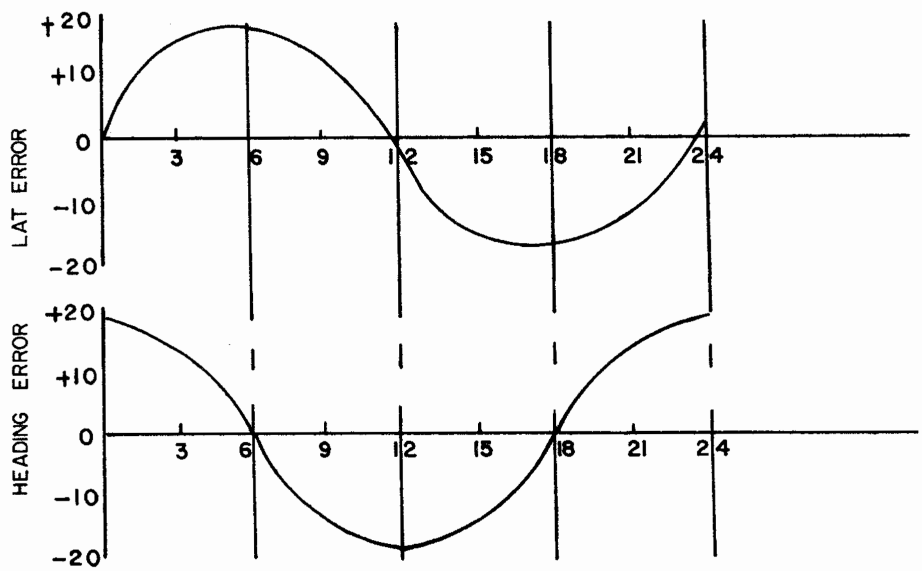


Figure 3-159. Heading vs. Latitude Error

hours the error will pass through zero, and at 18 hours it will be (-) 10 miles. Furthermore, since it is known that the HEADING error leads the latitude error, and that the latitude error is (+) miles (equal to 10 min) at 6 hours and will be zero at 12 hours, the HEADING error must be zero at 6 hours and will be (-) 10 min. at 12 hours (Figure 3-160). Actually, this occurs only at the Equator where the latitude is equal to 0°. At higher latitudes the Latitude Error = (Heading Error) (Cos Latitude). Therefore in the above example, at latitude 60°N, a latitude error of + 10 minutes will,

6 hours later, give a heading error of $\frac{+10 \text{ min.}}{\text{COS } 60} = +20$ min. In this case, the curve would appear like that shown in Figure 3-161. Nevertheless, the results are just as predictable. These curves follow a 24-hour oscillation pattern because a gyroscope tends to maintain rigidity with respect to SPACE. The earth turns in space at the rate of one revolution in 24 hours, therefore the predictability of SINS error propagation is as certain as the turning of the earth. Finally, a similar error plot of longitude will reveal that an incorrect longitude set into the SINS will remain as a constant error. This is illustrated in Figure 3-162.

3-8.6.6.7 Gyro Biasing

The errors discussed thus far were caused by the SINS not being properly aligned to start with, so that the SINS would generate incorrect information. However, the SINS will continue to do what it was programmed to do and will thus produce the characteristic 24-hour oscillations. If the oscillations are evaluated, initial errors can be detected and corrected. The gyroscopes used in SINS are finely

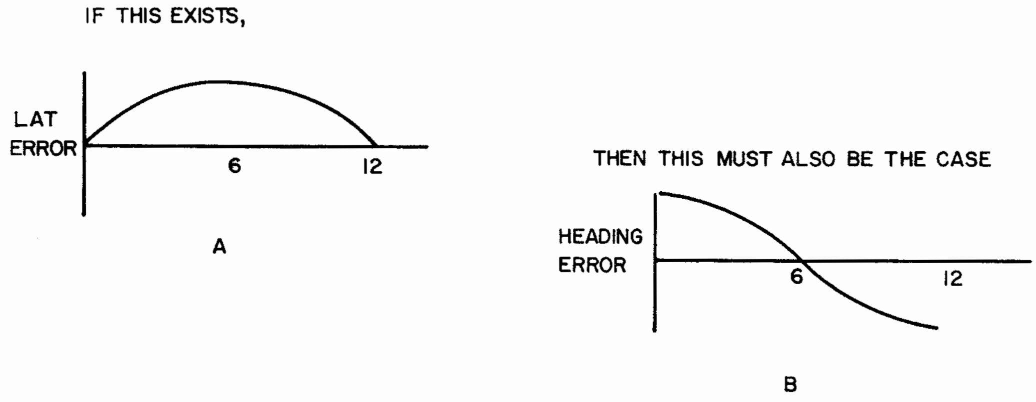


Figure 3-160. Predicted Heading Error

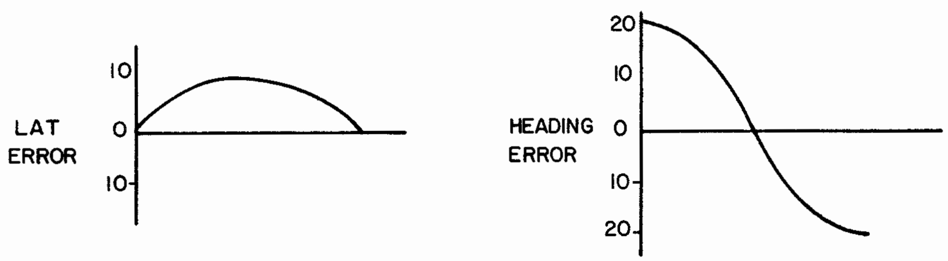


Figure 3-161. Latitude 60°N Predicted Heading Error Curve

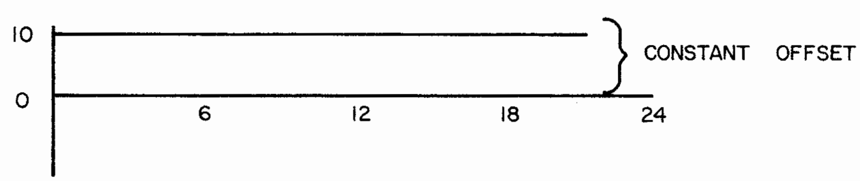


Figure 3-162. Longitude 24-Hour Error Plot

built precision instruments that are capable of operating without variation for long periods. Nevertheless, variations do occur. Every gyro built has inherent drift. Because of bearing drag or some other physical phenomena not completely understood, any operating gyro will tend to precess very slowly. A well-built gyro will precess at a constant rate and can therefore, be tested to determine the drift rate. This ascertained drift rate can then be fed into the computer, which can send torquing pulses to the gyro that exactly counter the known drift. This torquing is called "BIAS." Each new gyro installed must have its own particular drift rate ascertained and then set into the computer. Unfortunately, this drift rate changes randomly and unpredictably from time to time. These changes are called "BREAKS." If the latitude error curve is plotted over a period in which a break occurred, there would be a discontinuity (a break) in the curve (Figure 3-163). If, after the break, the error plot was continued, a new 24-hour oscillation would start such that magnitude of the sine wave (peak-to-peak) has changed (it may be larger or smaller). The whole curve would be offset above or below the zero axis of the plot, as illustrated in Figure 3-164. In previous discussions, the MEAN axis was said to be at the zero axis. However, the computer is now compensating for one drift

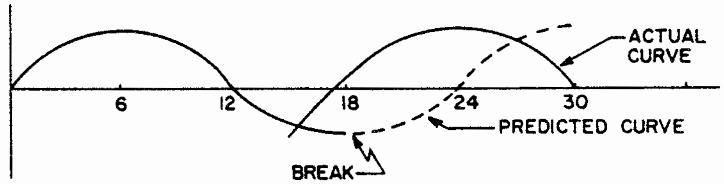


Figure 3-163. Breaks in 24-Hour Error Plot

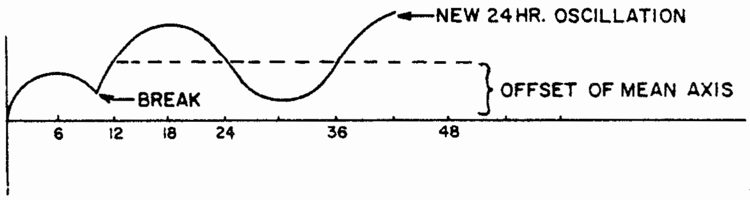


Figure 3-164. Gyro Break Curve

rate but the gyro is drifting at some new rate (i.e., BIAS is incorrect). The MEAN axis is now off the zero axis. It is important to note that the gyro always oscillates about the mean axis within a 24-hour period.

3-8.6.7 Resetting SINS

Two kinds of errors must be completely corrected for in the SINS. They are:

1. Misalignment errors (the ones the operator put in).
2. Gyro bias errors (the gyro's change in drift rate).

3-8.6.7.1 Latitude Misalignment

If, after plotting the error curve long enough to determine the characteristic sine wave, the amount of latitude error is determined and corrected for, the platform would be realigned. If this is done properly, the platform would be properly aligned and the 24-hour oscillation would be eliminated as illustrated in Figure 3-165.

3-8.6.7.2 Gyro Bias Error

Since gyro bias error causes a shift in the MEAN axis, a bias correction should shift the mean back to the zero axis as illustrated in Figure 3-166. This time the same oscillations continue

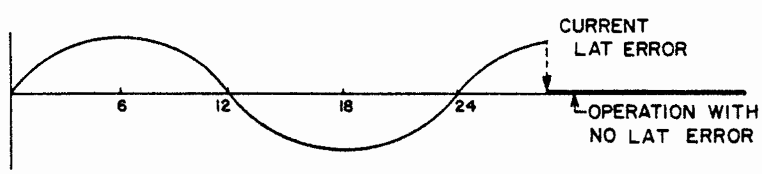


Figure 3-165. Latitude Corrections

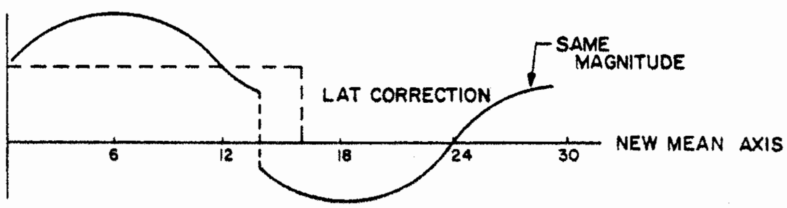


Figure 3-166. Bias Correction

but in a different place. So to completely eliminate latitude error the computer must be given a new latitude to work with and new torquing signals for the gyro drift rate as illustrated in Figure 3-167. Both corrections are necessary to eliminate all error. If latitude error were corrected without correcting bias errors, the 24-hour oscillation could be made to look like Figure 3-168. It would be possible to eliminate the 24-hour oscillation without correcting the bias error provided that the operator could determine where the mean axis is located. That is, in Figure 3-168 he could correct the latitude in the other direction as in Figure 3-169.

3-8.6.7.3 Heading

Heading is interrelated with latitude such that the heading error curve leads the latitude error curve by 6 hours, as illustrated in Figure 3-170. These relationships are not altered by bias errors, but each curve may have a different mean axis, as illustrated in Figure 3-171. Thus, the whole discussion of biasing latitude also applies to HEADING. Since they are dynamically interrelated, both must be considered when making corrections.

3-8.6.7.4 Longitude

As previously stated, an initial misalignment in longitude results in a constant standoff (constant error) (Figure 3-172). However, if the bias is incorrect for longitude, the error will continue to grow at a rate equivalent to the error in bias. This gives a simple ramp function (Figure 3-173.) If a BREAK occurs, the slope will change (Figure 3-174). Making a bias correction will cause a zero slope, and

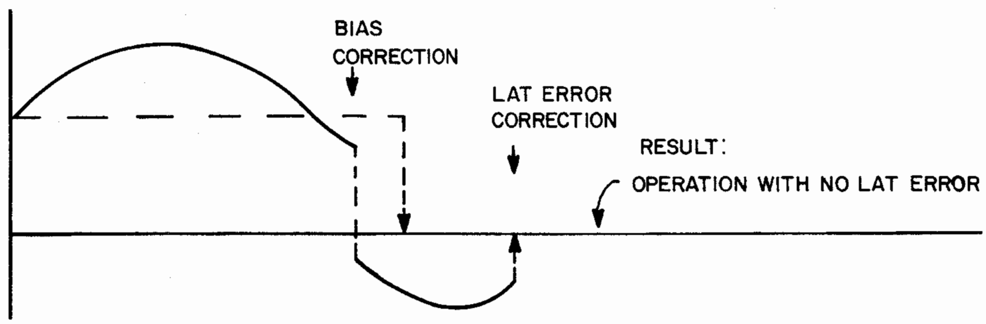


Figure 3-167. Latitude Error After Reset

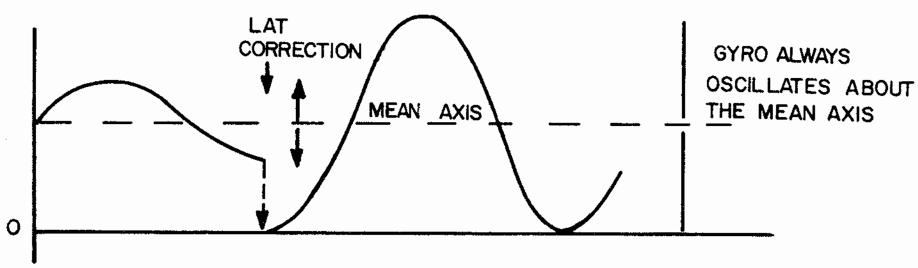


Figure 3-168. Latitude Error With No Bias Correction

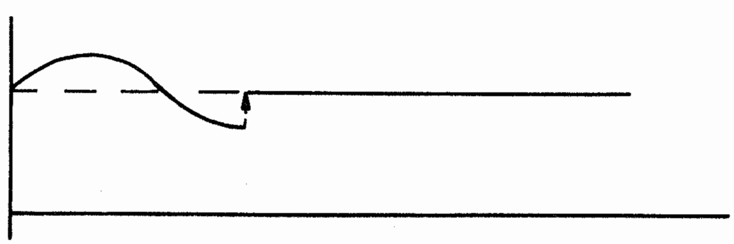


Figure 3-169. Latitude Error Corrected to Bias Mean Axis

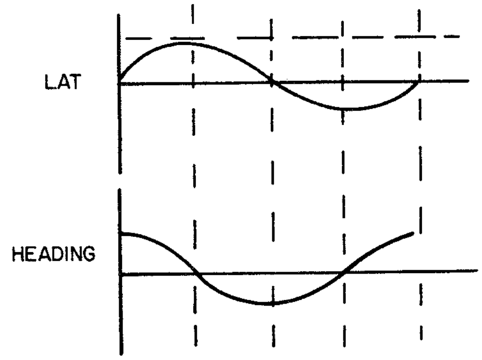


Figure 3-170. Latitude and Heading Error Curve

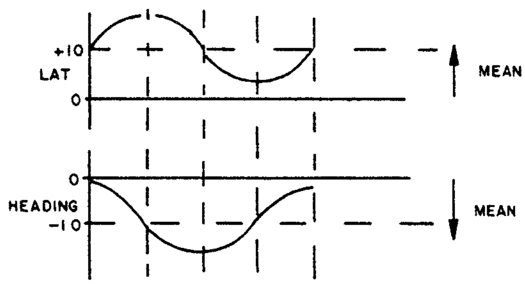


Figure 3-171. Latitude and Heading Mean Axis

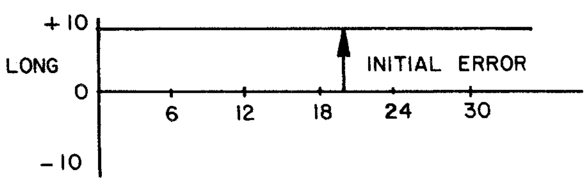


Figure 3-172. Longitude Standoff

whatever longitude error exists at that time will remain constant. The longitude error can then be removed (Figure 3-175). If the gyro is not biased to remove the slope, longitude corrections will only produce temporary results (Figure 3-176).

3-8.6.8 General Scheme of Operation

At sea, after leveling the stable platform in the Alignment Mode, it is necessary to get an accurate fix from some external source such as LORAN-C, NAVSAT, or OMEGA. The proper latitude and longitude can then be set into SINS and the set can be placed in the Navigate Mode. Thereafter, a plot of latitude and longitude error is maintained by plotting fixes about every three to four hours (Figure 3-177). After about 8 hours, the latitude error sine wave can be faired in and a longitude ramp drawn to predict the errors at time of Reset. By correcting the latitude and longitude by the amount of predicted error, and biasing the gyro to bring the mean to zero and thus remove the longitude slope, the SINS should be capable of operating for a long period with no appreciable error.

3-8.6.9 SINS - Detailed Theory

3-8.6.9.1 Gyros

A coordinate system made up of three mutually perpendicular axes is a three-dimensional system that can represent any rotation of the ship. Any particular rotation of the ship can be represented as a single vector, or can be represented as three separate component vectors on the system axis (Figure 3-178). These three component vectors can represent any particular motion in space. The SINS gyroscope, referred to earlier, is actually comprised of three gyros, mounted one each on the three mutually perpendicular axes of a coordinate system. The coordinate is related to the earth, as shown in Figure 3-178. The axes are lettered X, Y, Z. Since the Z axis points straight down (toward the center of the earth), it follows that X and Y axes must lie in the horizontal plane (Figure 3-179). If these three gyros are so arranged that they sense only rotation about their respective coordinate axis, the three gyros together can sense any rotation in space that the platform should make (Figure 3-180). The expanded view of this gyro, shown in Figure 3-181, is a single gimbal arrangement - it is pivoted in only one axis. Since it can turn only about the one single axis (in this case the verticle line through the base) it is termed a "single degree of freedom." To briefly review gyroscope physics, the two principles that make gyroscopes useful are, 1) Rigidity in Space, and 2) Precession. When a gyro like that shown in Figure 3-181 is "spun up" to a very high speed, it tends to continue to point to the same spot in space

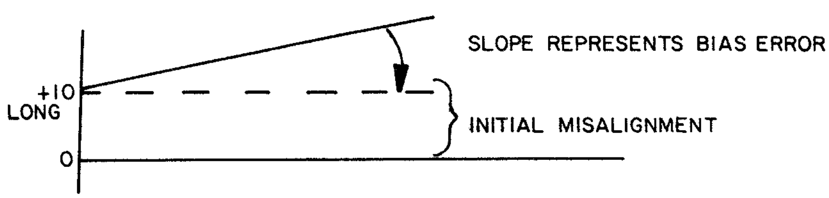


Figure 3-173. Incorrect Bias Ramp Function

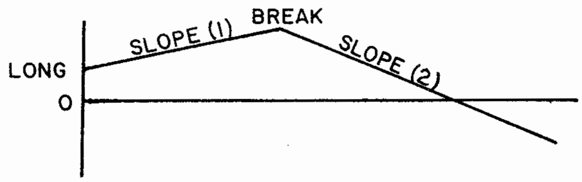


Figure 3-174. Longitude Break

(it could be thought of as pointing to a distant star). In freshman physics, a hand rule served to determine which way it would precess if a force tried to change its orientation. In other words, the gyro precesses in a direction 90° (in the direction of spin) from the direction force is being applied. This is shown in Figure 3-182. The three mutually perpendicular axes are shown through the gyro. One is the spin axis, about which the gyro rotor spins. Another is the sensitive axis. 1) If the ship moves about this axis (rotation), it moves the Base as shown. 2) This puts a couple force on the spin axis, pushing up on the right, pulling down on the left. 3) Looking at the top of the spinning wheel, each point has two forces acting on it, one is tangential in the direction of spin, the other at right

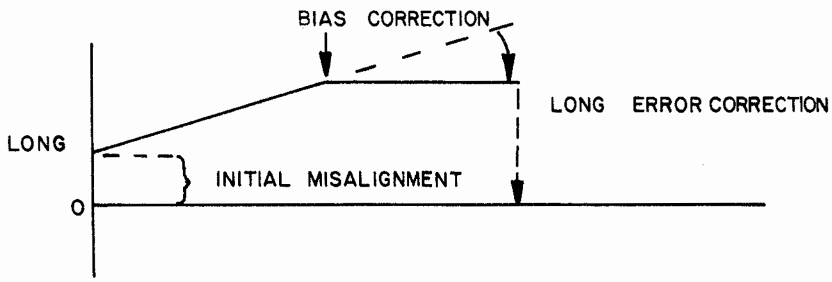


Figure 3-175. Eliminating Longitude Error

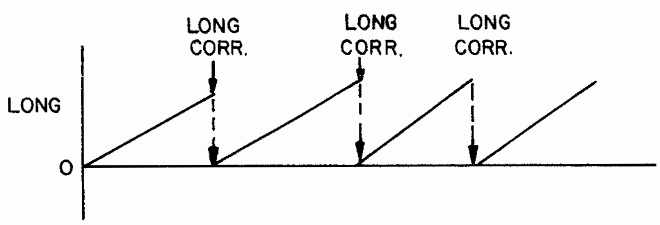


Figure 3-176. Correcting Longitude Error Without Correcting Bias

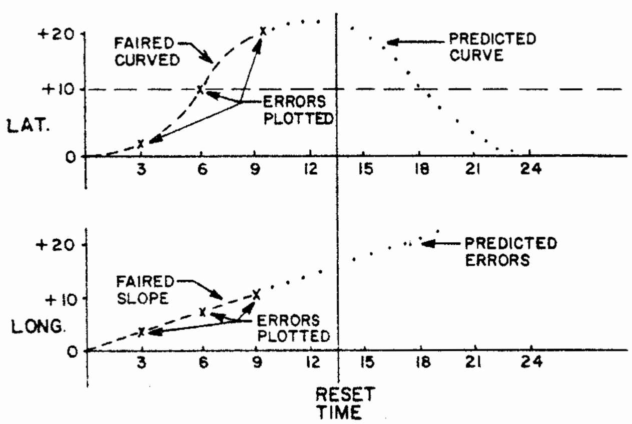


Figure 3-177. Plotted Errors

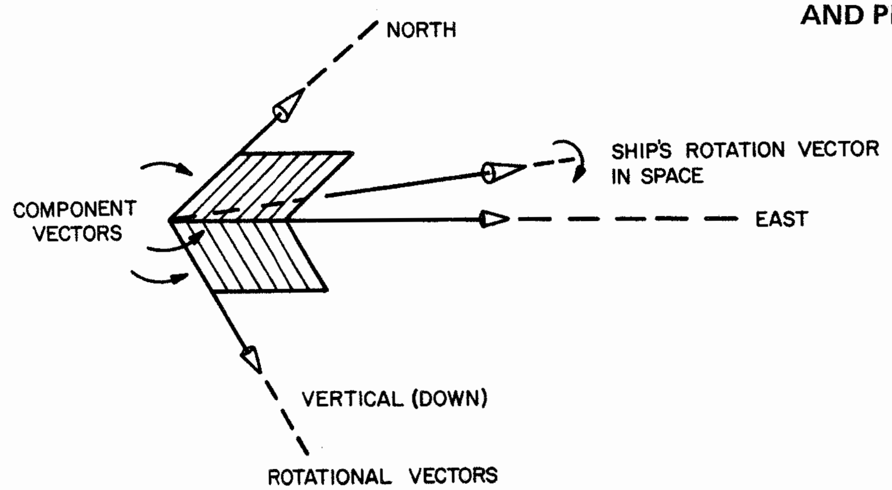


Figure 3-178. Rotational Vectors

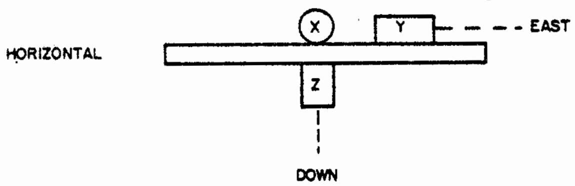


Figure 3-179. SINS 3-Gyro Arrangement

angles caused by the couple force of 2 above. The resultant of these two vectors is shown. 4) In order for each partial of the wheel to follow that resultant vector, the gyro must rotate about the base (the precession axis). If the ship rolled back again, the base would move in the opposite direction, the couple would reverse, and the resultant vector would be to the right. Thus, the gyro would precess in the opposite direction. Therefore, each rotation about the sensitive axis causes a precession in direct relation to it. Note that if the ship rotates about the spin axis there is no

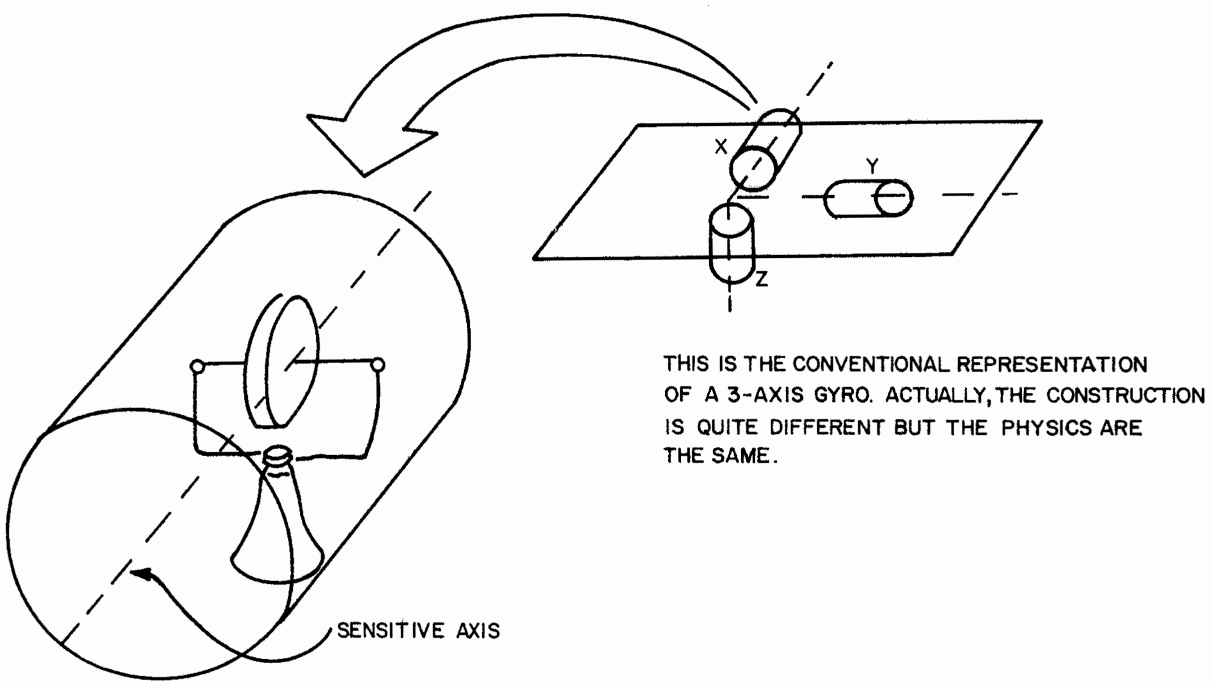


Figure 3-180. Gyro Axis

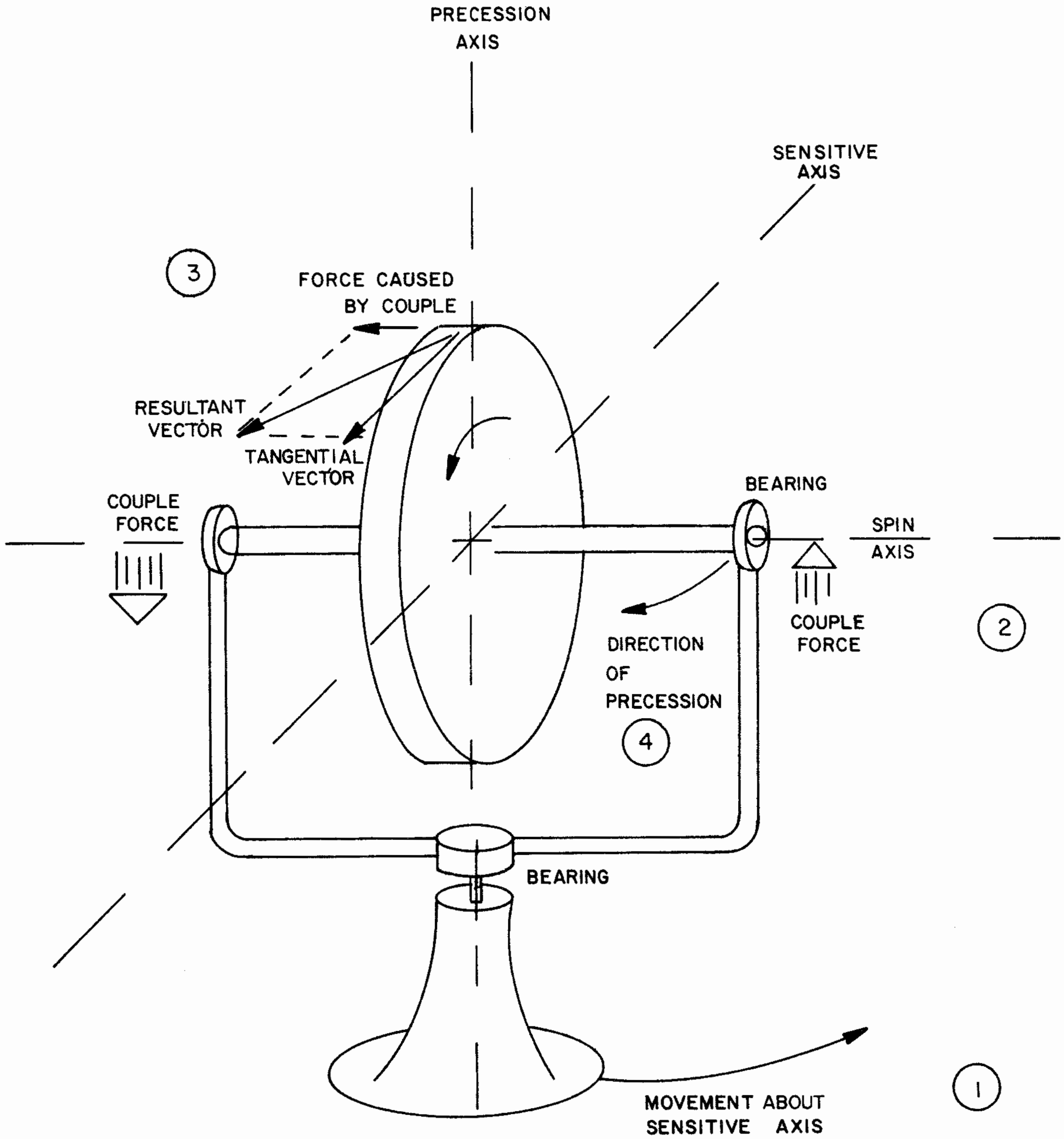


Figure 3-181. Gyro Expanded View

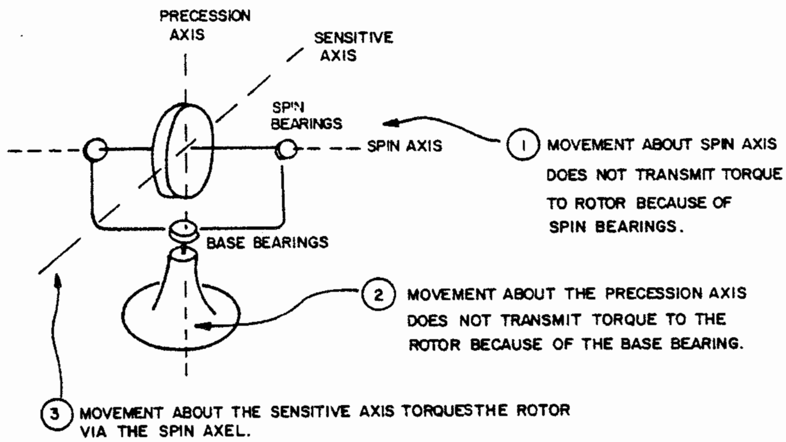


Figure 3-182. Gyro Perpendicular Axes

couple force applied to the gyro, therefore, no precession. Likewise, a rotation about the precession axis cannot put a couple force on the gyro. Therefore any precession represents only a rotation about the sensitive axis, and no other axis. Because any rotation of the ship can be represented by component rotations about the three axes, only that component part of the rotation that is about the sensitive axis is represented by a precession. Thus, three of these gyros, mounted so that their sensitive axes are mutually perpendicular, will sense their component of any complex rotation of the ship (and therefore the platform) and represent that component with precession. It is simple to incorporate an electronic pickoff that detects the precession and puts out a signal. These signals, fed to the Platform Electronics Cabinet, are turned into followup motor currents which cause the platform to be returned to its original alignment in space, thus nullifying the ship's rotations. Simplified drawings will show only a gyro case with the sensitive axis as portrayed in Figure 3-183.

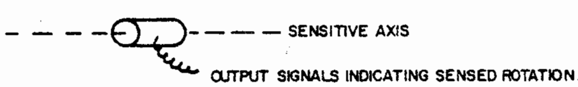


Figure 3-183. Simplified Gyro Drawing

3-8.6.9.2 Platform Followup

There have actually been two coordinate systems discussed. The platform coordinates just discussed were X, Y, Z - with the X axis pointing north, and Y axis pointing east, and the Z axis pointing straight down. The other set mentioned was Roll, Pitch and Yaw, the ship coordinate. These two systems rarely coincide because the ship is constantly in motion. Coincidence would occur when the ship was heading true north and had zero roll and zero pitch (Figure 3-184). Connecting the two systems are three gimbals which are geared together. These gimbals change their relation to each other only when the torque motor changes them on command from Platform Electronics as followup signals. Therefore, when the ship rolls, the platform tends to roll until the followup counters the movement. But the gyros sense all rotations in terms of X, Y, and Z, so there must be a way of relating the two systems. This is done with precision Resolvers between each gimbal (Figure 3-185). These Resolvers measure the angles between the gimbals and electrically divide an X axis rotation up into components of Roll, Pitch and Yaw. This cycle is rapid enough to actually prevent the platform leaving the horizontal so that the accelerometers will not sense gravity from normal ship's rotation as explained in Figure 3-186. When properly aligned to earth coordinates (latitude, longitude, heading) the SINS maintains its alignment only by sensing every ship's rotation and correcting for it immediately.

3-8.6.9.3 Compensating for Spherical Earth

Now consider the problem of moving the ship over the curved surface of the earth. As discussed in paragraph 3-8.6.9.2, movement over a curved surface, either N-S or E-W, actually resulted in rotation (see Figure 3-154 and 3-155). Considering the idea, again using the gyro arrangement discussed above, and referring to Figure 3-187, note the E-W

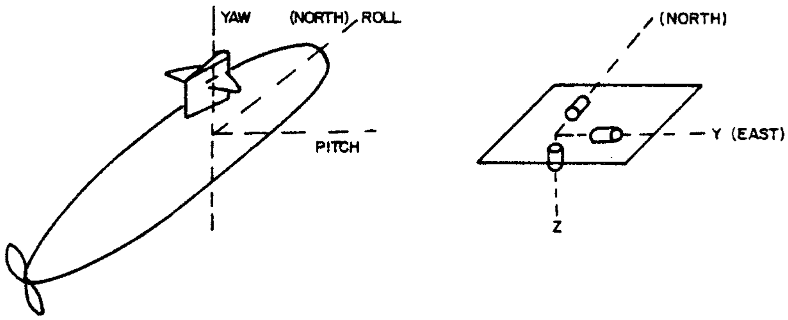


Figure 3-184. Ship Versus Platform Coordinates

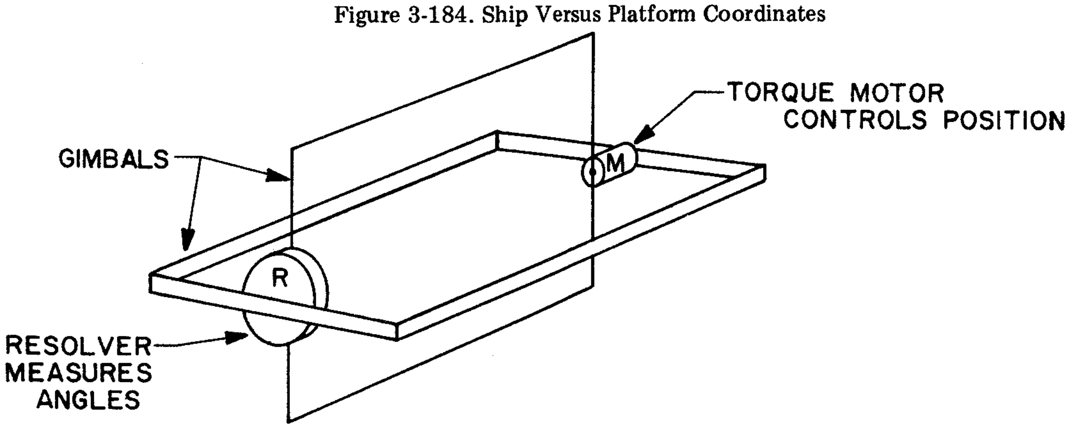


Figure 3-185. Gimbal and Torque Motor Arrangement

movement is sensed by the X gyro, and N-S movement is sensed by the Y gyro. First, looking up from the south pole, observe an easterly movement. While the ship traveled east, the X gyro sensed a rotation about its sensitive axis. The platform electronics followed up to maintain the platform orientation in space. The gyros sensed a rotation: the followup system corrected for the rotation. However, when the ship moved east, the east-pointing accelerometer sensed a movement east and transmitted data to computer. Thence, the computer used the velocity information to compute the change in longitude. Then an additional torque was generated for the platform about the X axis in exactly the right amount to obtain a new computed horizontal (Figure 3-188). The same analysis applies to N-S movement (Figure 3-189). Thus, an additional torque is applied to the platform about the Y axis to compensate for distance moved N-S, i.e., the change in latitude. The platform always remains horizontal to local vertical, accomodating changes in latitude and longitude by torquing to the platform. These compensating torques are generated for the X and Y axis. From the discussion of the followup system converting X, Y, Z coordinates to Roll, Pitch and Yaw, it can be seen that it is more convenient to send pulses to the X and Y gyros. These will cause the gyros to put out false signals of the right magnitude to make the followup system orient the platform to the new latitude and longitude.

3-8.6.9.4 Oscillations

The part of SINS theory of principal concern from day to day is platform dynamics. One of the basic shortcomings of the SINS is the tendancy to develop an undamped 84.4 minute oscillation. This oscillation will result from any disturbance that may cause the platform to be tilted from the horizontal momentarily. How far off horizontal the platform is tilted will determine the magnitude of the oscillation, but it will continue indefinitely. These disturbances could result from entering Resets, from switching to Navigate before settling the platform completely, from local disturbances in gravity, or from many other reasons. When the platform is tilted by disturbances, it is no longer horizontal (see Figure 3-190). Gravity is now pulling on the accelerometer so that it is displaced to the left. This is the same displacement that

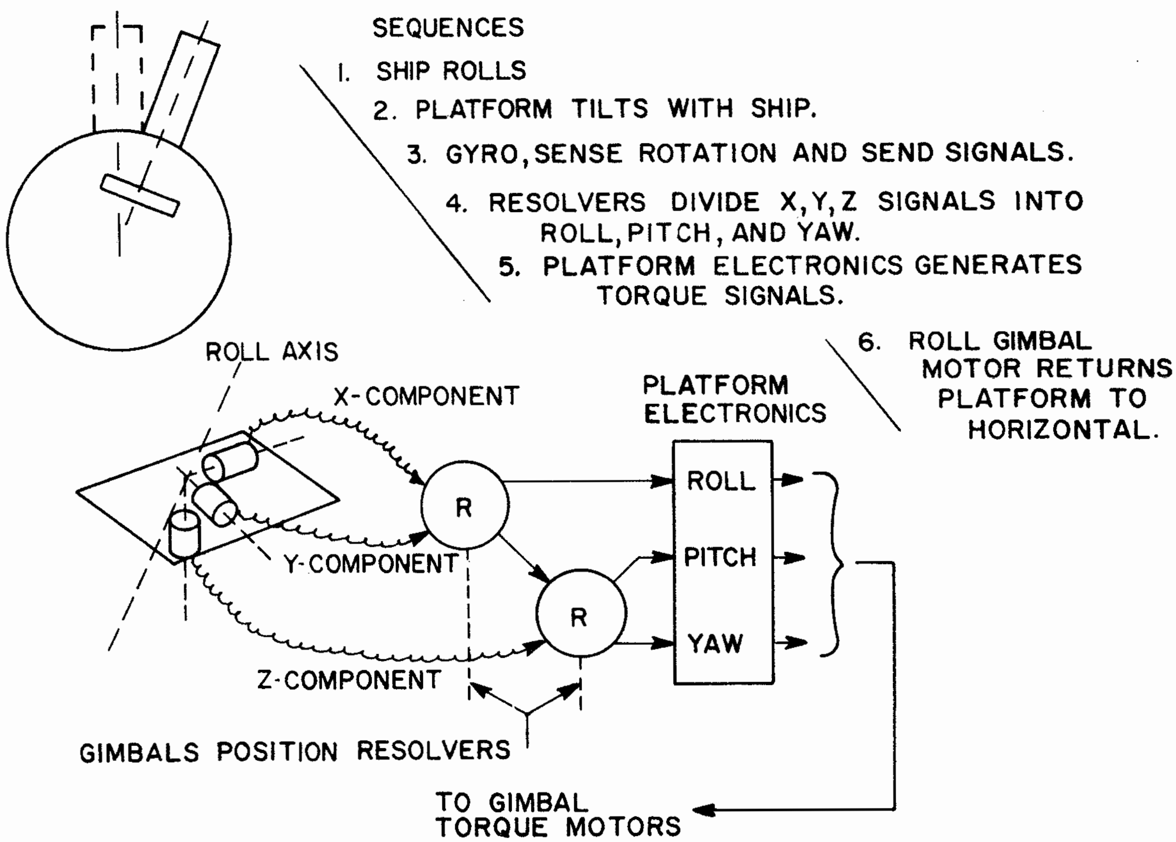


Figure 3-186. Platform Sensing Rotation

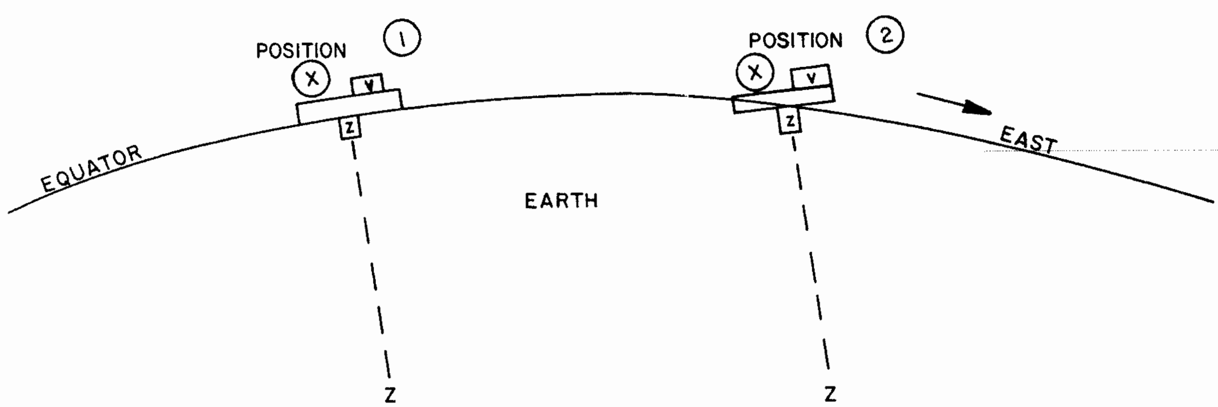


Figure 3-187. East Movement Looking From the South Pole Without Y Axis Compensating Torque

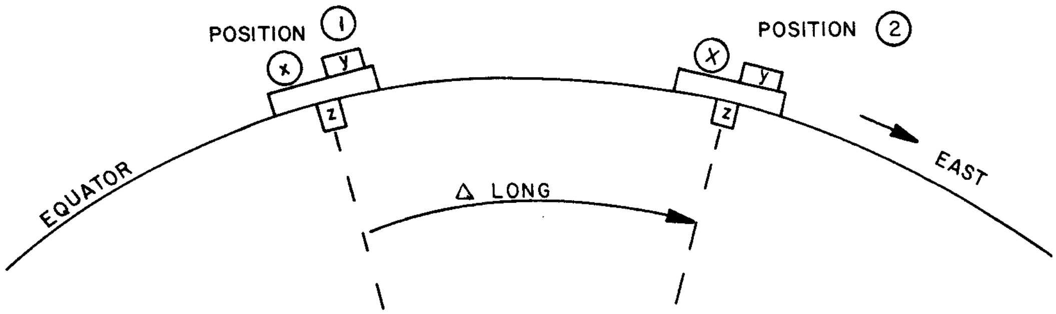


Figure 3-188. East Movement Looking From the South Pole with Y Axis Compensating Torque

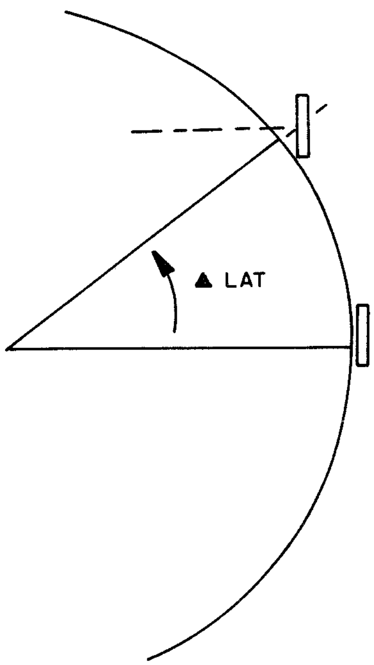


Figure 3-189. North South Movement

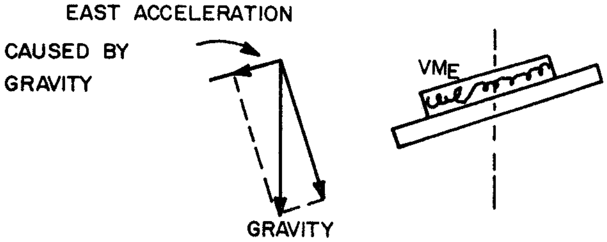


Figure 3-190. Disturbed Platform

would be caused by an easterly acceleration. The electrical sensing circuits therefore take this to be an acceleration and send the information to the computer. One of the jobs of the computer is to compute the new longitude for this sensed easterly movement and to compute a new horizontal. Then it sends out signals to orient the platform to the newly computed horizontal which tends to move the platform toward the actual horizontal. When the accelerometer finally reaches the horizontal and zero acceleration, the computer is still using the easterly velocity that it built up from the easterly acceleration. There has been no deceleration. Consequently, the computer continues to reorient the platform to move easterly longitudes until the platform is tipped in the opposite direction (Figure 3-191). In this position, however, the tipped platform now produces acceleration in the opposite direction, or deceleration. This stops further tipping, but the existing tip produces a westerly acceleration. Now the whole process repeats itself in the other direction until the platform is tipping the other way again, that is, the original tilt that started the oscillation. An undamped, 84.4 minute oscillation in longitude results (Figure 3-192). If this happened at sea while the ship was going in an easterly direction, the path might look something like Figure 3-193. Exactly the same problem occurs with the north-south accelerometer, so a northerly transit with an oscillation would give the same result (Figure 3-194): comparing the Schuler oscillations with the 24 hour oscillation previously discussed, it can be seen that they are easily differentiated (Figure 3-195). Damping an oscillation such as this results in an exponential reduction in magnitude so that it will all but disappear in one and one-half cycles (Figure 3-196). The problem in damping is to find a reference that will damp the oscillations but not damp out legitimate changes in speed. The best source of such a reference was taken as the EM log because it reflected changes in speed.

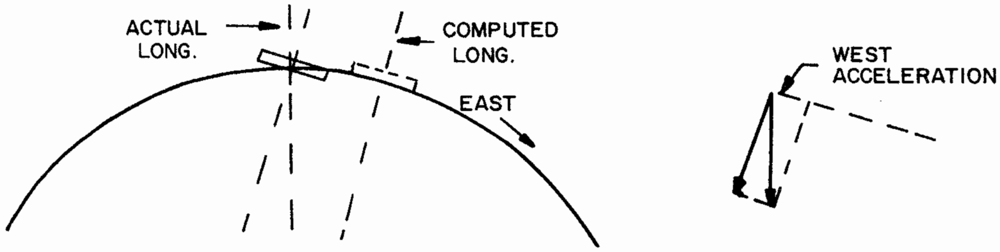


Figure 3-191. Disturbed Platform

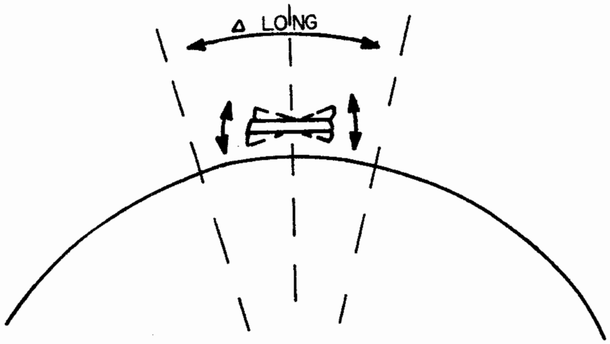


Figure 3-192. Schuler Oscillating Platform

By comparing the SINS velocity and the EM log velocity, the difference is used as an error signal. Only 20% of that signal is used in order that EM Log fluctuations do not cause serious disturbances. This signal is entered into the system as a false acceleration that tends to counteract the oscillation described above (Figure 3-197).

3-8.6.9.5 Damping Signal Error

If a constant error exists in the EM Log because of poor calibration of the presence of a current, a small but constant damping signal will result. Since this signal is put into SINS as a false acceleration, the platform will assume a small but constant tilt such that the offset of the accelerometer will just equal the damping acceleration. This will result in a very small position error. The error in position is considered preferable to the 84 minute oscillation. Using the EM Log, a damping signal is developed only when one velocity changes with respect to the other (Figure 3-198).

3-8.6.9.6 Rotating Earth and the 24 Hour Oscillation

Rotating earth and the 24 hour oscillation are the heart of the operating SINS. Almost all the problems and techniques of operation are concerned with the platform interaction with earthrate, i.e., the rate at which earth turns in space. To introduce these effects, consider a classical gyroscope such as the ship's gyro. It has gimbals so that it is free to rotate in any direction. As stated in the previous discussion of gyros, (see Figure 3-199) the most important property of gyroscopic action is rigidity in space. If the gyro were pointed at a distant star, no matter where it was taken on earth, it would continue to point at that star (Figure 3-199). As shown in Figure 3-200, if standing with this gyroscope at a fixed position on the equator, the gyro would appear to do a complete 360° turn in 24 hours (Figure 201). However, if the gyroscope were oriented so that the spin axis was parallel to the Earth spin axis, no change would be apparent (Figure 3-102). If the gyro were slightly misaligned with the earth's polar axis, as the earth turned, the gyro would appear to gyrate slightly in a circle (Figure 3-203), but still maintains its rigidity in space. Misaligned slightly to the left at position 1, after 6 hours in position 2 it would appear to point slightly into the earth. Twelve hours later, in position 3, it would appear to be misaligned to the right. In position 4, it would appear to be tilted away

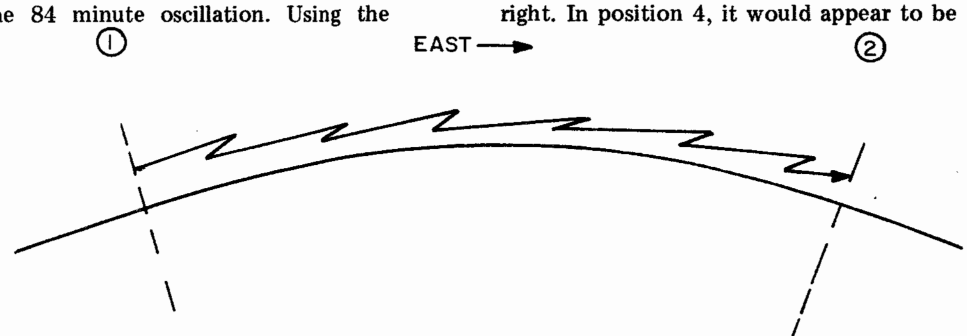


Figure 3-193. Result of Schuler Oscillation While Traveling East

These are called Schuler Oscillations. It is necessary to provide some system to damp out these oscillations.

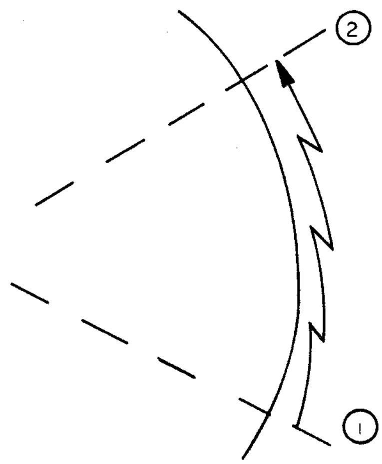


Figure 3-194. North-South Schuler Oscillation

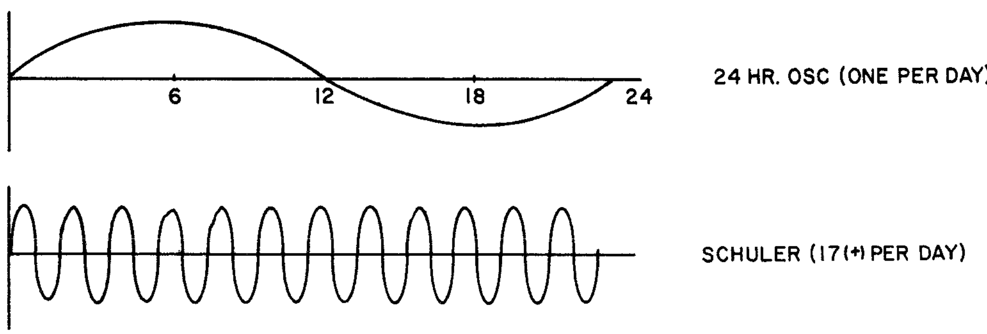


Figure 3-195. Schuler and 24 Hour Oscillations

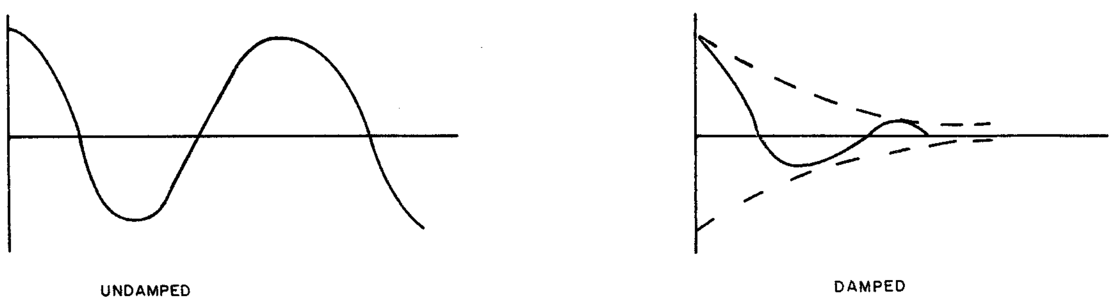


Figure 3-196. Damping

from the earth; and finally, after 24 hours, it would again be misaligned to the left. The tip of the spin axis for 24 hours would thus appear in Figure 3-204. This gyro was free to turn right or left, and into and away from the earth. That is, it had two degrees of freedom or two sensitive axes. As shown in Figure 3-205, the same effect is obtained with two gyros having single sensitive axes, such as the SINS gyro. Two axes crossing at right angles establish a plane, and their mutual perpendicular can be thought of as the polar axis. As the earth rotated, it would look like that shown in Figure 3-206. However, a constantly

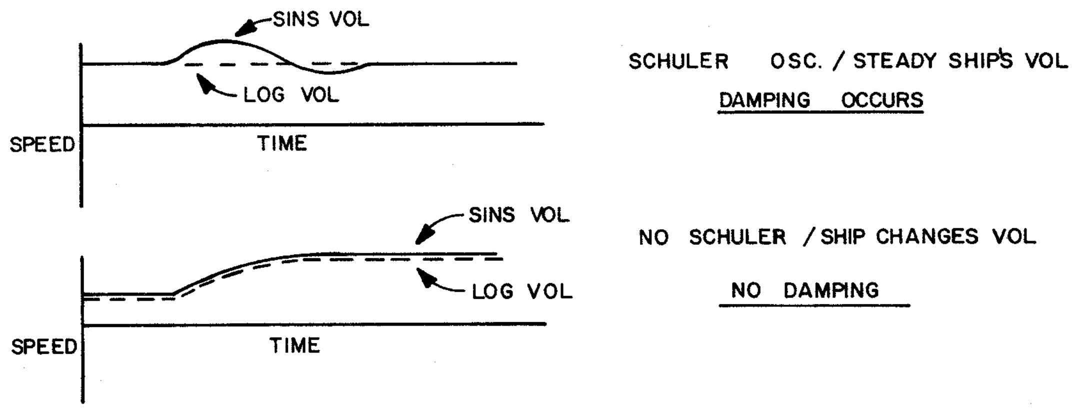
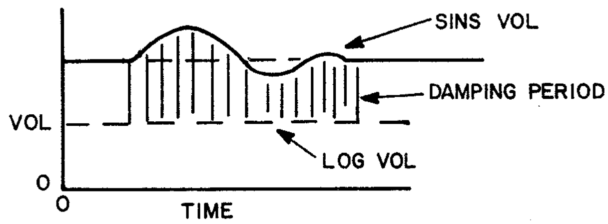


Figure 3-197. Second Order Damping



1. Damping occurs only while velocity DIFFERENCE is changing.
2. SINS Vel is still most accurate.
3. Constant Log error has NO EFFECT.

Figure 3-198. 3rd Order Damping

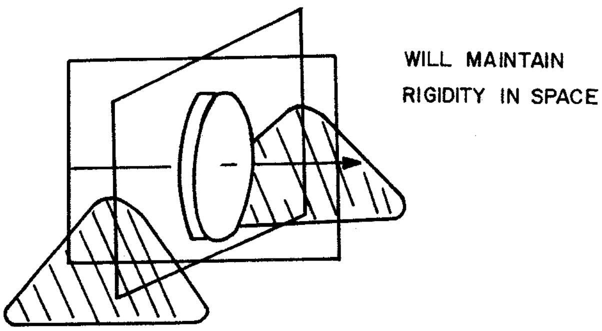


Figure 3-199. Gyro Gimbals

turning gyro will not serve this illustration because the observer is not interested in where he is in space, but rather as to where he is on earth. This can be ascertained by torquing the system shown about the polar axis at the same rate that the earth is turning. Now the 24 hour period would look like that shown in Figure 3-207. Note that the Y axis always points East. The axis labeled E is called the equatorial axis because it always lies in a plane parallel to the plane of the equator (Figure 3-208). This configuration proves to be a very convenient way of working with and understanding error analysis. Some SINS systems

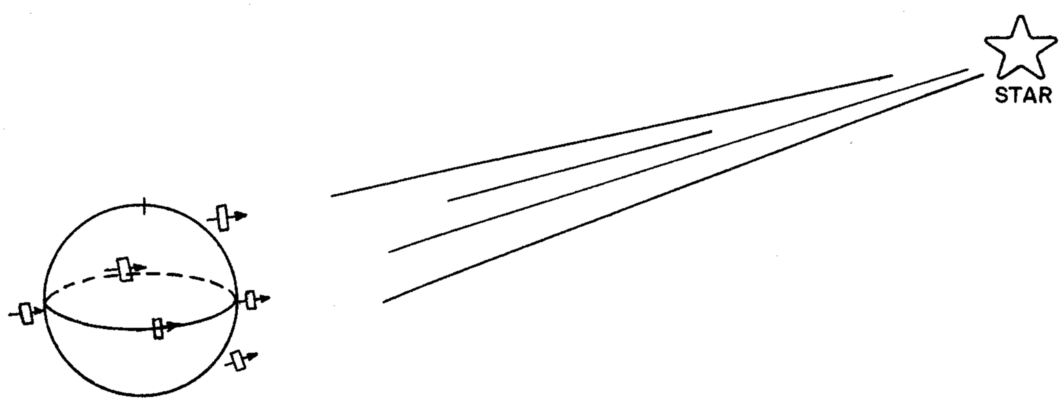


Figure 3-200. Space Oriented Gyro

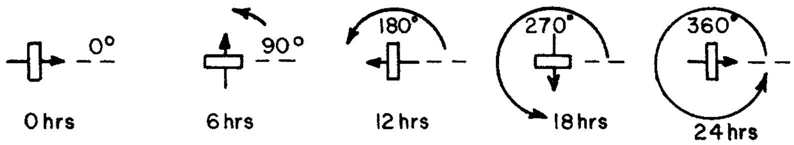


Figure 3-201. Gyro Rotation in a 24 Hour Period

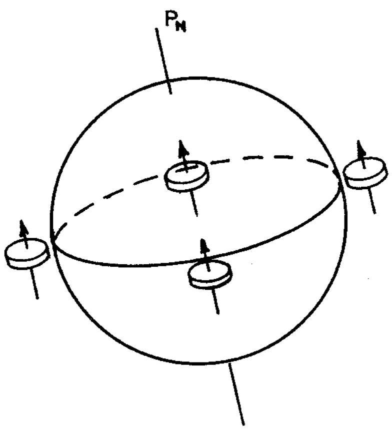


Figure 3-202. Earth Referenced Gyro

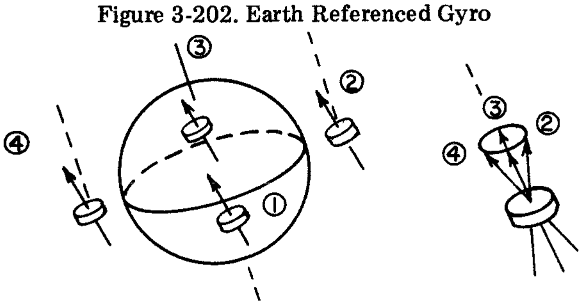


Figure 3-203. Misaligned Gyro

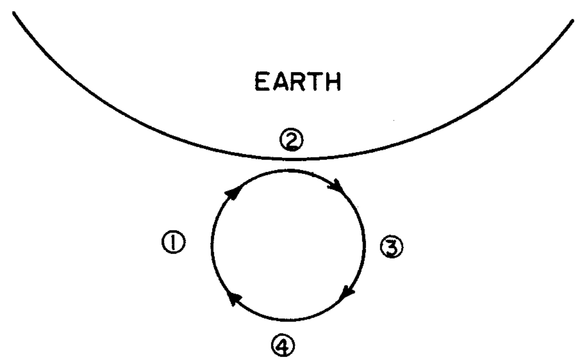


Figure 3-204. Misaligned Gyro Showing Spin Axis For 24 Hour Period

use this arrangement for the three gyros. Referring again to Figure 3-203 and 3-204, it was noted that when the polar axis is not perfectly aligned (parallel) to the earth spin axis, it still maintains its rigidity in space. Standing on earth and watching the polar gyro's axis, it would appear to describe a circle. Actually, it is holding a steady direction and the ship is rotating in a circle as the earth rotates in space. Consider this again while comparing the text with Figure 3-209. The polar axis is misaligned and the projection of the point of the arrow on the equatorial plane described a 24 hour circle. Note the action of the Y and E gyros. At time zero, the E axis is not properly aligned in the equatorial plane and this is the cause of the

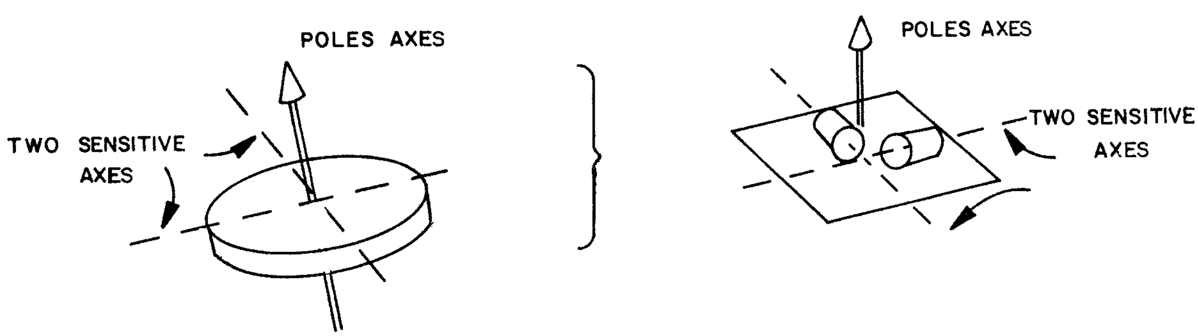


Figure 3-205. Defining the Polar Axis

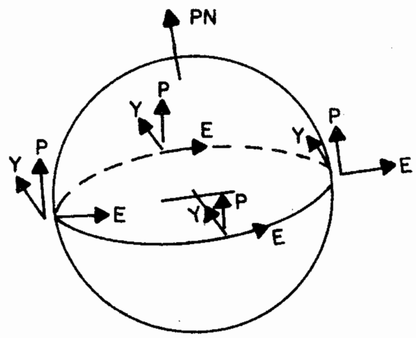


Figure 3-206. Polar Axis in Relation to the Earth

polar axis misalignment. As the earth revolves 1/4 turn in 6 hours, the system maintains its rigidity in space, but the system is torqued with earthrate so that the E axis points out and the Y axis points easterly. The result can be seen. The E axis is now in the equatorial plane where it should be, but the Y axis is now below the plane. Continuing to rotate, at (+) 12 hours, the Y axis has come back to the equatorial plane but the E axis is now below it. At (+) 18 hours, the E axis came back up to the plane, but Y has now moved above it.

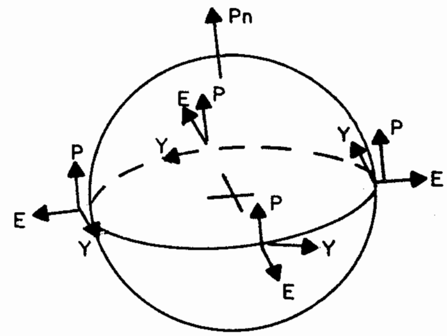


Figure 3-207. Polar Axis Torquing

Finally, after 24 hours (one revolution) the system has returned to the original configuration. The apparent motion over a 24 hour period is shown in Figure 3-210. If the sine wave of the Y and E axes were plotted, it could be seen that they are the driving function that produces a circle (the polar axis) as shown in Figure 3-211.

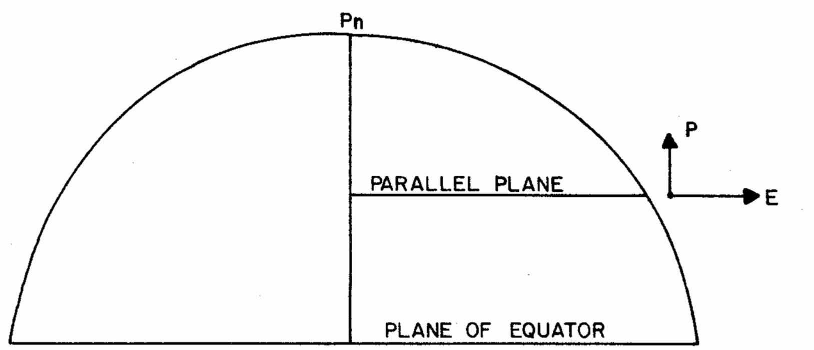


Figure 3-208. Defining the Equatorial Axis

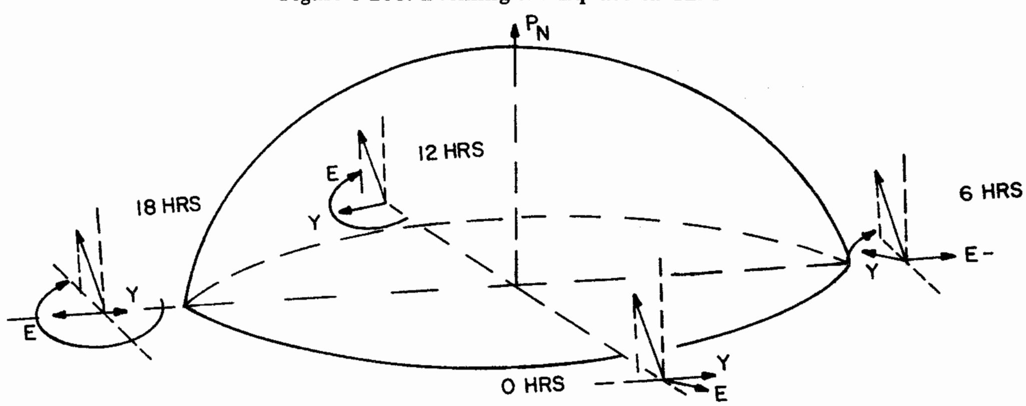


Figure 3-209. Polar Axis in Relation to Space

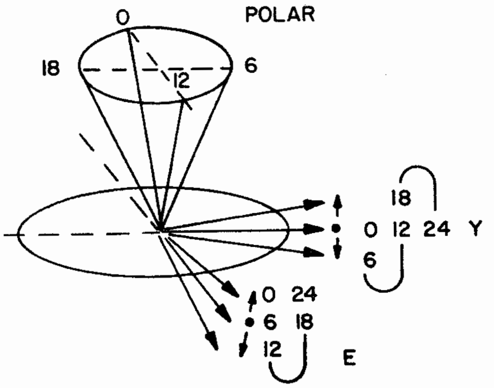


Figure 3-210. 24 Hour Rotation of Polar Axis

Points on this circle could be thought of as the top of the polar axis. If the tip of this axis were seized and moved back to the center of the circle, it would then be properly aligned with the earth spin axis. Then no more oscillation could develop. This same effect could be accomplished by moving the Y and E axes back to the equatorial plane. If both these axes are in the plane, then the polar axis must be parallel with the earth axis. This is actually what is done when SINS is reset. Remember, as the ship moves N-S on the earth surface, it is actually rotating in space. Consider the three gyro system in Figure 3-212. The axis sensing this rotation (into the paper, away from us) is the Y axis. Therefore, the Y gyro senses change in latitude. Again referring to Figure 3-212, a change in longitude is a rotation also. This time it is about the polar axis. Therefore, the P gyro senses change in longitude. The polar axis, when parallel to the earth spin axis, points at a distant star known as the North Star. Then if the system rotated slightly about the E axis, the polar axis would point to the right or left of north. Therefore, the E gyro senses change in heading. Also note from Figure 3-213 that a rotation about the E axis causes a misalignment of the Y axis. Therefore, if the Y axis is misaligned, there has been a rotation about the E axis and there is a heading error. Similarly, a misalignment of the E axis means a rotation about the Y axis and a latitude error. Shown in Figure 3-210, as the E axis oscillates up and down, a corresponding latitude error exists (rotation about Y). As the Y axis oscillates up and down, a corresponding heading error exists. In Figure 3-211, the misalignments of Y and E axes were plotted. It can be seen that the Y axis plot corresponds to heading error and the E axis plot corresponds to latitude error. To show these plots together (mindful of signs) refer to Figure 3-214. Compare these curves with a 24 hour period (6 to 30) in Figure 3-211. These are characteristic curves. All of this is summed up in Figures 3-209 and 3-210.

Y Gyro	SENSES	CHANGE IN	LATITUDE
Y Gyro	MISALIGNMENT	CAUSES	HEADING ERROR
E Gyro	SENSES	CHANGE IN	HEADING
E Gyro	MISALIGNMENT	CAUSES	LATITUDE ERROR

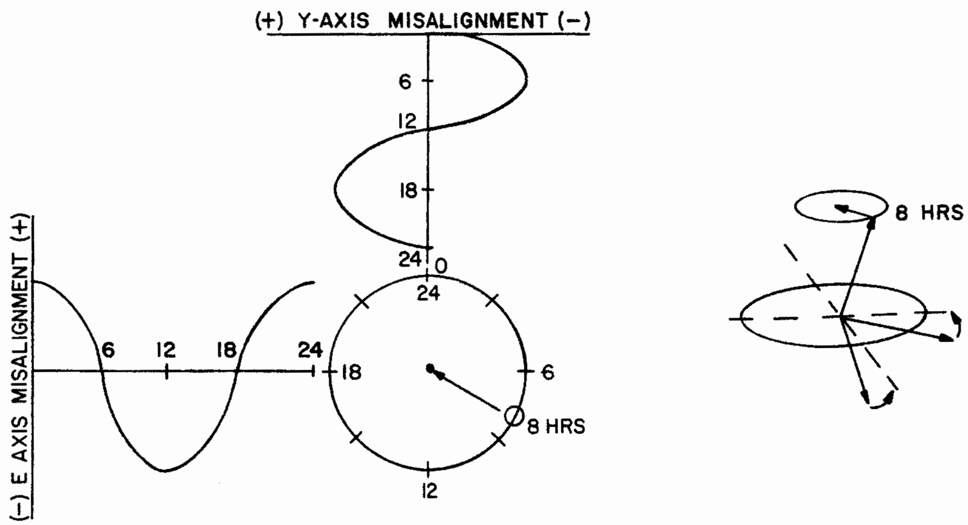


Figure 3-211. Y and E Axis Plots

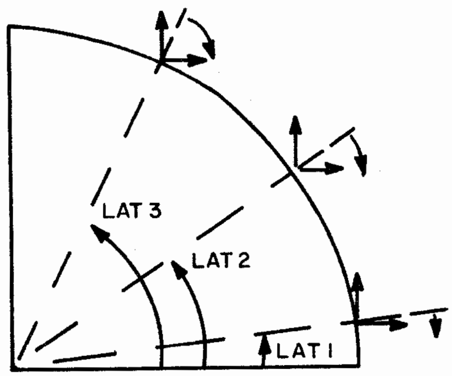


Figure 3-212. 3 Gyro System

3-8.6.9.7 Error Prediction

Figure 3-159 shows that if the latitude error curve is known, the heading error would also be known. Latitude error is E gyro misalignment and heading error is Y gyro misalignment (Figures 3-209 and 3-210). In Figure 3-209, the operator might reset latitude to zero error from a good NAVAIDS fix at time (+) 6 (E axis alignment as shown) but no heading fix was obtained. At time (+) 12, another NAVAIDS fix shows latitude is again in error (E axis misaligned as shown). What can now be said about the heading error (Y gyro alignment) at times (+) 6 and (+) 12? The only logical answer is that the polar axis was tilted as shown so that the Y axis was misaligned (negative) and heading had a negative error at time (+) 6. Therefore, heading error must be zero at time (+) 12 in order to maintain the polar axis rigid in space. The

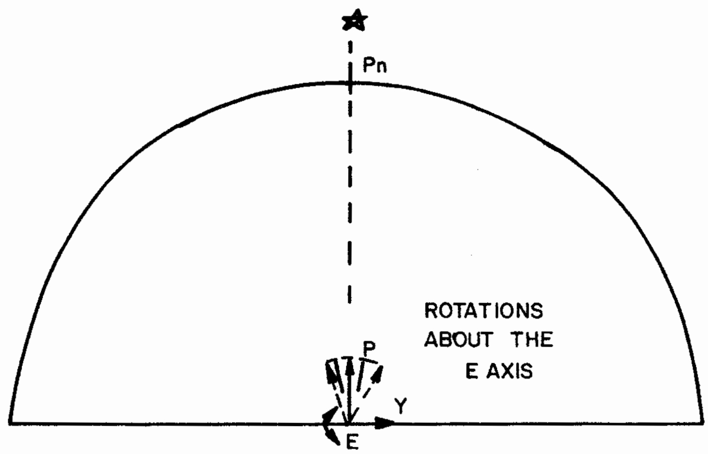


Figure 3-213. Y Axis Misalignment

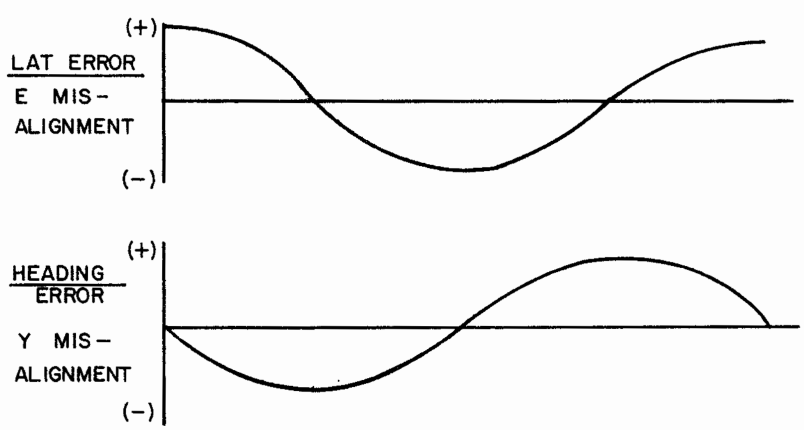


Figure 3-214. Y and E Axis Misalignments

heading error leads the latitude error. In other words, whatever the heading error is now, the latitude error will be 6 hours from now. The same logic can be used backwards: knowing what the latitude error is now, the heading error was that value 6 hours ago (provided of course, there is any significant misalignment).

3-8.6.9.8 Platform Dynamics

In the discussion of platform dynamics, or how the platform actually accomplishes what has been discussed, one of the most important concepts is introduced: that of the effects of earth rate on the gyros. Consider a properly aligned gyro system sitting on the equator: the polar axis is exactly parallel with the earth spin axis. Because this polar axis senses the earth turning in space, this must be countered by torquing the polar gyro at a rate equal to earth rate. This keeps the Y gyro pointing east (Figure 3-215). Since all of the earth is turning about its axis, only the polar gyro senses it. The Y and E gyros sense no earth rate: when properly aligned, Y and E gyros do not sense earth rate. Now consider what happens when they are not aligned properly so that there is a latitude or heading error (Figure 3-209). Consider the case of a heading error (Y gyro misalignment) being set into the SINS (Figure 3-215). Although the Y gyro senses no rotation when aligned in the equatorial plane, when it is above the plane it has a component vector parallel to the polar axis. The polar axis does sense earth rotation, so now the Y gyro will sense some small component of earth rotation. Because the gyro senses rotation, it puts out a signal and the platform electronics cause an immediate followup counter rotation and tilts the platform to the south. Gravity now acts on the accelerometer which puts out a north acceleration. The computer in turn, receives the north acceleration and computes a new latitude to the north, and thus a new horizontal. Since the computer does not know that the ship did not move north, it torques the platform to a new computed horizontal (Figure 3-216). This, in effect, is tipping the platform back toward the north, cancelling out the effect of the earth rate. Nevertheless, the Y gyro is still sensing earth rate, so the process is repeated again and again. These small cycles continue

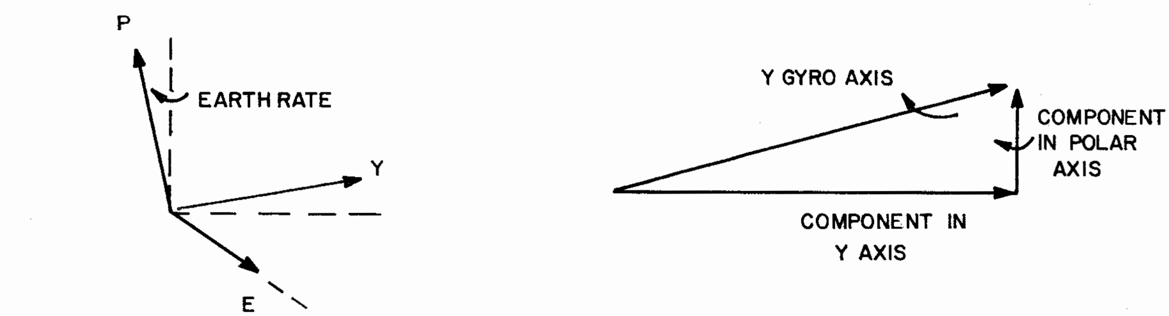


Figure 3-215. Heading Error

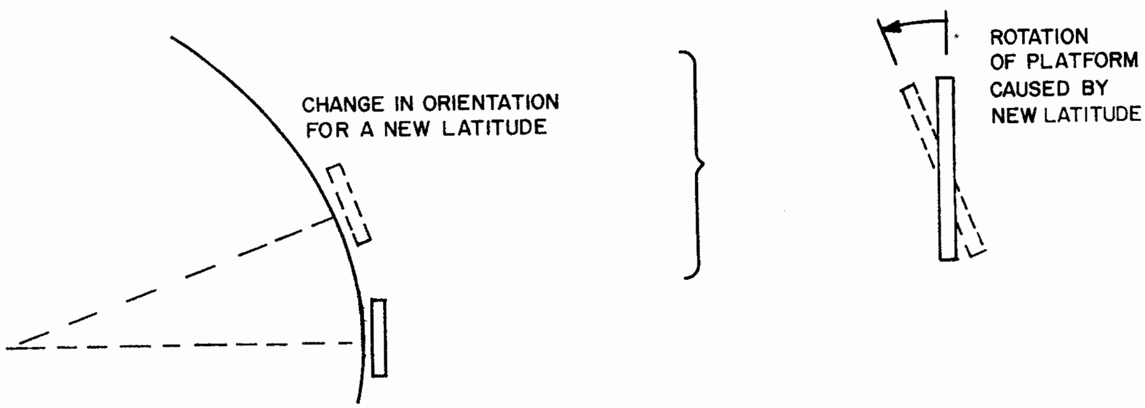


Figure 3-216. Latitude Change

to result in latitude errors further to the north, the result of torquing about the Y gyro. But latitude error is E gyro axis misalignment. This means that as the latitude error builds up, the E gyro moves about the equatorial plane (Figure 3-217). Now the E gyro has a component in the polar axis and it begins to sense a small component of earth rate. Since the earth rate is a sensed rotation, the platform followup system torques the platform to counter the sensed rotation. As shown in Figure 3-217, this tends to drive the Y gyro back toward the equatorial plane. As the Y gryo slowly approaches the equatorial plane, it senses smaller and smaller components of earth rate so that the rate of latitude error build-up (and consequently, the rate that the E gyro moves away from the E axis) reduces. After 6 hours, when the Y gyro has been driven back to the equatorial plane, it no longer senses earthrate and the latitude error ceases to build up (Figure 3-218). However, at this point, the E gyro is sensing a large component of earth rate, so that it continues to drive the Y gyro down. As the Y gyro goes below the equatorial plane, it begins to sense earth rate again, but in the opposite direction (Figure 3-219). This, of course, results in a build-up of latitude toward the south, reversing the process described before. Now the E gyro begins to move toward the equatorial plane. As it approaches the equatorial plane, it senses less and less earth rate until, when in the plane 6 hours later, it senses no earth rate and ceases to drive the Y gryo down (Figure 3-220). Again, the Y gyro senses maximum earthrate and continues to build up latitude to the south, driving the E gyro down. This process continues with nothing to stop it. It is an undamped 24 hour oscillation (Figure 3-221). Arranging these curves in the usual order, the heading error leads the latitude error by 6 hours (Figure 3-222). Note these points:

1. The effect of earth rate is to drive the misaligned gyro back into alignment, at the same time driving the other out of alignment.
2. The gyro will oscillate the same amount on each side of the zero axis.

Longitude misalignment is a much more simple matter. If something caused the platform to become tilted to the west, the E-W accelerometer would sense gravity and indicate easterly movement. Based on the easterly movement signal, the computer would count up a new longitude to the east and torque the platform to a new calculated horizontal. This tends to remove the tilt (Figure 3-223). When the longitude has moved far enough east, the platform will be horizontal and no further tilt will occur. The error will remain constant (Figure 3-224).

3-8.6.10 Gyro Drift

A gyro rotor is suspended by bearings (Figure 3-225). No matter what the actual shape of the rotor, and no matter what type of bearing medium is used, there will be some drag opposing the turning. This and other effects cause any gyro to precess. In the discussion of gyro biasing, it was noted that every gyro has its own particular drift factor which must be entered into the computer. The computer in turn constantly torques the gyro the exact amount to counter this drift. The effect of drift is to cause the gyro to sense a rotation that is not there. Thus, the gyro will put out a false signal. As long as the computer-set drift factor for that gyro continues to exactly counter the gyro drift, there will be no errors developed. Because the

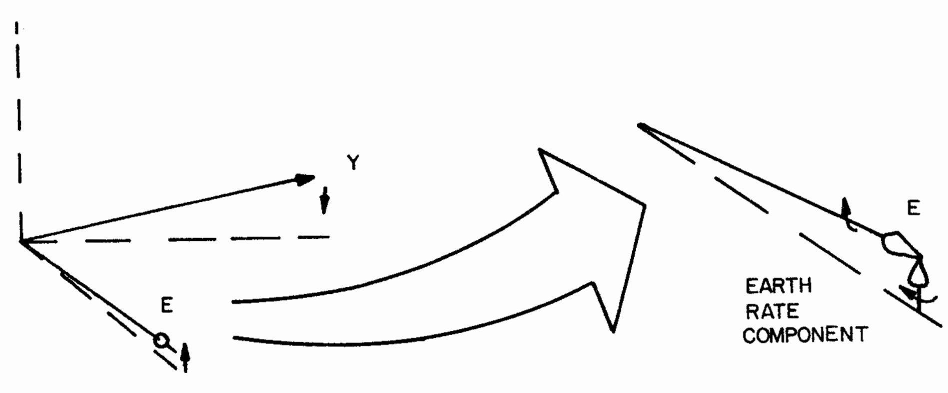


Figure 3-217. Latitude Change and the E Gyro

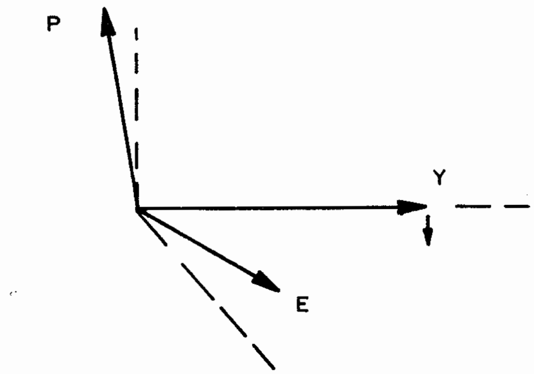
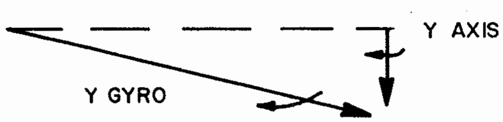


Figure 3-218. Earth Rate and the Y Gyro



Earth rate in opposite direction.

Figure 3-219. Y Gyro and Earth Rate in Opposite Direction

precision gyros used in SINS can maintain the same drift for long periods, it results in good performance. In other words, the SINS will continue to follow the characteristic forms of errors and allow us to make corrections. Unfortunately, a gyro's drift rate (usually

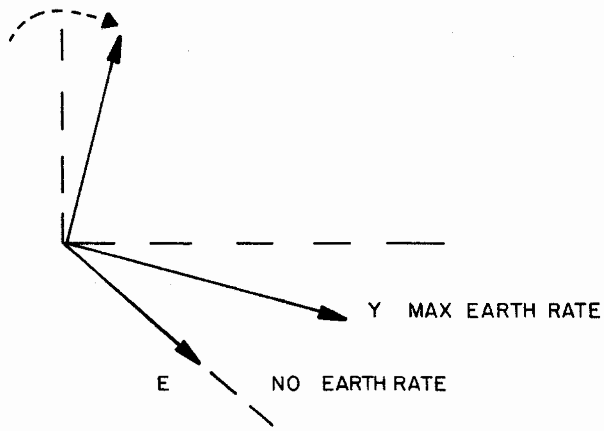


Figure 3-220. Earth Rate and the Y Gyro Approaching the Earth Plane

expressed in degrees/hour) will occasionally change radically. After operating with one drift rate for days, it may suddenly and randomly change. Since there is no way of immediately detecting this, the computer continues to compensate (called bias) for the old drift rate. These radical changes in drift rates are called breaks. Plotting the drift rates that one gyro has might appear as in Figure 3-226. This plot would cover a period of many weeks, so there would be long periods of uninterrupted service. Notice that after the break, the drift rate tends to remain steady.

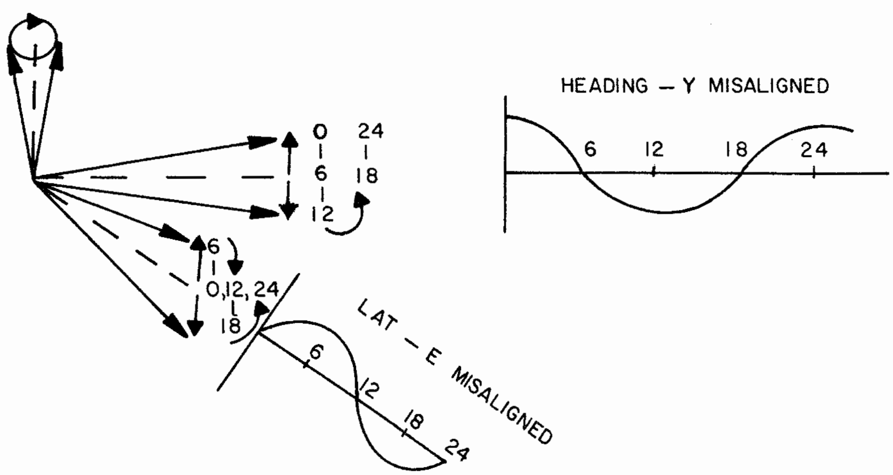


Figure 3-221. Undamped 24 Hour Oscillation

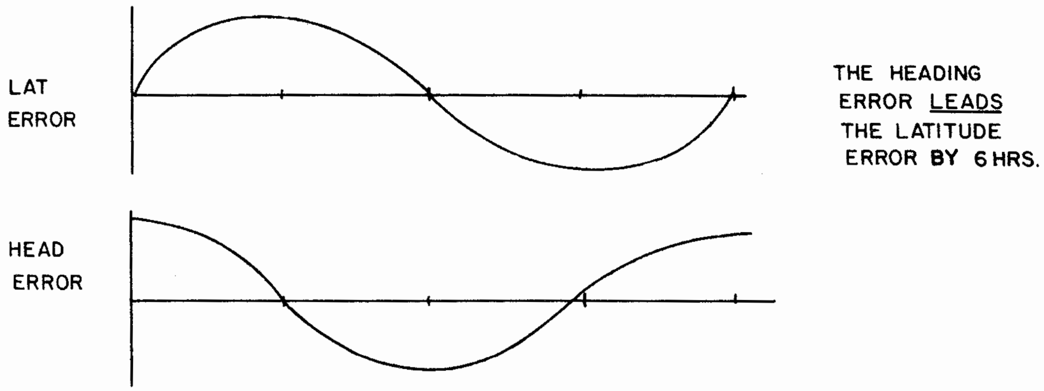


Figure 3-222. Heading vs Latitude

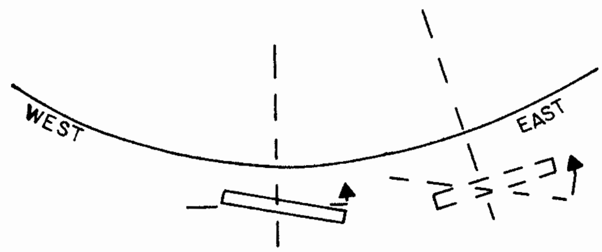


Figure 3-223. Longitude Error

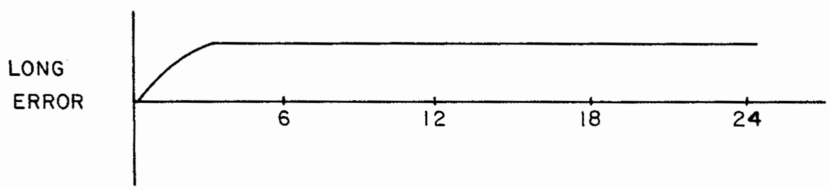


Figure 3-224. Longitude Standoff

This can happen to any one of the three gyros. Very seldom does it happen to more than one at the same time. Consider what happens when one gyro "breaks". For example, look at the Y gyro. If this gyro has a change in drift rate, it will put out a false signal indicating a rotation about its sensitive axis. Now, only part of that drift is compensated for by the computer so that the followup system will tilt the platform. Thus it will start the 24 hour oscillation that was described in the previous section. However, this drift rate is constant and continues until some other force balances it. In discussing the dynamics of the 24 hour oscillation, it was shown that as a gyro left alignment with the coordinate axis, it began to sense earth rate that resulted in a tendancy to return it to the coordinate axis. Now, there are two

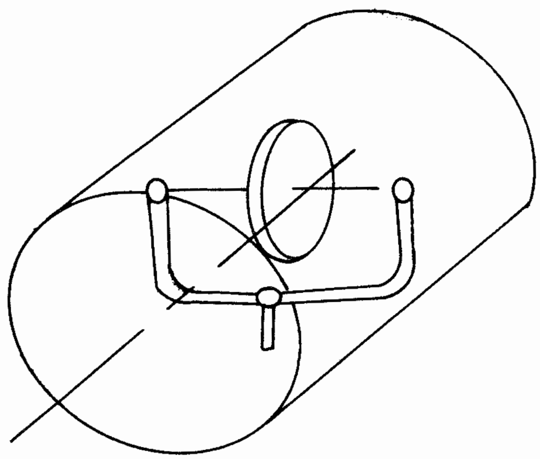


Figure 3-225. Gyroscope

opposing forces: 1) drift rate driving it away, and 2) earth rate driving it back. Since as the gyro moves farther away from the coordinate axis, it senses more and more earth rate, and it eventually finds a place where they are exactly equal and opposite (Figure 3-227). Note now, the gyro has two counter-balancing forces, so that it senses no rotation. Looking back to Figure 3-221, it was noted that the SINS 24 hour oscillation was caused by a misalignment of either Y or E gyros with their coordinate axes. This resulted in these gyros oscillating, undamped, equally on each side of the coordinate axis. Thus, the reason they oscillated about the coordinate axis was because at that axis they felt no earth rate. Now, with an uncompensated drift rate in the Y gyro, the point at which the gyro senses zero rotation is not on the coordinate axis (Figure 3-288). This is called the gyro drift center. The 24 hour oscillation actually takes place about this drift center. Therefore, if the gyro is reset back to the coordinate axis, only the magnitude of the oscillation would increase (Figure 3-229). Referring again to Figure 3-221, note that as the result of the Y and E gyro 24 hour oscillation, the polar axis described in a circle about the true earth spin axis. With the Y gyro uncompensated drift rate, the same circle develops, but it is no longer centered on the true earth spin axis. Note that the same curves have been developed as before. The heading error curve still leads the latitude error curve, but the heading curve is offset from the zero error axis (Figure 3-230).

3-8.6.11 3rd Order Damping

This method of damping (See Figure 3-230) which is now used in all SINS systems is preferable to the 2nd order damping because with 3rd order, a constant error in EM Log will not result in a platform constant tilt and therefore, a constant position error. Nevertheless, it should be realized that any such error is very small and disturbances to the platform will not be noticeable to the SINS unless entering or leaving a large current area (e.g., the Gulf Stream).

3-8.6.12 Monitor Gyro

Some SINS have another gyro mounted on the platform. This gyro, which is exactly the same as the other three, is mounted in another gimbal (Figure 3-231) and is called a monitor gyro. While the platform remains stable, the monitor gyro can be rotated to align with either the X or Y gyro, i.e., on all the cardinal headings (Figure 3-232). This gyro can be used to determine the drift of the X and Y gyros so that they may be biased, thus eliminating drift errors. No platform rotation is needed. The Z gyro biasing and the elimination of the 24 hour oscillations must still be accomplished using the normal reset methods requiring position fix information. However, the operator now has the ability to completely reset and bias the SINS at sea without external heading information, and without disturbing the platform with

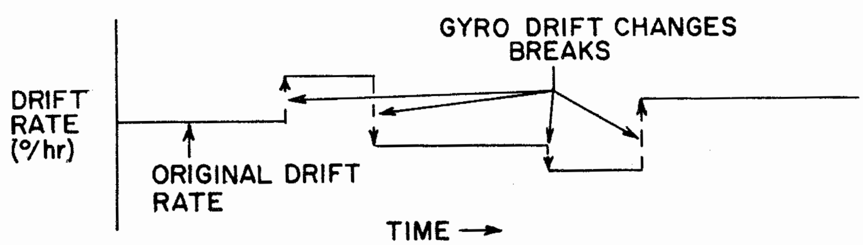


Figure 3-226. Gyro Breaks

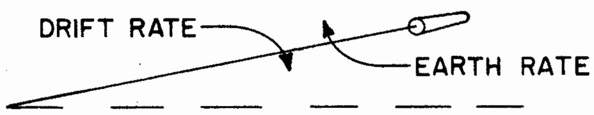


Figure 3-227. Drift Rate vs Earth Rate

rotations. The Monitor gyro is first aligned to the Y gyro axis on the platform. Any movement of the platform will also move the monitor gyro. Now, if both the Y gyro and the monitor gyro have the same amount of uncompensated drift, they would tend not to stay at rest on the Y (east) coordinate axis (Figure 3-233). That is, they would sense no rotation only at the equilibrium point, i.e., only where the component of earth rate exactly counters drift rate. The farther the gyro moves from that equilibrium point, the greater the output signal will be. The monitor gyro is not free to rotate, but is locked on the Y gyro axis, so that the above picture would physically look like this (Figure 234). The M gyro must be torqued continuously to prevent bottoming because it senses the same earth rate as Y, plus its own drift rate. Therefore, the torquing signal to the M gyro is proportional to the combined drift rates of M plus Y. Now the M gyro is rotated 180°. The M gyro equilibrium point is still to the south because earth rate must still be sensed in the same direction as before (assuming the M gyro drift did not change) (Figure 3-235). If the M gyro is above the E-W line, it would sense earth rate in the opposite direction. As before, the M gyro is actually torqued to maintain alignment with the Y axis.

3-8.6.13 System Testing

Because of the mechanically different systems and the complexity of SINS, the technician should follow the maintenance and testing procedures outlined in the test manual for that particular system. A good set of error plots will be a great aid in recognizing trouble and system degradation before it significantly affects the mission of the ship.

3-8.7 TACAN EQUIPMENT

TACAN (derived from TACtical Air Navigation) is a radio air navigation system of the polar-coordinate type. It provides a properly equipped aircraft with bearing and distance from a shipborne or ground beacon selected by the pilot. This equipment incorporates 126 two-way operating channels spaced 1 MHz apart within the bank of 1025 to 1150 MHz for air-to-ground transmission. A like number of channels for ground to air transmission are provided in the bands 962 to 1024 MHz and 1151 to 1213 MHz. The distance from the ground beacon is visually displayed in the aircraft on a meter calibrated in nautical miles. Another meter indicates direction in degrees with respect to magnetic north. With this information, the pilot can fix his position at all times. A single TACAN beacon can provide distance

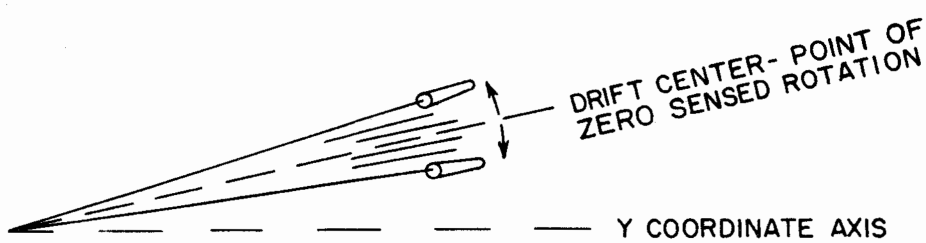


Figure 3-228. New Drift Center

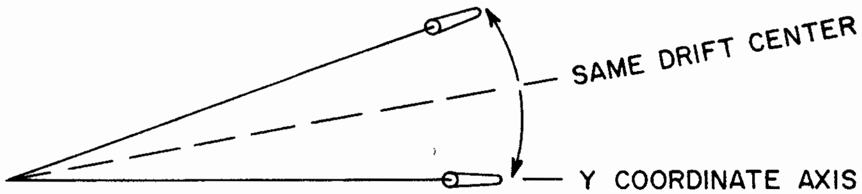


Figure 3-229. Drift Center After Reset

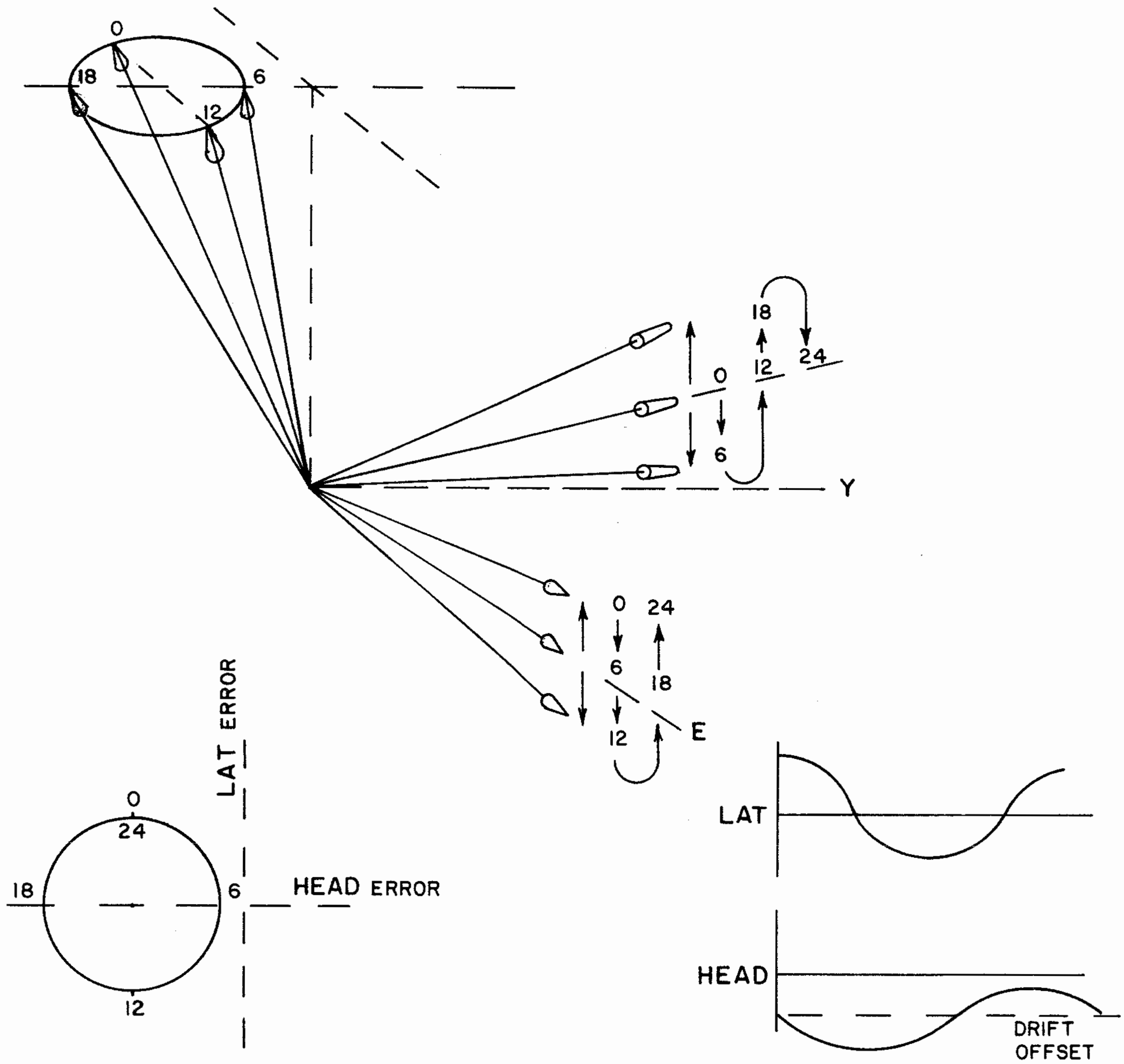


Figure 3-230. Latitude and Heading Drift Vectors

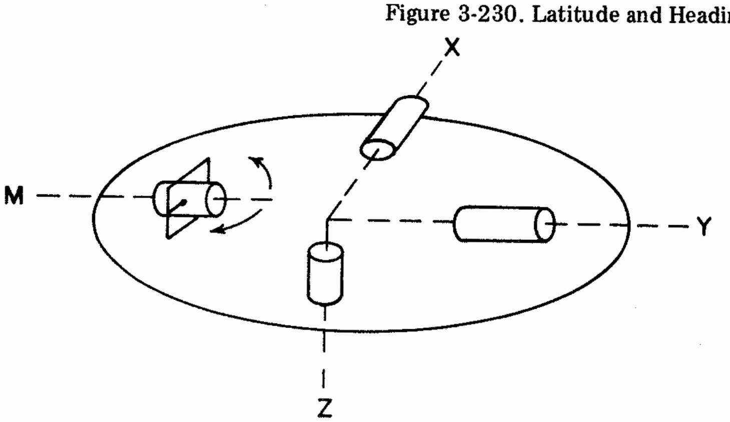


Figure 3-231. Monitor Platform

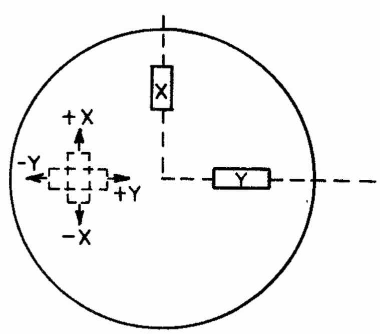


Figure 3-232. Monitor Platform Rotation

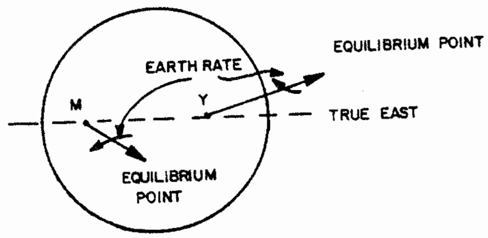


Figure 3-233. Monitor Gyro Axis Alignment

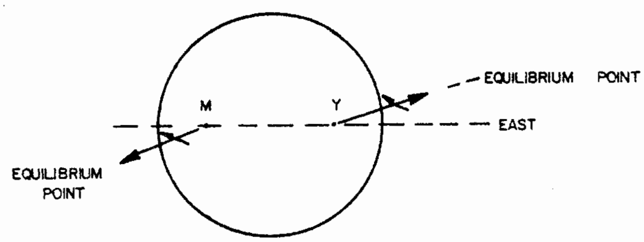


Figure 3-234. M Gyro Aligned With X Gyro

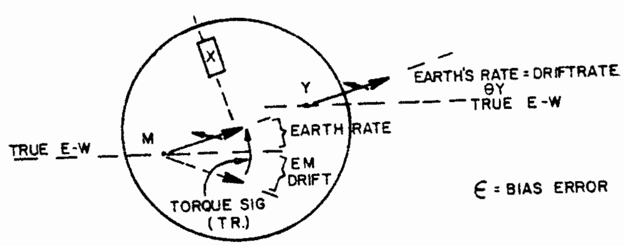


Figure 3-235. M Gyro Aligned With Y Gyro

information simultaneously to as many as 100 aircraft and bearing information to an unlimited number of aircraft within a line-of-sight range of 200 nautical miles. The operating principles of TACAN are based on the time required for a radio pulse signal to travel to a given point and return. In distance operation, the airborne transmitter repeatedly sends out interrogation pulses that are picked up by the ground beacon receiver, which in turn triggers the associated transmitter to send out reply pulses on a different frequency. The reply pulses are picked up by the airborne receiver. Timing circuits automatically measure the interval between the interrogation and reply pulses and convert the time interval into electrical signals that operate the distance indicator. Bearing information is supplied by amplitude modulating the train of pulses generated by the transmitter; modulation is generated by the antenna equipment. The antenna consists of a vertically polarized stationary array around which two sets of parasitic elements, contained in cylinders, rotate in a clockwise direction. The inner cylinder contains one parasitic element and modulates the pulse amplitude of the receiver at 15 hertz. The other cylinder contains nine elements and superimposes a 135 hertz modulation of the fundamental 15 hertz modulated signal. Accordingly, as a result of the rotation of this pattern, the signal received at any given direction from the beacon goes through corresponding cyclic variations in field strength as a function of time. When demodulated, the cyclic variations produce components of 15 hertz and 135 hertz in fixed phase relation to each other in the airborne equipment. Bearing information is derived from the electrical phase of these audio signals with respect to reference pulses. The TACAN radio beacon, with its associated antenna groups and accessories and aircraft radio set, together provide means through which the position of an aircraft can be accurately determined. As many as 100 aircraft may simultaneously obtain navigational information from a signal TACAN installation. The TACAN beacon is capable of receiving on any one of 126 frequencies in the range of 1025 to 1150 megahertz. The set can transmit on any one of 126 frequencies in the ranges of 962 to 1024 megahertz and 1151 to 1213 megahertz. Two types of antenna are available for use. Each antenna can operate on 63 channels, either in the low band of frequencies or in the high band of frequencies. Low band installations transmit at frequencies between 962 and 1024 inclusive, and receive at frequencies between 1025 and 1087 megahertz, inclusive. High band installations receive in the range of 1088 to 1150 megahertz and transmit in the range of 1151 to 1213 megahertz. Two frequencies are used in each channel, one for receiving and the other for transmitting. The frequency used for receiving in low band installations is 63 megahertz above the frequency used for transmitting in the same channel. In high band installation, the receiving frequency is 63 megahertz below the transmitting frequency. The radio beacon output consists of the beacon's identification call. Distance information signals, bearing information signals and random pulses used to make up the constant duty cycle. The radio beacon periodically transmits its identifying call in International Morse Code, thus enabling the aircraft to determine with which radio beacon it is in contact. The characters of the code consist of a train of pulse pairs generated at a fixed rate of 1350 hertz. A mechanical keyer accomplishes the coding. The pulses are phase-locked with the antenna rotation. The aircraft receiver detects these regularly occurring pulse pairs and reproduces the code as a keyed 1350-hertz audio tone. Identification call signals have priority over the distance information signals. Bearing information reference bursts (described below) have priority over the identification call. The relative durations of these signals are such, however, that effectively there is no interruption of distance information or identification call. The relative durations of these signals are such, however, that effectively there is no interruption of distance information or identification call. The aircraft radio set transmits distance interrogation signals and receives replies from the radio beacon. The aircraft radio set sends out on the assigned channel pairs of pulses. The pulses of each pair are spaced 12 microseconds apart. The radio beacon receives these pulse pairs, termed distance interrogation pulse pairs, and in reply transmits back to the aircraft pulse pairs on the assigned channel. The time delay of the distance interrogation pulse pair to the corresponding distance reply pulse pair in the radio beacon is adjusted to exactly 50 microseconds, and the aircraft radio set deducts 50 microseconds from the total time elapsed between interrogation and reply. The distance between the aircraft and the radio beacon is thus determined by measuring the total time elapsed between initial transmission of the distance interrogation pulse pair and reception of the corresponding radio beacon reply. Since all aircraft using the radio beacon operate on the same pair of receiving and transmitting frequencies, all radio beacon replies to all aircraft distance interrogations are received by all of these aircraft. It is therefore necessary for each aircraft to seselect those radio beacon replies which result from its own distance interrogations. This is done as follows. Transmission of aircraft distance interrogation pulse pairs is continuous, and, in turn, aircraft reception of distance replies is also continuous. Transmission of the distance interrogation pulse pairs is semi-random; that is, the number of pulse pairs per second transmitted by a particular aircraft remains fairly constant, but the intervals between pulse pairs vary. The variation in time spacing is peculiar to each aircraft, and permits the aircraft to pick out the replies to its particular interrogations. The time-spacing pattern of reply pulse pairs is continuously compared with the time-spacing pattern of interrogation pulse pairs. Only those pulse pairs which lie in matching patterns operate the distance meter of the aircraft radio set. This meter displays continuously the distance of the aircraft from the radio beacon. Bearing information originates within the radio beacon. The radio beacon antenna modulates the total pulse-pair output with sub-audio frequency components, and the radio beacon transmits special bursts of pulse pairs which are inserted in the radio beacon output at appropriate intervals. The bearing of the radio beacon relative to the aircraft, as measured between a line connecting the aircraft to the radio beacon and a magnetic meridian, is directly proportional to the occurrence of a special burst of pulse pairs and the phase of the sub-audio frequency components modulated on all the pulse pairs. In determining its bearing the aircraft utilizes 15-hertz and 135-hertz modulation on the total received signal, and 15-hertz and 135-hertz reference bursts which are components of the received signal. Figure 3-236 represents a simplified version of the relative time occurrence of the components of the signal at the radio beacon. The 15-hertz and 135-hertz reference bursts are actually a train of RF pulse pairs. The 15-hertz burst consists of 12 pulse pairs spaced 30 microseconds apart. The 135-hertz burst consists of six pulse pairs spaced 24 microseconds apart. It is the number and spacing of the pulse pairs which identify each of the bursts. The aircraft radio set first reconstructs the 15-hertz components from the envelope of the received pulse pairs. The 15-hertz and 135hertz bursts serve as marker signals. The aircraft radio set compares the time occurence of the 15-hertz burst with the phase of the 15-hertz component, and the time occurrence of the 135-hertz component. One hertz of the 15-hertz wave represents 360 degrees of bearing. One cycle of the 135-hertz wave represents 40 degress of bearing. The beginning of the cycle on the 15-hertz wave represents magnetic north. Thus, if the 15-hertz reference burst occurs when the 15-hertz wave is one third of the way to its first positive peak, and the first 135-hertz reference burst occurs when the 135-hertz component is at its first negative peak, the bearing is established as being 330 degrees to the beacon. The total beacon output consists of RF pulse pairs transmitted at a rate of 3600 pulse pairs per second, 900 pulse pairs of which are 15-hertz and 135-hertz reference bursts. The additional 2700 pulse pairs per second which are required to keep the rate constant are either distance

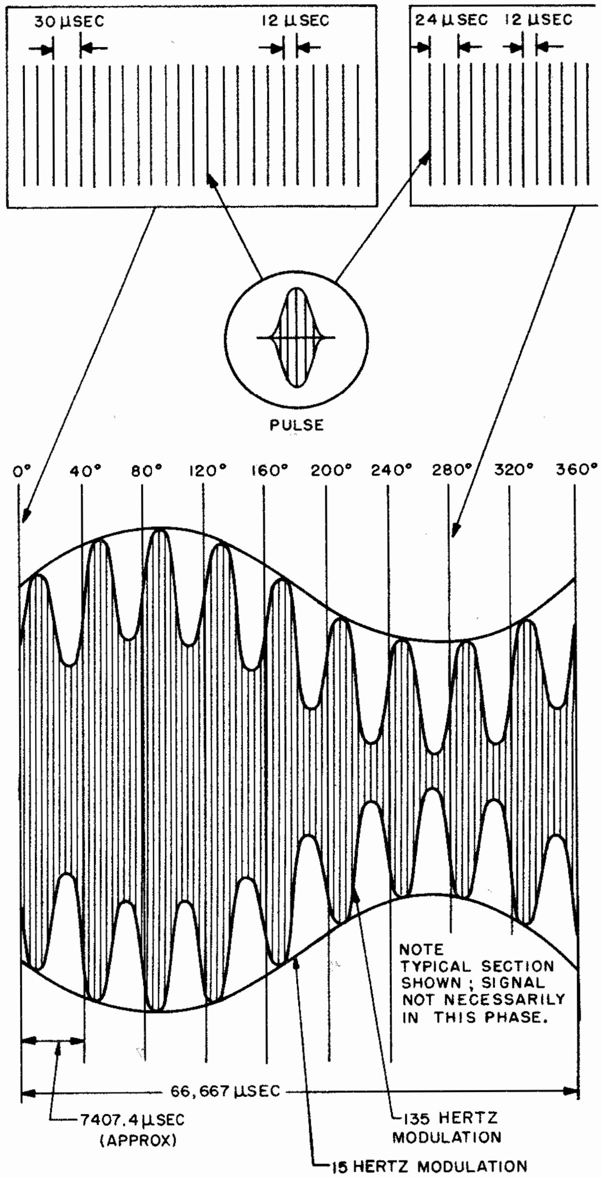


Figure 236. Relative Occurrence of TACAN Radio Beacon Signals

reply pulse pairs or random noise pulse pairs. The radio beacon generates these lest the number of replies become insufficient to maintain the required rate, since the number of replies is dependent on the number of interrogations received by the radio beacon. Should the number of interrogations exceed 2700 pulse pairs per second, the radio beacon will reply only to that number, thus maintaining the constant output pulse-pair rate. Identification call pulse pairs are substituted for distance replies or random noise pulse pairs, and consequently do not affect the output rate. A constant output rate is necessary in order that the aircraft may extract the 15-hertz and 135-hertz modulation. Because of the importance and complexity of the TACAN beacon, built-in test equipments are included in the console which greatly facilitate performance testing and trouble isolation of the unit. The test equipment includes a pulse analyzer, pulse counter, pulse sweep generator and an oscilloscope. Specialized and simplified procedures are included in technical manuals for performing tests such as receiver sensitivity, transmitter power output, and spectrum analysis. Since the reference bursts, coding, and other pulses and their spacings are highly critical, frequent checking of these waveforms is mandatory for optimum performance of this equipment. Some of the trouble encountered with the beacon is first detected by the aircraft and reported to the radio beacon location. Defects in the antenna output radiation, as indicated by failure of the aircraft to receive correct bearing information, distance information replies, or identification codes, may have their origin in units of the radio beacon other than the antennas. Ordinarily, trouble does not originate in the antenna unless it has suffered mechanical damage or has been dismantled for some reason. Because of the complex nature of the antenna group, the following tests can be performed only at a repair depot:

1. Harmonic analysis.
2. Measurement of percentage modulation and sideband tracking in the vertical plane.
3. Measurement of the relative phase of the 15- and 135-hertz components.
4. North calibration of the antenna.
5. Vertical pattern measurement.

The specific procedures for performing these measurements are included in the technical manual for the equipment.

3-8.8 HYDROGRAPHY

Hydrography may be defined as that science which deals with the measurement of the physical features of waters and their marginal land areas, with special reference to the elements that affect safe navigation, and the publication of such information in a form suitable for the use of navigators. Hydrographic surveying in the strict sense is defined merely as the surveying of a water area; in common practices, however, it has come to include such matters as the study and measurement, in a given survey area, of the magnetic variation and dip, the tides, currents, and meteorological characteristics. Probably the most important requirement in any survey is to establish the position of each feature measured. To do this, the location of a reference point must first be very precisely determined, for all other positions are located relative to this reference point, which is called the origin.

3-8.8.1 Accuracy

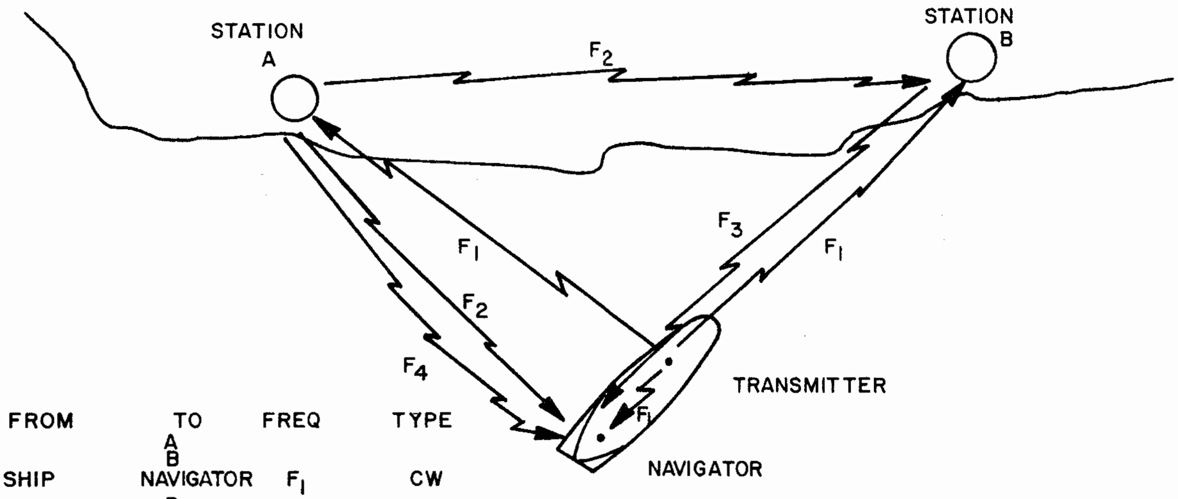
The navigator often places almost unquestioning reliance on his nautical chart. Consequently the hydrographer needs to use the utmost accuracy when conducting surveys. In hydrographic surveying it is extremely important to be able to comb an area, following lines with a fixed spacing in order to insure that no possible underwater obstacle has been missed, and that the area has been systematically covered. An error or inaccuracy in charted information may result in a marine disaster and possible loss of life and property. The accuracy of a given survey depends on the purpose for which it is intended. Accuracy ranges from one part in 5,000 to one in 100,000; the latter extremely high standard being necessary for missile and space craft control. The principal function of a hydrographic survey is to determine depths of water; the positions at which soundings are obtained can best be determined by accurate reference to known points on shore. This locates the points at their correct geographic positions, and also results in the land and marine features being placed in correct relationship to each other. This is an important consideration as the mariner, when near the coast, also uses reference positions on land to locate himself.

3-8.8.2 Controls

In hydrographic surveying, the term "control", when referring to a survey vessel, is used to mean the accurate location of the vessel within the survey area. Visual control is the determination of position by visual reference to conspicuous landmarks whose positions are established relative to the origin. The method most commonly used is to obtain horizontal sextant angles between landmarks or accurately located beacons on the shore. Position is then plotted by means of a three-arm protractor. At present, visual control is supplemented by electronic control, which permits operation in periods of low visibility, as well as beyond visual range. To be acceptable for this use, the electronic equipment must have great inherent accuracy; it must be operated within the closest practicable tolerances. Any electronic system meeting the requirements of accuracy might be used for control. Radio acoustic ranging, radar in conjunction with transponder beacons, and the electronic position indicator have been employed for control. Some of the electronic systems which are currently widely used in the United States are Raydist, Decca, Shoran and Hiran. Numerous other systems, operating with similar techniques, are also available. Various modes of operation are therefore possible for surveying. These systems, named from the curves of equal value generated on the chart when the signals received from the transmitting stations in the system are charted, and, in some cases, a combination of these. The range or distance measuring systems offer the advantage that they require no special charts. A hyperbolic curve represents a locus of points where the difference of times required for the signals from two transmitting stations to be received is equal. The chief limitation on the accuracy of these systems is caused by the geometry of the hyperbolae. The lane is the unit of measurement in these systems; its width is equal to half a wave length. When using a phase comparison system the lane is further subdivided. The best resolution that can be expected along the base line is about one one-hundredth of a lane; at 2 MHz, this is about three feet, and at 10 kHz it is about 450 feet. As the vessel departs from the base line, the lanes expand and are successively wider. In a circular system, the vessel uses the signals from only one station at a time; the range to the station is based on the length of time of signals travel. There is no deterioration in signal accuracy due to position relative to the transmitter, as the signals spread outward over the earth's surface as a series of equispaced concentric rings. In an elliptical system, the sum of the times of transmission of signals from two stations would be used. A line representing a locus of points where these sums are equal would take the form of an ellipse.

3-8.8.3 Raydist

Raydist is the name given to a series of transportable and highly accurate positioning systems in which the phases of two continuous-wave (CW) radio signals are used to determine the position of a mobile station. Various modes of operation are available differing according to the system. Raydist operates in the 1.6 to 5.0 MHz frequency range, which permits the use of small efficient portable transmitters and antennae. The systems are not limited to line-of-sight operations, and may be used at distances in excess of 150 nautical miles with a sensitivity of approximately 1 meter when sufficient power is available. The design of the Raydist systems permits presentation of positioning data in various forms: range, elliptical, and hyperbolic or in a combination of any two of these. The DM Raydist using the range-elliptical combination has been widely employed by the Coast and Geodetic Survey. Distance measuring systems using circular coordinates have a greater potential for accurate positioning than do the hyperbolic systems. This is because of the geometry involved; lane widths remain constant with increased range in a circular system, while they increase very rapidly with range in a hyperbolic system. This is a system requiring four radio frequencies for use by three stations as indicated in Figure 3-237. Two of these frequencies are distance-measuring CW frequencies, and the two others are amplitude modulated (AM) to transmit information back to the ship concerning the phase relationship of the CW transmissions received at each station. The AM frequencies do not enter into the determination of distance; however, interference on any of the four frequencies will adversely affect operation of the lane counters in the phase meter. A CW transmitter located at the A station operates at a frequency 200 Hz higher or lower than half the frequency used by the ship. A special purpose, dual-frequency receiver is also used at both the A and B shore stations. The receiver at each shore station accepts the transmission from both the vessel and from the A shore station, doubles the lower frequency and heterodynes it with the higher frequency to generate 400 Hz audio beat note. Separate audio notes are then returned to the mobile station by the A and B stations as a modulation of two additional and completely independent frequencies (the AM frequencies). These signals are phase-compared with an internally generated beat note, by an instrument called the "navigator" aboard the survey vessel. A comparison of the A signal with the local mobile frequency yields a reading on the phase meter of the range (distance) to the A station. The B station compares the phase relationship of the signals from the mobile and A stations yielding a measurement in elliptical coordinates which the phase meter then converts into distance of the B station from the mobile station. The "navigator" receives the data from all stations and converts them to a usable output of position coordinates. Since the system is completely automatic, the distance from the shore station can be read directly off the dials. By heterodyning the transmitted frequencies the equipment has computed the distance traveled by both transmitted frequencies, and properly located the position of the vessel. Since with the DM Raydist, land or other obstacles between the two shore stations can cause sufficient interference to limit the accuracy of the system, a DR Raydist system is often employed. With this system the A and B stations are identically equipped, each having its own CW transmitter operating so closely in frequency as to be considered a single frequency. In this range-range system the distance between the shore stations, plus the features of the terrain, are no longer a hindrance, all-weather, around-the-clock information for positional data. The latest DR system is the DR-S with the single sideband transmitters requiring only two frequencies. The most widely used Raydist systems - the DM, DR, and DR-S - are capable of providing position information simultaneously to a number of users. They are high-accuracy ranging systems which operate over the horizon with multi-user capability. The heterodyne tone, as it is received at the two shore stations, is returned to the "navigator" and compared phase-wise with the same tone as it is detected directly in the "navigator". The tone being relayed back from the A station is compared with the tone detected locally in the "navigator" to determine the range to the A station; this process is concurrently repeated to determine



FROM	TO	FREQ	TYPE
SHIP	A B NAVIGATOR	F_1	CW
A	B NAVIGATOR	F_2	CW
B	NAVIGATOR	F_3	AM
A	NAVIGATOR	F_4	AM

Figure 3-237. Raydist Overview

the range of the B station. Since the heterodyned frequency is derived from a fixed transmitter and the mobile transmitter, its phase, as it is detected at the shore stations, will be dependent upon the location of the mobile transmitter. Therefore, as the mobile station moves, the relative phase of the heterodyne tone changes, causing the phasemeter or Position Indicator dials to rotate. Any phase-comparison system produces position within a lane; as in most surveying systems, Raydist does not identify the lane. A starting position must therefore be known. A lane counter, an integral part of the equipment, then counts the lanes crossed from the starting position. The base stations for the new DR-S Raydist system, as well as some of the older Raydist systems, are designed to operate from conventional automobile-type storage batteries; the high-powered systems are generally operated from 110 volt, 60 cycle sources. The conventional batteries can operate up to 24 hours without recharging and provide a reliable power source. For longer operations, larger capacity batteries are used.

3-8.8.4 Shoran

Shoran (for SHOrt RAnge Navigation) is a secondary radar system using two transponder beacons, located ashore, and a single indicator aboard the vessel to measure the distance from each beacon. This system is a pulse-modulated, ultra-high frequency (230 to 310 MHz) distance measuring system used primarily in survey work. Its operation is based on the assumption that the velocity of electromagnetic waves is known to within a few miles per second. In operation the two transponder beacons are positioned and the distance between them established. Then the vessel transmits pulse type signals alternately on frequencies of 230 and 250 MHz. These signals actuate the transmitters at the transponder beacons and, after a preset time delay, the stations transmit the signal back to the vessel, pulse for pulse, on a frequency of 310 MHz. The returning pulses are matched with a fixed index, thus automatically converting the elapsed time for the round trip to distances in statute miles. The distance readout is calibrated in hundredths of a mile; distance can be read with an accuracy of from 30 to 60 feet, except at extreme ranges. After reception of the distances from both shore stations the navigator plots distance circles and the intersection of these circles marks his location. Due to the high frequency used, the distances at which the system can be utilized is limited by the curvature of the earth. A higher precision version of Shoran, called Hiran (From HIgh precision RAnge Navigation) was developed to meet the most exacting survey requirements. Hiran is similar to Shoran but signals are kept at a constant amplitude to prevent an error in the time of triggering the signal from the transponder. Hiran also has a greater accuracy, the smallest division of the counter being one five-hundredths of a mile, rather than one one-hundredth as in Shoran.

3-8.8.5 Decca HI-FIX

Decca HI-FIX, as used in survey work, is lightweight and compact, thus providing high mobility for erecting stations for survey work. HI-FIX has a maximum range of about 100 miles, the range depending upon the radiating power of the station, and provides lane identification, in addition to the phase comparison, in one of the configurations available. It can be used either in a two-range version or in a hyperbolic system. A master and two slave stations share a common carrier frequency on a time-multiplex basis in the 2 MHz frequency region.

3-9 SONAR SYSTEMS - OVERALL PERFORMANCE CHECKS

3-9.1 EXTERNAL FACTORS AFFECTING PERFORMANCE

Numerous conditions other than the performance of the actual sonar equipment affect the overall performance of a sonar system. The following factors should be considered when evaluating an overall performance check.

3-9.1.1 Water Conditions

The ocean is not an ideal medium for the transmission of sound. It is not infinite in extent, being bounded by the bottom and the surface. Furthermore, the ocean is not the same in composition at all levels and locations; the upper layers are usually warmer than the lower layers, and near large rivers the composition may be less saline. For both reasons, the water is less dense in the upper regions. The temperature and salinity may also change in a horizontal direction. Thus, a sound wave propagated through the ocean does not propagate in a straight line. Other less obvious properties of the ocean contribute to making the propagation of sound waves difficult. As a sound wave travels outward from a source in the sea, some of its energy is converted into heat by friction due to the viscosity of the water. This process is called "absorption." Another portion of energy goes into the production of secondary wavelets which travel in directions other than that of the primary wave. This is the phenomenon called "scattering." "Attenuation" is a more general term, embracing both absorption and scattering.

3-9.1.2 Turbulence

"Turbulence" is the general term applied to noises which affect the reception of sonar targets. These noises can be divided into two general classifications: background, and reverberation noises, as discussed below.

3-9.1.2.1 Background Noise

Background noise includes extraneous sounds of various kinds. These are subdivided into four categories, as follows:

1. Self-Noise. Self-noise is produced at or on own ship, and stems from the following causes: rotation of the ship's propeller; motion of the ship and/or motion of gear relative to the water (e.g., the movement of water past the transducer); mechanical vibrations incident to operations aboard own ship, producing sound waves that are picked up by the hydrophone; and stray noises caused by operation of the ship's power system and electrical machinery that that may be picked up by direct vibration or by propagation into the water.

2. Median Self-Noise. Median noise is defined as the middle value of a series of readings taken 10 degrees apart in the search arc from 215 to 115 degrees relative in the forward direction with the ship steaming on a steady course under normal steaming conditions.

3. Ambient Noise. Ambient noise is produced in the ship's environment, and stems from the following causes: sea noise (principal cause unknown), heard even in deep water; rain, surf, and whitecaps are also occasional sea-noise causes; biological noise, made by certain shrimp, fish, etc., that produce sounds that are audible on sonic and ultrasonic equipments; traffic noise, caused by ship traffic and, in or near harbors, by industrial operations.

4. Target Noise. Target noise is produced at the target. This is a desired sound for listening equipments but an unwanted background when echo-ranging, and may prevent accurate determination of the range of the target.

3-9.1.2.2 Reverberation

Reverberation noise is inherent in all echo-ranging operations, and is caused by the scattering of the emitted sound as a result of irregularities of the sea bottom, irregularities of the sea surface, and unknown scattering elements widely distributed throughout the sea.

3-9.2 INTERNAL FACTORS AFFECTING PERFORMANCE

Various conditions within the equipment itself contribute to the overall performance of a sonar system. A complete performance check should include the checks listed below.

3-9.2.1 Driver Power Output

This check should be made to ascertain the amount of RF power being delivered by the driver (transmitter) to the transducer. Various methods are available for performing this test, depending upon the specific equipment upon which the test is to be made.

3-9.2.2 Receiving Sensitivity

Sensitivity measurements furnish accurate information as to the overall performance of the receiving system. A decrease in efficiency to the point of lowered performance would be apparent in a sensitivity measurement, and steps should be taken to restore the equipment to peak efficiency.

3-9.2.3 Keying Circuits

Accurate keying, controlling, and timing circuits are vital to the satisfactory performance of a sonar system. Operation of a complete sonar system at times involves the use of numerous companion units in addition to the basic sonar equipment, such as resolvers, computers, range recorders, and attack directors. In addition to controlling the driver, the keying circuits must provide synchronization of all the various functions of the equipment, including indicating circuits, ranging and depth-determining circuits, etc. The circuits which control training, and raising or lowering of the transducer, are vital to the operation of a sonar system. These circuits must be tested to complete an overall performance check.

3-10 TELEVISION SYSTEMS AND EQUIPMENT

The purpose of television is to extend human vision beyond its normal limits. By this means an observer can sit comfortably in front of a television monitor and yet watch the interior of security spaces, observe closely the launch of an untested space booster from every angle, explore the bottom of the ocean, view objects under infrared or ultraviolet light. Video tape recording allows an almost immediate second look at any unusual event and a permanent record of it when desired. Modern-day users of closed-circuit television range from supermarket personnel to astronauts. The potential applications are only as limited as our imagination.

3-10.1 BASIC TELEVISION SYSTEMS

A television system in its simplest form consists of a camera and a monitor joined by a cable (Figure 3-238). As a next step, more monitors may be added. Finally, the most extensive systems consist of many cameras and monitors with complex switching and distribution systems. The camera consists of a lens which projects an image of the subject on the sensitized surface of a camera tube, deflection circuits that perform a systematic dissection of the projected image (scanning), and a video amplifier that boosts the feeble signal from the camera tube to a usable level. The image is divided into bits of information (video signal) that are transmitted in sequence to the monitor. Here the video signal is amplified further and fed to a cathode-ray tube, called a kinescope or picture tube. Deflection circuits move a beam of electrons inside the kinescope in step with the scanning process in the camera tube. Consequently, an image which is made up of brightness levels corresponding to those on the sensitized area of the camera tube appears on the face of the picture tube. In virtually all systems, scanning progresses a line at a time from left to right and from top to bottom (as the picture is viewed). A complete picture consists of 525 lines in most American systems. Figure 3-239 illustrates the pattern followed by the scanning beam. When this pattern appears on the face of a picture tube, it is called a raster. Each time the beam traces this pattern, one complete still picture is transmitted. A succession of 30 or more still pictures per second gives a person the illusion of viewing a moving object. Since the normal vertical scan rate is 60 pictures per second, only alternate lines need be scanned each time the beam passes from the top of the picture to the bottom. Each complete picture, recurring at a 30-hertz (cycles per second) rate, is called a frame. The half-picture, recurring at a 60-hertz rate, is called a field. Therefore, two fields make up a frame. Since a frame consists of 525 lines and is repeated 30 times a second, the horizontal scanning rate is 30 times 525 - 15, 750 hertz. Note that the line rate is an odd multiple of the frame rate; that is, a fractional number of lines (262 1/2)

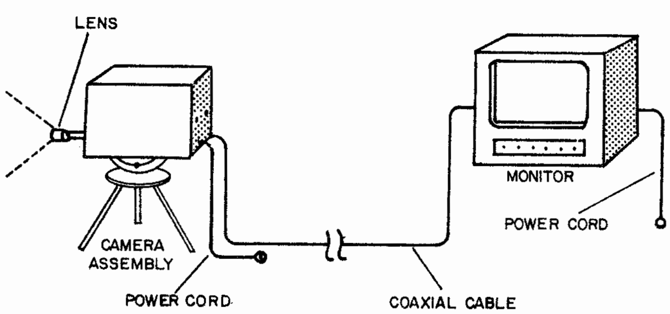


Figure 3-238. Simple Television System

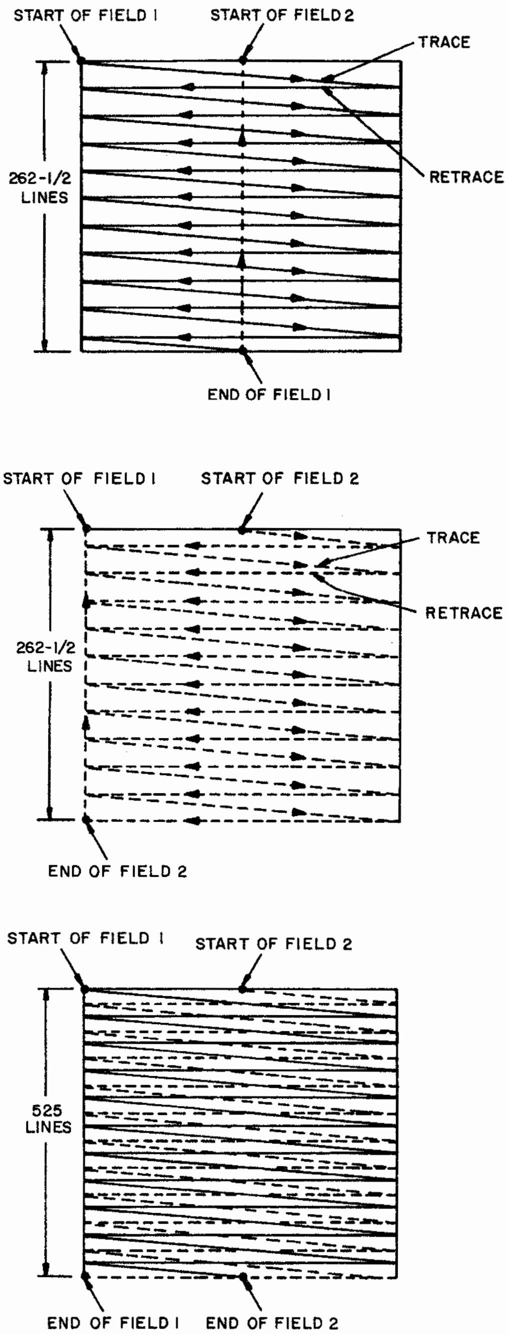


Figure 3-239. Interlaced Scanning Sequence

are scanned per field. This characteristic causes the lines of one field to fall midway between those of the other (interlace). Its purpose is to increase the apparent flicker rate from 30 to 60 hertz. The higher rate is much less noticeable that the lower; in fact, it cannot be detected by the naked eye under normal viewing conditions. Not all closed-circuit television systems maintain a fixed phase relationship between the horizontal and vertical scans. In some of the least expensive systems the horizontal- and vertical-deflection generators operate independently of each other. The result is a random relationship between the two generators, or random interlace. Scanning rates other than the foregoing are frequently used in closed-circuit television. Where higher resolution is necessary or desirable, the number of lines in the picture may be increased. Since these systems are virtually always interlaced, an odd number of lines will make up the picture. Some rates commonly employed are 729, 945, 1023, and 1203. Where higher resolution, increased sensitivity, or decreased bandwidth is required, the vertical-scan rate may be decreased. This type of operation, commonly called slow-scan, is suitable when a continuous display of motion is not required and photographs are to be taken of the television frames. It also applies if other recording devices, such as facsimile recorders, are used to display the picture. This type of scanning was successfully employed by the Ranger, Mariner, and Surveyor spacecraft in photographing the moon and the planet Mars where it made possible extremely long-distance transmission with very low-powered transmitters. The narrow bandwidths permit the use of much more sensitive receivers than can be done using standard scan rates and standard bandwidths. The Mariner system allows television signals to be recorded on a light-weight video tape recorder suitable for installation aboard a modest-sized spacecraft. Slow-scan systems may employ frame rates as low as one over a period of several minutes. Interlace is normally dispensed with in these systems, since its purpose is solely to reduce flicker. The aspect ratio (ratio of width to height) of the active picture area of broadcast-type signals is 4 to 3: four units of width to three units of height. This is usually maintained for systems using other scanning standards. During the active (picture) portion of the raster the beam of the picture tube is modulated in intensity according to the signal from the camera. The usual polarity of signal transmission is black-negative, with the sync and blanking portions of the signal more negative than the video portion — blacker-than-black. The most positive portions of the signal correspond to the whitest areas of the picture. The monitor scan is maintained in synchronism with the camera scan through the transmission of pulses along with the video signal. Two types are used: the horizontal-synchronizing and the vertical-synchronizing pulses. The former are located between successive scanning lines while the latter are placed between successive fields. Their function is to initiate retrace in the monitor. These pulses must have some distinguishing features in order that the monitor may identify them. This is normally accomplished by making the vertical pulses several times as long as the horizontal pulses. The scanning beam in the monitor must return from the right side of the picture to the left at the end of each scanning line, and from the bottom to the top at the end of each frame. To prevent a retrace pattern from appearing in the picture, the monitor beam is cut off during these intervals. This is accomplished by including a negative square wave, called the blanking signal, as part of the video information to the monitor. It appears as a blacker-than-black signal. The blanking pulses in a 1.0-volt composite signal are approximately .07-volt in amplitude. This level is variously called the pedestal, black, set-up, or blanking level. Both horizontal- and vertical-blanking pulses are used; they also serve as sync pulses in some industrial equipment. In this case their amplitude is increased to approximately that of the usual sync signal. The horizontal-blanking pulses in a 525-line system ordinarily have a width of about 11 microseconds. Vertical-blanking pulses have a duration equal to about 20 horizontal lines and recur at the field rate, 60 hertz. The vertical blanking pulses are not interrupted by the horizontal-blanking pulses. Television cameras are ordinarily fed a composite signal consisting of both horizontal- and vertical-blanking pulses. The simplest form of synchronization that is likely to be encountered is the so-called "industrial" or "simple" sync. In this form the same pulses are utilized for both synchronization and blanking. Typically, the horizontal-sync pulse has a duration of about 10 microseconds, while the vertical pulse has a duration of about 20 horizontal lines. These values are somewhat dependent on the number of lines in the picture, and typical of a 525-line system. Figure 3-240 illustrates a video signal using this form of synchronization. Pulses of the same width may or may not be used in driving the camera scans and for camera blanking. Interlace may be fixed or random, though the former is generally regarded as superior. The objection to the latter is the appearance of "line crawling" in the picture. Figure 3-241 shows a more advanced type of industrial synchronization. Note that the horizontal-sync pulses follow a short blanking interval, called the front porch. This allows capacitors in sync-separation circuits to discharge to a uniform value before each pulse, thereby establishing more precise horizontal timing than is possible with the simplest type of sync. Note that the vertical-sync pulse is also delayed slightly for the same reason. The foregoing systems of synchronization are only two of many. Other variations are possible and do exist; each of them may have an interlaced and noninterlaced form. The system shown in Figure 3-242 has certain disadvantages, even in its interlaced form. Horizontal synchronization is not maintained in the monitor during the vertical-sync interval. This causes a "hook" at the top of the picture which occurs while the horizontal-deflection circuits of the monitor are regaining synchronization. Another difficulty lies in the many types of use and in the different component values required in monitors for optimum performance with each. More serious problems arise when video tape recorders are used and it is necessary to switch between the various types. For these reasons the EIA signal, used in commercial broadcasting, has increasing favor with industrial users. The EIA Television Standard RS-170 provides uniform values for the television synchronizing signal, specifies signal polarity and voltage, and defines the input and output characteristics of equipment. This assures compatibility between units of various manufacturers. Thus, elaborate switching systems may be employed using assorted makes of equipment, and video tapes may be exchanged between different organizations which adhere to these standards. The latter is not always possible when industrial-type signals are used. Figure 3-243 illustrates the EIA signal. In addition to the refinements incorporated in the signal of Figure 3-242, it provides equalizing pulses immediately before and after the vertical-sync interval. These are approximately half the duration, but twice the frequency, of the horizontal-sync pulses. They allow capacitors associated with synchronization to assume equal charges at the start of each vertical-synchronizing interval, regardless of picture content. This assures that the vertical-deflection cycle will start at the same time, every time, with respect to the vertical-sync signal. In addition, the equalizing pulses permit synchronous operation of the horizontal-deflection circuit during this interval. The vertical sync signal is interrupted (serrated) in five places. The serrations are differentiated in the sync-separation circuits of the monitor so their trailing edges will trigger the horizontal-deflection circuits and maintain horizontal sync during the vertical-sync interval. In the time that follows the first field, the horizontal oscillator of the monitor is triggered successively by the first, third, and fifth equalizing pulses, the leading edge of the vertical-sync pulse, and the trailing edges of the second and fourth serrations within the vertical-sync pulse. After the second field the horizontal oscillator is triggered by the second, fourth, and sixth equalizing pulses, and the trailing edges of the first, third, and fifth serrations of the vertical-sync pulse. Pulse widths and other time measurements in television are sometimes given in terms of the time taken by one horizontal line ("H"). If one complete picture is transmitted

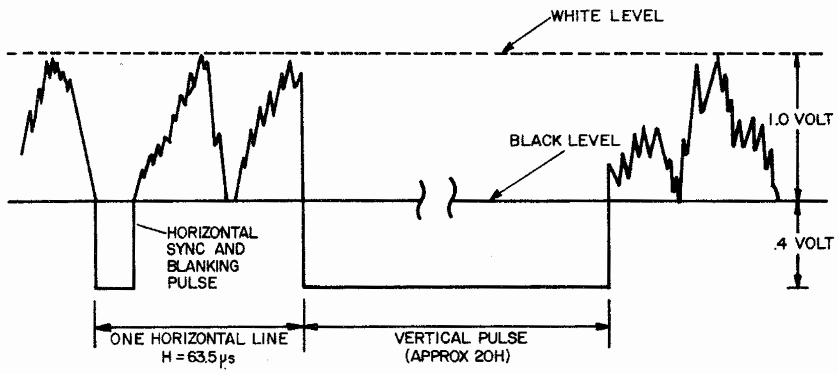


Figure 3-240. Simple Industrial Synchronization

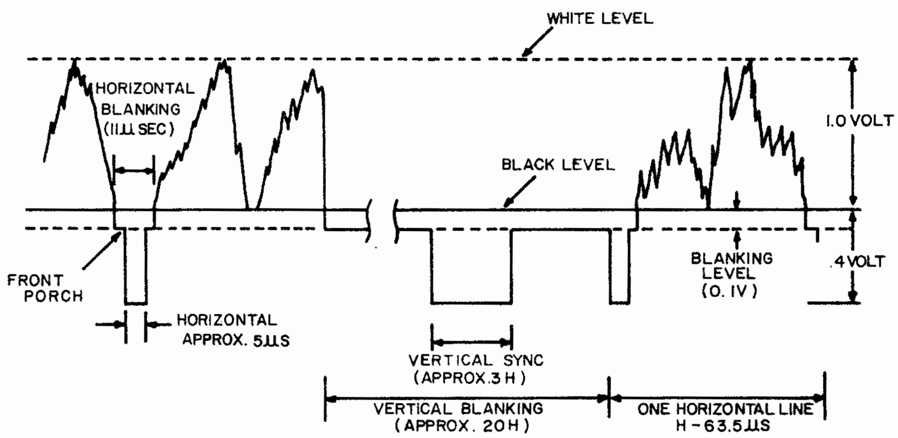


Figure 3-241. Industrial Synchronization with Blanking

in 1/30 second, then H equals approximately 63.5 microseconds if the picture is comprised of 525 lines. Longer times are occasionally given in terms of the time taken by one vertical field ("V") — 1/60 second (16.7 milliseconds) in most systems. EIA standards also specify that the whitest portions of the signals shall be positive in polarity. The video portion of the signal shall have a maximum amplitude of .714-volt, while the sync portion of the signal shall have an amplitude of .714-volt, while the sync portion of the signal shall have an amplitude of .286-volt, measured from the reference black level. The input and output impedances are to be 75 ohms and unbalanced to ground. The standards also specify synchronization pulses not part of a composite video signal are to be negative in polarity and 4 volts in amplitude. In November, 1966, EIA adopted RS-330 — a new standard for closed circuit television cameras. Figure 3-243 shows the composite signal waveform recommended. Note that interlace, front porches, and back porches are required for all signals conforming to this standard. It permits omission of the vertical serrations and equalizing pulses, but the RS-170 signal is compatible with this standard. RS-343, adopted in 1967, is similar to RS-330, but it applies to nonstandard scan rates.

3-10.2 TELEVISION SYSTEMS

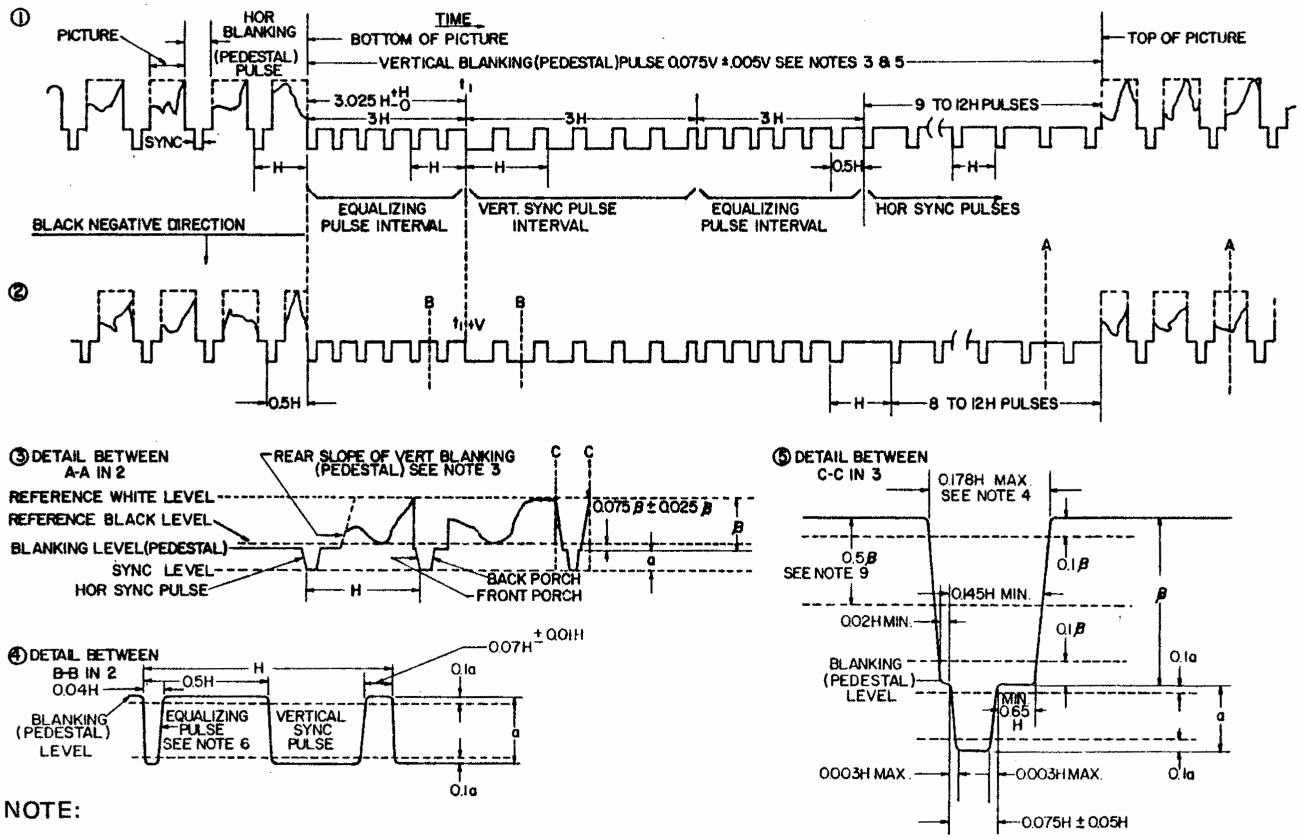
Television signals are normally distributed at a 1.0-volt level for composite signals. In older systems a 1.4-volt value was used. Normal sync levels are 0.286-volt and 0.4-volt, respectively, for the two standards. Video signals are measured with an oscilloscope equipped with a graticule graduated in IRE units. This simplifies the setting of sync, blanking, and video levels (see Figure 3-244). Where reference white is present in the signal, the overall level is 140 IRE units. It makes no difference whether the 1.0-volt or the 1.4-volt composite signal is used. Only the

MONOCHROME TELEVISION
PICTURE LINE AMPLIFIER STANDARD OUTPUT (VISUAL TRANSMITTER INPUT SIGNAL WAVEFORM)

SYNC SIGNAL AMPLITUDE α SHALL BE HELD CONSTANT WITHIN

$\pm$ 4% DURING TRANSMISSION
$\beta = 1.0 \pm 0.05$ VOLTS
$\alpha = 0.4\beta \pm 0.05\beta$

DRAWINGS NOT TO SCALE

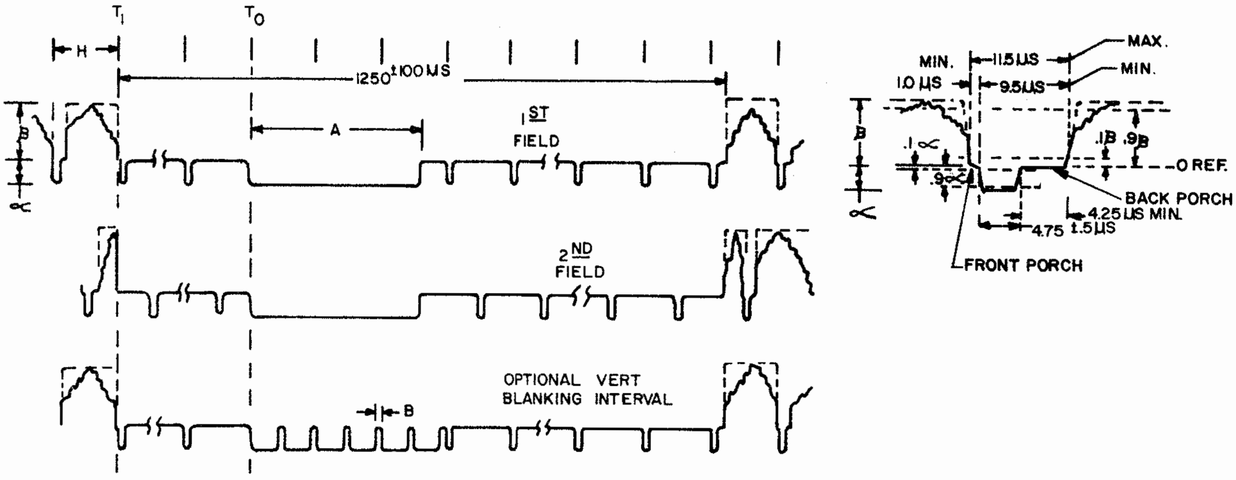


NOTE:

1. H = time from start of one line to start of next line.
2. V = time from start of one field to start of next field.
3. Leading and trailing edges of vertical blanking (pedestal) should be complete in less than 0.1H.
4. Leading and trailing edges of horizontal blanking (pedestal) shall be steep enough to preserve min. and max. values of durations under all conditions of picture content.
5. All tolerances and limits shown in this drawing are permissible only for long-time variations.
6. Equalizing pulse area shall be between 0.45 and 0.5 of the area of a horizontal sync pulse.
7. All rise and decay times shall measure between 0.1 and 0.9 amplitude reference lines.
8. The overshoot on blanking (pedestal) signal shall not exceed 0.02β at the beginning of the front porch and shall not exceed 0.05β at the end of the back porch. The overshoot on sync signals shall not exceed 0.05α. Any other extraneous signals on the blanking (pedestal) signal shall not exceed $.02\beta$
9. For setting aspect ratio, the horizontal blanking (pedestal) duration should be set to 0.175H at the 0.5β point.
10. Rise time and decay time of horizontal blanking shall not exceed 0.003H.

RISE TIME & DECAY TIME OF ALL PULSES 0.003H MAX
(SEE NOTE 7)

Figure 3-242. EIA Standard RS-170 Television Waveform



COMPOSITE VIDEO WAVEFORM 525/60 INTERLACED 2:1

NOTES:

1. β = 0.714 ± .1 volts (100 IRE Units).
2. $\propto$ = 0.286 (40 IRE Units) nominal.
3. Sync to total signal ratio $\left(\frac{\propto}{\beta + \propto}\right)$ = (28.6 ± 5)%.
4. Blanking = 7.5 ±5 IRE Units (2.5% to 12.5% of β).
5. Horizontal Rise times measured from 10% to 90% amplitudes shall be less than 0.3μ/s.
6. Overshoot on horizontal blanking signal shall not exceed 0.02β at beginning of front porch and 0.05β at end of back porch.
7. Overshoot on sync signal shall not exceed 0.05β.
8. T_0 = start of vertical sync pulse.
9. T_1 = start of vertical blanking.
10. $T_1 = T_0 \,^{+0}_{-250\mu/s}$.
11. A - vertical sync pulse = 150 ± 50μ/s measured between 90% amplitude points.
12. Rise and fall times of vertical blanking and vertical sync pulse measured form 10% to 90% amplitudes, shall be less than 5μ/s.
13. Tilt on vertical sync pulse shall be less than 0.1$\propto$
14. If horizontal information is provided during the vertical sync pulse it must be at 2H rate and as shown in the optional vertical blanking interval waveform.
15. B - vertical serration = 4.5 ± .5μ/s measured between the 90% amplitude points. Rise times measured from 10% to 90% amplitudes shall be less than 0.3μ/s.
16. If equalizing pulses are used in the vertical blanking interval waveform they shall be 6 in number preceding the vertical sync pulse and be at 2H rate.

Figure 3-243. EIA Standard RS-300 — Waveform for Closed-Circuit Television

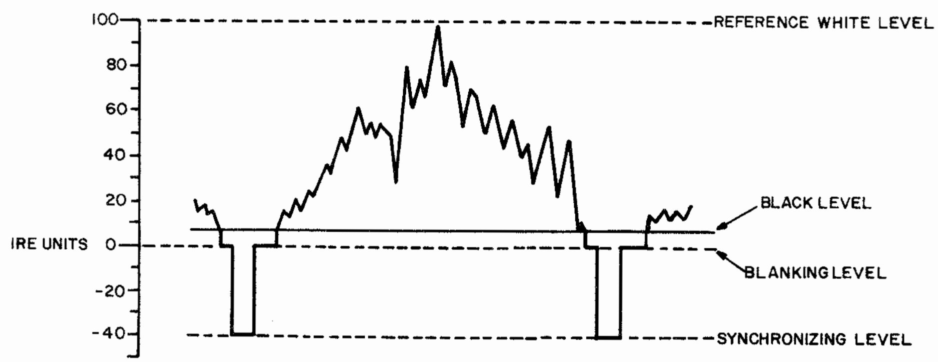


Figure 3-244. IRE Scale for Monitoring Television Signals

lower frequency components of the video signals are measured when video levels are set. These correspond to the broad areas rather than the fine detail of the picture. To simplify this measurement, oscilloscopes designed for television-waveform monitoring are equipped with a low-pass filter, called an IRE or IEEE roll-off filter. Two response curves of television-waveform monitors are shown in Figure 3-245. The difference between the two is not significant except in situations where color signals are being measured.

3-10.3 SYSTEM INTEGRATION

The simple closed-circuit television system, consisting of a television camera and a monitor as illustrated in Figure 3-238, is suitable for surveillance of a single area. This easily constitutes the majority of installations. The addition of a remote-controlled pan-and-tilt unit and zoom lens adds immeasurably to the flexibility of such a system. Often it is desirable to feed a signal to more than one location simultaneously - perhaps to two or more monitors, or to a video tape recorder and two monitors. Figure 3-246 shows the usual method of doing this, using a camera having a 75-ohm output. Addition of equipment to this extent causes no serious distortion of the signal because of the high-impedance inputs available on television equipment. In the example shown, if the monitor inputs are 10,000 ohms each, there is no appreciable effect on the 75-ohm impedance of the video tape recorder, and the total load on the camera remains 75-ohms - the impedance for which it was designed. The loop-through connections shown in Figure-3-246 are possible because most television equipment includes dual input connectors. Either connector may be used for the signal input; the second connector may be used to feed the signal on to other equipment. Figure 3-247

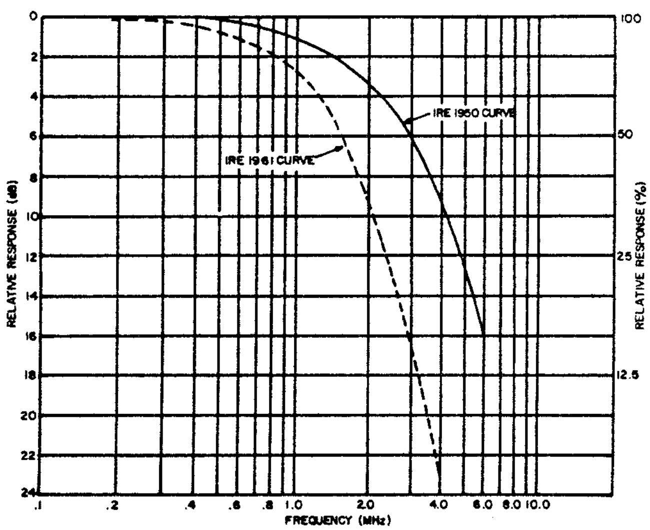


Figure 3-245. Oscilloscope Frequency-Response Characteristics for Television Signal Monitoring

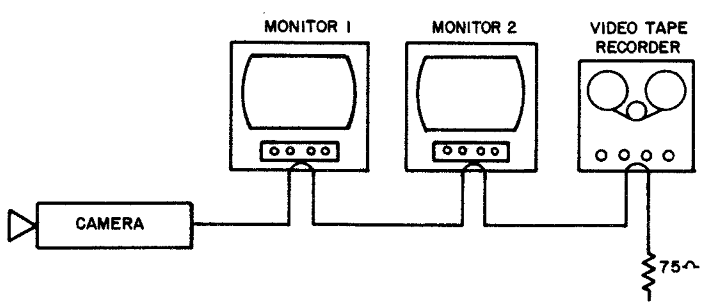


Figure 3-246. Simple Closed-Circuit Television System

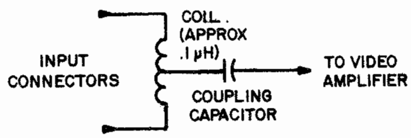


Figure 3-247. Loop-through Input Circuit

shows such a typical circuit. The small coil, consisting of only a few turns of wire, is included on most equipment to minimize the discontinuity of the input. Its value is chosen to equal the inductance of a length of coaxial cable having the same capacitance as the input of the equipment. The last piece of equipment in the string must be terminated in 75 ohms. This is ordinarily done by attaching a 75-ohm resistor to the unused connector. Commercial terminations are available that fit popular connector types, or they may be built as shown in Figure 3-248 to fit uhf connectors. Some equipment has internal resistors which may be switched in or out. Other units, particularly simple video switchers, do not permit loop-through connections since they have single 75-ohm input connectors. If video-distribution amplifiers are not to be added, connect this equipment to the last place in the string. The number of inputs which may be looped together with no noticeable distortion is usually six, although this varies somewhat with equipment characteristics. More can be added if distortion can be tolerated. This appears as light ghosting due to the discontinuities in the coaxial line and loss of fine detail due to combined capacitances of the equipment and cable. The addition of video-distribution amplifiers can extend almost indefinitely the number of monitors fed from a single camera. Figure 3-249 illustrates an improbable distribution network, but it does show how three distribution amplifiers with four outputs each can quickly multiply the outputs of a single camera. Figure 3-250 shows the block diagram of another simple television system. In previously described systems no source of sync pulses has been shown, although there must be

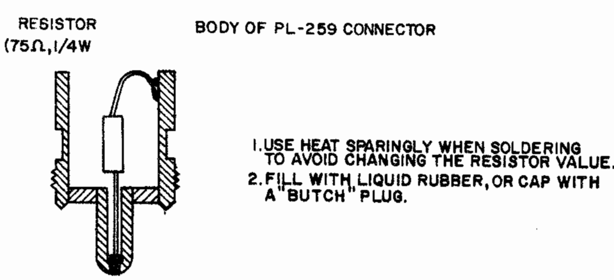


Figure 3-248. Construction of 75-ohm Terminating Plug

one. In this example, the sync generator feeds the blanking and drive signals directly to the cameras. Where a large number of feeds from a single source is required, pulse-distribution amplifiers may be used. When there is a camera-control unit separate from the camera, the signals from the sync generator feed the control unit rather than the camera. Sync is added to the video signal at the video-distribution amplifier in the example shown. An alternate connection would be to feed sync directly to the cameras. However, the connection shown is preferred when a switcher is used because the sync portion of the signal is not interrupted during the switching interval. Thus, a vertical roll at the time of switching is avoided. Alternately, the sync signal may be fed directly to an external-sync input jack on the monitor.

3-10.4 COMPLEX SYSTEMS

A television system may consist of two cameras, each feeding a separate monitor. In many installations, however, one camera may be of interest to viewers at both monitors at a given time. The use of video switchers gives this flexibility. Figure 3-251 shows a possible way of accomplishing this with cameras having dual outputs. Additional monitors, video tape recorders, and other equipment may be added in any of the video paths by using the loop-through connections. If only single outputs are available, it is necessary to add video distribution amplifiers at the outputs of the cameras. When two or more cameras are installed in the same area, it is economical to use one sync generator to supply pulse signals, as illustrated in Figure 3-250. Where a large number of cameras are installed in the same area, a single sync generator may still supply signals to all of them through the use of pulse-distribution amplifiers. These increase the number of outputs of a single sync generator practically without limit. The rules for interconnecting them are the same as for video-distribution amplifiers, except that a larger number of loop-through inputs is now possible. Where large numbers of cameras and monitors are employed, a suitable switching system can add versatility to the system by allowing cameras to feed monitors in whatever combination desired. Of course, video tape recorders or RF transmitters could be connected to some outputs if they were required. External signals, rather than local cameras, may also feed some inputs of the switcher. Figure 3-252 illustrates such a system.

3-10.5 SPECIAL SYSTEMS

The video-insert amplifier makes it possible to distribute timing or other data along with the video picture. Figure 3-253 shows a typical hookup.

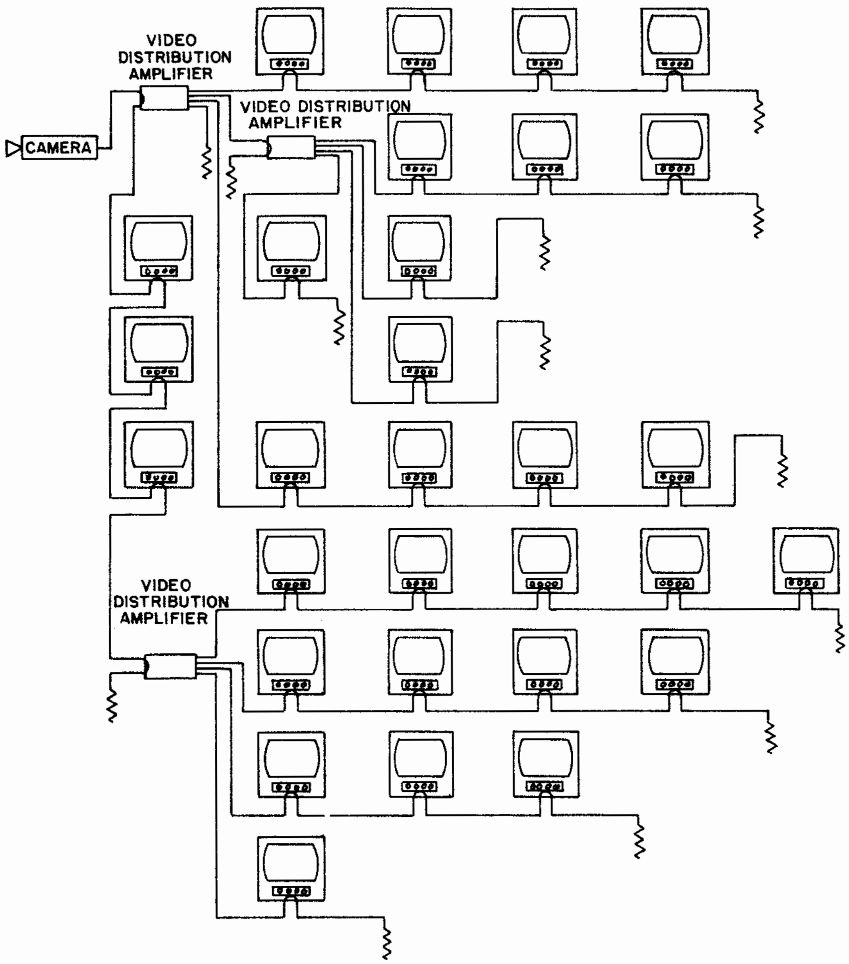


Figure 3-249. Network Using Video Distribution Amplifiers

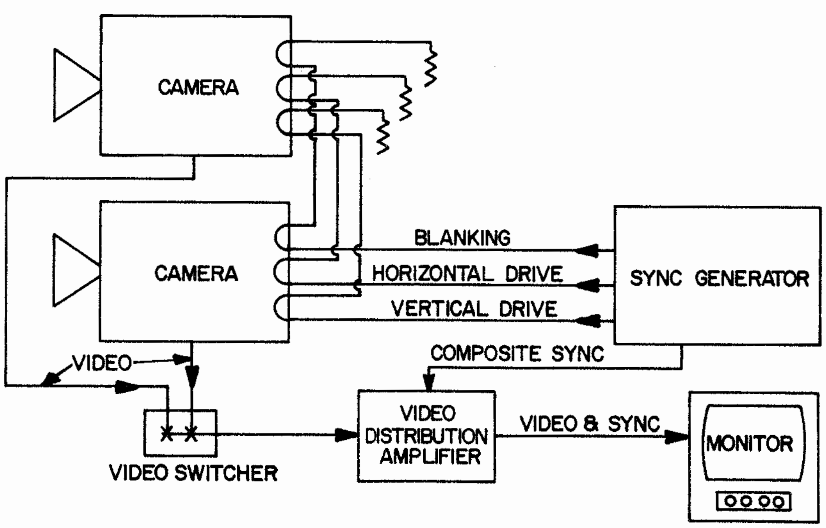


Figure 3-250. Television System Showing Sync Distribution

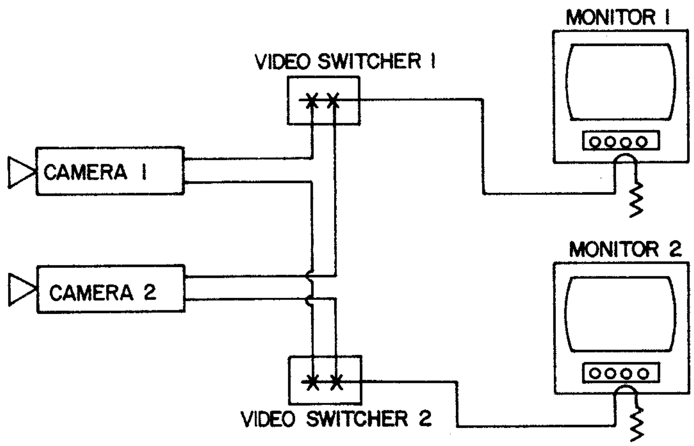


Figure 3-251. Television System with Flexible Video Switching

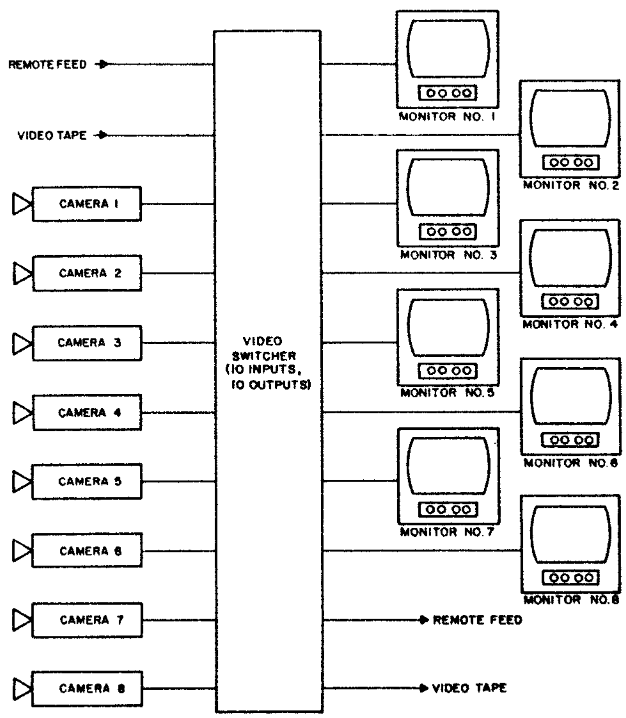


Figure 3-252. Complex Television System

Note that sync is not fed to the camera chains. Signals to the insert amplifier are therefore noncomposite, which is what many of them require. The video-distribution amplifier is used here only to accomplish sync mixing. Alternately, sync could feed the external-sync jack of the monitor. Most video tape recorders, however, require composite video signals. It is important that signals from the two cameras arrive at the insert amplifier in phase. If the distances between the cameras and the pulse-distribution amplifier (or sync generator) differ appreciably, it is desirable to add a delay in the pulse lines feeding the shorter path. Ordinary passive delay lines are suitable for this. Another use of the insert amplifier is to allow the simultaneous display of two

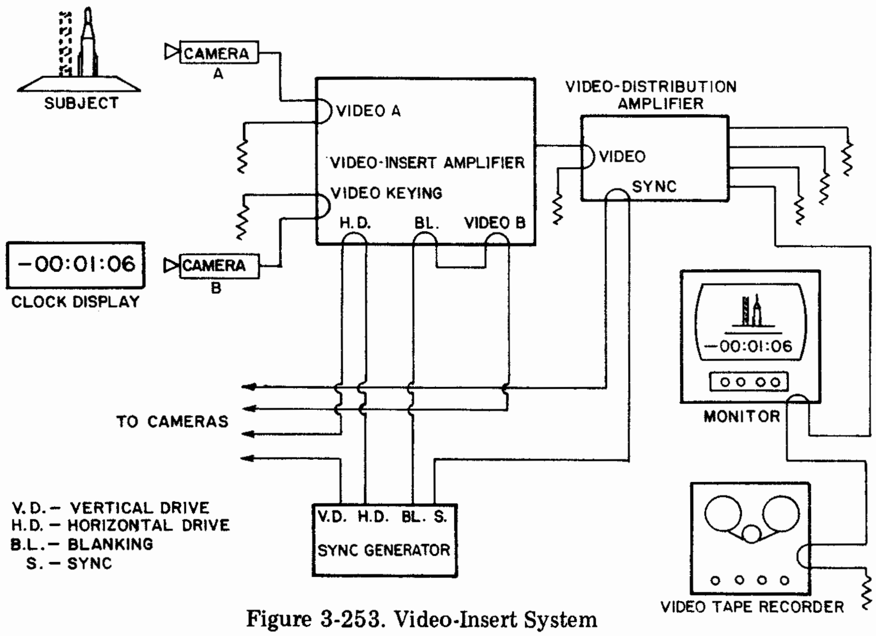


Figure 3-253. Video-Insert System

pictures on the same screen. One reason for so doing would be to preserve positive timing correlation when video taping. Figure 3-254 shows how the various units may be connected. R1, R2, and C1 form a 1-volt sawtooth from the 4-volt horizontal-drive signal. The screen can be split into whatever proportion is desired by adjusting the key level on the insert amplifier. The sawtooth network would be omitted if the insert amplifier has an internal waveform generator.

3-10.6 HIGH-RESOLUTION SYSTEMS

Where more detail is required in the picture than is possible with a 525-line system, the number of lines may be increased to 1200 or more. This, of course, results in a system that is incompatible with those having the usual scan rates. The width of the horizontal-sync and blanking pulses are decreased in proportion to the change in duration of a horizontal line ("H"). When expressed in terms of "H", the values for the particular system are similar to those of the

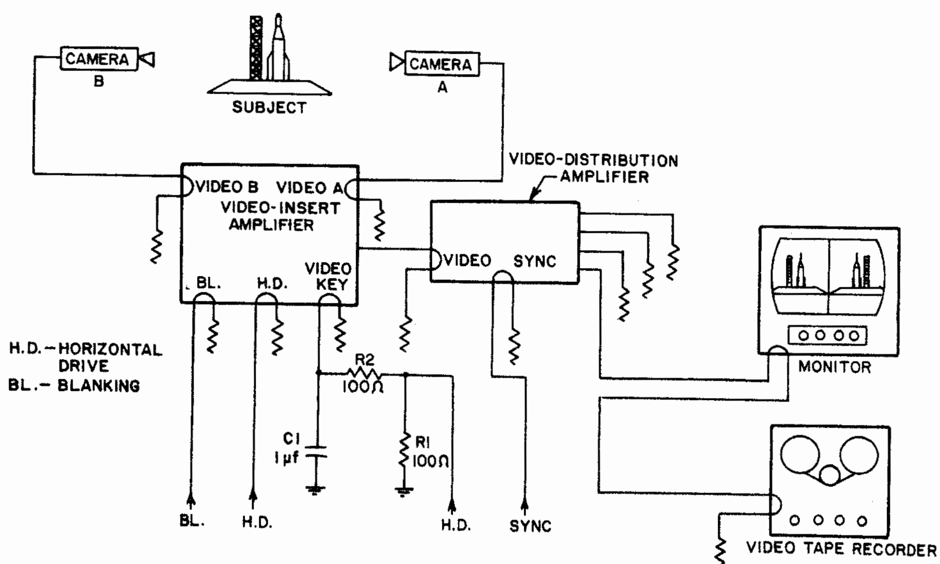


Figure 3-254. Split-Screen Hookup

EIA system. Since the number of vertical fields per second usually remains at 60, the number of lines during the vertical-blanking interval is usually increased to allow sufficient retrace time for the monitor. Thus the interval, measured in absolute time, will remain about the same as in the EIA system. When the number of scanning lines in the system is increased, the bandwidth of the video amplifiers must be increased in order to take advantage of the resolution available. For example, if the number of lines in the picture is doubled, it is necessary to double the video bandwidth in order to maintain the original horizontal resolution. To see horizontal resolution proportionately as improved as vertical resolution, the bandwidth would have to be doubled again. Therefore, if the number of scanning lines doubles, the video bandwidth must be quadrupled to take full advantage of the increase.

3-10.7 SLOW-SCAN SYNCHRONIZATION

Slow-scan, employed where severe bandwidth limitations are necessary, requires a still different synchronization system. These may, and often do, have the number of horizontal lines making up the picture increased far beyond that of the EIA RS-170 system. Since interlace is superfluous here, the field and frame rates become the same. The aspect ratio may be altered in slow-scan systems. One of the best-known slow-scan television systems is that employed in the Ranger moon probes. The wideangle camera in these systems produced pictures comprised of 1152 lines and transmitted one picture every 1.6 seconds. Of course, many variations of slow-scan are possible and are in use. No standard has been established, since these systems are usually tailored to suit their particular requirement.

3-10.8 VIDEO PATCH CORDS AND JACKS

The inclusion of video patch panels provides a high degree of flexibility in signal routing and in bypassing defective equipment. It results in in sets of centrally located test points for maintenance. The plugs and jacks of video patch panels are coaxial. They are available in a variety of types, including normal-through, self-normalling, self-terminating, and simple jacks. These can be arranged in a convenient pattern on panels, with intermixed types if desired. Two rows of from 12 to 24 jacks may fit on a 3 1/2-inch panel. The "simple" jack is just that. It can be used to connect one circuit with another. Extracting the plug breaks the connection - nothing more. When this type of jack is used in video outputs, a 75-ohm termination designed for this purpose should be placed in the jack unless other equipment is connected. A normal-through connector is a pair of jacks in a single assembly. The spacing is such that they fit vertically between two holes in the panel. There are two connections on the rear of the jack - the upper section usually connects to a video source, while the lower section feeds a monitor or other load. When no patch cords are inserted in these jacks, the source feeds internally through the jack to the monitor. Placing a patch cord into either portion breaks the connection internally. To connect the video source to a different monitor, plug into the top section. Conversely, if a different video source is to be connected to the monitor, plug into the lower section. In this case, insert a terminating plug into the upper jack (the output for a video source). An alternate to the normal-through connector is the self-normalling jack. It has a single plug-in connection on the front with two connections on the back. A pair of these is required for a two-way normal-through connection as described previously. Where the self-normalling jack is the output for an unused video circuit, connect a 75-ohm termination to the lower jack which would normally feed the monitor. Simple jacks may also be used in normal-through connections by inserting jumper plugs ("hairpins"). The jumper plug fits neatly between two vertically adjacent jacks. Of course patch cords would serve the same purpose, but the jumper plugs make a much neater physical appearance. The self-terminating jack resembles the simple jack externally, but it has an internal 75-ohm resistor. This is cut out of the circuit when a plug is inserted into the jack. When the plug is removed, the resistor is internally switched across the jack. This ensures that the video source feeding the jack has a suitable termination across it at all times. The foregoing types of jacks, and probably every possible variation, are available from one of the many manufacturers. Two basic sizes may be purchased: the WECO and the RCA. Neither type of plug fits the other type of jack although the appearance is very similar. Figure 3-255 shows a few of the commonly used types. These may be installed on panels available from manufacturers.

3-10.9 GROUNDING

Exposed surfaces of frames, racks, chassis, camera pedestals, and pan-and-tilt units should be connected to an earth ground as a safety precaution. Though most equipment is grounded through the third wire of the ac power cords, most installers prefer not to rely entirely upon this. They connect separate ground cables between equipment racks and an earth

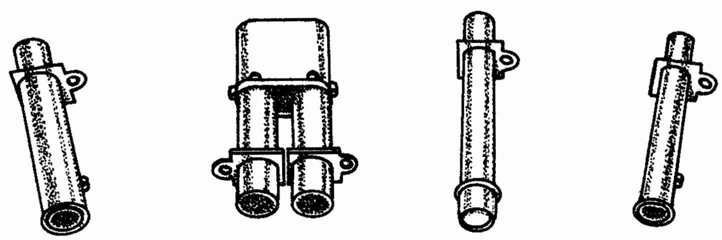


Figure 3-255. Some Types of Video Jacks

ground of known quality. How to eliminate ground loops when video signals are transmitted over unbalanced coaxial cables that must have their shields connected at both ends to maintain signal paths? Here are three ways:

1. Short out the ground loop by connecting a heavy ground wire from each piece of equipment to a common ground point. This may prove impractical where equipment is spread over a large area.
2. Use monitors with differential inputs to cancel the hum signals.
3. Use a clamper amplifier at the receiving end of the video line. The clamping feature is provided in stabilizing amplifiers also.

Camera cables pose a more complex problem. Where camera assemblies must be mounted on a steel framework, ground paths may exist through both the framework and the coaxial-cable shield. Isolating the camera housing from the camera circuitry will break the ground loop, but this may result in electrostatic pickup of 60-hertz or other interference by sensitive components of the video amplifier. Internal shielding or a dual housing usually provides a satisfactory solution.

3-10.10 TYPES OF TELEVISION CABLE

Television video and pulse signals are normally transmitted over 75-ohm unbalanced coaxial cables. The RG types of coax are used most often because of their uniformity, wide availability, and the many standard types of connectors available. Figure 3-256 shows the physical characteristics for two of the most popular coaxial cables. Most RG coaxes have jackets of black polyvinyl chloride. Types of I and IIa

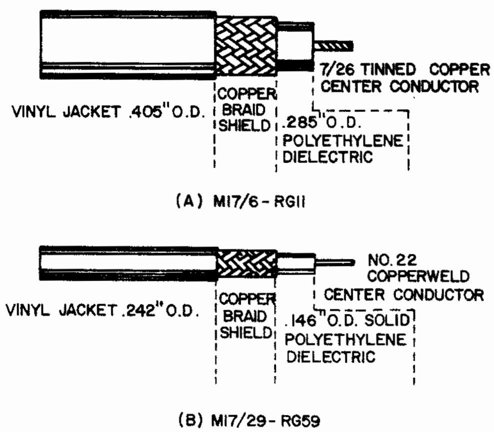


Figure 3-256. Coaxial Cable Dimensions

are most commonly used, with Type IIa being preferred. The plasticizer of Type I eventually evaporates, leaving the jacket stiff and brittle. Some of this plasticizer finds its way through the coaxial braid to the dielectric, contaminating it and increasing attenuation. The noncontaminating type - IIa - retains its flexibility and attenuation characteristics indefinitely in spite of constant use. The M17/6-RG11 cable is favored for critical video paths. Loss at 10 MHz is about 0.7 dB for 100 feet. The M17/29-RG59 coaxial cable is physically smaller than the M17/6-RG11 cable. The smaller size, lighter weight, and greater flexibility of this cable make it a popular choice for wiring between racks, especially where equipment slides are used. It is the cable most favored for video patch cords and is used in high-density areas for the same reasons. Attenuation is about 1.1 dB per 100 feet at 10 MHz.

3-10.11 COAXIAL CABLE CONNECTORS

Two types of coaxial connectors are used for virtually all video and pulse cables - the BNC and the so-called UHF connectors. Which is to be used depends on the equipment. (Some equipment manufacturers offer either style). Neither type of connector has the 75-ohm characteristic impedance of the coaxial cable, but this is not significant at video frequencies. Both BNC and UHF connectors are available in a number of variations suitable for RG-11 and RG—50 cable. Both come in crimp and solder types; the former forms the more rugged and reliable connections. The smaller physcial size of the BNC connector makes it a popular choice for high density installations, such as patch panels and distribution amplifiers.

3-10.12 COAXIAL ADAPTERS

Coaxial adapters may be required for some test setups or for interfacing certain test equipment with the system equipment. Figure 3-257 illustrates some typical coaxial adapters. The most frequently used adapters are:

1. Right-angle adapters. These have a male fitting on one end and a female on the other. They are used to avoid sharp bends in the cable when it approaches the receptacle from the side.

2. Double female connectors. These are used to join two pieces of coax together. The UHF type is commonly called a "barrel" because of its appearance.

3. T connectors. These are often required for test purposes, but should be avoided in installations.

4. 75-ohm terminations. These are simply male connectors that contain a 75-ohm resistor between the center pin and the shield. They are used to terminate loop-through connections, to terminate unused video outputs, and to provide connecting points required in test setups.

5. Adapters between series. These allow the connection of a BNC-terminated cable to a UHF jack or vice versa. They are sometimes used in installations when it is desired to standardize on a single type of connector for all coaxial cables. A supply of these is necessary for test setups since test equipment may have either type of connectors.

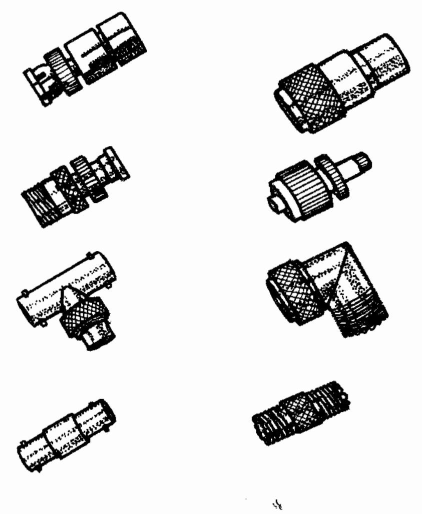


Figure 3-257. Types of Coaxial Adapters

3-10.13 CAMERA CABLES

Camera cables are available in a number of standard types plus several proprietary types used by camera manufacturers. Typically, camera cables contain 12 or more wire conductors in addition to two or three coaxes of 50- and 75-ohms impedance (occasionally 92 ohms). These may be all the same or a mixture of types. A portion of the single conductors is usually arranged in one shielded bundle which is reserved for mechanical functions, such as pan-and-tilt control and zoom-lens control. These minimize the coupling of motor noise to sensitive video and control circuits. Finally, the entire bundle is shielded and covered with an insulated jacket which may be vinyl, neoprene, or a special material, depending on whether light weight, durability, or environmental compatability is of prime importance. Table 3-23 gives the physical characteristics and dimensions of several camera cables. Video and horizontal-drive signals are always transmitted through coaxial conductors. Vertical drive may utilize a coaxial conductor or a twisted pair of conductors included for this purpose. The remaining single-wire conductors are used for camera power, control voltages, communications, pan-and-tilt, zoom-lens, and light-control circuits as required. Most standard cameras use either Amphenol MS or Cannon FK connectors. In special cases, other types are sometimes chosen; for example, Bendix QWL connectors are favored for rugged use and severe climates. Special types are required for underwater use.

Table 3-23. Characteristics of Camera Cables

Cable Type	Coaxials	Control Wires					Overall Shield	Jacket Type and Diameter
		Shielded Groups			Single Wires			
	Qty. Type	No. of Wires in Group	Wire Type (AWG)	No. of Groups	Qty.	Type (AWG)		
Boston Insulated Wire & Cable Co. BIW TV-25TN	3 50-ohm				4 18	14 22	85%	Black Neoprene, .840" max.
BIW TV-33P	3 75-ohm	4	19	3	18	24	85%	Black vinyl, .800" max.
BIW TV-33N	3 75-ohm	4	19	3	18	24	85%	Black Neoprene, .810" max.
Belden Corporation Beldon 8280	3 51-ohm				4 21	18 22	85%	Chrome vinyl, .750" nom.
Belden 8282	1 72-ohm 1 50-ohm				9 2	22 18		Chrome vinyl, .470" nom.
Belden 8284	3 53.5-ohm	3	22	1	14 4	22 16	85%	Chrome vinyl, .800" nom.
Belden 8286	3 75-ohm	4 7	18 22	1 1	14	22	85%	Black Hypalon, .730" nom.

Most camera manufacturers supply cable assemblies of any predetermined length. However, in many cases the camera cable must run through conduit. Since connectors have a much larger diameter than the cable itself, it may be necessary to run the cable through the conduit before the connectors are attached. The latter is a tedious job, but it should be done with the utmost of care since a poor job will result in poor reliability and may cause picture deterioration. Coaxial conductors which have relatively stiff dielectrics are especially difficult to work with since they may become noncoaxial as they pass through pin-tube interconnections. If only enough shielding is stripped so the cable can flex, distortion and crosstalk will be held at a minimum. The use of "hyring" connections for shields assures good electrical contact without the risk of damaging the cable by the heat of a soldering iron. Likewise, crimp-type pins give electrical connections equal to soldered ones and usually result in a neater, more reliable assembly and a quicker installation.

3-10.14 CATHODE-RAY TUBES

The cathode-ray tube is a special kind of vacuum tube. Electrons emitted by the cathode are formed into a narrow beam, deflected by electric or magnetic fields, and directed against a specially prepared surface that may be the fluorescent screen of a picture tube or the sensitized surface of a camera tube. Because the electron beam is so light in weight, it reacts readily to external fields. This property makes it useful for the high-speed scanning required by the television system; it is also taken advantage of in oscilloscopes, radar, and microwave tubes.

3-10.14.1 Construction

The principal features of cathode-ray tubes are: a glass envelope, an electron gun, a focusing method, a deflection method, and a screen (target). Electron guns of all CRTs are similar in principle. Tubes may use electromagnetic or electrostatic methods of focusing the beam. Those using the former nearly always use electromagnetic deflection, while those using the latter may utilize either electromagnetic deflection, or electrostatic deflection. In picture tubes and oscilloscope tubes the beam lands on a fluorescent screen. The beam of a camera tube strikes a photosensitive surface (in the vidicon and the plumbicon) or a glass target (in the image orthicon).

3-10.14.2 The Electron Gun

Figure 3-258 shows a simplified cathode-ray tube. The directly heated cathode emits rays of electrons that are attracted to the positively charged anode disc. Some of the electrons pass through a small aperture and continue on to the screen with no appreciable loss in velocity, but most strike the anode. The electrons tend to repel one another since they are carrying equal negative charges. However, very little actual scattering takes place due to the high velocity they have. A practical cathode is shown in Figure 3-259. It consists of a cylindrical sleeve with an end cap and is heated by an internal filament that is electrically insulated from the sleeve. The principle source of electrons is an oxide coating at the end of the cathode. The addition of a control grid substantially improves the efficiency of the cathode and provides a means of controlling the intensity of the electron

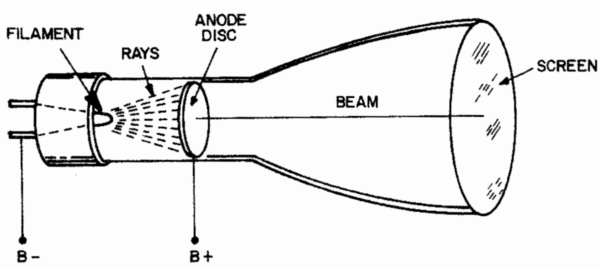


Figure 3-258. Simple Cathode-Ray Tube

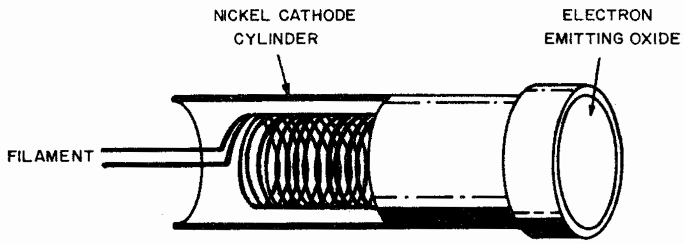


Figure 3-259. Cathode of Cathode-Ray Tube

beam. In the cathode-ray tube of Figure 3-258 only a small fraction of the electrons emitted by the cathode passes through the aperture. In Figure 3-260 a control grid consisting of a small metal cylinder placed around the cathode has been added. The end is covered by a disk having a small opening which controls the direction in which electrons are emitted. Figure 3-260 shows the relationship of the control grid to the cathode and the anode, and the lines of force between them. The field of the first anode extends through the control-grid aperture to the cathode. If an electron is emitted in any direction other than directly toward the center of the opening, its path will be changed by the electrostatic field. It turns in the direction of the arrows on the lines of force which it crosses. An electron which tends to move in direction KA will be curved along path KCP. Likewise, an electron that tends to move in direction LB will follow path PDP. The paths cross at point P. This holds true for all electrons that pass through the opening. The concentration of the electrons at the crossover point in many respects resembles the focusing of light by an optical lens system. For this reason, the combination of the cathode, control grid, and anode is often called an electrical lens. The control grid determines the number of electrons in the beam. If it is made more negative, the effect of the anode is lessened, and fewer electrons will follow paths which take them through the opening. Likewise, as the control grid is made more positive, more electrons will pass through the opening, and the intensity of the beam increases. Intensity-modulated tubes, such as picture tubes, use this principle to control the brightness of the spot on the screen. In camera tubes the beam is not modulated except for the blanking signal.

3-10.14.3 Electrostatic Focus

The control gird and the first anode cause the electron beam to focus at a point very close to the cathode. However, the beam diverges after passing this point. Additional focusing is needed to cause the beam to converge again at the screen. Two such methods are possible: electrostatic, and electromagnetic focusing. In its simplest form, the former

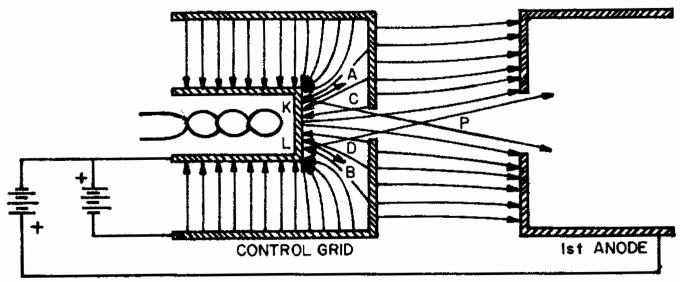


Figure 260. Action of Control Grid and First Anode

requires only the addition of a second anode, as shown in Figure 3-261. This anode is charged positively with respect to the first and has a hollow cylindrical shape. Figure 3-262 shows the relative position of the two anodes and the lines of force between them. An electron introduced into this field will tend to follow the lines of force toward the second anode. If the electron entering the field has sufficient velocity, it will pass out of the field before it is pulled completely over. The field will divert the electron slightly from its original path; the extent depends on the distance from the electron to the center axis of the anodes. The farther the electron is from the center axis, the more it will be diverted. For example, if an electron comes from the crossover point along line XAB, it will cross lines of force a through c, and each will push it toward the center axis. The electron will thus change from AB to the path AC. As it passes point c it is traveling nearly parallel to the lines of force, so it will be accelerated rather than deflected by the relatively weak field at the right side of the second anode. The forces acting on the electrons farthest from the center axis are greater than those acting upon electrons traveling along the center axis. Consequently, all electrons will converge again at point G - on the screen of the CRT. The focus of a tube of this type is controlled by adjusting the voltage on the first anode. This varies the amount of force which the field exerts upon the electron beam,

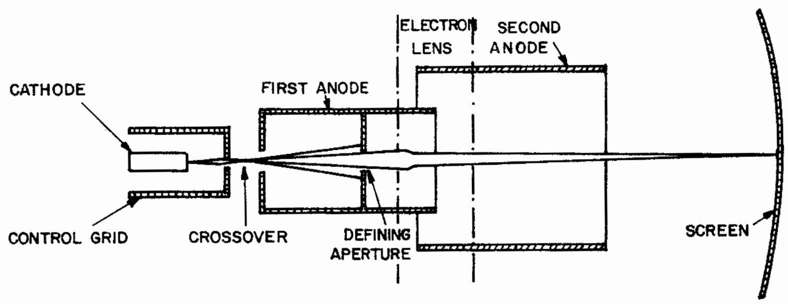


Figure 3-261. Electrostatic Focusing Elements

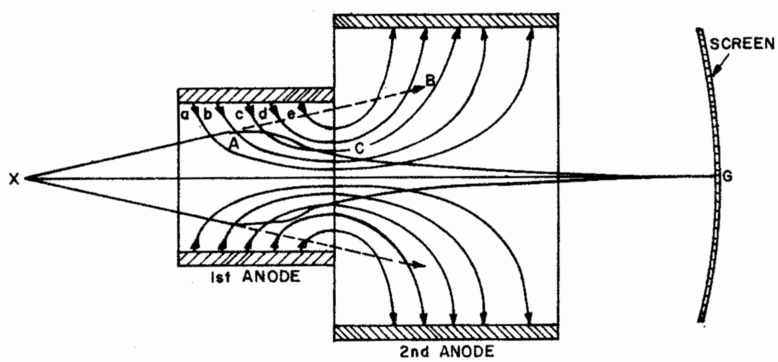


Figure 3-262. Electrostatic Field Between First and Second Anodes

and, therefore, the point at which the beam will converge. The first anode normally contains one or more defining apertures - internal discs containing small holes. Their purpose is to block electrons which are outside the main stream of the beam. This keeps the diameter of the beam within a prescribed limit. Figure 3-263 shows the arrangement of focusing elements in a practical cathode-ray tube employing electrostatic focus. The additional electrode, called the accelerating grid, reduces interaction between the focusing and control functions. It is connected internally to the second anode. The inner surface of the front portion (face) of picture tubes is normally coated with aquadag - a thin layer graphite. This is connected electrically to, and is considered part of, the second anode. Connection is normally made through a terminal located in the flared portion of the glass envelope. In addition to the accelerating function, the coating collects secondary electrons emitted from the screen. If these were allowed to accumulate, they would create a negative charge that would interfere with the operation of the tube. Other functions of the coating are to provide electrostatic shielding for the electron beam and to reduce light reflection from the walls of the tube. Most picture tubes also have an outer graphite coating which is normally grounded. This and the inner coating form a capacitor with the glass envelope as a dielectric. This provides filtering of the ultor (second anode) voltage in addition to that normally found in the power supply.

3-10.14.4 Electromagnetic Focus

Electromagnetic focus is used in many cathode-ray tubes, particularly camera tubes and the earlier types of kinescopes. Here an external electromagnetic coil is used to create a magnetic field that focuses the beam. As with electrostatic tubes, the first crossover point is refocused on the screen. Figure 3-264 shows a magnetic focusing field inside the neck of a cathode-ray tube. The focus coil for kinescopes is usually wound inside a soft iron ring having an annular groove. The presence of iron makes it possible to set up the desired field strength in the tube with a minimum of current, and allows the field to be concentrated in the area where it is useful. The air gap is

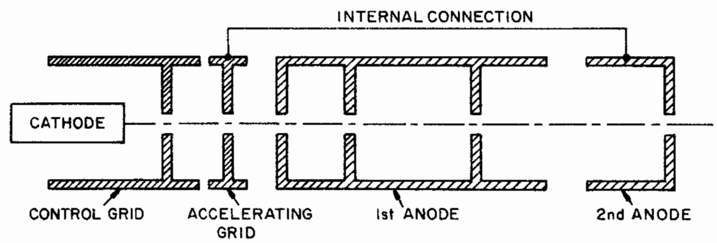


Figure 3-263. Practical Electrostatic Focus System

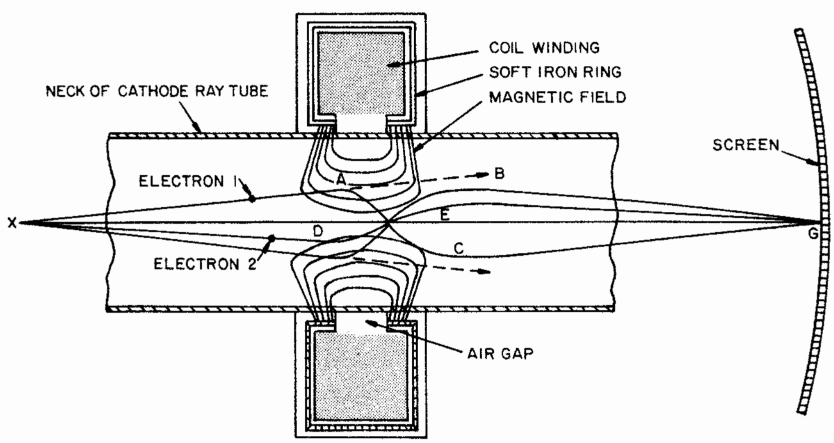


Figure 3-264. Electromagnetic Focusing

necessary in order to direct the field to the inside of the tube. If there were no gap, the iron would shield the field from the tube. Direct current passes through the coil to create the desired field. Normally there is a means of adjusting the coil along the axis of the tube for rough focusing. Fine focusing is controlled by adjustment of the current through the coil. Permanent magnets are also suitable for establishing the focus field and are extensively used in such applications especially for picture tubes. The permanent-magnet focusing unit usually consists of three or four bar magnets mounted between two iron rings and placed over the neck of the tube. The strength of the magnetic field is adjusted either by diverting the field from the tube with an adjustable magnetic shunt, or by controlling the width of the air gap with an adjustable sleeve under the assembly. Consider, for example, an electron entering the field along line XAB of Figure 3-264. If its diagonal motion is separated into horizontal and vertical parts, the horizontal component is parallel to the tube axis and to the magnetic field. The vertical component is at right angles to the horizontal and cuts across the magnetic field. In the instance shown, the electron, which has a vertically upward component, will tend to come up out of the page. As it moves in this direction it remains in the magnetic field and is continually deflected, describing a circular path through the field. The effect of the magnetic field on the motion of the electron, therefore, is to make it travel in a helical path, like the thread on a screw. The effect of the field is strongest on electrons that are farthest off the axis. As a result all electrons are directed toward a single point. The position of the point depends on the speed of the electrons and the strength of the magnetic field. In camera tubes that have the magnetic focusing field extending the full length of the tube, the process is a continuing one; there may be several points at which the electrons converge. Any of these may be chosen as the point of focus. An end view of the paths that the electrons follow is shown in Figure 3-265. Electron 1 of Figure 3-264 enters the magnetic field at A. From X to A its motion is along a straight line. While the electron is under the influence of the magnetic field (from A to C), it follows a circular path. At C the electron leaves the magnetic field and travels in a straight line again until it reaches G on the screen. Electron 2 enters the field at a smaller angle to the axis. However, it takes this electron just as long to pass through the field as it did the first. All electrons are heading for the same point when they leave the field. If the strength of the field is adjusted so that this point is at the screen of the tube, the beam will be focused. A second method of focusing the beam is to adjust the speed of the electrons so they will

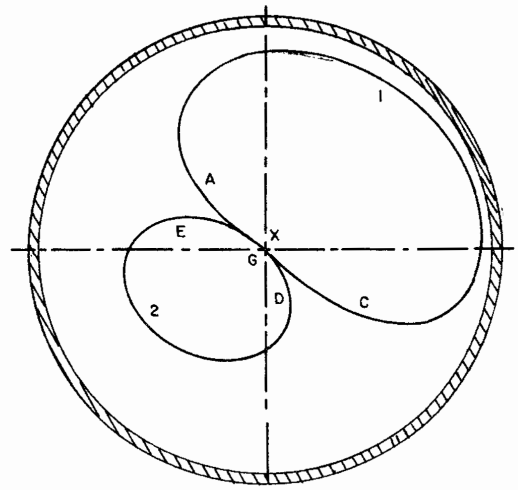


Figure 3-265. End View of Electron Paths in Uniform Magnetic Field

converge at the screen. This method is usually used in camera tubes. In a practical cathode-ray tube, an additional electrode (accelerating anode) is added ahead of the ultor (second anode). This isolates the focus and beam-modulation functions, and may be used in the ion-trap function.

3-10.14.5 Electrostatic Deflection

Since an electron has a negative charge, it is therefore repelled by negatively charged objects and attracted to positively charged ones. Figure 3-266 illustrates an electrostatic field between two charged plates. If an electron is shot into this field, it will be deflected in the direction of the arrow on the line which it is crossing. If the electron is moving fast enough, it will pass out of the field without striking the positive plate, but its path will be changed. The amount of deflection depends on the speed of the electron and the strength of the field between the two plates. A slow-moving electron is in the field for a

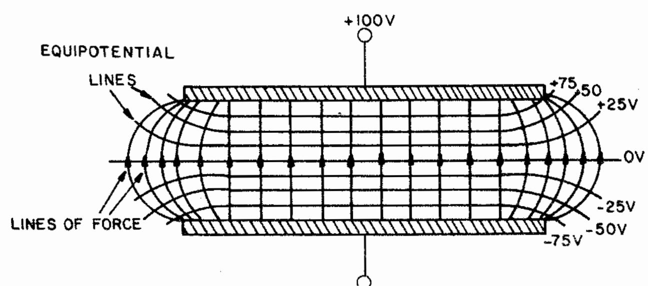


Figure 3-266. Electrostatic Field Between Two Parallel Plates

relatively long time; therefore, the field has a relatively long time in which to act. It will pull this electron closer to the plate than it will the high-speed electron that is in the field for a relatively short time. As the strength of the field increases, the amount the electron is deflected increases because the higher voltage exerts a greater force. An electrostatic cathode-ray tube has two sets of deflection plates mounted at right angles to each other (Figure 3-267). The pair of plates at the top causes the beam to move in a vertical direction; the pair in the middle causes it to move in a horizontal direction. At the bottom of the illustration the two are shown combined. The plates normally diverge at the right to prevent the beam from striking the plates at maximum deflection. A television raster can be produced by applying the proper sawtooth voltages to the deflection plates. Electrostatic deflection is used only in the very smallest picture tubes. Cathode-ray tubes of oscilloscopes and a few vidicons are equipped with this type of deflection.

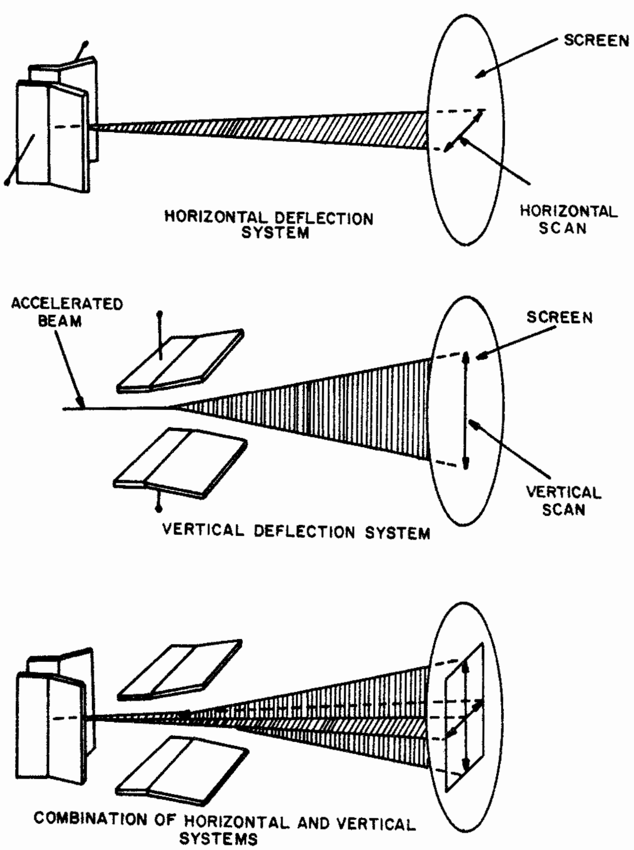


Figure 3-267. Electrostatic Deflection System

3-10.14.6 Electromagnetic Deflection

A beam of electrons has a magnetic field surrounding it similar to a wire carrying electrons. An external magnetic field will affect the beam as though it were a wire carrying a current. Figure 3-268A shows a uniform magnetic field. If an electron beam is placed in it, the result of normal interaction between their fields is shown by the distorted lines of force in Figure 3-268B. The force acting on the beam tends to move it in the direction shown because the lines of force try to take the shortest possible path. In Figure 3-268C, the polarity of the magnets (and direction of the field) is reversed, so the beam is moved in the opposite direction. Figure 3-269 shows a common arrangement using two paris of deflection coils. In the absence of other magnetic fields, the pair above and below the neck of the tube controls the position of the beam horizontally, and the pair to the left and right controls its position vertically, since an electron beam is deflected at right angles to a magnetic field. Where the deflection coils are located within the field of a focus coil, as is the case with most camera tubes, the positions of the coils are rotated 90 degrees. The amount of deflection depends on the strength of the magnetic field; this, in turn, depends on the amount of current through the coil. Where a linear scan is required and the deflection coil contains a significant amount of inductance (as in the horizontal-deflection circuit of a television monitor), the shape of the deflection-voltage waveform will deviate from a sawtooth. It will become more or less trapezoidal as shown in the proportion of inductance to resistance changes in the deflection coils. The deflection-coil assembly is commonly called a yoke.

3-10.14.7 The Ion Trap

An electron beam in a cathode-ray tube will inevitably contain ions. These are caused by either residual gas molecules trapped in the tube or by molecules of cathode material. A negative ion possesses the same charge as an electron and is sensitive to the same accelerating voltages. For tubes having electrostatic deflection, the ions are deflected the same as the electrons and cause no serious problem. Where electromagnetic deflection is employed, ions are deflected very little; consequently they strike the screen in a steady stream near the center. In time the screen will become damaged at this point by these heavy particles. Many modern picture tubes use a thin coating of aluminum behind the screen to prevent this damage. To prevent damage to the fluorescent screen of older type picture tubes, two types of ion traps are used. The first consists of a bent electron gun as shown in Figure 3-270. Both ions and electrons are accelerated

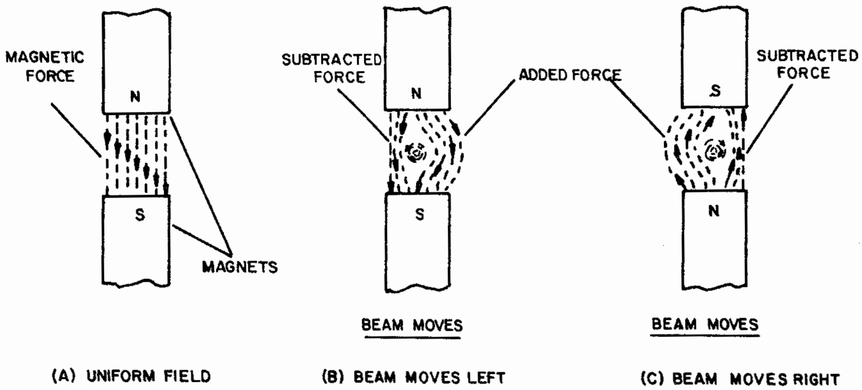


Figure 3-268. Electron Beam in a Magnetic Field

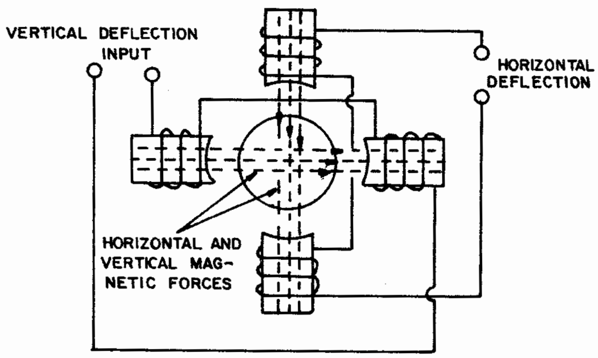


Figure 3-269. Arrangement of Deflection Coils

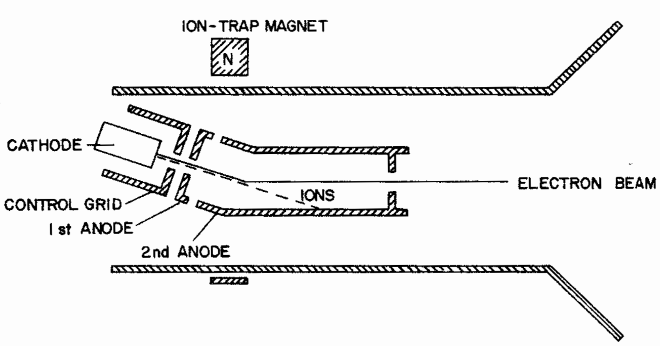


Figure 3-270. Bent-Gun Ion Trap

by the first electrostatic lens. If no other forces acted upon them, all would strike the side of the second anode. The small ion-trap magnet (beam bender), external to the tube, deflects the electrons so that they travel down the center of the tube, but it has little effect on the ions. Thus, the ions are eliminated from the beam. A second ion trap, called the tilted-lens type, is shown in Figure 3-271. Here, the electron gun is on the axis of the tube, but the gap between the first and second anodes is at an angle. The uneven electrostatic field at this point causes both ions and electrons to be deflected downward. If no other forces acted on them, all would strike the wall of the second anode. However, the ion-trap magnet causes the electrons to be deflected upward but has little effect on the ions. They travel harmlessly to the second anode, while the electrons proceed down the center of the tube. This type of ion trap may utilize one or two small magnets. Where two magnets are used, the one nearest the tube socket is the principal magnet. The second magnet, weaker than the first, is located just past the gap between the two electrodes. Its function is to center the beam in the gun structure, giving improved beam geometry. The ion-trap magnet should be positioned longitudinally and angularly for the brightest possible picture. Improper adjustment will cause a dim picture or no picture at all.

3-10.14.8 Picture Tubes

The picture tube (kinescope) is the cathode-ray tube used in television monitors and receivers for picture display. In it the beam strikes a fluorescent screen, and a portion of the energy contained in the beam is converted to light. The remaining energy is expended as heat, in knocking secondary electrons off the screen, and in exciting x-rays from the screen. The light-emitting materials used in coating the screen are called phosphors. The property of emitting light when bombarded by electrons is called phosphorescence, afterglow, or persistence. Television screens normally have both fluorescent and

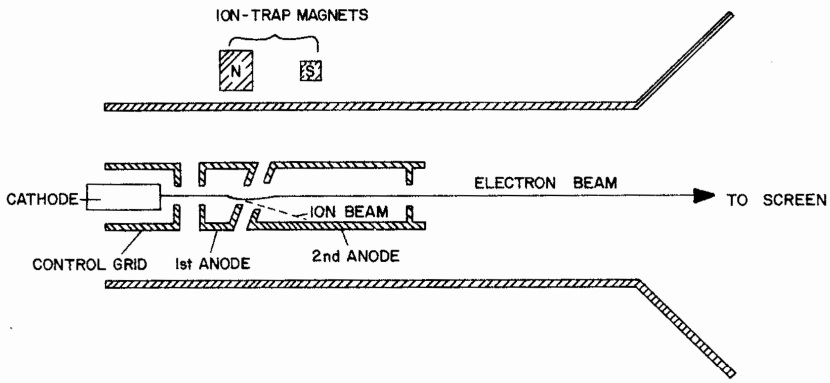


Figure 3-271. Tilted-Lens Ion Trap (2-Magnet Type)

phosphorescent properties. Some of the phosphors used in cathode-ray tubes for television and other purposes are:

1. Willemite (zinc orthosilicate), for predominantly green light.
2. Zinc oxide, for predominantly blue light.
3. Zinc beryllium silicate, for predominantly blue light.
4. Zinc sulphide mixed with cadmium zinc sulphide or zinc beryllium silicate, for white light.

All fluorescent materials have some phosphorescence. The duration (persistence) depends on the type of material and the energy in the beam causing the emission of light. Radar screens use tubes having a comparatively long persistence. Most oscilloscope screens use phosphors of a shorter persistence. Television display screens use phosphors having a still shorter persistence. Some of the most common phosphors are listed in Table 3-24. The intensity of the spot on the screen depends on the speed of the electrons and the number of electrons striking the screen at a given point. This leads to two possible methods of controlling intensity. One is to vary the speed of the electrons. This is not practical, since other adjustments are necessary when the speed of the electron beam is changed. The second and customary way of controlling the intensity of the spot is to vary the number of electrons in the beam, normally at the cathode or the control grid. Type numbers for picture tubes are coded to indicate the approximate size of the tube and the type of phosphor. The first number gives the approximate diagonal measurement; the last numbers (starting with the letter P) give the type of phosphor. For example, 23KP4 indicates that the tube has an approximate diagonal measurement of 23 inches and uses a P4 phosphor. The letter K (on some types there are two letters in this position) simply distinguishes this tube from others of the same size which use the same phosphor. The brightness and contrast of the picture-tube display can be enhanced by the addition of a thin film of aluminum behind the fluorescent screen. The film is thin enough to allow the electron beam to pass through but thick enough to act as a light reflector. The fluorescent screen generally radiates light in all directions from a given point when bombarded by electrons, so the light energy which would normally go into the tube is reflected out the front by the aluminum backing. It also conducts secondary electrons away from the screen and eliminates the need for an ion trap.

3-10.14.9 The Vidicon

The number of vidicons used in television easily surpasses all others combined. Reasons for the popularity of the vidicon are:

1. The most rugged tube.
2. The most compact tube.
3. Available in several sizes.
4. Simple operation.
5. Lowest initial cost.
6. Simple installation.
7. Excellent signal-to-noise characteristic.
8. Long life.
9. Ruggedized types available for severe environments.
10. Less sensitive to temperature than the image orthicon.
11. Lends itself easily to automatic sensitivity control.

Table 3-24. Phosphors for Cathode-Ray Tubes

TYPE	CHARACTERISTICS			
	FLUORESCENCE	PHOSPHORESCENCE	PERSISTENCE	USE
P1	green	green	medium	oscilloscopes
P2	blue-green	green	long	oscilloscopes
P4 sulfide	white	white	short	picture display
P4 silicate	white	blue	medium	picture display
P4 silicate-sulfide	white	blue	medium	picture display
P5	bluish	bluish	very short	kinescope recording
P7	blue-white	yellow	long	radar displays, slow-scan signals
P11	blue	blue	very short	kinescope recording
P15	blue-green	blue-green	very short	flying-spot scanners
P16	violet	violet	very short	flying-spot scanners
P22	See Note 2			color display
P24	blue-green	blue-green	very very short	flying-spot scanners

1. High ultraviolet content.
2. Tricolor phosphor: red, blue, and green.

12. Moderate sensitivity — usable under normal room lighting.

13. X-ray, infrared, and ultraviolet types available.

14. Standard motion-picture lenses usually suitable — 16mm for one-inch tubes, 35mm for 1 1/2-inch tubes, and 8mm for 1/2-inch tubes.

15. Good resolution — the 1 1/2-inch type offers resolution equal to that of any other popular type of tube.

The foregoing list nearly precludes the use of any other type of tube, ever. When another type is employed, it is generally for one of the following reasons:

1. Lag (sometimes called persistence or retentivity). The vidicon has a longer lag time than the other types. This is noticeable as a smear effect when moving objects are viewed in dimly lighted scenes. Lag is generally not troublesome in industrial or scientific applications.

2. Sensitivity. Image orthicons and SEC tubes have appreciably more sensitivity than vidicons; however, this is substantially canceled by the generally faster lenses available for vidicon cameras. Special image orthicons are available for extremely low light conditions (such as moonlight).

Physical construction of the vidicon can be seen in Figure 3-272. The tube consists of an electron

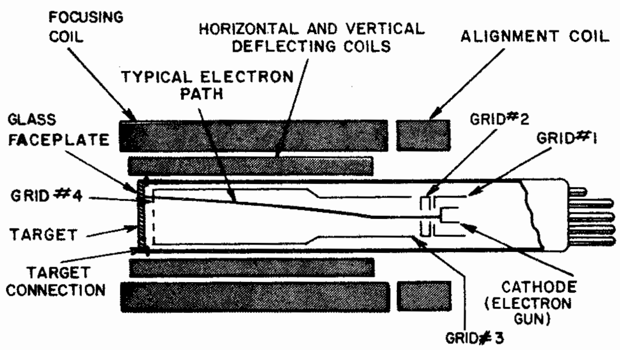


Figure 3-272. Physical Construction of Vidicon

gun, focusing electrodes, and a photoconductive target. Photoconductivity is a property whereby the electrical conductivity of a material increases more or less in proportion to the amount of light falling upon it. Many such materials exist. Most vidicons employ antimony trisulfide as the photoconductive material. The target consists of two layers. That nearest the glass faceplate (conductive coating) is made of a material chosen for both electrical conductivity and transparency to light: nesa (zinc oxide) is usually used. The signal connection for the tube is made at this point. The layer nearest the gun is the photoconductor. In operation, the nesa layer (conductive coating), is held at a positive potential with respect to the cathode. This may range from a few volts to about 100 volts - the higher the voltage, the less light required for a given output (within the limitations explained later). The scanning beam deposits a negative charge on the back surface of the photoconductor. Where the photoconductor is dark, its resistance is high. The negative charge accumulates here until this portion is at the same potential as the cathode. Further electrons are rejected. (These form a return beam which is not utilized.) The dark portions of the photoconductor become charged like a capacitor. Where light falls on the photoconductor, its conductivity increases, more or less in proportion to the intensity of the light. The photoconductor cannot hold a charge here; it becomes a leaky capacitor. The leakage leaves gaps in the negative charge across the photoconductor. When the beam passes this point again, it deposits additional electrons which flow through the photoconductor, into the nesa layer, and through the load resistor of the video amplifier. This is the signal current of the tube. The photoconductor has the ability to retain its charge from one scanning cycle to another - to "store" an image. This property makes the tube useful in motion-picture pickup. It also makes it useful for slow-scan television of moving objects when a shutter mechanism is incorporated. The photoconductor can store an image for several minutes without appreciable loss of detail or amplitude. The signal output of the vidicon is roughly proportional to the voltage applied to the target. However, this voltage cannot be increased indefinitely, since at a given level a leakage signal (dark current) will become nonuniform. This signal is not particularly difficult to compensate for as long as it is uniform for all parts of the picture, but when it becomes excessive, it also becomes generally nonuniform and appears as a shading signal. In any event, as the target voltage is increased, a point will be reached where only the dark current increases; there will be no significant change in the signal itself. This is the practical maximum of the sensitivity of the tube under any circumstances. The magnitude of the dark-current signal is somewhat dependent upon the temperature of the target; it tends to increase as temperature increases. For best performance, the target should be held at a constant temperature. Where this is not feasible, occasional adjustment of the target voltage will probably be necessary. Increased temperature is generally accompanied by an increase in level of the signal proper. Dark current appears in the picture waveform as a setup signal. As dark current increases, the apparent setup signal increases. Under ideal lighting conditions the target voltage is reduced to a point below which the dark current is not noticeable in the signal waveform. If the scanning beam is insufficient to drive the target surface to cathode potential, the highlights will appear clipped. This is because the entire beam is able to land. Therefore, the signal current cannot possibly exceed the maximum available beam current. This is easily corrected by either decreasing the amount of light falling on the target, or by decreasing the negative control-grid potential. Figure 3-272 shows the physical construction of a vidicon that has electromagnetic focus and deflection. The cathode, control grid, and grid No. 2 form the gun structure. This is not different in principle from the gun of a picture tube. Grid No. 3, which usually extends for the remaining length of the tube, is an electrostatic-focus electrode. Normally, focus is obtained by adjusting the voltage on this electrode, since it controls the speed of the electrons through the tube. When a crossover point of the beam coincides with the back surface of the target, the beam is in focus. Grid No. 3 corresponds to the ultor of a kinescope. It is sometimes called the "wall" electrode. Grid No. 4 consists of a very fine mesh screen located near the photoconductive layer. This is connected internally to the wall electrode (grid No. 3) in many vidicons. The screen provides a uniform decelerating field between itself and the photoconductor. It appears to the beam as a nearly invisible, uniformly charged plane. The potential difference between the screen and the target is the same at all points. Electrons are accelerated to the screen, but after they pass through it, they decelerate to the target because the target has a lower potential. The screen, therefore, prevents the low-voltage charge at the target from having any appreciable effect on the velocity of the electron beam until the electrons have passed through it. The decelerating field helps the electrons to strike the target perpendicularly. This is necessary if the target is to be driven to cathode potential. Should the electron beam arrive at other than a right angle, it would cause secondary electrons to be emitted from the target. The resulting nonuniform charge at the target would cause a shading signal in the tube output. An electron beam that does not strike the target perpendicularly also causes the beam spot to be enlarged with a resultant loss of resolution away from the center of the picture. In some types of tubes the screen is not connected to the wall electrode (grid No. 3); it is brought out to a separate connection. Then the screen voltage is normally set at about 1.7 times that of the wall electrode. This increased voltage results in improved corner focus. The mesh of vidicons is normally oriented so that it lies at an angle to the raster lines. This is the best position for minimizing its appearance in the television picture. Most vidicons use a mesh screen having 750 openings per inch; a few have 1000 openings per inch. Figure 3-272 shows the path of an electron through an electromagnetically-focused tube. Note that the beam deviates from its axial path only while it is in the field of the deflection coils. When the beam leaves this field, it remains in the field of the magnetic-focus coil, so it returns to a path parallel to the tube axis. Thus the full-length magnetic-focus field also helps the beam electrons to approach the target at a right angle. Since this prinicple does not apply to electrostatically focused tubes, they usually have a separate mesh connection. Figure 3-273 shows the internal construction of one type of electrostatically-focused vidicon. This tube has two more electrodes than the magnetically-focused version. Grids No. 3, 4, and 5 form an electrostatic lens similar to that of the electrostatic kinescope. Grid No. 5 is connected internally to grid No. 3, both normally operating at a considerably higher voltage than grid 4. Typical voltages are 300 volts at grid No. 2, 50 to 100 volts at grid No. 4, 300 volts at grid No. 3 and 5, and 500 volts at grid No. 6. Grids No. 3 through 6 may be operated at higher voltages in order to obtain an increase in resolution. Grid No. 1 is the control grid of the tube and operates at a slight negative voltage. The electrostatic-focus vidicon was introduced several years after the all-electromagnetic tube. It has certain advantages over its predecessor - generally improved corner resolution, excellent geometrical characteristics, and substantial savings in the weight, size, and power consumption of the camera. These are made possible because the focus coil is eliminated, and the tube requires considerably less deflection power than the electromagnetic type. Figure 3-272 shows an alignment coil located just outside grid No. 2. This assembly normally consists of two coils at right angles to the beam axis of the tube and 90 degrees apart from each other. Their purpose is to compensate for manufacturing tolerances which might cause the beam to leave the gun at a small angle to its axis. Should this happen, the beam would not enter the focus field at its center. This would have to be compensated for later through centering adjustments, and there would be a lack of symmetry as the beam passed through the tube. This would decrease resolution and increase shading signals. This condition is easily remedied by passing a current of the sufficient magnitude to recenter the beam through either or both of the alignment coils. Many cameras use small permanent magnets that are mounted so that they can be rotated to produce an accumulative field of the proper direction to center the beam.

3-10.14.10 The Image Orthicon

The image orthicon has the highest sensitivity of the familiar camera tubes, therefore it is useful where lighting conditions are poor. Special versions are available that produce credible pictures by the light of a quarter moon or a match at a distance of twenty-five feet. This tube is more complex to operate than the other types discussed; more expensive than the other types; and requires the most

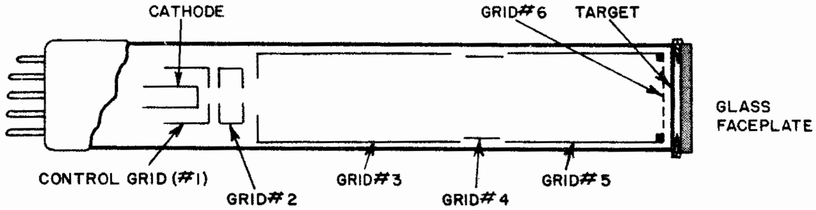


Figure 3-273. Physical Construction of an Electrostatic-Focus Vidicon

expensive deflection components, the largest housing, the most complex power supplies, and the most expensive lenses. Automatic-sensitivity operation is more difficult, the tube has a shorter life than the other types, is the most sensitive to temperature and vibration, and has a poorer signal-to-noise ratio under normal conditions. Ruggedized versions sell for around 5000 dollars. Microphonics are inherent in the tube. For these reasons the tube does not find wide application in industrial and scientific television systems. The image orthicon is used extensively in broadcast operations and in industrial television where the increased sensitivity is of consequence. Figure 3-274 illustrates the internal construction of the image orthicon. The tube consists of three basic parts: the image section, the scanning section, and the multiplier section. The image section consists of a photocathode, an accelerating anode, and a glass target. The photocathode is a thin, nearly transparent photoemissive coating on the inside of the tube face. The usual material is cesium silver oxide. When light strikes the photocathode, it emits electrons. These are attracted to the target, where they form an electrostatic image. The accelerating electrode (grid No. 6), operating in conjunction with the external focus field and the negative charge (-400 volts) on the photoconductor itself, causes the electrons to be attracted to the target, which is nominally at ground potential. Adjustment of the negative voltage on grid No. 6 (about -300 volts) provides focus. The target consists of a very thin (.0002 inch or less), low-resistivity glass membrane. When the electrons from the photocathode stike the target, they cause more electrons to leave the target than arrived. The result is an electrostatic image consisting of positive charges. Areas corresponding to the lightest picture elements are the most positive. The target is designed to retain this charge for the time of one frame. (Over a longer period the charge will diffuse due to lateral conductance in the target.) The scanning section consists of the reverse side of the target, grid No. 5 (decelerator), and the deflection fields. The reverse side of the target is scanned by a beam from the electron gun, similar to that of other cathode-ray tubes. Image orthicons usually are both magnetically deflected and focused. The focus electrode (grid No. 4) consists of a conductive coating inside the glass envelope. Focus voltage is normally set at approximately +200 volts. Grid No. 5 is a decelerating electrode and normally is set at approximately +25 volts. Its purpose is to slow the beam and reduce or eliminate secondary emission due to the beam hitting the target. Voltage is normally set to give the best corner focus consistent with good shading. As the low-velocity electron beam scans the target, only enough electrons are deposited to neutralize the charges. The excess is turned back and returns to the electron gun. Areas on the target corresponding to the whitest areas in the scene have the most positive charge and accept the most electrons. Darkest areas reject the most. An externally applied magnetic field aligns the beam, correcting for minor tolerances in the orientation of the electron beam. Figure 3-275 illustrates the multiplier section of the image orthicon. Electrons returning from the target strike grid No. 2 which serves as the aperture disc of the electron gun and the first electrode of the image multiplier (first dynode). This is coated with material especially selected to emit secondary electrons when hit by electrons from the target. Secondary electrons strike the second dynode, dislodging more electrons. This

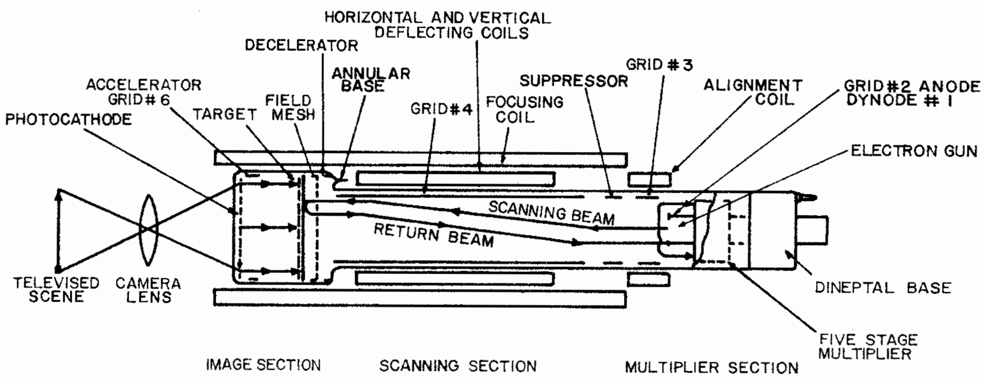


Figure 3-274. Internal Construction of an Image Orthicon

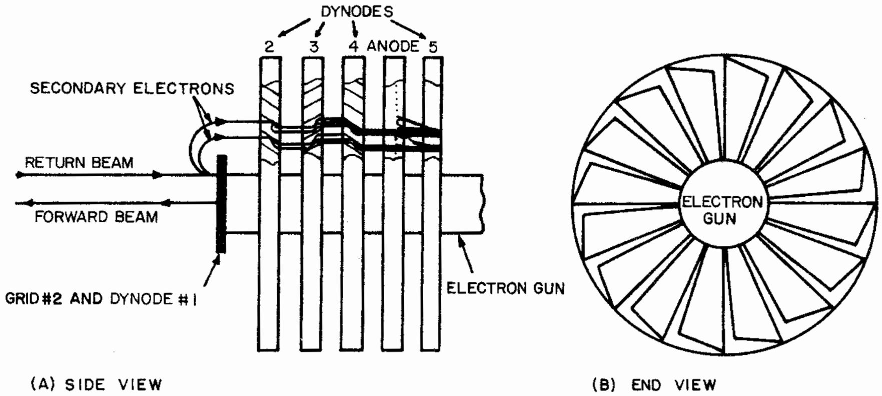


Figure 3-275. Multiplier Section of an Image Orthicon

process continues through all five dynodes until the stream finally reaches the anode. Each dynode has a higher positive potential than the preceding one, and therefore attracts its electrons. Since more secondary electrons leave each dynode than incident electrons strike it, the multiplier section normally has a gain of 200 to 500. The signal taken from the anode is termed a "black-negative." The usable area of the image-orthicon photoconductor is about 1.6 inches in diameter, giving a picture area of .96 x 1.28 inches. This requires a lens comparable to that of a 35-mm still camera. The tube has long life, partially due to the absence of a thermionic cathode. It is not temperature sensitive. The image dissector has a rather high output (about 0.5 volt) and is capable of excellent pictures where light levels are adequate.

3-10.14.11 The SEC Tube

The Secondary Electron Conductance (SEC) tube has similarities to both the image orthicon and the vidicon. Its sensitivity, physical size, ease of operation, and complexity lie between those of the older types. It has a storage characteristic that makes it useful for high-sensitivity operation with a low duty cycle, and its lag characteristics are excellent. Figure 3-276 shows the internal construction of the SEC tube. The image section is similar to that of the image orthicon, and the same photocathode materials are suitable. The scanning section more closely resembles that of the vidicon. The target, however, is different from that of any other type of tube. Figure 3-277 shows the construction of the SEC target. It consists of a very thin supporting layer of aluminum oxide, a central layer of aluminum, and a low-density layer of potassium chloride (KC1). The KC1 layer faces the electron gun. The aluminum layer serves as the signal plate and operates at a DC potential of about +20 volts. Electrons from the photocathode penetrate both the aluminum oxide and the aluminum layers and dissipate most of their energy in the KC1 layer where they loosen many secondary electrons - as many as 200 for each primary electron. Most of these pass through the voids of this low-density layer to the signal plate. This leaves a positive charge in this spot of the KC1 layer which is periodically discharged by the scanning beam. This discharge results in a current through the signal plate which develops a signal voltage across a load resistor.

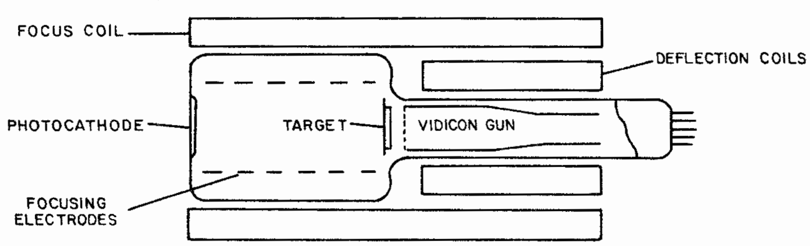


Figure 3-276. Electrode Arrangement of the SEC Tube

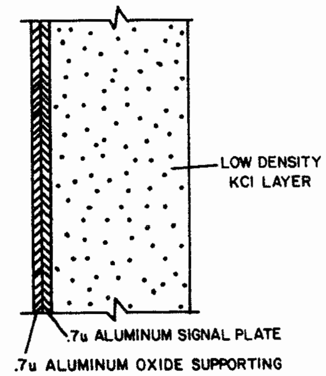


Figure 3-277. SEC Target

3-10.15 SYNC GENERATORS

The synchronizing (sync) generator is the central timing device of a television projection system. Its purpose is to provide reference signals through which the scanning circuitry of the camera and of the monitor maintain synchronism. The sync generator normally feeds signals directly (or through distribution amplifiers) to the camera, but the synchronizing signal for the monitor most often appears as part of the video signal. Sync signals may also be supplied to other television equipment such as video-insert amplifiers, video switchers, and stabilizing amplifiers. In a simple television projection system the sync generator may be an integral part of the television camera. In more elaborate systems a single generator usually feeds all cameras simultaneously since this allows them to be synchronized with each other. When electronic inserts (signals from several sources displayed on a single screen) are used, or when roll-free switching is desired, all cameras must use a common source of sync. The sync generator is often built into a simple industrial type of television camera. In its simplest form the sync-generator functions are performed by free-running horizontal- and vertical-deflection generators. This always results in a non-interlaced signal.

3-10.15.1 Sync Signals

The usual outputs of the sync generator are: 1) horizontal drive; 2) vertical drive; 3) composite blanking; and 4) composite sync. Refer to Figure 3-278 for details of standard signals as recommended by the Electronic Industries Association. The horizontal-drive signal is a negative-going rectangular pulse that recurs at the line rate - 15,750 hertz in a 525-line system. The duration of the pulse is 0.1 H (about 6.3 microseconds), where H is the time from the start of one line to the start of the next. The pulses are used in the horizontal-deflection and clamping circuits of the camera; they may also be used by other equipment employing clamp circuits, such as stabilizing amplifiers and insert amplifiers. The vertical-drive signal is a negative-going rectangular pulse recurring at the field rate of 60 hertz. The nominal duration of the pulse is 666 microseconds. The pulse is used in the vertical-deflection circuit of the camera; it may also be used by video switchers which operate during the vertical-blanking interval. The composite-blanking signal consists of rectangular pulses recurring at both the horizontal and the vertical rates. The vertical-blanking portion of the signal consists of a single pulse which lasts about 20H (approximately 1270 microseconds). The leading edge coincides with the leading edge of the vertical-drive pulse. The horizontal blanking pulse has a duration of approximately eleven microseconds. Its leading edge is timed to coincide with the leading edge of the horizontal-drive pulse. The composite-blanking signal is mixed with the video signal within the camera in order to blank out the retrace lines of the television monitor. It may also be used in video processing equipment such as insert and stabilizing amplifiers. Blanking signals for the camera tube are rarely, if ever, derived from the composite signal of the sync generator since the horizontal- and vertical-drive signals can serve this purpose. The composite-sync signal consists of horizontal -sync pulses, vertical-sync pulses, and equalizing pulses. It may be mixed with the video signal either at the camera or in succeeding video-processing equipment. The purpose of this signal is to supply to the television monitor a means of maintaining its horizontal- and vertical-scanning signals in synchronism with those of the camera. The leading edge of the horizontal-sync pulse follows the leading edge of the blanking pulse by the duration of the front porch - 1.3 to 2.0 microseconds by EIA Standard RS-170. The leading edge of the first vertical-sync pulse follows that of the vertical-blanking pulse by 3H (about 190 microseconds). The equalizing pulses occur just before and after the vertical-sync pulse.

3-10.15.2 Types of Sync Generators

Most sync generators are either of the pulse-counter or the binary type. Each of these types may also be divided into two classes: simple industrail or EIA. Pulse-counter generators are usually somewhat less complex than the binary type, and the simple-industrial units are less complex than the EIA.

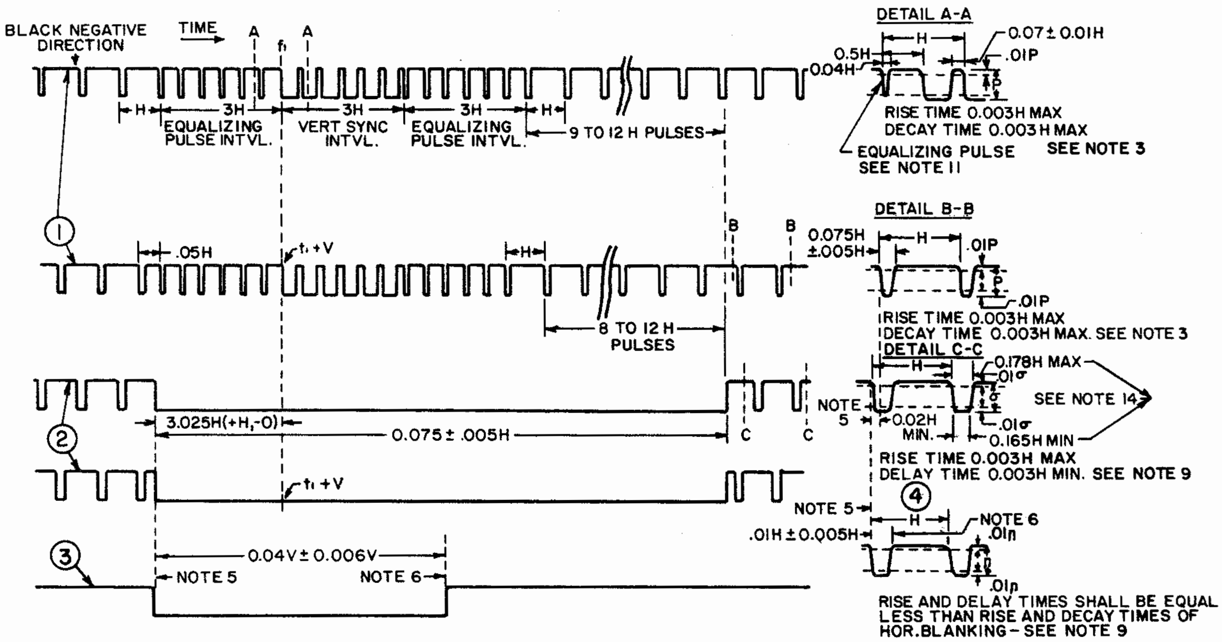
3-10.15.2.1 Pulse-Counter Sync Generators

Figure 3-279 presents a block diagram of the sync-generator section of a simple-industrial

MONOCHROME TELEVISION
STANDARD SYNC
GENERATOR WAVEFORMS

(1) SYNC SIGNAL
(2) BLANKING SIGNAL
(3) VERTICAL DRIVING SIGNAL
(4) HORIZONTAL DRIVING SIGNAL

ALL SIGNAL AMPLITUDES SHALL BE ADJUSTABLE OVER THE RANGE FROM 3.5 TO 4.5 VOLTS ACROSS A LOAD IMPEDANCE OF 75 OHMS ± 5%. NEGATIVE SIGNAL POLARITY SHALL BE AVAILABLE FOR ALL PULSES. SOURCE IMPEDANCE FOR ALL OUTPUT CIRCUITS SHALL BE 75 OHMS ± 10%.



NOTE:

1. H = time from start of one line to start of next line.
2. V = time from start of one field to start of next field.
3. Leading and trailing edges of vertical driving and vertical blanking signals should be complete in less than 0.1H.
4. All tolerances and limits shown in this drawing are permissible only for long time variations.
5. Time adjustment, if any, shall include this condition.
6. The vertical driving pulse duration shall be 0.04V, ± 0.006V. The horizontal driving pulse duration shall be 0.1H, ±0.005H.
7. The time relationship and waveform of the blanking and sync signal shall be such that their addition will result in a standard RETMA signal. The time relationship shall be adjustable in order to satisfy this relationship for the condition where the blanking signal is delayed with respect to the sync signal over the range from 0.0H to 0.05H.
8. The standard RETMA values of frequency and rate of change of frequency for the horizontal components of the sync signal at the output of the picture line amplified shall also apply to the horizontal components of the output signals from the recommended sync generator.
9. All rise and decay times shall be measured between 0.1 and 0.9 amplitude reference lines.
10. The time of occurence of the leading edge of any horizontal pulse "N" of any group of twenty horizontal pulses appearing on any of the output signals from a standard sync generator shall not differ from "NH" by more than 0.0008H where H is the average interval between the leading edges of the pulses as determined by an averaging process carried out over a period of not less than 20 nor more than 100 lines.
11. Equalizing pulse area shall be between 0.45 and 0.5 of the area of a horizontal sync pulse.
12. The overshoot on any of the pulses shall not exceed 5%.
13. The output level of the blanking signal and the sync signal shall not vary more than ±3% under the following conditions:

    A. The a.c. voltage supplying the sync generator shall be in the range between 100 V and 120 V and must not vary more than ±5V during test.
    B. A period of 5 hours continuous operation shall be considered adequate for this measurement after suitable warm-up.
    C. The room ambient shall be in the range between 20 deg. and 40 deg. C and shall not change more than 10 deg. C during this test.

14. Adjustment shall be possible between minimum and maximum limits so that aspect ratio can be set to the normal value.

Figure 3-278. EIA Standard RS-170 Sync Generator Waveforms

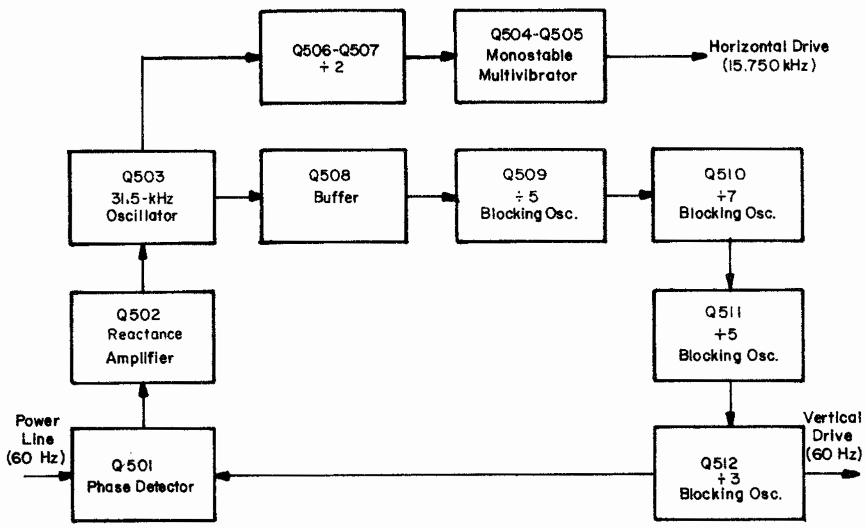


Figure 3-279. Sync Generator of a Typical Television Camera

type of camera. (This type produces interlaced scanning.) In its normal mode of operation the operating frequencies are referenced to the 60-hertz power line. The output of the 31.5-kHz oscillator is translated to 60 hertz by electronic counters. A phase-detector compares this frequency with the power line and, when drifting occurs, feeds a correction signal to the 31.2-kHz oscillator. The oscillator is thus stabilized by the power line, which is maintained at a high degree of accuracy. Since both horizontal- and vertical-drive outputs are derived from the same oscillator, their relationship remains constant. The interlace characteristic results from the 15.75-kHz signal being an even submultiple of 31.5-kHz (1/2), while the 60-hertz signal is an odd submultiple (1/525). The Q503 circuit is an LC-tuned oscillator having two outputs. One feeds a flip-flop (binary-divider) circuit which has an output (15.75-kHz) that is one-half the input frequency. This triggers a monostable multivibrator that cycles in a period equal to the desired duration of the output pulse. In many cameras this pulse is used for both horizontal sync and horizontal blanking. The second output of the 31.5-kHz oscillator feeds, through buffer Q508, a four-stage frequency-divider chain. The individual stages, dividing successively by 5,7,5, and 3, provide a total countdown of 525. The divider stages in virtually all pulse-counter sync generators are blocking oscillators. The buffer stage is desirable because blocking oscilators load the stage that drives them. The final stage in the frequency-divider chain (Q512) has two outputs. One is the vertical-drive signal, and in many instances this will be used as a vertical-blanking and vertical-sync signal. The second output is fed back to the phase-detector (AFC) circuit. This compares the phase of the vertical drive with a second input consisting of a 60-hertz reference signal from the power line. Should the frequency or phase of the feedback signal drift, the phase-detector provides a DC correction signal to the 31.5-kHz oscillator through the reactance amplifier. The latter, being in parallel with the oscillator tank circuit, causes the frequency of the oscillator to shift according to the magnitude of the correction signal. This changes the frequency of the 60-hertz vertical-drive signal, correcting it as necessary to match the 60-hertz reference.

3-10.15.3 Binary Sync Generators

Figure 3-280 shows the block diagram of a binary sync generator that produces signals in accordance with EIA standards. By making minor circuitry changes this sync generator can be converted to higher scanning rates. The explanation which follows applies to the 525-line version. (Waveforms for this circuit are shown in Figure 3-281. The mode switch, located in the master-oscillator circuitry, permits controlling the master oscillator by any of the following methods:

1. AFC. In this position the master oscillator is referenced to the power-line frequency through the circuit (Q503, Q509).

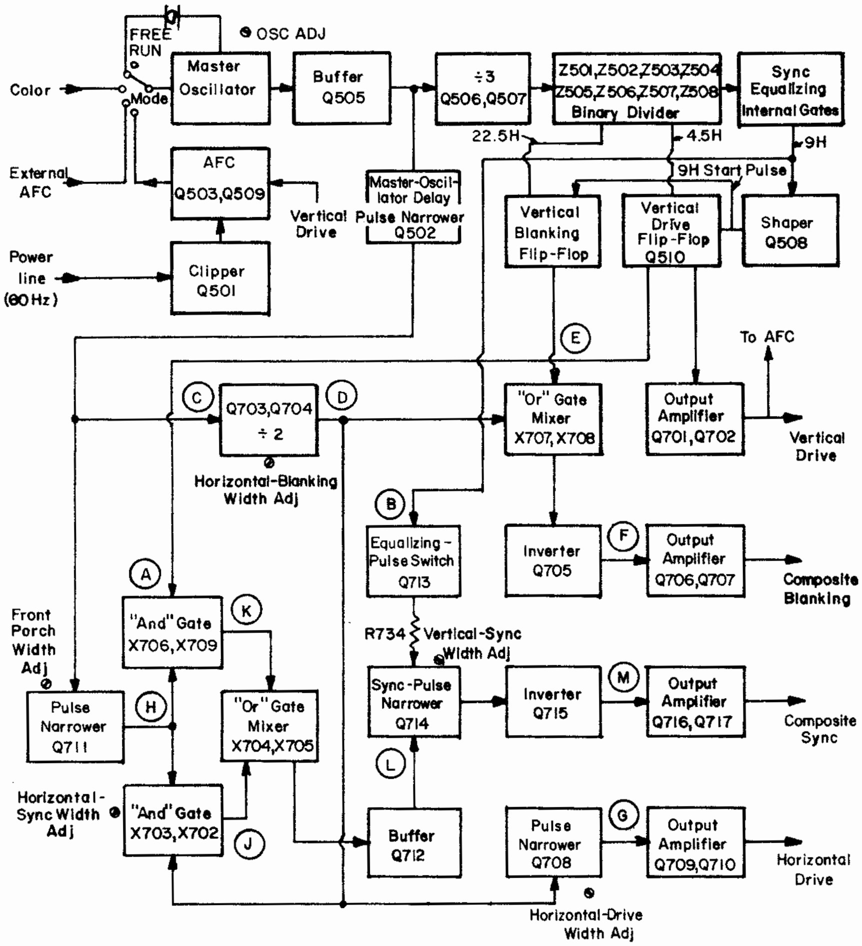


Figure 3-280. Binary Sync Generator

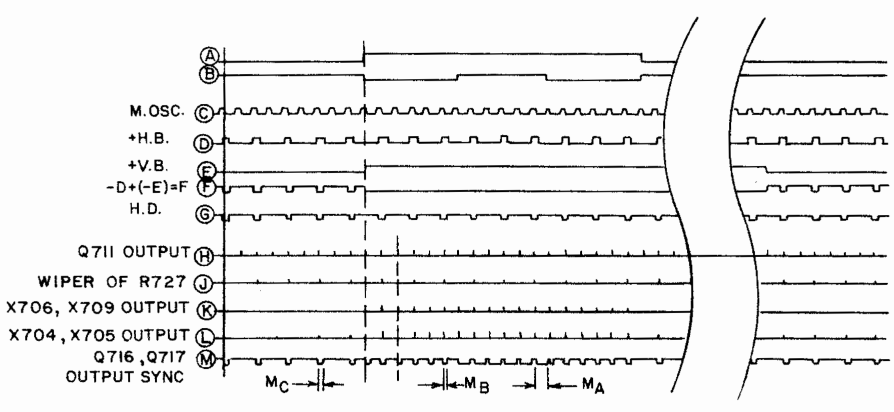


Figure 3-281. Waveforms for Generator

2. External AFC. In this position the signals of the sync generator are compared in frequency and phase with those from a remotely-located generator, and a correction signal is applied to the master oscillator. The sync generator thus operates as a slave unit held in phase with a remote unit.

3. Color. In this position the signal from an external color-subcarrier generator, is substituted for that of the master oscillator (which is hence not used).

4. Crystal. In this position the master oscillator operates as a crystal-controlled oscillator without external reference.

5. Free Run. In this position the master oscillator operates as an adjustable LC oscillator without external reference. The vertical-sync signals remain a submultiple of the horizontal drive, therefore interlace is maintained. This mode of operation is generally employed only for initial adjustments.

The signal from the master oscillator feeds through buffer Q505 to astable multivibrator Q506-Q507, which operates as a frequency divider. Its output frequency is one-third the input - a 10.5-kHz square wave in a 525-line system. The astable multivibrator feeds a multistage divider circuit (Z501-Z508) that produces pulses to control the timing of the various parts of the composite-sync and blanking signals. This eight-stage binary divider is connected in such a manner that it divides by 175 (Figure 3-282). An eight-stage binary divider chain has a basic count of 2^8 (256); a seven-stage chain has a count of 2^7 (128). Therefore, division by 175 is not possible using standard connections. Feedback connections must therefore be inserted that will delete 81 from the number of pulses required for the chain to go through a complete cycle. This is accomplished by resetting binaries Z507, Z505, and Z501 - giving a reduction of 64 + 16 + 1, or 81. These stages are flipped over at the start of each overall count, which is the equivalent of feeding 81 pulses to the chain. Consequently, 256 minus 81 (175 totoal) pulses are required for the chain to complete its full cycle. The binary divider chain has three outputs:

1. The equalizing-interval gate (9H). This is a double negative-going pulse having a total duration of 9H, with a 3H gap at the center. It is formed by combining the right-hand output of Z502 with the left-hand outputs of Z504, Z506, and Z508 through the five-diode mixing network shown in Figure 3-282. The two negative-going portions of the equalizing gate establish the two intervals during which the equalizing pulses occur. The positive-going portion of the pulse determines the vertical-sync-pulse interval. The equalizing interval gate also establishes the start time of the vertical-drive and vertical-blanking signals.

2. The signal from Z503. This signal, in combination with the 9H signal, establishes the timing and duration of the vertical-drive signal. The 9H

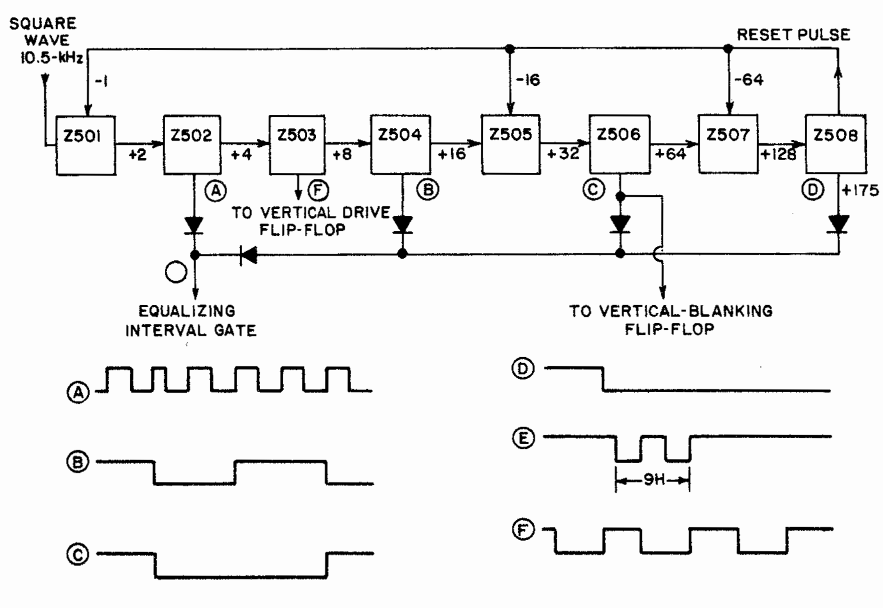


Figure 3-282. Binary Divider Circuit

pulse determines the start time, and the signal from Z503 determines the stop time.

3. The signal from Z506. This signal, in combination with the 9H signal, establishes the timing and duration of the vertical-blanking signal. The 9H pulse determines the start time, and the signal from Z506 determines the stop time.

The vertical-drive signal is generated by flip-flop Z510 (Figure 3-280). Start-time is established by the leading edge of the 9H signal, the stop-time by the signal from the right side of Z503. The signals feed the two separate inputs of Z510. The first positive-going swing of the signal from Z503 returns Z510 to its quiescent state, where it remains until the next 9H signal initiates another pulse. The signal from Z510 feeds output amplifier Q701-Q702. A second output of Z510 is used in generation of the composite-sync signal. Generation of the vertical-blanking signal occurs in flip-flop Z509. The process is similar to that for the vertical drive except that the signal from Z506 establishes the stop time. The horizontal-blanking signal is also derived from the master oscillator. The output of buffer Q505 is a 31.5 kHz square wave. Pulse-narrower Q502 converts this to a five-microsecond negative-going rectangular pulse which feeds astable multivibrator Q703-Q704, which is a divide-by-two pulse counter. The timing and phase relationships of the input and output signals are shown in Figure 3-283. The width of the horizontal-blanking pulse is established by RC circuits in the multivibrator. This signal is also used in the generation of horizontal-sync and horizontal-drive signals.The horizontal- and the vertical-blanking signals are both fed to the OR gate (X707, X708) where they are mixed to form the composite-blanking signal. Inverter Q705 changes the polarity and clips the signal to a constant level, eliminating the horizontal-blanking signals superimposed on the vertical-blanking signal (Figure 3-283). Output amplifier Q706-Q707 feeds the signals to the external circuits. The horizontal-drive signal is derived from the horizontal-blanking signal in this sync generator. The latter is fed through pulse-narrower Q708 where it is also inverted. The time-constant of the pulse-narrower circuit determines the width of the horizontal-drive signal. This is adjustable. Output amplifier Q709, Q710 feeds the signal to external systems. Generation of composite sync involves the master-oscillator, sync-equalizing-interval, vertical-drive, and horizontal-blanking signals. The output of pulse narrower Q711 has a width equal to that of the front porch of the composite-video signal. This is derived from the master oscillator (waveform C, Figure 3-282), and is shown as signal H. It has a repetition rate of 31.5 kHz. Pulse-narrower Q711 has two outputs, one of which feeds AND gate X702-X703. A second input to this AND gate is provided by the horizontal-blanking pulse. Since both signals must be present for the AND gate to conduct, only those pulses occuring during the horizontal-blanking intervals pass through. The trailing edges of these pulses trigger the horizontal-sync pulses. The horizontal-sync-width control adjusts the amplitude of this pulse. The second output of Q711 feeds AND gate X706-X709, which also receives the vertical-drive pulses. The output of this gate is in the form of 31.5-kHz pulses occuring during the vertical-drive (9H) interval. The trailing edges of these pulses trigger the equalizing and vertical-sync pulses. The signals from the two AND gates are mixed in OR gate X704-X705 and fed through buffer Z712 to sync-pulse narrower Q714. A second input to this circuit is provided by the 9H signal which drives equalizing-pulse switch Q713. To better understand the operation of the sync-pulse-narrower circuit, refer to Figure 3-284. The gated front-porch-delay pulses are differentiated by C709, R735, and R736. The negative-going trailing edges of

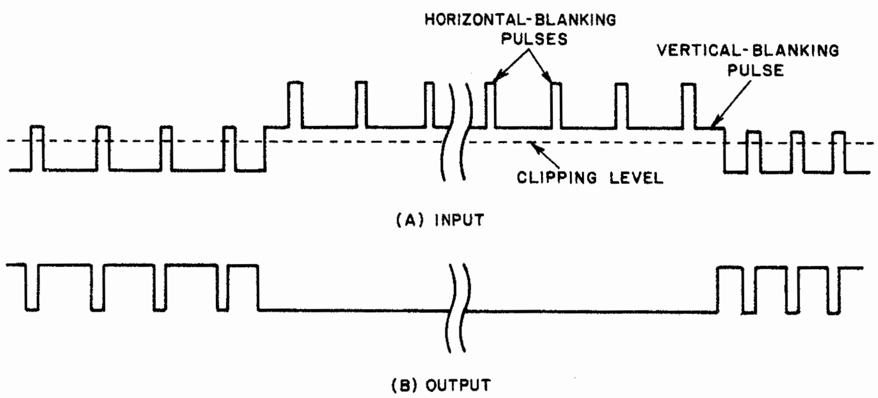


Figure 3-283. Waveforms at Inverter (Q705).

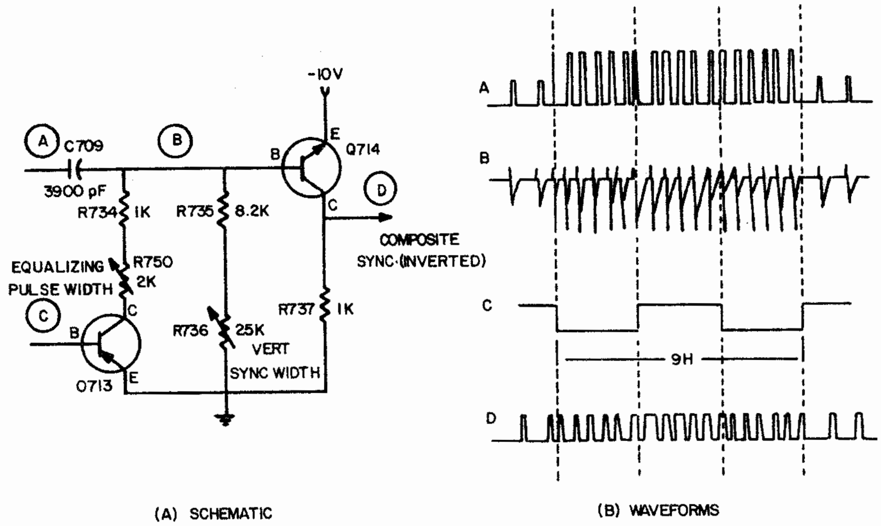


Figure 3-284. Sync Pulse Narrower Circuit

these pulses trigger Q714, cutting it off. The length of time during which Q714 is cut off, and therefore the width of the output pulse, is determined by the width of the driving signal at the level which cuts off Q714. Thus, the width of the output pulse may be set by either the amplitude or the width of the driving signal. During the vertical-sync interval the amplitude of the driving pulses is several times greater. This causes the widths of the output pulses to greatly increase to that of the vertical-sync pulses. The time-constant of the RC network is adjustable at R736 to give an output pulse of the correct duration. During the horizontal and vertical sync time, Q713 is cut off by the 9H signal, hence has no effect on the circuit. During the equalizing-pulse intervals Q713 is turned on on by the negative-going portions of the 9H pulse, placing R734 in parallel with R735 and R736. This sharply decreases the time-constant of the RC network, causing the conduction time of Q174 to likewise decrease to the duration of the equalizing pulses. The exact duration of these pulses is adjustable at R750. The signal from Q714 feeds Q715 (Figure 3-280) where further clipping takes place, and the signal is inverted. Output amplifier Q716-Q717 feeds the composite-sync signal to external systems.

3-10.15.4 Circuit Analysis

Sync generators consist of only a few basic electronic circuits that are repeated many times in different areas. This section describes the circuits utilized most frequently.

3-10.15.4.1 Master Oscillator

Figure 3-285 is a simplified schematic of the 31.5-kHz master oscillator in a representative sync generator, with Q3 as the oscillator stage. Its

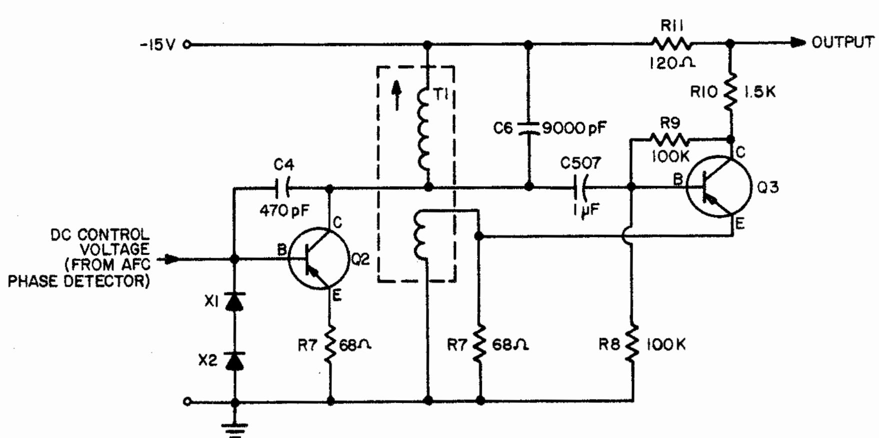


Figure 3-285. Master Oscillator

free-running frequency is determined by C6 and T1, and is adjustable by moving the slug of T1. The circuit is a variation of the Armstrong oscillator. Reactance stage Q2 changes the frequency of the oscillator in accordance with a DC control-voltage applied to its base. Diodes X1 and X2, acting as varicaps, form a capacitive voltage divider with C4 that is part of the 31.5-kHz tank circuit (along with C6 and T1). The reactive voltage at the base of Q2 is proportional to the magnitude of the DC correction voltage from the afc phase-detector. This is amplified by Q2, also in parallel with T1. Since the correction-signal voltage from Q2 lags the signal current, its effect on the signal at T1 is reactive and causes a shift in the frequency of oscillation. The direction of the shift is such that the frequency of oscillation is corrected.

3-10.15.4.2 Astable Blocking Oscillator

An astable blocking oscillator also may be used as a master oscillator. Figure 3-286 shows a circuit of this type, a free-running oscillator whose frequency can be varied by a DC control-voltage from the AFC circuit. The circuit operates in the following manner. Because Q1 is forwarded biased, it conducts when power is applied to the circuit. The collector current of Q1 is regeneratively coupled through T1 to the base of Q1, causing Q1 to conduct heavily. Capacitor C2 becomes negatively charged due to the voltage drop acrosss R3 and R5, and transformer T1 rapidly reaches saturation. When this occurs, no signal is coupled to the base of Q1, hence regeneration stops. The large negative voltage at the emitter now cuts off Q1. Capacitor C2 now discharges through R3 and R5 until Q1 can again conduct, starting another cycle. Potentiometer R5 controls the discharge rate of C2. Since this determines the off-time of the transistor, it regulates the spacing between pulses which, in turn, affects the oscillator frequency. The output of the AFC circuit provides a correction voltage that supplies bias to Q1; its value also affects the spacing between pulses because it helps determine the time at which Q1 can conduct. Clipper diode X1 limits the output to pulses of a single polarity. This signal is transformer coupled to the following stage through a tertiary winding on T1.

3-10.15.4.3 AFC Phase-Detector

Figure 3-287 shows the circuit of an AFC phase-detector. The 60-hertz power supply feeds the emitter of Q1 while the vertical-drive pulse feeds the base. Transistor Q1 is normally cut off, but is turned on by the negative-going vertical-drive pulse. If the phase relationships are normal, as shown in Figure 3-288, the stage has a given output. Normal is when the vertical-drive pulse coincides with the 60-hertz sine wave as it passes through zero. If the vertical-drive signal should drift, it will coincide with a different portion of the sine-wave reference signal. Then the instantaneous voltage at the emitter, when Q1 is turned only the vertical-drive pulse, is different from the normal timing. If the emitter swings positive, the voltage-drop across R4 and R6 will increase since Q1 will conduct more current. If the emitter swings negative, the voltage-drop across R4 and R6 will decrease. The pulsating signal at the collector of Q1 is

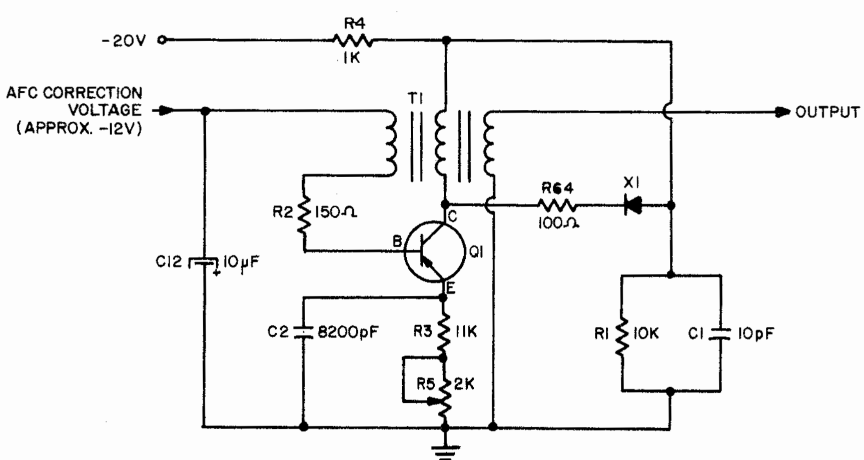


Figure 3-286. Master Oscillator Circuit

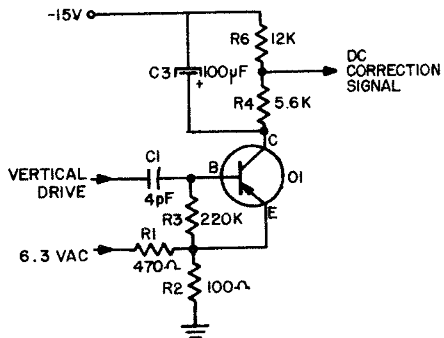


Figure 3-287. AFC Phase Detector

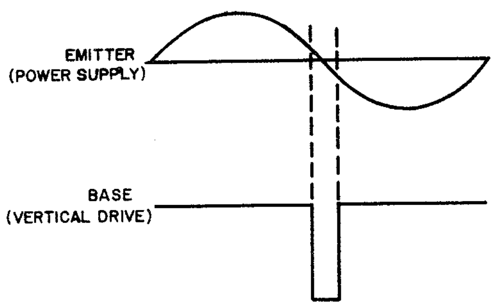


Figure 3-288. Phase Relationship in AFC Phase Detector

smoothed into a DC signal by the large value of capacitor C3. Thus, if the phase of frequency of the vertical-drive signal tends to drift, it causes a change in the magnitude of the signal at the collector of Q1 and at the junction of R4 and R6. This is fed as a correction signal to a reactance stage that controls the frequency of the master oscillator.

3-10.15.4.4 Blocking-Oscillator Frequency Divider

Figure 3-289 shows the schematic of a blocking oscillator used as a frequency divider. Transistor Q9 is normally cut off. Positive-going input pulses at the emitter of Q9 gradually build up a charge on C17. In a divide-by-five circuit the values of C17, R31, and R30 are chosen such that every fifth pulse charges C17 sufficiently to turn Q9 on. When Q9 conducts, the signal at the collector is coupled to the base through T2. Because this effect is regenerative, Q9, goes into saturation, and C17 then discharges rapidly through

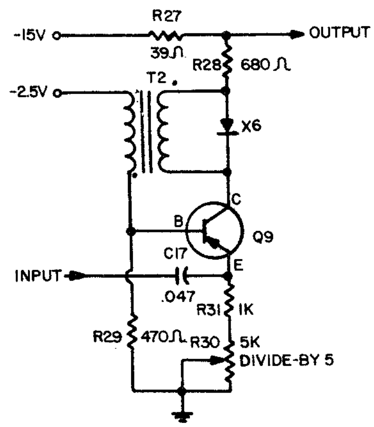
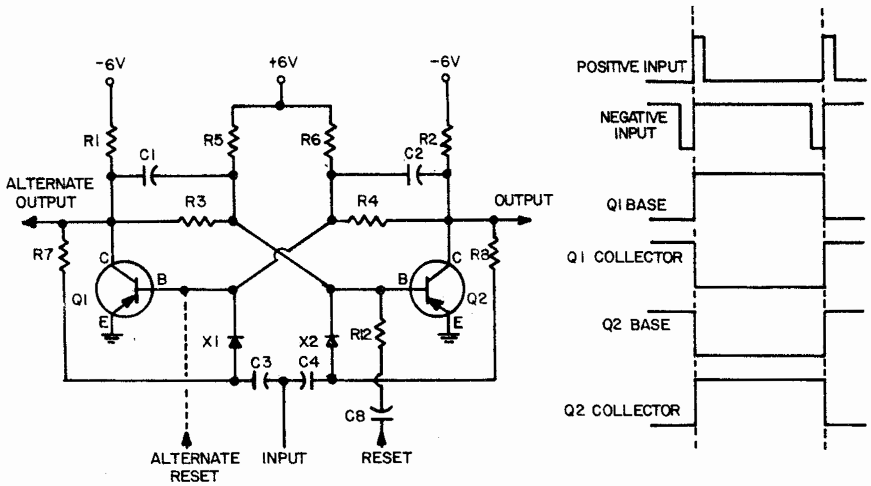


Figure 3-289. Blocking Oscillator Frequency Divider

the transistor. Transistor Q9 stops conducting abruptly when C17 has discharged sufficiently because of the feedback coupling through T2. Diode X6 damps further oscillation which might occur due to resonance of T2. Potentiometer R30 allows adjustment of the time-constant in the emitter circuit of Q9. This time-constant determines the number of pulses necessary to cause Q9 to conduct.

3-10.15.4.5 Binary Divider

The binary-divider circuit is also called a flip-flop and a bistable multivibrator. It requires two input pulses (triggers) to complete one cycle; for this reason it is useful where a division by two is required. The circuit is symmetrical, very stable, and usually has no adjustments. It is used extensively in television sync generators. Figure 3-290 shows a representative binary-divider circuit. Both transistors are forward biased through resistive voltage dividers, therefore they can conduct. When the circuit is initially energized, one transistor will conduct slightly more than the other due to minor variations in its components. The transistor that conducts more heavily, as controlled by Q1, has a slight decrease in voltage at its collector. This is coupled to the base of Q2, causing a slight decrease in current through Q2 and a consequent increase in voltage at its collector. The latter is coupled to the base of Q1, causing an increase in forward bias with a consequent decrease in collector voltage. The action is regenerative. It continues in this way until Q1 is driven to saturation and Q2 is driven to cutoff. The circuit will then remain in this state, with



(A) SCHEMATIC.

(B) WAVEFORMS.

Figure 3-290. Basic Binary Divider Circuit

either transistor conducting, until triggered by a signal from another circuit. While Q1 is conducting, diode X1 is forward biased by the collector voltage of Q1 through resistor R7. At the same time, X2 becomes reverse-biased. If a positive-going pulse is applied to the input, it will be blocked by X2 but not by X1 because the anode of X2 is so much more negative than its cathode at this time. It is not possible for X2 to conduct during triggering under these conditions. However, the same pulse will cause the conducting transistor, Q1, to have a decrease in current, and its collector voltage will increase. This increase is coupled to the base of Q1, causing a further decrease in the current through Q1. Regenerative action continues until Q2 is driven completely into saturation and Q1 is driven to cutoff. Each succeeding input pulse causes a reversal in the state of the two transistors. The leading edge of a positive pulse or the trailing edge of a negative pulse may be used to trigger the circuit. Positive feedback pulses may be used to reset a particular binary stage in a counter chain to alter the overall count. They are applied to the base of the conducting transistor. Two possible connections for reset pulses are shown in Figure 3-290. Capacitors C1 and C2 speed the turnover time of the circuit by coupling the leading edge of the output waveform to the base of the opposite transistor. This avoids a slowing down that would take place if it were necessary to charge the distributed circuit capacitance through R3 or R4.

3-10.15.4.6 Monostable Multivibrator

The monostable (one-shot) multivibrator is useful in generating pulses of a predetermined width from pulses of another width. It is also employed in certain types of time-delay circuits. Two types are used: collector coupled; and emitter coupled. Figure 3-291A is a schematic of a collector-coupled monostable multivibrator. Transistor Q1 is biased beyond cutoff through R5, and Q2 is biased through R4 such that it conducts heavily. A negative-going pulse at time A, Figure 3-291B, causes Q1 to conduct. The negative voltage at the Q1 collector decreases (changes to a more positive value). This is coupled through C1 to the base of Q2, which is then cut off, causing the voltage at its collector to swing more negative. This change is coupled to the base of Q1, causing its current to decrease further. The effect is regenerative and continues until Q1 is completely saturated and Q2 is completely cut off. The action is practically instantaneous and causes a steep negative-going signal to appear at the collector of Q2. During the interval from A to B, capacitor C1 is charging through Q1 and R4. At time B, the charge on the side of C1 connected to the base of Q2 becomes sufficiently negative to cause Q2 to conduct so that the voltage at its collector will swing positive. This is coupled to the base of Q1 through C2 and R3 and cuts Q1 off. The voltage at its collector now swings positive. This action is again regenerative and continues until Q1 is completely cut off and Q2 is completely saturated. Since the collector of Q2 is direct-coupled to the base of Q1, the circuit will maintain this condition indefinitely, or until it receives another trigger pulse, as at C. The signal at the collector of Q2 is a negative pulse whose duration is determined by the charging time of C1. Adjustment of R4 will vary the pulse width. A positive-pulse

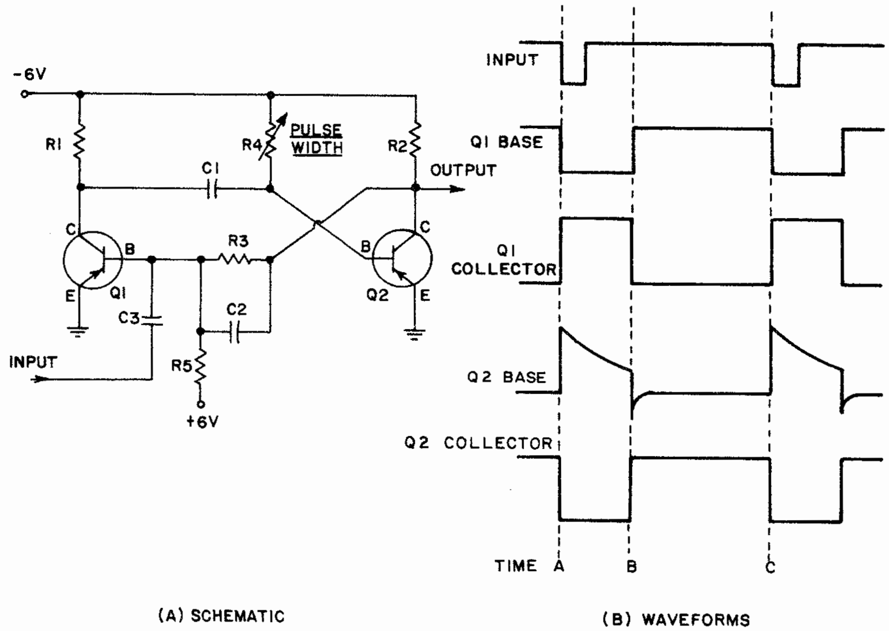


Figure 3-291. Collector-Coupled Monostable Multivibrator

output may be taken from the collector of Q1 instead of, or in addition to, the negative pulse at the collector Q2. The circuit may be triggered at the base of Q1 by the leading edge of a negative-going pulse, as shown, or by the trailing edge of a positive-going pulse. If the input is applied to the base of Q2 instead, the circuit may be triggered by the leading edge of a negative-going pulse, as shown, or by the trailing edge of a positive-going pulse. If the input is applied to the base of Q2 instead, the circuit may be triggered by the leading edge of a positive-going pulse or the trailing edge of a negative-going pulse. Figure 3-292 shows an emitter-coupled monostable multivibrator. This is employed extensively in television sync-generators. In the absence of a trigger pulse, positive bias supplied by the voltage-divider network (R3, R4, R7) causes Q2 to conduct heavily. The voltage drop across R6, caused by current through Q2, keeps Q1 cut off. A positive-going trigger pulse causes Q1 to conduct. A negative-going signal appears at the collector of Q1 and is coupled to the base of Q2 to cut it off. The latter then remains cut off until C1 has discharged through R3, R7, and R1. When Q2 begins to conduct, the voltage drop across R6 increases and rapidly cuts off Q1. The circuit may be triggered by the leading edge of a positive pulse or by the trailing edge of a negative pulse at the base of Q1. It may also be triggered at the base of Q2 by the leading edge of a negative pulse or by the trailing edge of a positive pulse.

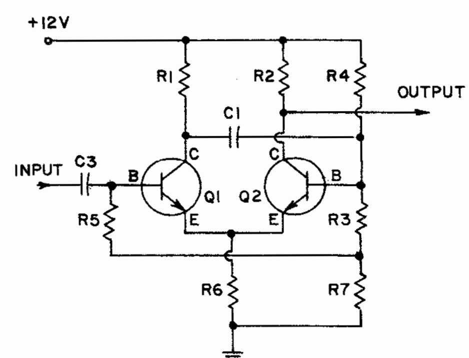


Figure 3-292. Emitter Coupled Monostable Multivibrator

3-10.15.4.7 Astable Multivibrator

The distinctive feature of an astable multivibrator is that it is free running, therefore the circuit has a rectangular-wave output with no input signal. However, the circuit may be synchronized by input pulses to produce a stable output. If fed synchronizing pulses at a multiple of a natural frequency of the multivibrator, the circuit can act as a frequency divider. It is sometimes used for this purpose in television sync-generators. This circuit is useful where an output is desired in the event of loss of input pulses. Astable multivibrators may be collector-coupled or emitter-coupled. The former is usually favored by circuit designers when a symmetrical or nearly symmetrical waveform is required. The emitter-coupled version is preferred when the output is to be unsymmetrical. Figure 3-293 presents the

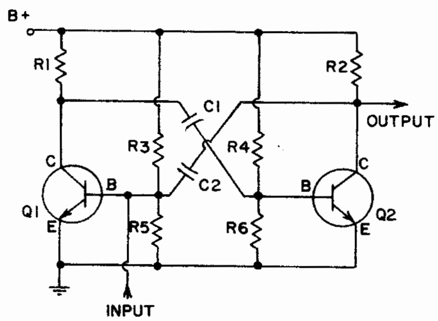


Figure 3-293. Collector Coupled Astable Multivibrator

schematic of a collector-coupled astable multivibrator. Although both transistors are baised to conduct, cross-coupling by C1 and C2 prevents both transistors from conducting simultaneously. When power is applied to the circuit, one transistor will conduct more heavily than the other. Assuming Q1 is the one, the voltage drop across R1 is coupled as a negative signal to the base of Q2, causing less current to flow through it. The voltage at the collector of Q2 increases at the same time, and is coupled as a positive-going signal to the base of Q2, causing increased current through Q1. The action is regenerative, and occurs practically instantaneously until Q1 is conducting at saturation and Q2 is cut off. When a sufficient amount of the charge on C1 has leaked off, Q2 starts to conduct and the voltage at the collector becomes less positive. This output is coupled through C2 to the base of Q1 as a negative-going signal. It causes a decrease in current through Q1 and a positive-going signal to appear at the collector of Q1. Regenerative action continues until Q2 is conducting heavily and Q1 becomes cut off. The circuit remains in this state until a sufficient amount of the charge on C2 has leaked off to allow Q1 to conduct. The cycle then repeats. The length of time each transistor conducts is determined by the time-constants of the capacitor-resistor combinations in both base circuits. If a symmetrical waveform is desired, the corresponding capacitors and resistors will have equal values. If an asymmetrical waveform is required, the value will be unequal. Output signals may be taken from the collector of either transistor, depending on the polarity of the signal required. In some cases output signals may be derived from the collectors of both transistors. This circuit may be synchronized to an external waveform having a repetition rate slightly higher than the natural period of oscillation of the multivibrator. Sync may be fed to the base of either transistor, depending on the polarity of the pulse and whether the output pulse is to be synchronized to the leading or the trailing edge of the input pulse. The circuit shown in Figure 3-294

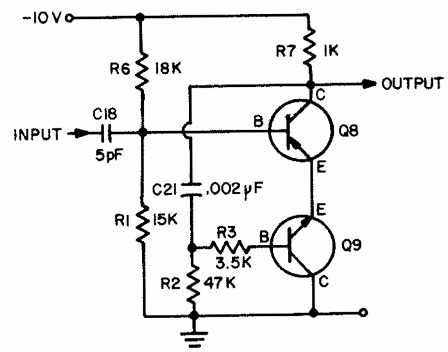


Figure 3-294 Emitter Coupled Astable Multivibrator

may also be used as a frequency divider. In this case the input-signal frequency will be slightly higher than a multiple of the natural frequency of the circuit. Figure 3-294 is a schematic of an emitter-coupled astable multivibrator employing complementary circuitry. As shown, it is a divide-by-three frequency divider. Both transistors are biased to conduct. When power is applied to the circuit, current flowing through both transistors causes a voltage drop across R7 that is coupled to the base of Q9 through C21. This positive-going signal causes an increase in current through transistor Q9 and a further increase in current through Q8. The action is regenerative and very quickly drives the transistor to saturation. At this point, no signal change is coupled through C21, therefore it discharges. With both transistors conducting heavily the voltage at the emitters and at the collector of Q8 drops to a more positive value. Transistor Q8 is now left with a reverse bias established by voltage divider R1-R6. The transistors cut each other off in turn. When the charge across C21 returns to its original value, the transistors conduct again, and the cycle repeats. As with other astable multivibrators, this one may be triggered by a signal having a slightly higher frequency than the natural period of oscillation, or it may be used as a frequency divider.

3-10.16 TELEVISION CAMERAS

Television cameras convert the optical image of a scene into an electrical signal through the process of scanning - translating the intensity of each successive point of the image into a value of voltage. This event takes place within the pickup tube at a regular rate as established by the sync generator and the deflection circuitry. The vidicon tube is most frequently used in industrial and scientific cameras. The camera lens imposes an image on the front of a sensitized surface (called a target) at the same time an unmodulated electron beam scans the reverse side. The target of the vidicon conducts best in areas where the most light is falling. As the beam scans the target, current through the target is highest in the most highly illuminated areas of the scene. This current passes through a series resistor, and the varying voltage that is developed is coupled to video amplifiers and other circuitry which process it to form the video signal. Figure 3-295 is the block diagram of a typical television-camera system. These circuits are frequently divided between the camera and the camera-control unit. The exact point of division varies from types having virtually all circuitry within the camera head itself, to those having little more than the vidicon and the deflection assembly within the camera head. The deflection generators use the drive signals, supplied by the sync generator, to form the sawtooth currents necessary to cause beam deflection in the pickup tube. The videoamplifier stages raise the relatively feeble signal from the pickup tube to a level more suitable for signal processing. The high-peaker stage compensates for shunt capacitance in the input circuit of the video preamplifier. It has a frequency and phase response characteristic which is the inverse of the video-preamplifier circuitry, thus the overall response of the channel through this point is flat. The aperture-correction circuit compensates for the finite size of the scanning beam of the pickup tube. This circuit has a response characteristic that increases with frequency in order to compensate for the high-frequency

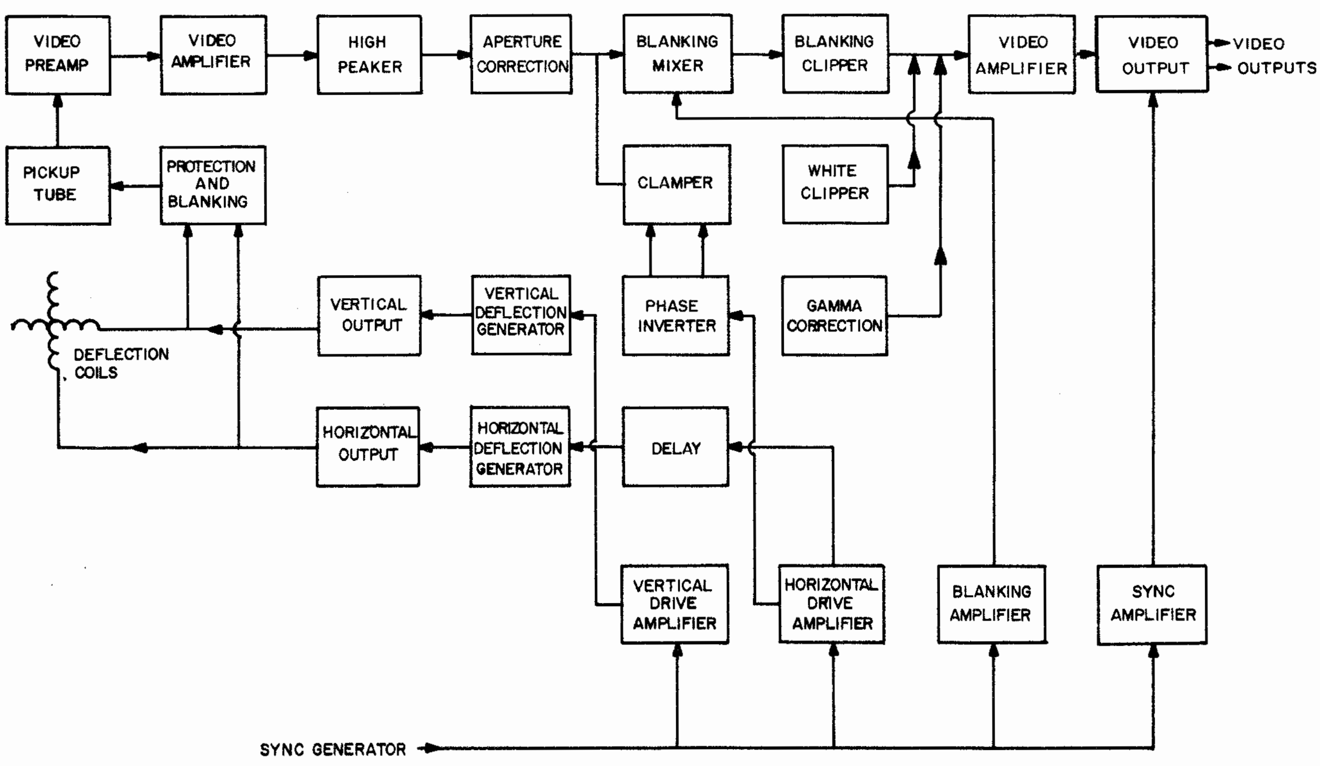


Figure 3-295. Typical Television Camera Circuitry

rolloff of previous stages. Ideally, it has uniform phase-response because no phase distortion was introduced in the process for which this stage compensates. The high-peaker and aperture-correction cicuits are always located ahead of the blanking-insertion circuitry, since the blanking is presumed to be a clean signal for which no correction is required. The high-peaker and aperture-correction circuits would distort the blanking signal if it passed through them. The next step in the processing of the video signal is the insertion of blanking. First, the video signal is clamped because it will have lost its DC component by the time it reaches this point due to capacitive coupling between video-amplifier stages. This DC component must be restored here in order to maintain a uniform blanking level among video signals that may have changing DC components. With a consistent relationship maintained between blanking level and the video signal at this point, they may be mixed. Subsequent DC-restorer circuits in monitors or other equipment may use the blanking level or the sync signal (to be added later) as a reference. The blanking clipper removes excess blanking signal strength from the mixed signal along with any spurious signals that may be present in the blanking interval. Blanking, pedestal, black, or setup level, as it is variously called, is established by setting the level at which the clipper operates. Gamma-correction and white-clipping circuitry (if used) will usually be located at this point in order to take advantage of the clamped and DC-coupled signal that these circuits require. The gamma-correction circuit modifies the light-to-video signal-transfer characteristic of the pickup tube. For example, if the points of principal interest in the signal are in the darker areas of the picture, the gain may be increased in this area and decreased in the lighter area. The white-clipper circuit determines an absolute maximum for the white signal. Peak clipping is desirable when the signal feeds a video tape recorder or an RF transmitter, and overmodulation must be avoided. The final circuit in the video-signal path is the video-output section which consists of one or more amplifier stages and the output stage. If sync is to be added within the camera, it will be done here. The output (power-amplifier) stage accommodates impedance matching, coupling the video signal into one or more 75-ohm coaxial cables for transmission.

3-10.16.1 Vertical Deflection

The vertical-deflection circuit causes the beam within the pickup tube to scan vertically in synchronism with the vertical-drive signal. In the standard EIA system the beam moves linearly from top to bottom, as viewed on a picture monitor, 60 times each second. To do this, an electromagnetic deflection coil requires a sawtooth current, and an electrostatic system requires a sawtooth voltage. When the vertical rate is only 60 hertz, the reactance of the vertical-deflection coil is very low compared to the resistance. Therefore in an electromagnetic deflection system an essentially sawtooth voltage waveform is required to develop a sawtooth current in the vertical-deflection coil. Figure 3-296 presents a simplified schematic of the vertical-deflection circuit of a currently popular television camera. The vertical-drive signal terminates in R28. Capacitor C8 provides DC isolation between the circuit supplying the signal and the base of transistor Q3. The latter is biased to be nonconducting in the absence of a vertical-drive pulse - the interval from B to C in Figure 3-297. During this interval C9 charges to approximately -0.4 volts. At time C, the vertical-drive pulse causes Q3 to conduct, such that its collector voltage drops to near its emitter voltage, and C9 discharges. At time D, Q3 returns to cutoff, C9 slowly charges to -0.4 volts, and the cycle repeats. Since C9 charges to only a small fraction of the power-supply voltage, it operates over a nearly linear portion of its charge curve; therefore, a sawtooth waveform appears across it. Resistor R27, in series with C9, prevents C9 from discharging completely, accounting for the step in the waveform at point B which corresponds to the top of the picture. The step compensates for the small amount of inductive reactance in the vertical-deflection coil. The value of R27 is chosen for best linearity at the top of the picture; in some cases this resistor will be variable. Vertical size is controlled by adjustment of the voltage to which C9 will charge. The sawtooth waveform from Q3 is fed to the base of mixer amplifier Q7, and the sawtooth waveform developed across R23, which is common to Q7 and the vertical-deflection coil, is fed to the emitter. The output of Q7 is proportional to the difference between the two signals. The signal developed across R23 is proportional to the current through the vertical-deflection coil and serves as a negative-feedback signal. This compensates for distortion and variations in gain in the intervening stages and for the increase in deflection-coil resistance that occurs with temperature changes in normal operation. Resistance R21 carries bias current to the emitter of Q7. Transistor Q4 is a conventional amplifier that feeds a complementary output stage consisting of Q5 and Q6. Since Q5 in PNP and Q6 is NPN, the two transistors are fed in phase. Stabistor diodes X3 and X4 allow coupling of the same signal to each of the output transistors although each has a different bias voltage. (Transistor Q6 is biased through direct coupling from Q4). Balance between the two transistors is maintained through a feedback network consisting of Zener diode X5, and C18, and C11.

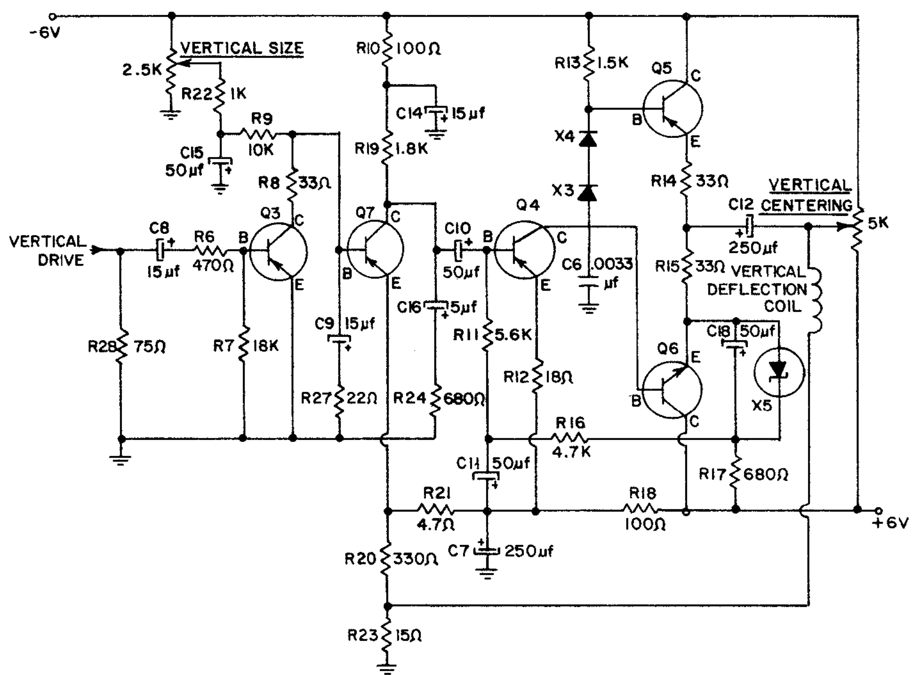


Figure 3-296. Vertical Deflection Circuit

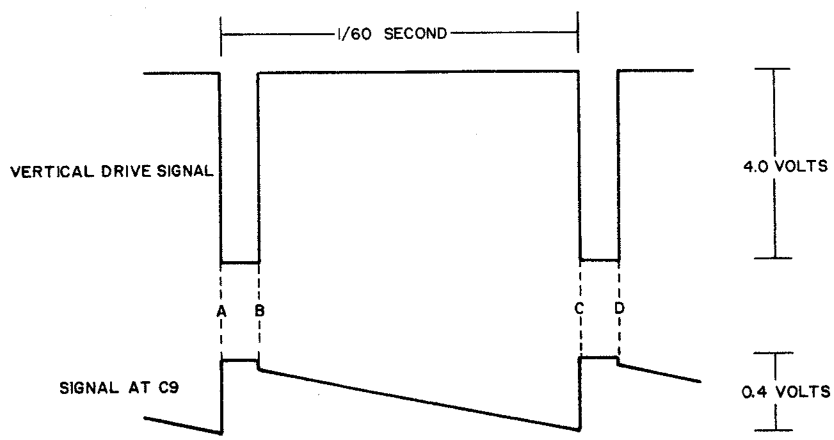


Figure 3-297. Vertical Deflection Timing

Transistors Q5 and Q6 have the low-impedance characteristic that the deflection-coil requires. With respect to the signal, the two transistors are parallel emitter-followers. The vertical-deflection coil is AC coupled through C12 to isolate the DC centering current from the emitters in the output stages. The centering control varies the DC bias current through the deflection coil. Figure 3-298 is a schematic of another popular vertical-deflection circuit. Transistor Q2, a conventional emitter follower, isolates the succeeding stages from other circuitry. Transistor Q3 is a conventional unijunction sawtooth generator. This normally free-running oscillator is locked in synchronism with the vertical-drive pulses. The sawtooth waveform results from the charging of capacitors C7 and C8 through R10, R11, and R12. The negative-going drive pulse at base 2 of Q3 causes it to conduct, and C7 and C8 then discharge through R5, R10, and Q3. Potentiometer R12 adjusts the free-running frequency of the oscillator. Output stage Q6 is driven by emitter follower Q5. A portion of the voltage

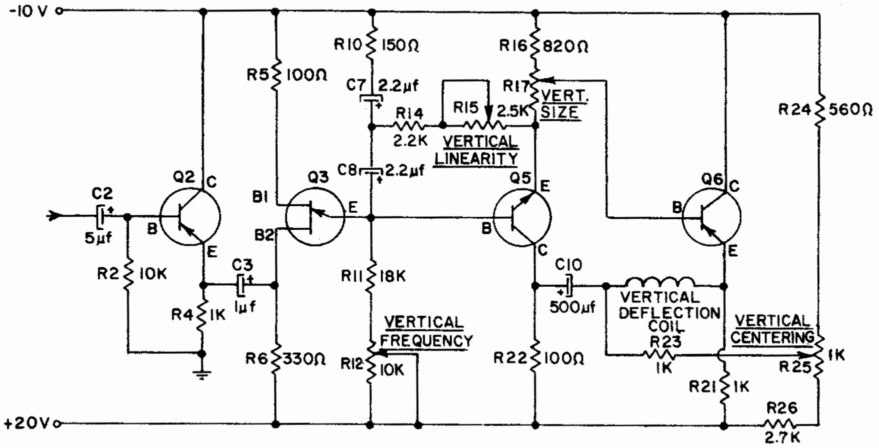


Figure 3-298. Vertical Deflection Circuitry

appearing at the emitter of Q5 is fed back to the junction of C7 and C8 in phase with the signal at the base of Q5, therefore it adds to the charge on C7. This makes the charging rate of C8 more nearly linear. Control R 15 adjusts vertical linearity by varying the amplitude of this feedback voltage. The output stage is also an emitter follower. Negative feedback from the lower end of the vertical-deflection coil is applied to the collector of Q5 through C10 and R22. The value of C10 is relatively large, providing negligible reactance to the sawtooth waveform. Resistor R22, in series with the vertical-deflection coil, will develop a sawtooth-voltage waveform is there is a sawtooth current through it and the coil. Thus distortion in the vertical-deflection coil appears as a negative-feedback signal at the collector of Q5. Vertical size is adjusted by controlling the amplitude of the signal at the base of Q6 through potentiometer R17. The vertical-centering control functions by setting a DC bias current in the vertical-deflection coil.

3-10.16.2 Horizontal Deflection

Because the horizontal-deflection coil in an electromagnetic-deflection system has a high ratio of reactance to resistance, square-wave pulses are necessary to cause a sawtooth current. In electrostatic-deflection systems a sawtooth waveform is used since no reactance of consequence is present. Circuitry is similar to that used in oscilloscopes. The circuits now described apply to cameras having electromagnetic-deflection systems, since it is used more extensively. An electro-magnetic-deflection coil requires a rectangular waveform to drive it. The horizontal-drive signal may therefore be amplified with a minimum of reshaping, and utilized as the horizontal-deflection signal. Figure 3-299 shows

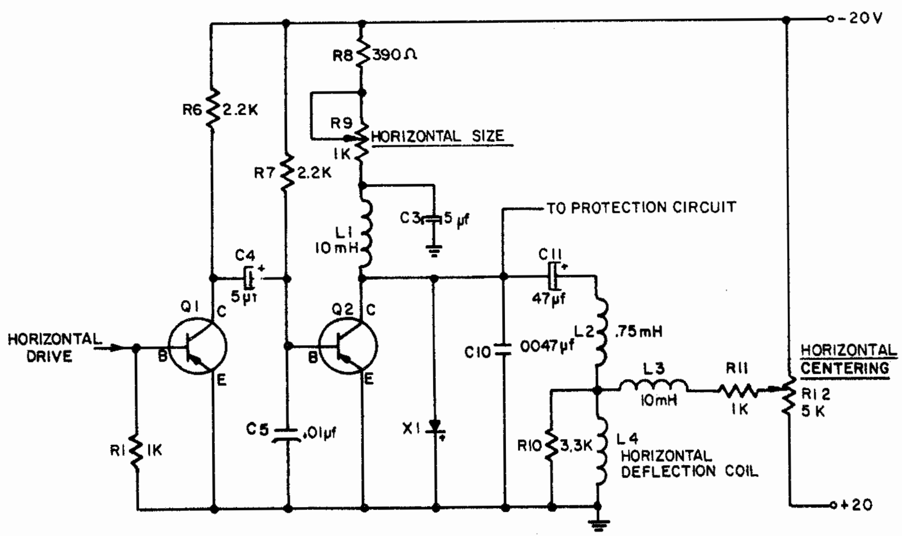


Figure 3-299. Horizontal Deflection Circuit

a circuit which uses this method. Transistor Q1 amplifies and inverts the horizontal-drive signal. This stage is normally biased beyond cutoff and therefore conducts only when a negative pulse is present. Capacitor C4 couples the signal from A1 to output stage Q2. Coil L2, in series with the horizontal-deflection coil, is a toroid having a very high Q; this increases the Q of the series combination of L2 and L4. The current in L4 therefore increases at a more uniform rate than it would in a lower Q circuit. In this way L2 improves the linearity of the current sawtooth waveform in L4. Inductor L2 resonates with the horizontal-drive pulse to shorten retrace time. Continued oscillation is avoided by damping diode X1. Capacitor C10 tunes the inductor to produce the desired retrace time. A DC current, adjustable at R12, centers the scan, while R11 limits the maximum current. The centering circuitry is isolated from the drive signal by L3 and from the deflection amplifier by C11. This circuit is often employed without a horizontal-linearity control. A linearity of one percent is attainable in the form shown. Figure 3-300 presents a different type of horizontal-deflection circuit. Transistor Q2 conducts during scan until the positive-going horizontal-drive pulse cuts Q2 off. The energy stored in the yoke (L4) collapses and resonates with Cll and C12 during retrace. Damper X2 prevents continued oscillation. Diode X1 clamps the horizontal-drive pulse to +2 volts. The flyback pulse is taken from the junction of C11 and C12 for use as a clamp in the video circuits; it is also used in the sweep-failure protection circuit. The linearity of the current sawtooth is enhanced by the insertion of inductor L2 in series with the deflection coil. Capacitor C18, in series with resistor R10, partially cancels the effect of L2 and thus provides a means of adjusting the linearity of the scan. Transistor Q4 stabilizes the voltage on Q2, and therefore stabilizes the width of the scan. Control R8 varies the width of the scan by adjusting the voltage at the collector of Q2. Network R12 and R13 can adjust the centering current to flow in either direction through the deflection coil since the low side of the yoke is connected to -6 volts.

3-10.16.3 Video Amplifiers

The feeble signal relayed from the pickup tube requires amplification between the pickup tube and the output jack of the camera-control unit. The video-amplifier stages boost this signal to a practical level for processing, and compensates for losses in the various processing circuits. Transistors are almost universally

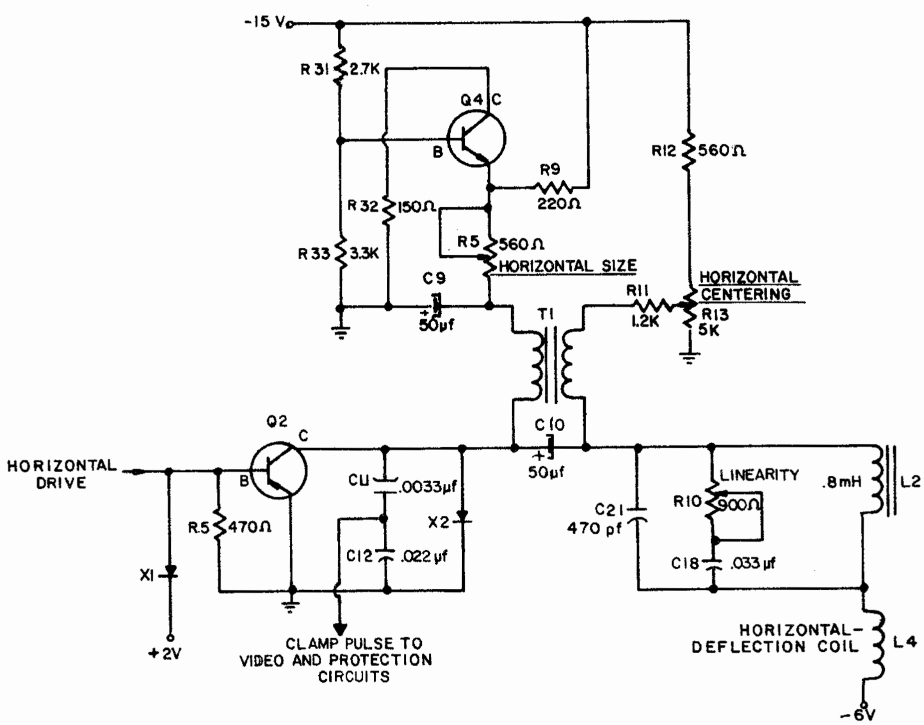


Figure 3-300. Horizontal-Deflection Circuitry

used for video amplifiers in modern cameras. Vacuum tubes, in the form of the nuvisitor, are the notable exception. These are frequently used in video-preamplifier stages and may be used in the video stages of cameras having exceptionally wide bandwidths. The principal difference between the video amplifier and the audio amplifier is in the bandwidth of frequencies that must be accommodated. While an audio amplifier that will uniformly amplify signals from 20 to 20,000 hertz is considered adequate for nearly any purpose, the video amplifier in the typical television camera or monitor must pass frequencies from 30 hertz to 10 megahertz. In addition, phase distortion that would go unnoticed in an audio amplifier would be unacceptable in a video amplifier. The video amplifier accomplishes its task of amplifying the broad band of frequencies with a minimum of distortion through a reduction in gain over a comparable audio amplifier. Usually more elaborate interstage-coupling networks are required. The distributed capacitance of an amplifier stage establishes its useful upper frequency limit. To reduce the shunting effect of this capacitance, smaller load resistors must be employed. Unfortunately, in the case of a video amplifier, the value must be sharply reduced to obtain satisfactory performance. This expedient sharply reduces the gain of the stage. However, a portion of the gain thus lost can be recovered through the use of inductive components in interstage-coupling networks.

3-10.16.3.1 RC-Coupled Circuits

Figure 3-301 shows a typical RC-coupled vacuum-tube video amplifier. This circuit differs from that of an audio amplifier only in that load resistor R_L, and grid resistor R_G have lower values. This reduces the shunting effect of the output capacitance of V1 (C_o), the distributed capacitance of the circuit (C_d), and the input capacitance of V2 (C_i). Reducing the values of the load resistors while the shunting capacitances remain fixed makes the reactances of the shunting capacitances relatively higher than the load resistances, and the shunting effect is consequently reduced. The reduction in load-resistor values, however, reduces the gain of the stage to the point where it will rarely be utilized in practical circuits. The transistor version of the RC-coupled video amplifiers shown in Figure 3-302 is similar to its audio counterpart except for the value of its components. As in a vacuum-tube circuit, the bandwidth is extended primarily through a reduction in the values of load resistors. Due to the inherent low impedance of transistors, the shunting effects of the distributed capacitance in transistor circuits are less severe than in a vacuum-tube circuit. As a result, the transistor version finds more frequent application than does its vacuum-tube counterpart.

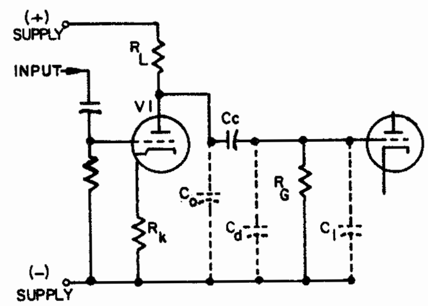


Figure 3-301. RC Coupled Video Amplifier

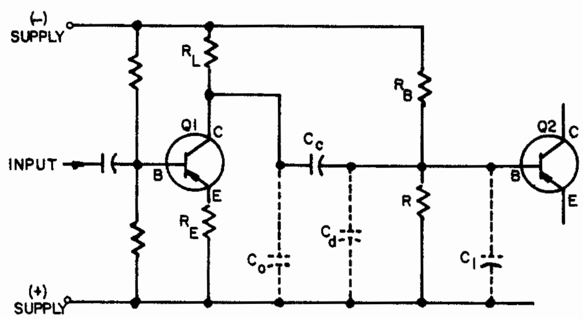


Figure 3-302. RC Coupled Transistor Video Amplifier

3-10.16.3.2 Shunt-Compensated Circuits

It is practical to build a video amplifier with a relatively large plate load and consequently more gain, provided the plate load remains constant for all frequencies of interest. Compensation for the rolloff in high-frequency response of amplifiers can be accomplished by adding a shunt peaking coil in series with the plate load resistor (L1 in Figure 3-303). The value of L1 is chosen to resonate with the distributed and input capacitances at a frequency about 1.4 times the highest frequency to be amplified. The total plate load of V1 has an impedance that increases with frequency, thereby increasing the gain of the stage at the higher frequencies. Ideally, this cancels the loss of high-frequency response due to the shunting effects of the various unavoidable capacitances in the circuit. Figure 3-304 shows the transistor equivalent of the shunt-compensated amplifier. Its principle of operation is similar to that of the vacuum-tube version.

3-10.16.3.3 Series-Compensated Circuits

A somewhat more effective method of overcoming the effects of undesired circuit capacitance

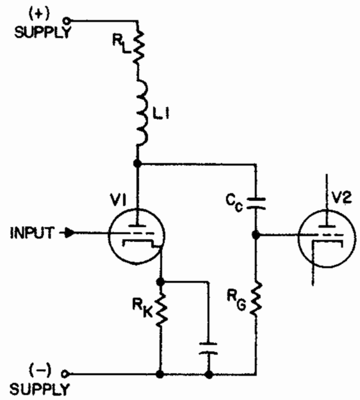


Figure 3-303. Shunt Compensated Vacuum Tube Video Amplifier

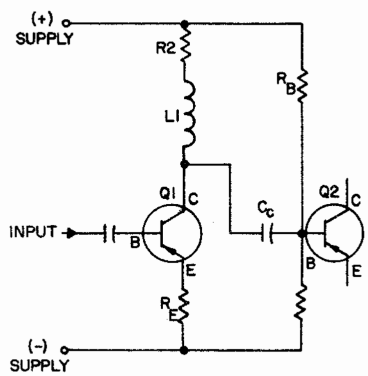


Figure 3-304. Shunt Compensated Transistor Video Amplifier

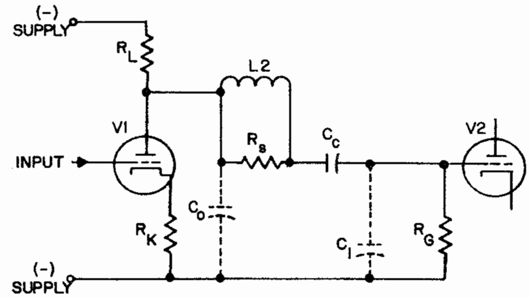


Figure 3-305. Series Compensated Vacuum Tube Video Amplifier

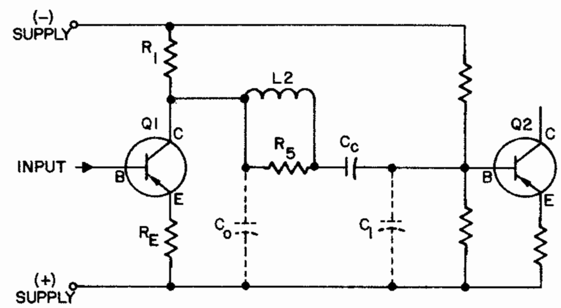


Figure 3-306. Series Compensated Transistor Video Amplifier

is to utilize them as part of the interstage coupling network. The requirement is that a constant-impedance load be placed on the stage feeding the network, and a constant-impedance load be placed on the stage feeding the network,and a constant-impedance source be supplied to the stage being fed. Any attenuation within the network must be uniform for all frequencies of interest. A suitable network is the low-pass pi filter. This is shown in Figure 3-305 and 3-306 where C_o and C_i form the input and output capacitors of the network. To form a constant impedance the value of one capacitor must be approximately double that of the other. The input capacitance of a vacuum tube or transistor is typically greater than the output capacitance, and in many cases the two have a correct ratio. Where the ratio is other than 2 to 1, Resistor R_S is added to lower the Q of L2 and smooth high-impedance peaks that may be present. R_L is also part of the filter network; its value is chosen to match the network impedance. The cutoff frequency of the network is chosen such as to be above the highest frequency to be amplified, and the resulting network offers almost no attenuation within the passband. The Delay of the signal through the circuit is greater than for the RC and shunt-compensated circuits. Since it is essentially uniform for all frequencies (in a properly designed circuit), this is ordinarilly of no consequence. The phase-versus-frequency characteristic of this circuit is superior to that of the circuits discussed previously. The interval base resistance of a transistor, appearing between the base terminal and the input capacitance lowers the Q of the network, therefore, this arrangement cannot be used efficiently in transistor circuits. For this reason, the sines-compensated amplifier will be found only occasionally in its transistorized version.

3-10.16.3.4 Shunt-Series Compensated Circuits

The most frequently used vacuum-tube video-amplifier circuit combines the two foregoing circuits to produce the shunt-series compensated amplifier, shown in Figure 3-307. This circuit can be considered as a low-pass filter having two sections. The addition of the dead end filter section (L2, R_L) to the single-section filter of the series-compensated circuit aids in maintaining a uniform plate-load impedance on V1. Addition of more filter sections to the right of L2 would result in some increase in gain over the network shown, but the two-section network approaches the practical maximum. Additional sections are therefore rarely, if ever, encountered in practical circuits. As in the series-compensated circuit, R_s is added when the ratio of C_o to C_i is not favorable. The network is reversible - the arrangement of Figure 3-305 is used when C_i is greater than C_o, and the input and output connections are reversed when C_o is greater than C_i. The arrangement shown is most usual, however.

3-10.16.3.5 Cathode Compensation

The cathode and emitter resistors have been left unbypassed in the foregoing circuits since this is often done in video amplifiers. There are two reasons. First, there is a decrease in distortion and an increase in stability of the stage due to the resulting negative feedback. Second, where very low frequencies are to be amplified, the value of the bypass capacitor must be very large. Placing a small capacitor across the cathode or emitter resistor will produce a response characteristic that rises with frequency and therefore compensates for high-frequency losses due to distributed capacitance elsewhere in the circuit. In Figure 3-308 the combination C_k and R_k have a time constant that is long for frequencies to be boosted and short for frequencies not requiring any boost. This circuit is extensively used in both its vacuum-tube and transistor versions. The approximate gains of the foregoing types of amplifiers are shown in Table 3-25.

3-10.16.3.6 Low-Frequency Compensation

The video amplifier may require low-frequency as well as high-frequency compensation if it is necessary to use a smaller value of coupling capacitor than would be ideal for passing the lowest frequency of interest. The smaller values are often used because electrically large capacitors are also physically large; they may introduce an excessive amount of stray capacitance into the circuit, causing rolloff in the high-frequency response of the stage. The most common method of low-frequency correction in both vacuum-tube and transistor versions is shown in Figure 3-309. At high frequencies C1 bypasses R2, leaving only R1 as the plate or collector load resistor. As the signal frequency decreases, the reactance of C1 increases. At lower frequencies the bypassing effect of C1 becomes insignificant, and R2 becomes a load

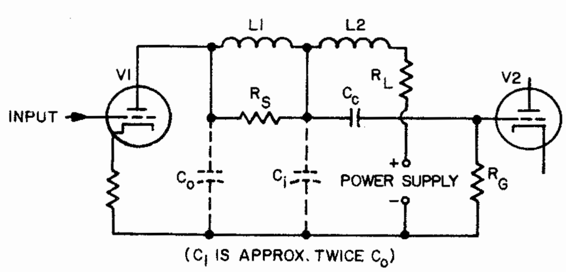


Figure 3-307. Series Shunt Compensated Video Amplifier

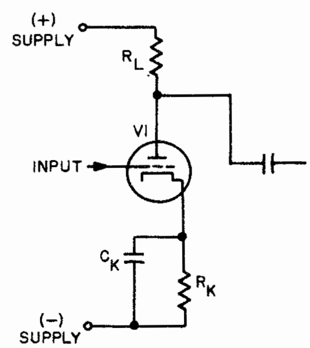


Figure 3-308. Cathode Compensated Video Amplifier

Table 3-25. Relative Gains of Video Amplifiers

TYPE OF AMPLIFIER COMPENSATION	RELATIVE GAIN OF CIRCUIT	
	VACUUM-TUBE	TRANSISTOR
None	1.0	1.0
Shunt	1.4	1.7
Series	2.0	1.4
Shunt-Series	2.6	2.0
Cathode	1.4	1.7

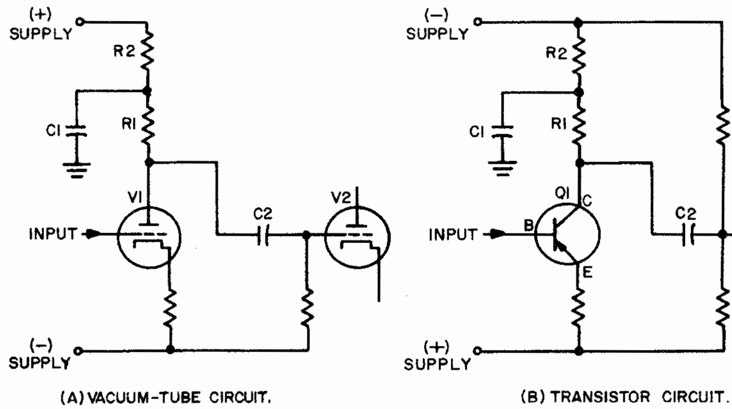


Figure 3-309. Low Frequency Compensation

resistor. This, added to R1, increases the gain of the stage.

3-10.16.3.7 Transistor Video Amplifiers

The approach to the design of transistor video amplifiers generally differs from that of the vacuum-tube type for several reasons. First, transistor circuitry has lower impedances. Second, internal characteristics of transistors are decidedly different from those of vacuum tubes. In addition, intermixing NPN and PNP types makes circuits practical that would be impossible using vacuum tubes. Finally, the forward bias required by linear transistor amplifiers affords opptunities for direct coupling that do not exist in vacuum-tube circuits. Negative-feedback circuits provide high- and low-frequency correction when required. Hence, peaking coils, found in nearly every vacuum-tube video amplifier, are used rather sparingly in transistor video amplifiers. Feedback networks often occur around transistors in pairs, or even trios. Such combinations are so common that they are usually regarded as single stages. A few of the basic forms should be explored before some of the unique applications are analyzed. Figure 3-310 shows a voltage-feedback pair that has a relatively high input impedance and low output impedance. A variation of the circuit may be used for a video-output stage. The DC coupling and feedback provide the principal DC stabilization. Both stages in the circuit are voltage amplifiers, and R3 provides a negative-feedback path from the output to the emitter of Q1. The feedback signal cancels distortion occuring in the circuit, flattens frequency response, and stabilizes gain. It also increases the input impedance of the circuit. Figure 3-311 shows the circuit of a current-feedback pair. Both transistors are common-emitter amplifiers, and the feedback path is through R3. The out-of-phase signal at the emitter of Q2 is mixed with the input signal at the base of Q1. DC coupling and feedback provide the principal DC stabilization, or it may be provided by

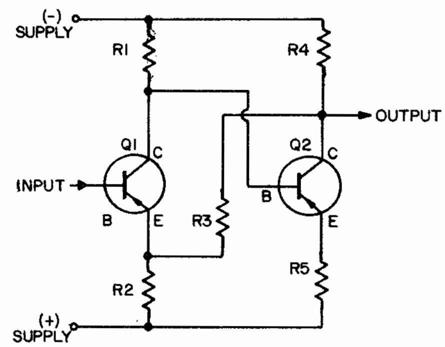


Figure 3-310. Voltage Feedback Pair

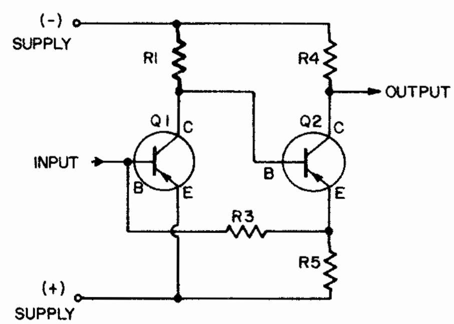


Figure 3-311. Current Feedback Pair

using AC coupled feedback and inserting a bypassed resistor at the emitter of Q1. AC coupled feedback may be employed to correct response at the lower frequencies. In this case a capacitor having a relatively high reactance at low frequencies and a low reactance at the higher frequencies will be placed in series with R3, giving more cancellation (less gain) at the higher frequencies. A small capacitor placed across R5 will give additional gain at high frequencies, should this type of correction be required. This circuit has a low-impedance input and a high-impedance output. The configuration of Figure 3-312 has an extremely

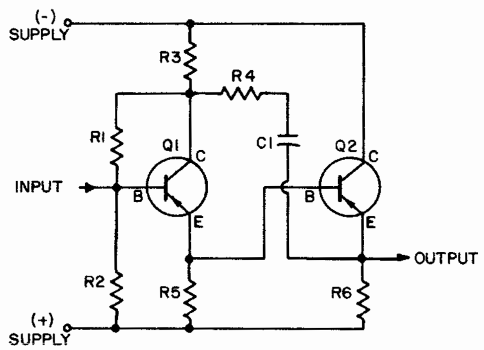


Figure 3-312. Direct Coupled Amplifier

high input impedance and a relatively low output impedance. It is most useful in loop-through video inputs where minimum loading of external circuits is desirable. Two negative-feedback paths appear in this circuit: the first is through R1; the combination of R4 and C1 form the second. These increase the input impedance of the circuit, cancel distortion which may occur, and stabilize gain. Direct coupling also aids gain stabilization. Figure 3-313 shows the voltage-feedback

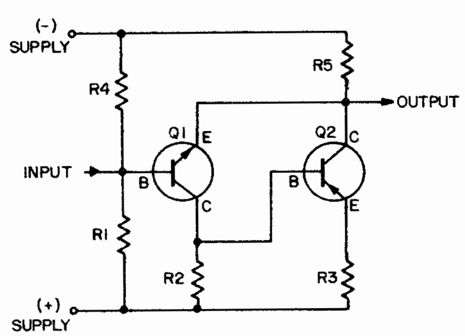


Figure 2-313. Complementary Voltage Feedback Pair

pair that uses complementary transistors. The circuit is similar in operation and characteristics to that of Figure 3-308. In the version shown the feedback path is a direct connection between the collector of Q2 and the emitter of Q1. An example of video amplifier in a practical circuit is shown in Figure 3-314. Transistors

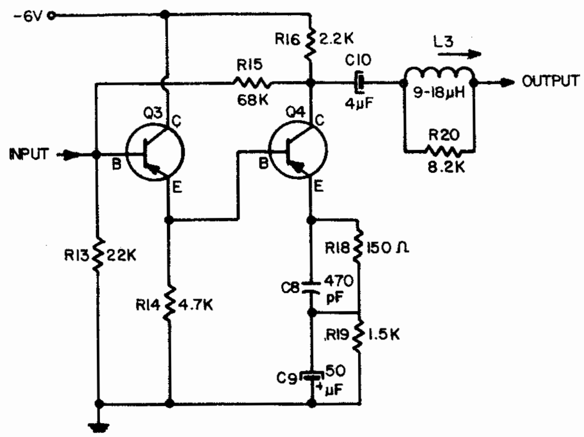


Figure 3-314. Video Amplifier Circuit

Q3 and Q4 function as a single stage. Since Q3 has a large current gain and a low output impedance, a large voltage gain in Q4 is possible. Feedback resistor R15 provides DC stabilization as well as improved frequency response and reduced distortion. Capacitor C8 aids in high-frequency correction since it bypasses R18 at the higher frequencies only. The series-compensation network, L3 and R20, provides high-frequency correction between Q4 and the following stage. An example of the variety of video-amplifier circuits that may be utilized in a single channel is shown in Figure 3-315. The first stage is an emitter follower that has current gain and a low-impedance output. It drives common-base voltage amplifier Q4. In this stage the combination of R76 and C36 form a high-frequency correction network. The circuit in the base of Q4 is degenerative at low frequencies only; at higher frequencies it is at AC ground, and the stage has full gain. The combination of L3 and R20 is a conventional series-peaking network. Transistors Q5 and Q6 form a current- and voltage-amplifier combination. The network of R77, R22, and C37 is for DC stablization only because C37 removes the signal from the network, and no negative feedback results. C15 and R27 are a high-frequency-compensation network for Q6. R25, C14, and C38 form a decoupling network.

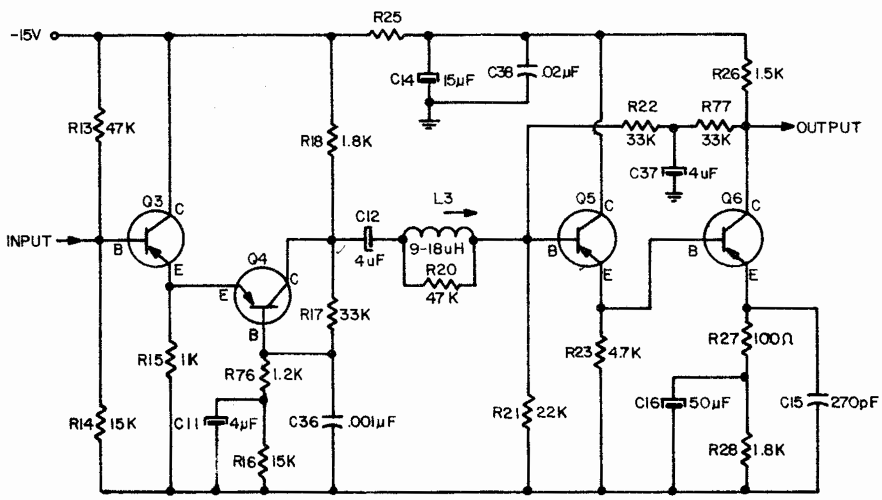


Figure 3-315. Video Amplifier Circuits

3-10.16.4 Video Preamplifiers

The video preamplifier is a special form of amplifier which couples the signal from the pickup tube to the following video stages. Because of the unique requirements of this stage and the many variations in use, it is discussed apart from the more conventional video amplifiers. Virtually all cameras now being marketed use transistor amplifier stages. They therefore require a matching circuit between the very high output impedance of the pickup tube and the relatively low input impedance of a transistor amplifer. A vacuum-tube (usually in the form of a nuvistor) cathode follower or a transistor emitter-follower will therefore often be a component of this circuit.

3-10.16.4.1 Vacuum-Tube Cathode Follower

Figure 3-316 shows the input circuitry of a camera using a vacuum-tube cathode follower for its video preamplifier. The stage has no voltage gain, but it does have considerable current gain. This is desirable, since the output current of the vidicon tube, for example, is ordinarily much less than one microampere. The employment of two tubes in parallel increases the transconductance of the stage and, therefore, increases the current gain. The low-impedance output of the stage provides a suitable match for succeeding transistorized gain stages. Components L2 and R20 form a prepeaking circuit designed to reduce high-frequency noise. Inductor L2 resonates

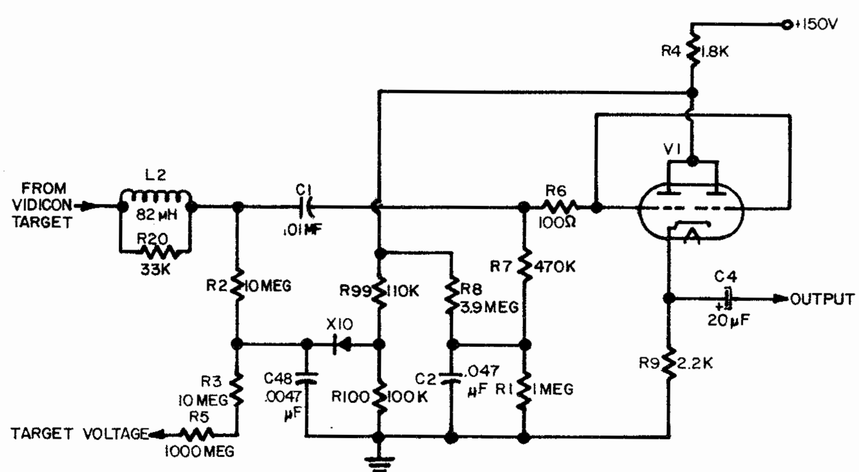


Figure 3-316. Cathode Follower Video Preamplifier

with the distributed capacitance of the circuit to cause a peak in the response around 7 megahertz. A succeeding stage has a response curve which complements that of L2 and R20, thus giving the amplifier as a whole uniform frequency response, if other corrections to the signal are disregarded. Resistor R20 broadens the resonance curve of L2. (See Figure 3-317). An

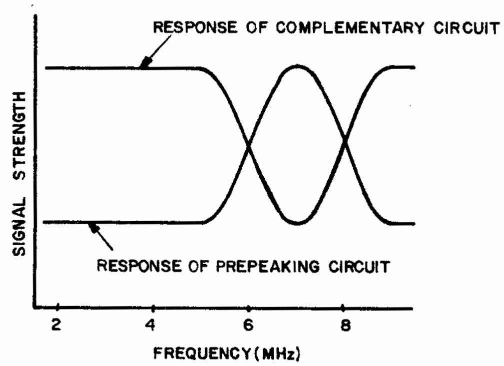


Figure 3-317. Response of Prepeaking Circuit

additional refinement is the inclusion of automatic sensitivity (target) control at two points in the preamplifier. One is the use of a very large resistor (R5-1000 meg) in series with the target voltage supply. As the amount of light falling on the vidicon increases, the target current increases. This causes the voltage drop across R5 to increase, thereby decreasing the voltage at the vidicon target and the signal output of the vidicon. The large value of R5 causes the signal output of the vidicon to remain relatively constant. The second form of automatic target-voltage control consists of the network R4, R99, R100, X10, and C48. A portion of the output of C1 is developed across resistor R4. This is rectified by X10 and fed to the vidicon target in opposition to its DC supply. As the signal from the vidicon increases, the signal across R4 increases and so does the DC output of X10. This, being in opposition to the normal voltage supply, causes the total target voltage to decrease and reduces the signal output of the vidicon.

3-10.16.4.2 Two-Stage Vacuum-Tube Amplifier

The input circuitry of a camera that uses a conventional two-stage vacuum-tube amplifier is shown in Figure 3-318. Note that the plate loads of the vacuum tubes are returned to ground, while the cathodes are connected to the -20 volt supply. This unconventional connection enables the vacuum-tubes to operate from the same power supply as the transistors, and permits direct coupling to the transistor emitter follower. The target voltage supply is decoupled by C1 and R2. Inductor L1 is a shunt-peaking coil for improving the high-frequency response of the stage. In this particular circuit the video signal undergoes considerable amplification before being fed to the transistor circuits. Emitter follower Q1 serves as an impedance-matching circuit between the vacuum-tube circuitry and the succeeding transistor amplifiers.

3-10.16.4.3 Transistor Emitter Follower

Figure 3-319 shows the circuit of a video preamplifier that uses a transistor emitter follower (Q1) as an impedance match between the vidicon tube and the succeeding voltage amplifiers. The

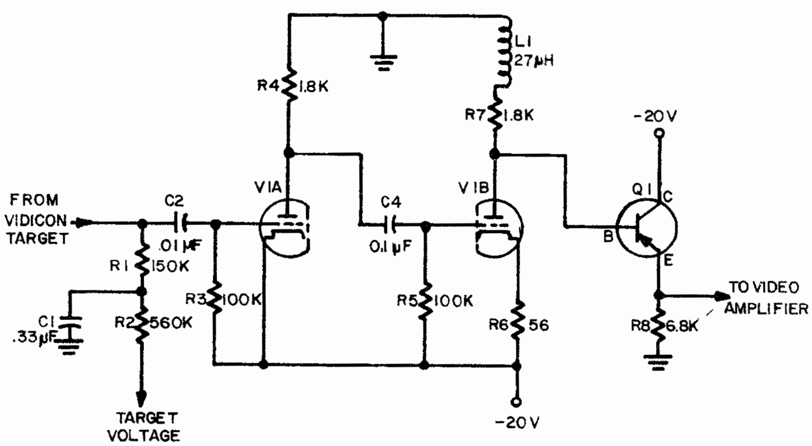


Figure 3-318. Two Stage Vacuum Tube Video Preamplifier

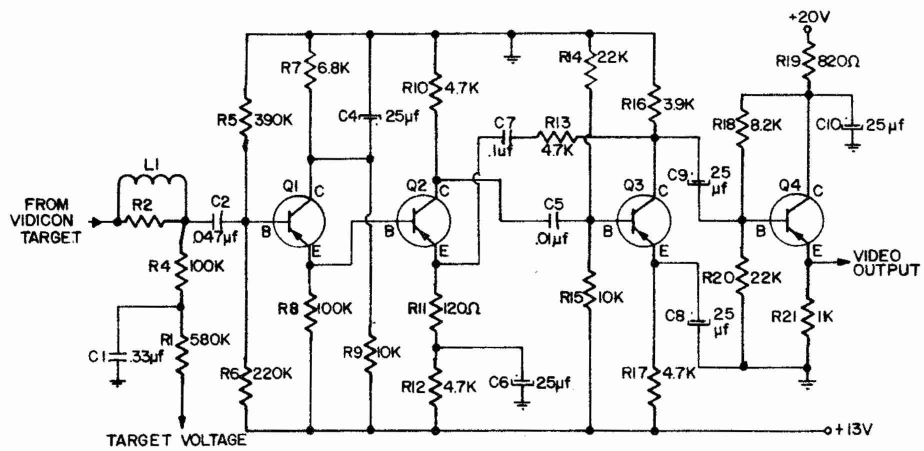


Figure 3-319. Emitter Follower Video Preamplifier

transistor also provides current gain. A prepeaking network (L1 and R2) boosts the high-frequency response of the circuit, thus minimizing the amount of correction required in later stages for the stray capacitance in the input circuit. Transistors Q2 and Q3 are a conventional voltage-feedback pair in which the voltage gain of the circuit takes place. A negative-feedback network (C7 and R13) stabilizes the gain of Q2 and Q3 and minimizes the distortion of these two stages.

3-10.16.4.4 Transistor Current-Feedback Pair

Figure 3-320 shows the input circuit of another type of television camera, this one having an all-transistor video preamplifier. The network consisting of R1, R2, C1, and C2 decouples the target voltage supply, while R3 serves as the vidicon load resistor. The DC target potential is isolated from the base of Q1 by C3. L1 and R4 form a prepeaking network. Transistor Q1 is an emitter follower that matches the high output impedance of the vidicon tube to the relatively low input impedance of voltage amplifier Q2. Resistor R7 forms a negative-feedback loop, stabilizing the gain and frequency response of the transistor pair. It also increases the input impedance of Q1. Components R9 and C7 are a decoupling network. Signals through emitter resistor

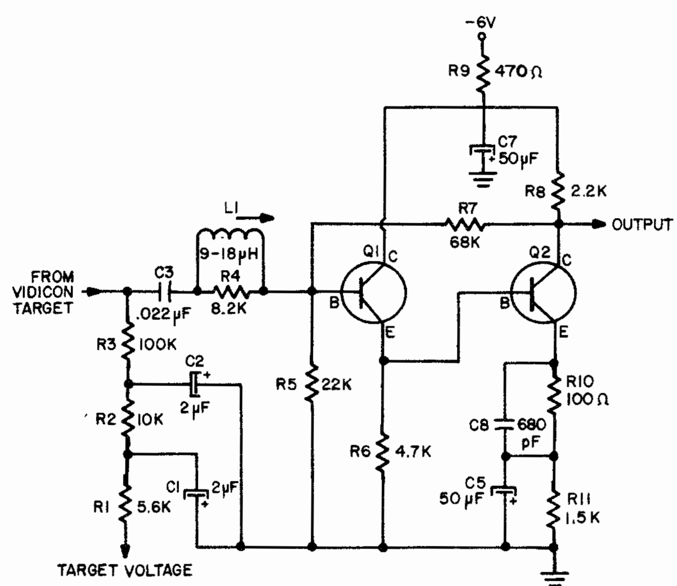


Figure 3-320. Transistor Current-Feedback Video Preamplifier

R10 are bypassed by small capacitor C8 at high frequencies only, therefore Q2 has more gain at high frequencies than at the lower frequencies. This compensates in part for the high-frequency rolloff caused by the distributed capacitance in the vidicon output circuit.

3-10.16.4.5 Nuvistor Cascode Amplifier

Figure 3-321 shows the input circuitry of a camera having an all-nuvistor video amplifier. The low-noise characteristic of the nuvistor is used to advantage in this circuit featuring a 32-megahertz band-pass, a high input impedance, and an unusally low noise figure. It is also characterized by high gain. Target voltage for the vidicon tube is fed through resistors R1 and R2. Test points TP1 and TP2 on either side of the 75-ohm resistor (R2) provide a convenient connection for a test-signal generator since it can be terminated in 75 ohms within the operating circuitry. Fixed bias is applied to V1 through voltage divider R4 and R5. Resistor R4 also functions as the grid resistor of V1. In this cascode circuit the plate load of V1 consists of V2 plus the plateload circuit of V2. The output of V1 is direct coupled to V2, which functions as a grounded-grid amplifier since the grid is held at AC ground by C7 and C37. Capacitor C37 is the principal bypass capacitor, while C7 provides a low-reactance path for high frequency signals. This is necessitated by a small amount of inductance present within electrolytic capacitor C37. Fixed DC bias for V2 is applied through voltage divider R7 and R8. Inductor L1 is a series-peaking coil; L2 is a shunt-peaking coil; and C9 and R10 form a decoupling network. The video signal is inverted by V1, but not by V2. Consequently, the signal undergoes a single inversion in a cascode stage.

3-10.16.4.6 Hybird Cascode Preamplifier

Figure 3-322 shows the circuit of a hybrid cascode preamplifier. The input of V1 is similar to the corresponding circuitry in Figure 3-321. Plate current is fed via decoupling network R5, R6, and C6. The emitter current of Q2 flows through R7 rather than through the nuvistor. The video signal is coupled from V1 to Q2 by C7. In circuits using NPN transistors, part or all of the vacuum-tube current may flow through the transistor. The base Q2 is bypassed to ground through C8. The DC bias for Q2 is supplied by a voltage divider consisting of R8 and R9. Inductor L1 is a shunt-peaking coil. This particular circuit is designed to have a rising frequency response which compensates for the distributed capacitance of the input circuit of V1; therefore, the slug of L1 serves as the high-peaker adjustment. Capacitor C9 couples the output to the next stage through L2, a series-peaking coil. The hybrid cascode circuit is used extensively as the video

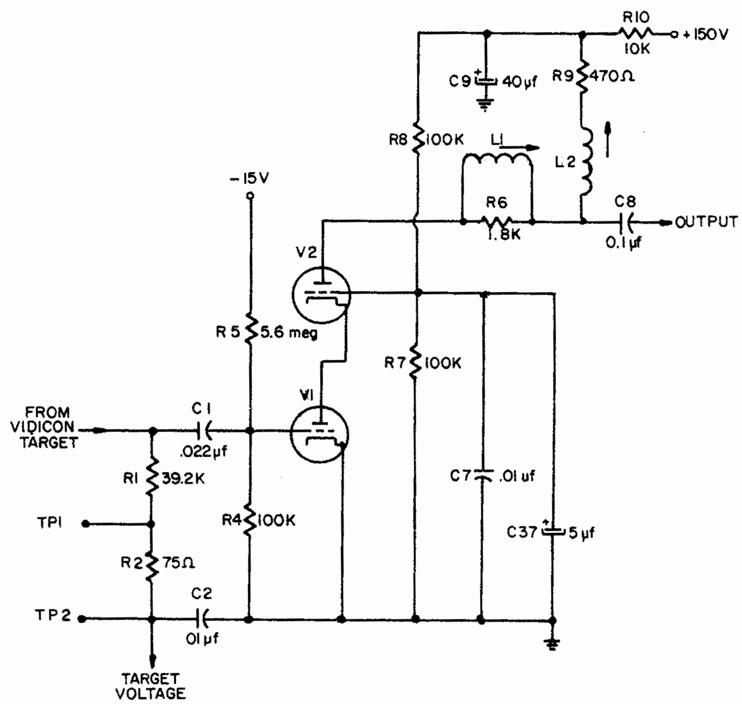


Figure 3-321. Nuvistor Cascode Video Preamplifier

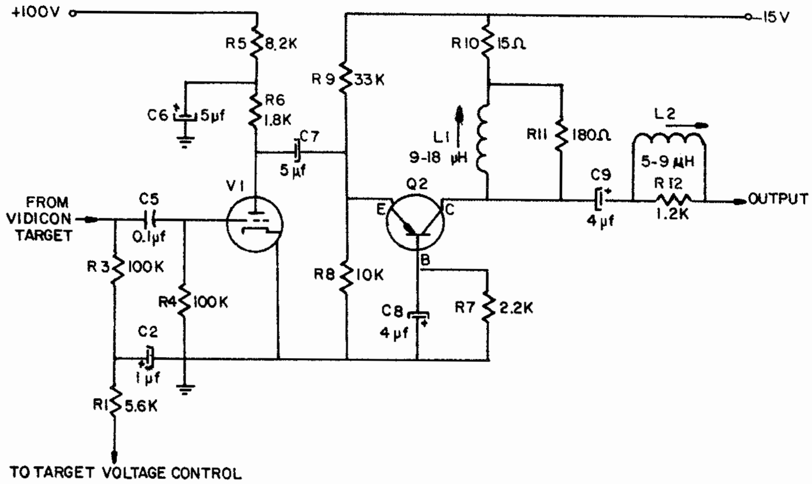


Figure 3-322. Hybrid Cascode Video Preamplifier

preamplifier of television cameras because of its exceptionally low noise figure.

3-10.16.5 Video-Output Circuits

Most major components of a television system are coupled by means of 75-ohm coaxial cables. Since serveral hundred feet of cable may be used on occasion for this purpose, an accurate 75-ohm impedance is required to feed such a line. The purpose of the video-output stage, therefore, is to provide coupling between the amplifier stages and one or more video lines. The simplest possible circuit for achieving this is the emitter follower; Figure 3-323 shows

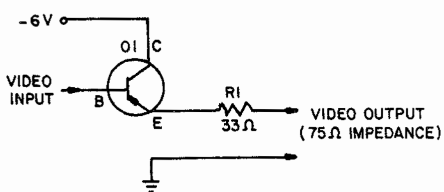


Figure 3-323. Simple Video Output Circuit

an example. Source-termination resistor R1 makes up the difference between the output impedance of Q1 and the actual load, and the stage output is therefore 75 ohms. In the circuit shown, a 75-ohm load resistor in the next unit forms a portion of the emitter load. This is often undesirable because the DC supply of the transistor must be fed through the load resistor. Also AC coupling with this circuit is difficult to achieve because of the large value of the coupling capacitor required. Other shortcomings are: the actual impedance of the circuit is somewhat dependent upon the value of load resistance used; and the output impedance of the circuit varies somewhat with changing signal levels. The emitter follwer is favored as the output stage of a camera head located separately from its control unit. In this case, the foregoing shortcomings are avoided since the circuit at the receiving end of the coaxial cable is designed specifically to operate from the emitter follower. A voltage-feedback pair may be adapted to feed a coaxial line. Figure 3-324 shows a circuit of this type. Here, negative feedback through C35 aids in lowering the output

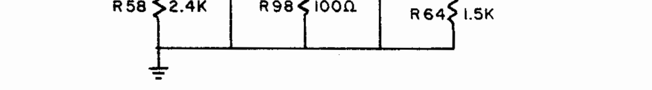


Figure 3-324. Video-Output Circuit

impedance of the circuit in addition to the usual benefits of negative feedback. Note that output-coupling capacitor C34 is included within the feedback loop; therefore, low-frequency losses within this component are compensated for (in part) by the feedback. Network C51 and R98 provides additional compensation for low-frequency losses through C34. A series amplifier used extensively in video-output applications -- commonly called the "totem-pole" circuit -- is shown in Figure 3-325. The two amplifiers acting in series are capable of a very high power gain, so they are suitable for feeding a video line. The circuit is characterized by a high input impedance and a low output impedance — 50 ohms or less for the vacuum-tube version, and only a few ohms for transistor circuitry. Where a 75-ohm impedance is desired, building-out resistors can be inserted to make up the difference. The low output impedance makes possible multiple isolated outputs from the same stage. Tube V1 is a phase splitter in the vacuum-tube version Figure 3-325A. The signal at its cathode forms part of the output. Signal from the plate of V1 feeds the grid of V2, which acts as a conventional plate-coupled amplifier. When a positive-going signal, as shown in Figure 3-325A arrives at the grid of V1, it causes the voltage at the plate to decrease and the voltage at the cathode to increase with respect to ground. The decrease at the plate of V1 is fed to the grid of V2, causing the voltage at the plate of V2 to increase. This is the same as the signal at the cathode of V1, and, since the voltage swings are in phase, they are greater than they would be with a single tube. The transistor version in Figure 3-325B functions similarly to the vacuum-tube version.

3-10.16.6 High-Peaker Circuits

While the interelectrode capacitance of the vidicon tube is not high by vacuum-tube standards (about 3 pF from the target to all other electrodes for a one-inch vidicon), the output impedance of the tube is so high that the input impedance of the first video-amplifier stage must also be kept high to avoid a serious loss of signal. For this reason the shunting effects of the interelectrode capacitance, the stray-wiring capacitance, and the input capacitance of the first amplifier stage become very significant. The result of this undesirable circuit capacitance is a roll-off of the video signal as frequency increases due to its shunting effect. In addition, the high frequencies are delayed longer than the lower frequencies, so there is phase distortion. Figure 3-326 shows the effect of shunt capacitance in the input circuitry. The loss of high frequencies results in a loss of fine detail in the picture; the phase shift appears as smear. The latter is the more noticeable of the two effects when a picture is viewed on a monitor. The high-peaker circuit compensates for the shunt capacities in the input circuit of the video preamplifier. It follows then that it should have frequency and phase characteristics complementary to those of the input circuit. In its simplest form the circuit consists of a shunt-peaking coil in a video amplifier stage, tuned for a rising frequency characteristic. Coil L1 in Figure 3-322 is an example of this type. Figure 3-327 shows the usual type of high-peaker circuit. Capacitor C1 is chosen

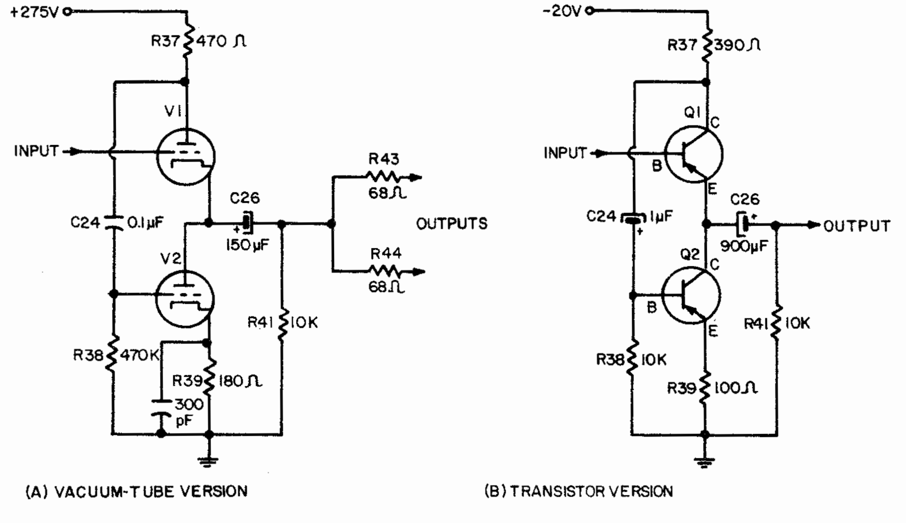


Figure 3-325. Totem Pole Video Output Circuits

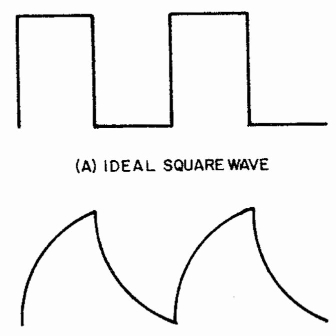


Figure 3-326. Effect of Shunt Capacitance in Input Circuit

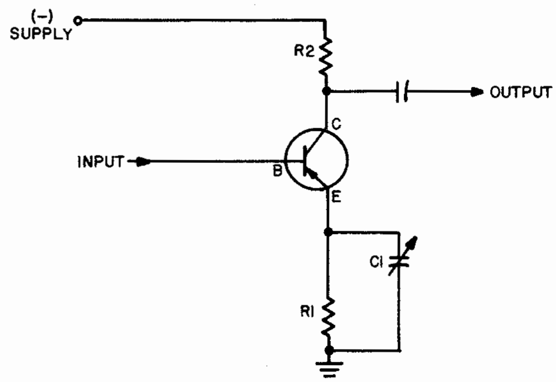


Figure 3-327. High-Peaker Circuit

to provide a low-reactance path, bypassing R1 at high frequencies. This gives the stage a gain characteristic that rises with frequency. Ordinarily, C1 is adjustable to allow for variations in input capacitance and, in some cases, cable length. In cameras where the control unit is located apart from the camera, compensation for distributed capacitance and inductance in the coaxial cable carrying the video signal is often required. The compensating circuit must have characteristics that are very similar to those of the high-peaker circuit; therefore, similar circuits are used. In a few cameras one circuit performs both functions.

3-10.16.7 Aperture-Correction Circuits

The finite size of the scanning beam within the pickup tube causes a decrease in the signal output as frequency increases due to the inability of the spot to resolve details smaller than itself. If the beam were infinitely small in diameter, the signal output of the tube would be a perfect square wave as the beam passed across the image of a vertical bar. Figure 3-328A illustrates this. Figure 3-328B shows how high-frequency components are lost as a beam of finite size passes across the image of a vertical bar. The gradual transition of the beam across the bar causes the signal current to change gradually rather than instantly. This distortion appears in the picture as a loss of fine detail. Since the process that causes this decrease in high-frequency signals does not introduce phase shift, compensation — if it is to be compensated for — must be done without phase shift. The usual amplifier circuit having a response characteristic which increases with frequency, such as the high peaker, is not suitable for this purpose. The aperture-correction circuit in Figure 3-329 uses a delay line that is one-half a wavelength at the frequency for which maximum boost is required — between 4 and 10 megahertz in 525-line systems. At these frequencies the signals at the two grids are 180 degrees out of phase. At lower frequencies the phase shift is somewhat less. Maximum phase shift occurs when the two potentiometers are in their maximum clockwise position. The plate of V1 then has no load resistor (the tube functions as a cathode follower), and the plate load of V2 consists of all of R1 plus all of R2. The V2 stage, therefore, has maximum gain. At the frequency of maximum correction the two grids are 180 degrees out of phase. The signal from the cathode of V1, in phase with the grid of V1, is coupled to the cathode of V2. The cathode and grid signals of V2 are 180 degrees out of phase with each other, so the stage has maximum input. At the lower frequencies the two grids are essentially in phase; therefore, the grid and cathode of V2 have in-phase signals, causing a decrease in gain at the lower frequencies. The stage has more gain at high frequencies than at the lower frequencies, but no phase-versus-frequency distortion was introduced in the circuit. The requirements for aperture correction are thus fulfilled. When the two potentiometers are at their maximum counterclockwise settings, the two tubes are in parallel. At low frequencies the signal currents through R2 add, giving the stage maximum output. At the higher frequencies, where the signals at the grids are out of phase, the signal currents through R2 cancel, and the circuit has maximum gain. Figure 3-330 shows the circuit of a transistor aperture corrector using a delay line. Its operation is similar to that of its vacuum-tube counterpart. A feature of this particular circuit is the addition of S1. When S1 is in the closed position, the circuit functions as a normal aperture-correction circuit. When S1 is open, the emitter coupling between the two transistors is broken, and the circuit functions as a normal aperture-correction

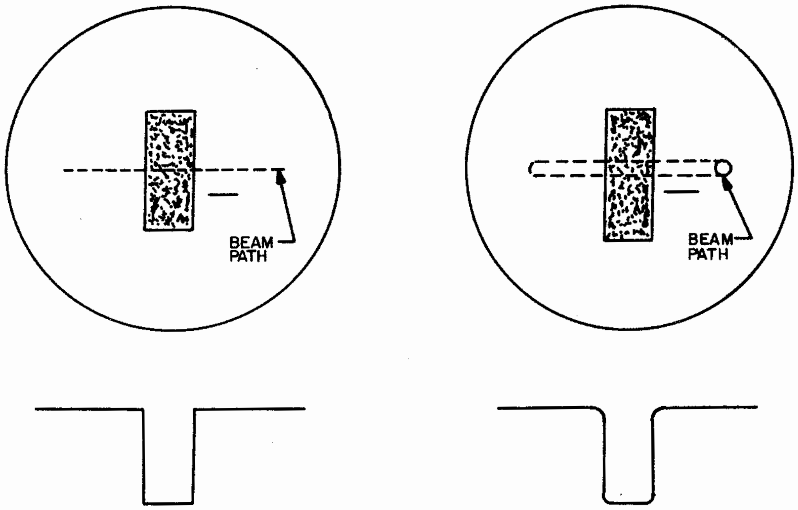


Figure 3-328. Output Waveform of Camera Tube

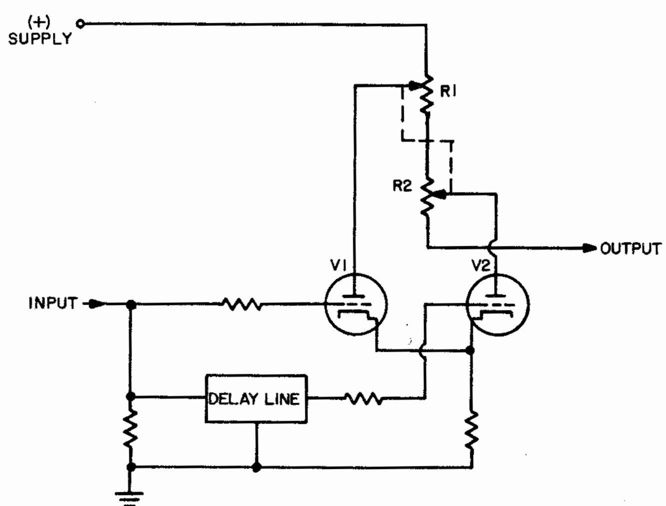


Figure 3-329. Vacuum Tube Aperture Correction Circuit

circuit. When S1 is open, the emitter coupling between the two transistors is broken, and the circuit functions as two amplifiers in parallel, but with the input signals out of phase at the frequency for which the delay line is one half wavelength long. Signals at this frequency cancel in the output circuit when the potentiometers are in their maximum counterclockwise position. As frequency decreases, the amount of cancellation decreases, giving an increase in putout. When the potentiometers are in their maximum clockwise position, only Q1 has an output and the circuit has flat response. The line in this particular circuit has a delay of .05 microsecond. Since this represents one-half a wavelength at 10 megahertz, maximum correction will occur at this frequency. Because of the low impedance of the base circuits of transistor amplifiers, more efficient use can be made of a delay line by placing it in the collector circuit. Figure 3-331 shows this arrangement. Here the signal is coupled from Q1 to the emitter of Q2 through R4 and R3. At low

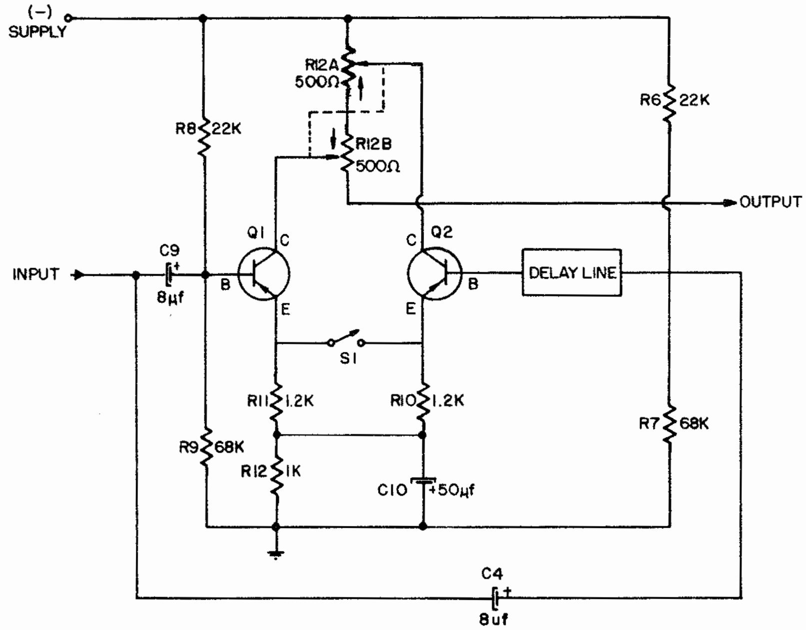


Figure 3-330. Transistorized Aperture Correction Circuit

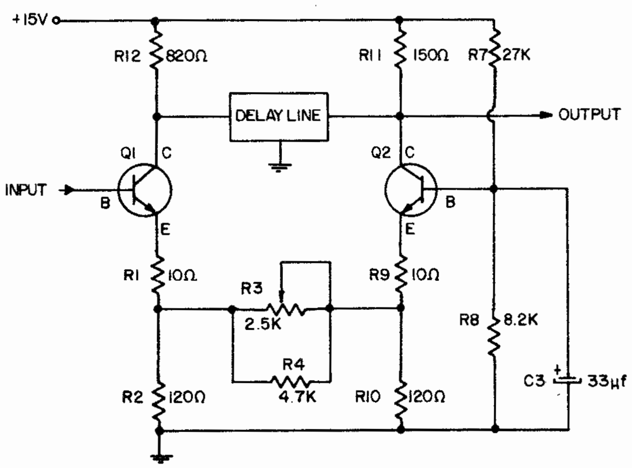


Figure 3-331. Aperture Correction Circuit with Improved Frequency Response

frequencies, where the effect of the delay line is insignificant, the collector currents of the two transistors cancel, and the stage has minimum gain. At the frequency where the delay line is one-half a wavelength, the currents through R11 are in phase, giving the circuit maximum output. The amount of correction is controllable at R3. When R3 is at its minimum setting, the amount of signal coupled to the emitter of Q2 is at a maximum, resulting in minimum low-frequency addition and minimum high-frequency cancellation. This gives the circuit its most uniform frequency-response characteristic. Figure 3-332 shows a different approach to aperture correction. In this circuit video signals that have been rounded off due to the finite size of the scanning beam are sharpened by a transformer that rings at the frequency for which maximum correction is required. To understand the operation of the circuit, refer to the waveforms. Video signals are coupled to the output from both the emitter and collector of Q1. A square-wave input (A) will be delayed in the emitter circuit (B) because of C1. The signal at the collector (C) will cause transformer T1, resonating with distributed capacitance C_D, to ring. The ringing signal is coupled to the secondary of T1, adding to waveform (B). The amount of correction is adjustable at R1. In the maximum-counterclockwise position the secondary of T1 is shorted, and the stage functions as an emitter follower with some high-frequency rolloff because of C1. At the opposite end of the control the maximum ringing signal is added from T1, giving maximum high-frequency boost. The slug of T1 adjusts the frequency of maximum correction. Waveform D shows some overcorrection of the signal. This setting may be used to give an added crispness to the picture or to compensate for deficient monitor resolution.

3-10.16.8 Blanking-Insertion Circuits

The next significant step in the processing of the video signal is the addition of blanking pulses. This always follows high peaking and aperture

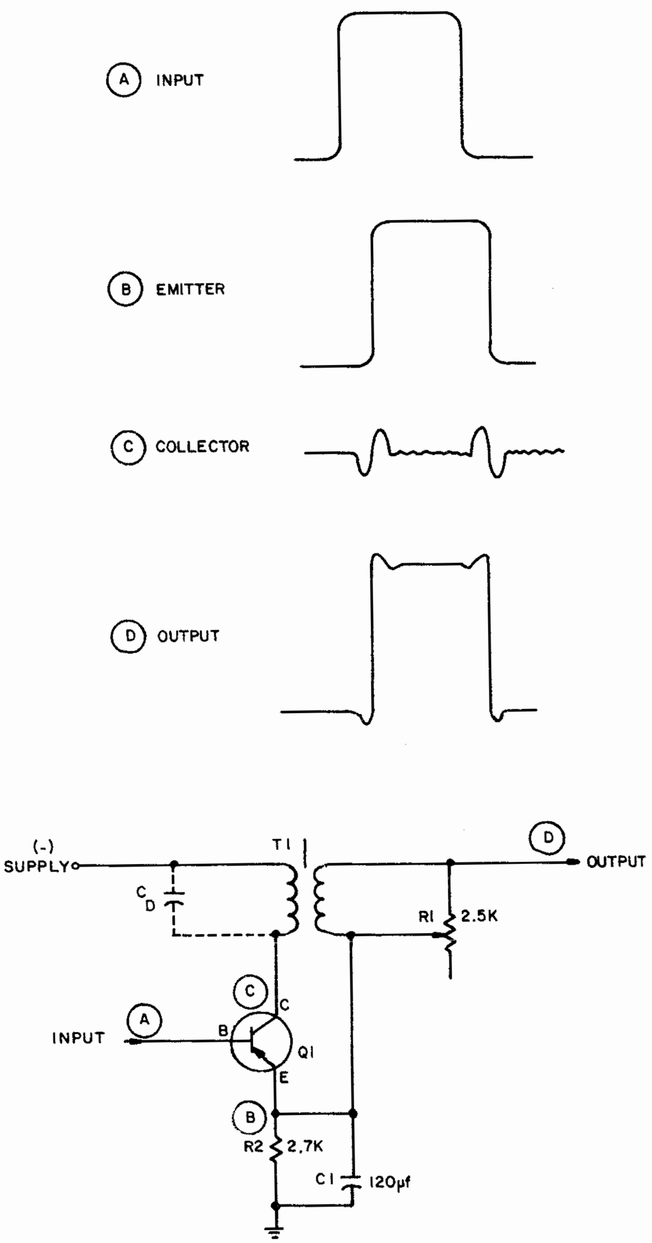


Figure 3-332. Aperture Correction Circuit with Illustrative Waveforms

correction, since only the video signal needs these changes. The addition of the blanking signal is always accompanied by clamping. The need for clamping arises from the inability of a coupling capacitor to pass a DC signal and its difficulty in passing low-frequency AC signals. This causes the video signal to assume an average level centered on the bias voltage of each stage, as in Figure 3-333B. Note how a very close approximation of the original signal can be regained by clamping Figure 3-33C. In all but the simplest cameras a keyed clamper circuit will be used - clamping action in either the positive or negative direction will take place only when a keying pulse is present. In the more inexpensive cameras a simple DC resistor circuit is sometimes used at this point. In a vidicon camera the beam is cut off during retrace by the blanking signal. During this interval the beam current is zero, corresponding to the blackest possible picture elements. Due to capacitive coupling this interval may lose its uniform level, as shown in Figure 3-333B. The task of the clamper is to return all camera signals during the blanking interval to a uniform reference level. In the usual arrangement, the blanking signal is added to the video signal immediately after clamping. Under all conditions, the video signal will be DC-coupled between these two points. To eliminate the necessity of adjusting monitor brightness each time cameras are switched, it is necessary to adjust all cameras to the same blanking level, usually 5 to 10 percent. Since lighting conditions and camera tubes may vary considerably, the input signal to the blanking mixer may have a higher or lower percentage than this value. The blanking-level control will be capable of increasing or decreasing the blanking level present in the video signal fed to the circuit. Figure 3-334 is representative of the blanking circuits in vacuum-tube cameras. The clamping circuit consists of X2 and X7. The positive portion of the clamp pulse causes a positive charge (+2 volts) to build up at the junction of X2 and X6, holding both diodes at cutoff. Negative clamp pulses cause both diodes to conduct, bringing grid 1 of V2 to the voltage established by the blanking-level control each time there is a clamping pulse. The voltage at grid 1 is the bias voltage established by the clamping circuit pulse the instantaneous value of the video signal. In this circuit the horizontal-drive pulse is used for clamping. The composite-blanking signal is fed to grid 3 of V2. The mixed signal at the plate of V2 is now black-positive, with the blanking portion excess of what is actually required. Control R11 determines the bias at grid 1, which sets the plate voltage of V2. The blanking-level control thus determines the level at which series diode X4 will clip the excess blanking signal. Diode X3 clips waveform transients introduced in the blanking-clipping process. Since the video waveform at the plate of V2 is normally only about 0.7 volt peak-to-peak, even a very small variation in the plate or screen voltage of V2 will

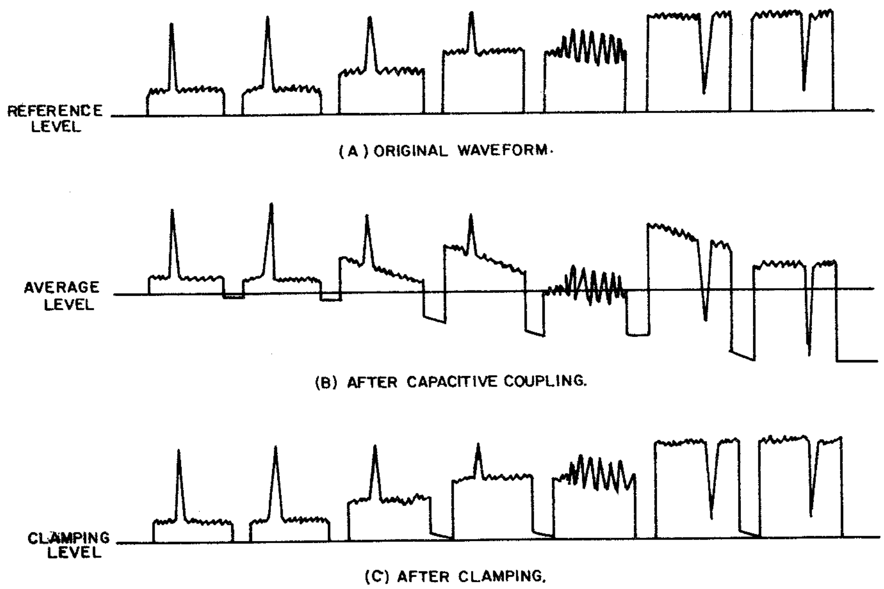


Figure 3-333. Clamping Waveforms

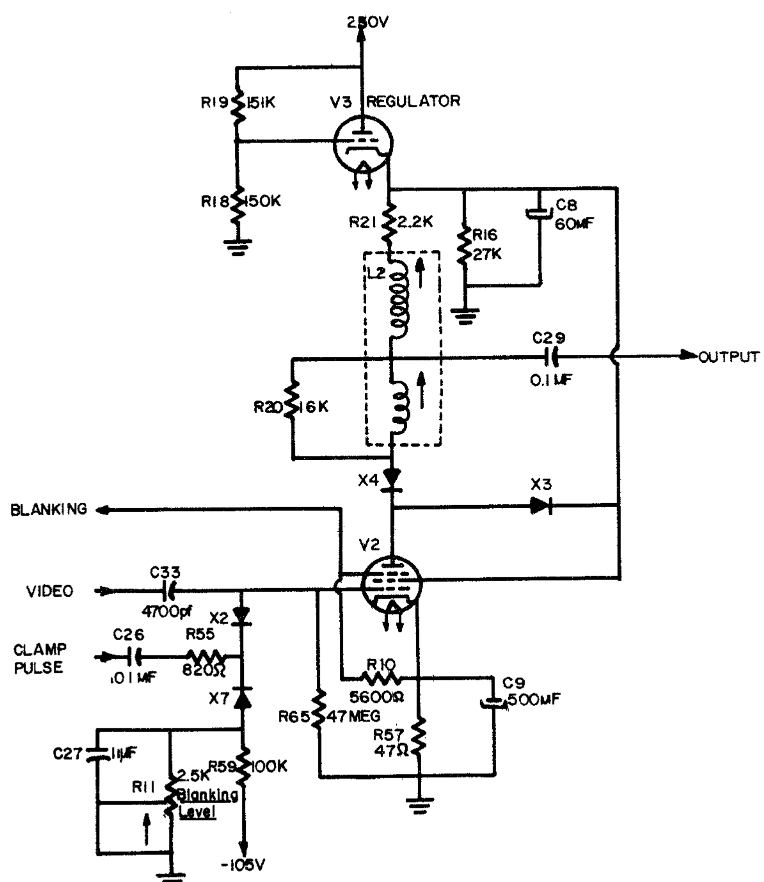


Figure 3-334. Vacuum Tube Blanking Insertion Circuit

cause a very noticeable change in the level at which X4 will clip. For this reason, series regulator V3 is included in this circuit. The circuit of Figure 3-335 takes an entirely different approach to blanking insertion. Clamping is accomplished through transistor Q6 rather than through diodes. This transistor is normally cut off. The clamp pulse, coinciding with vidicon blanking pulses, causes Q6 to conduct, bringing the voltage at the base of Q7 to the value established by voltage divider R26 and R27. Capacitor C16 then charges to this value, and has sufficient capacitance to retain this charge until the next clamp pulse arrives. All bias current through Q7 flows through Q6. Capacitor C19 maintains a constant voltage at the collector of Q6. Blanking is inserted at the emitter of Q7. Since blanking amplifier Q8 is normally cut off, a blanking signal at the base of Q8 causes it to conduct. This shifts the voltage at the emitter of Q8 (and the emitter of Q7) to approximately the value established by voltage divider R28, R29, and R30. When Q8 returns to cutoff, the voltage across R33 returns to the value set by the clamper at the base of Q7. The difference in voltage across R33, when Q8 is conducting and when it is not conducting, is the blanking level. This is adjustable at R29. Figure 3-336 illustrates another approach to blanking insertion. Negative clamp pulses cause transistor Q8 to conduct. This sets the video signal during the camera-blanking interval at the DC level established by R31 (the blanking-level control). The clamped signal is direct coupled

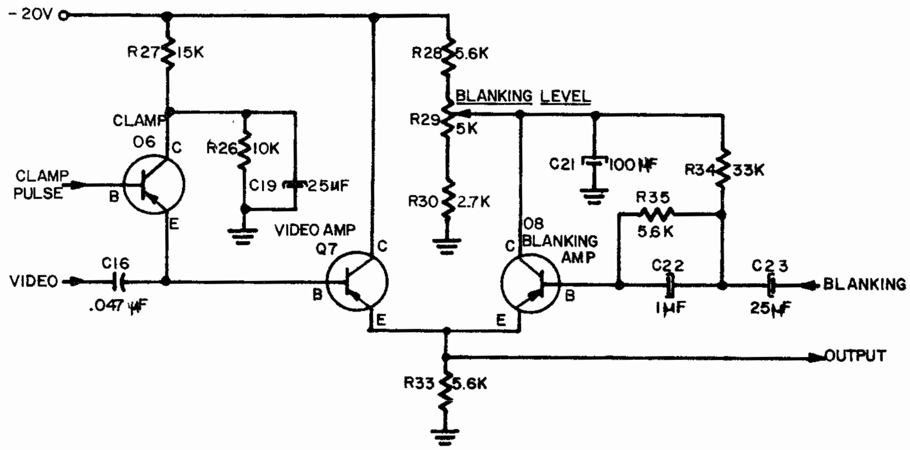


Figure 3-335. Transistorized Blanking Insertion Circuit

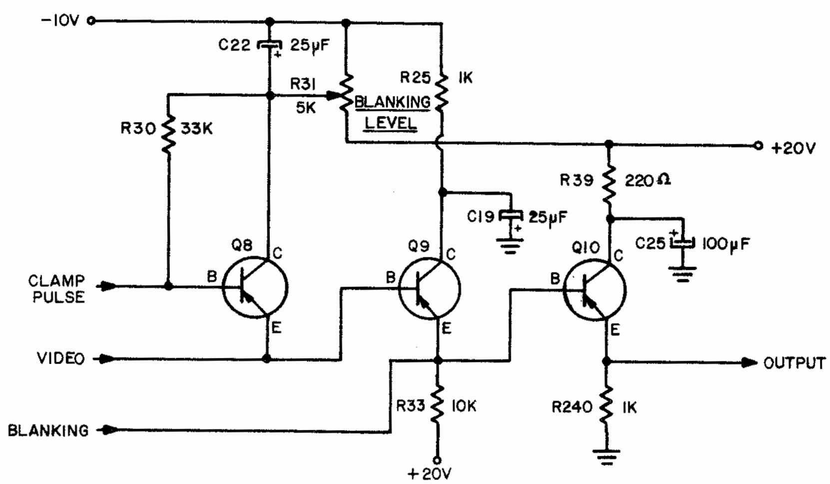


Figure 3-336. Blanking Insertion Circuit

through emitter follower Q9 to the base of Q10 where it is mixed with the blanking signal. Because the latter is of much greater amplitude than is required, the excess is clipped by limiting action in Q10. The bias level established at the base of Q9 by the blanking control determines the bias at the base of Q10 and, therefore, the level at which Q10 will clip. In the circuit of Figure 3-337 blanking and video signals are mixed prior to clamping. Under these conditions the blanking superimposed upon the video signal follows the line-to-line variations of the video signal. Blanking tips are then clamped to a reference level (ground in this case). In a subsequent stage (Q7) the blanking signal is clipped to the desired level. Video and blanking signals are mixed at the base of Q4, amplified, and coupled to the emitter follower Q6. Clamp pulses cause Q5 to conduct, holding the base of Q6 at ground potential at the start of each horizontal line. Blanking-level control at the emitter of Q7 adjusts the bias of Q7, establishing the operating point and the point at which positive blanking pulses will drive Q7 to saturation. Figure 3-338 shows the blanking-insertion circuit

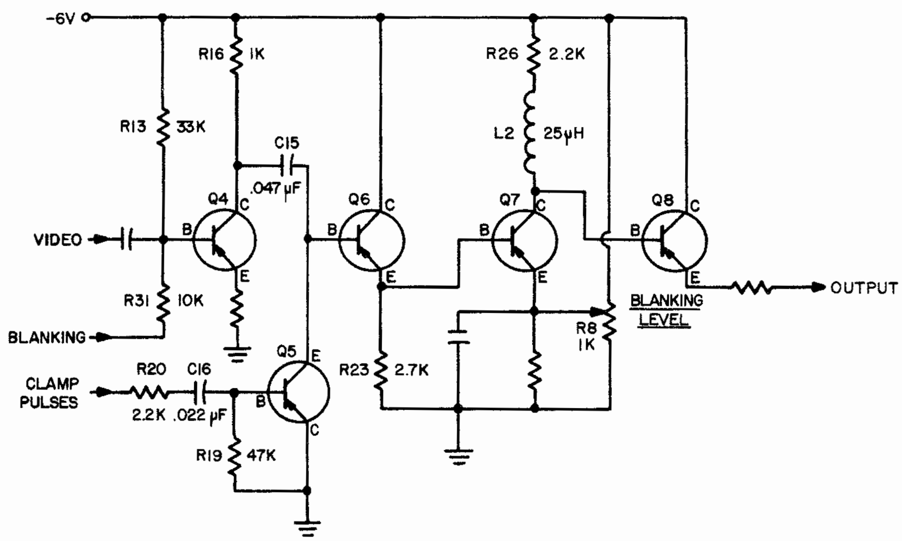


Figure 3-337. Blanking Insertion and Clamp Circuit (Simplified)

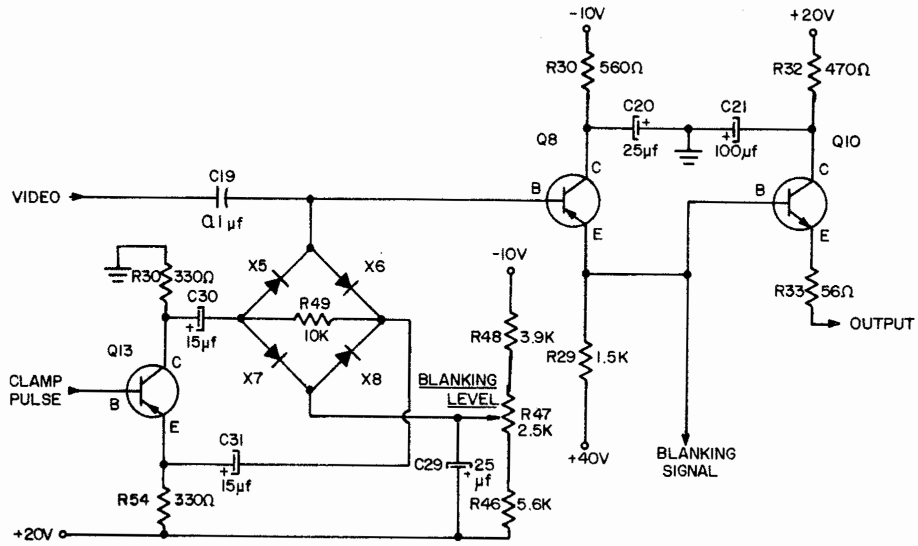


Figure 3-338 . Blanking Insertion Circuit of another popular transistorized camera. The clamper circuit, consisting of paraphase amplifier Q13 and diodes X5 through X8, clamps the signal at the base of Q8. The clamper diodes are normally cut off, but all diodes conduct when a clamp pulse is received. Thus the charge on the right side of C19 is adjusted to agree with the setting of R47. In this way the bias level of Q8 is corrected at the start of each line. Blanking and video signals are mixed at the base of Q10, and excess blanking is clipped through the limiting action of Q10. Control R47 sets the clamping level at the base of Q8. This, in turn, sets the bias level at the base of Q10 and establishes the level at which it will clip.

3-10.16.9 Gamma Correction

Gamma (γ) refers to the transfer characteristic of a device such as a kinescope tube, pickup tube, or amplifier. It tells whether the output of a device is uniformly proportional to its input, or if it has another relationship. The gamma figure for an amplifier is derived by the formula:

$$\gamma = \frac{d\,(\log E_o)}{d\,(\log E_i)}$$

where

E_i is the amplitude of the input signal

E_o is the amplitude of the output signal.

If the amplifier is perfectly linear, the value of gamma is one. If it has a gain characteristic like curve A in Figure 3-339, the gamma is greater than one. Curve C relates to an amplifier having a gamma of less than one. The waveforms in Figure 3-340 show the shape of a stairstep signal after passing through amplifiers having different gamma characteristics. The transfer characteristic of a camera tube can likewise be

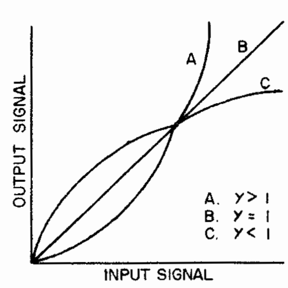


Figure 3-339. Curves Showing Various Values of Gamma

defined in terms of gamma. The formula is similar to that of an amplifier:

$$\gamma = \frac{d\,(\log E_o)}{d\,(\log Y_o)}$$

where

Y_o is the intensity of the light falling on the sensitized surface of the tube

E_o is the amplitude of the output signal.

If the camera has a gamma of one, it will produce a linear stairstep signal, Figure 3-340B, when viewing a linear reflectance chart. For other values of gamma the output will resemble Figure 3-340A or C. The typical vidicon tube has a gamma of about .65. Remember that the gray scales supplied with the standard EIA reflectance chart are logarithmic, but a linear scale is required to estimate the gamma of a tube. Camera-tube manufacturers usually supply curves with the tubes showing the transfer characteristic plotted logarithmically. Figure 3-341 shows typical curves for different values of gamma. If they are uniform throughout the operating range of the tube, the curve is a straight line. The value of gamma determines the steepness of the slope. Besides video amplifiers and camera tubes, a third component affects the overall gamma of a television system - the kinescope. A typical value here is 2.2. To calculate the gamma of an entire television system, multiply the gammas of each component of the system through which the signal passes. Thus, the value for the pickup tube is .65, the amplifier is 1, and the kinescope is 2.2, the gamma of the system is .65 X 1 X 2.2, or about 1.4. This figure is nearly ideal for normal viewing conditions, and no correction is ordinarily required. There are conditions under which a change in overall gamma is desirable: e.g., to compensate for external light falling on the viewing screen; to improve detail when the area of most interest is in a dark or light portion of the picture; or to compensate for the gamma of motion-picture film when a kinescope recording is being made. The most convenient place for gamma adjustment is in the video amplifier of the television camera. It may be accomplished by operating an amplifier stage over a nonlinear portion of its characteristic curve, or by using an impedance that varies with signal level for the load of an amplifier stage. The latter method is the more satisfactory of the two and is nearly universally used in modern equipment. The circuit ordinarily consists of biased diodes. Because the video

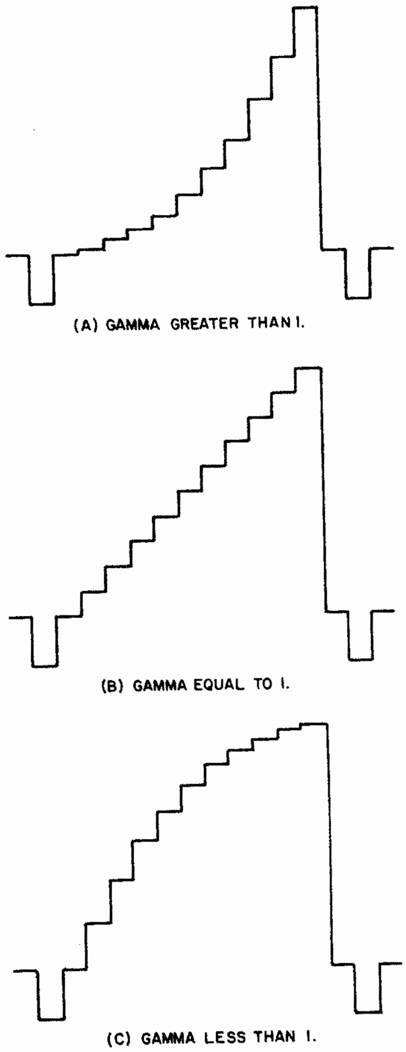


Figure 3-340. Effect of Gamma on a Linear Stairstep Signal

signal fed to the gamma-correction circuit must maintain its DC component, this circuit is normally associated with the blanking mixer circuit. This allows a single clamper to serve both functions. If the gamma-correction circuit is located elsewhere, an additional clamp is required. Figure 3-342 shows a biased-diode gamma-correction circuit. It forms a variable load for the blanking-clipper stage. With S1 in the unity position, the diodes are out of the circuit, and the stage functions linearly. When S1 is in the .7 position, a portion of the plate current flows through R1 and R2. As the signal goes negative (white), plate current also flows through the diodes and R5. (R7 and R5 determine

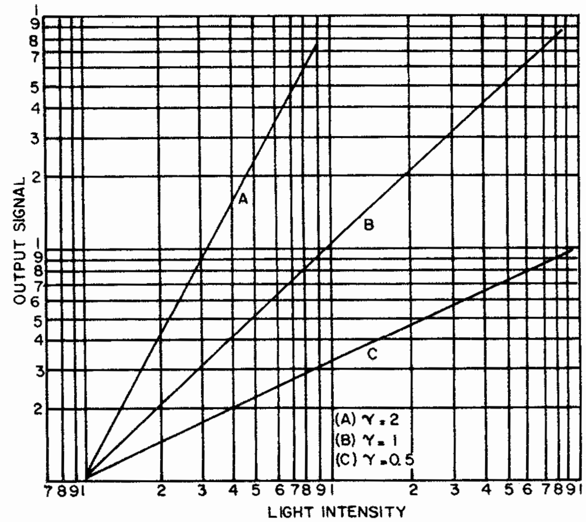


Figure 3-341. Gamma Curves on Logarithmic Plot

the bias level of the diodes and therefore the level at which they will conduct.) When the diodes are conducting, R5 is shunted across plate-load resistors R1 and R2, reducing the gain of the stage. When the signal voltage swings positive, the diodes are cut off, and all plate current must flow through R1 and R2. This increases the total value of the plate load and therefore increases the signal output. In the .5 position of the switch, the circuit functions as in the .7 position, but R1 is the only plate-load resistor when the diodes are cut off, and there is a smaller shunting resistor across R1 when the diodes are conducting. Consequently a greater amount of correction occurs in the .5 position than in the .7 position. Since the diodes are not perfect switches, the change from cutoff to conduction is not abrupt. This results in a smooth curve, resembling the desired gamma curve. Figure 3-343 shows the gamma-correction circuit of a transistorized camera. Transistor Q10 serves as a blanking clipper. The clamped incoming video signal retains its DC component throughout the circuitry. Diode X5 is biased by R48 and R49 to conduct at 0.7 volt. When the signal swings below this level (corresponding to the darker regions of the picture), the diode is cut off, and the signal is unaffected. When the signal swings positive, the diode conducts so that X5 and C28 are in parallel with the video output. The internal resistance of X5 is sufficiently higher, in parallel with the 75-ohm load resistance, that the signal is not completely shorted to ground, but is significantly

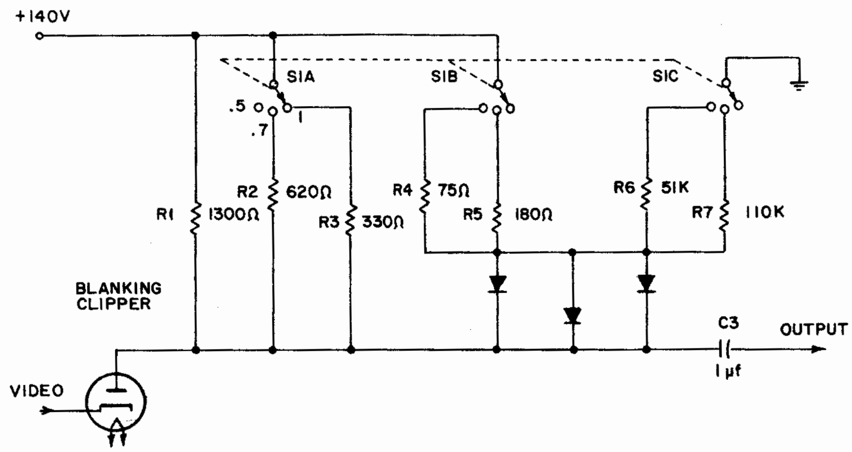


Figure 3-342. Biased Diode Gamma Correction Circuit

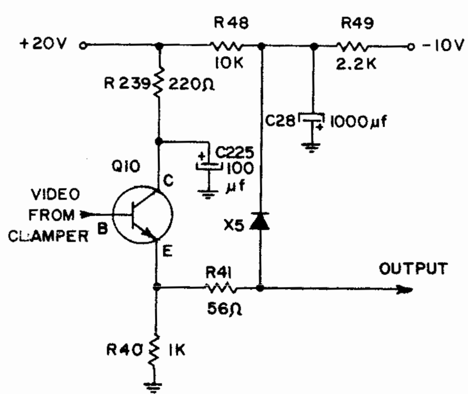


Figure 3-343. Gamma Correction Circuit

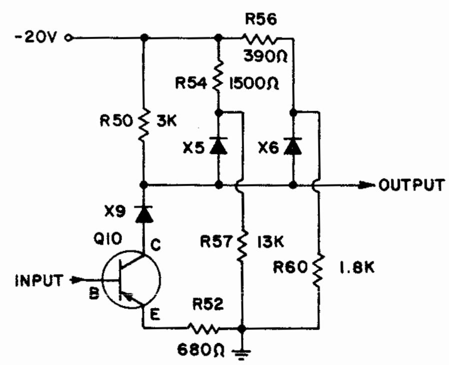


Figure 3-344. Gamma Correction Circuit Using Differently Biased Diodes

reduced in level. Thus, the gain of the circuit is decreased when the signal amplitude goes above 0.7 volt, compressing the white signal. This causes an overall decrease in gain, which may be reset to normal with the video-gain control. The result is black stretch and white compression. In the circuit of Figure 3-344 two diodes biased at different levels provide a smoother gamma-correction curve than would result with two diodes operating at the same level. On negative swings of the video signal the diodes are cut off, and R50 alone is the collector load of the blanking mixer (Q10). As the signal swings positive, X5 (and later X6) conducts, placing R54 and R56, in turn, in parallel with R50. The overall gain may be reset to normal with the video-gain control. The result again is black stretch and white compression (a decrease in gamma).

3-10.16.10 White-Peak Clippers

Many television cameras include a white-peak-clipper circuit that limits the maximum signal level to a predetermined value. It is useful where a predetermined maximum should not be exceeded, such as when feeding a transmitter or video tape recorder. Figure 3-345 shows the usual circuit of a white-peak limiter. Since the DC component of the video signal must be maintained if clipping is to be uniform, the circuit is direct coupled from the clamper circuit. The setting of potentiometer R37 establishes the bias level of X4, which conducts when the video signal at its cathode exceeds this level, thus loading the signal feeding Q10. The result is a clipping action. When X4 conducts, a relatively high impedance to ground at the higher video frequencies is maintained through reactor L1. This allows fine detail to remain

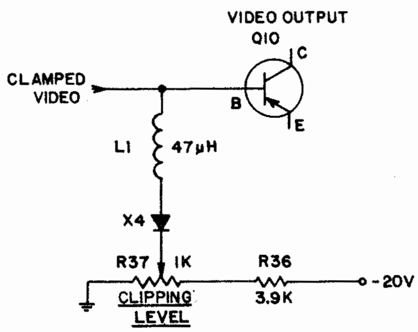


Figure 3-345. White Clipper Circuit

visible in areas when clipping is taking place. Figure 3-346 illustrates the effect when a portion of a stairstep signal is clipped.

3-10.16.11 Sync Addition

Where sync is to be added to the video signal within the camera-control unit, it must take place following the insertion of blanking. Most often it is added in the video-output stage. The usual circuit is a simple additive mixer. Figure 3-347 shows the circuit in a camera-control unit having two video outputs. Sync is added to the video signal at the video-output terminals through R35 and R36. Transistor Q14 is an isolation amplifier that prevents video from being fed out of the sync-input terminals. The high input impedance of this amplifier allows a loop-through connection at the sync-input terminals. Potentiometer R56 controls the sync amplitude by adjusting the gain of Q14. Transistor Q10 is controlling in the video-output stage.

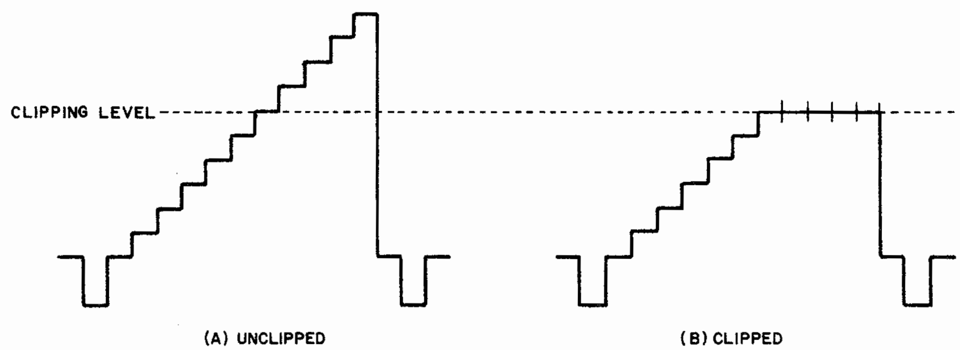


Figure 3-346. White Clipping of Linear Stairstep Waveform

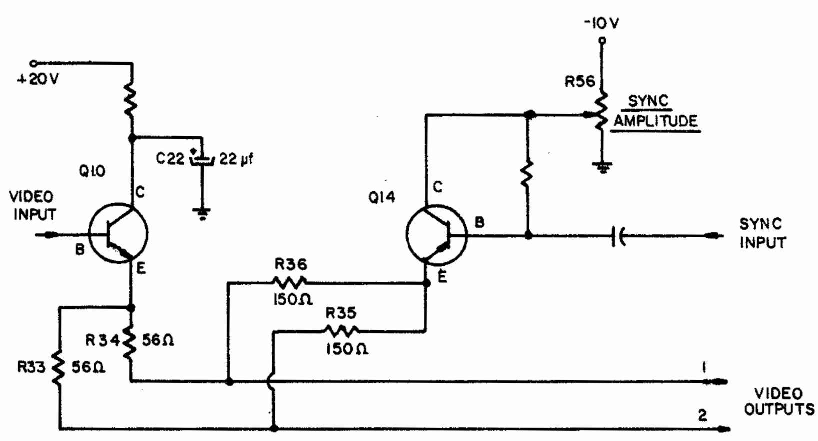


Figure 3-347. Sync-Adding Network

3-10.16.12 Automatic Sensitivity

The need for automatic-sensitivity circuitry in television cameras arises from the varying light conditions under which television cameras are often required to operate. A camera may be required to function unattended in bright sunlight, rain, and artificial light at night. An automatic-sensitivity device will provide a continuously usable signal without manual adjustments being made to the camera. Most widely used in this context is the automatic-target-voltage circuit. Some designers claim nearly uniform output for light-level variations as great as 10,000 to 1 with this circuit. The automatic-sensitivity control may also take the form of an automatic iris, a shutter mechanism using neutral-density filters, or a combination of either. Figure 3-348 shows a version of a popular automatic-target-voltage control. Video signals from the camera are amplified by Q11 and the output is rectified by voltage-double X2-X3. This produces a DC voltage directly proportional to the signal from the camera on the base of Q12. As the amplitude of the incoming video signal increases, the voltage at the base of Q12 becomes increasingly positive, causing the collector voltage to decrease. The latter is the voltage that is fed to the target of the vidicon tube. Consequently, as the signal output of the vidicon increases, the automatic-sensitivity circuit causes its target voltage to decrease. In this way the signal output of the vidicon tube remains relatively constant. Zener diode X4 limits the collector voltage of Q12 to 82 volts, providing protection for both Q12 and the vidicon target. The voltage doubler is biased through potentiometer R42 to provide a means of adjusting the video level the circuit is to maintain. In the circuit of Figure 3-349, improved low-frequency stability and fast response are obtained by operating the vidicon target at a fixed DC potential (ground); controlled voltages are applied to the remaining elements of the tube. In this way the automatic-target-voltage control is isolated from the video preamplifier. The video signal is fed to the three-stage video amplifier, then rectified by X2 and X3 to form a DC signal at the grid of V3B. This signal is proportional to the strength of the video signal. A strong signal causes the DC voltage at the grid of V3B to go more negative, causing the plate voltage to increase with respect to ground. Since the plate of V3B is connected to the junction of R20 and R21 of the floating power supply, this increase causes all voltages to become more positive with respect to ground. The relative target voltage decreases and the video-signal level is also reduced. A very simple but effective automatic-target-voltage circuit is shown in Figure 3-350. This circuit utilizes the increase in voltage-drop across a large series resistor as vidicon current increases with light level. When the amount of light falling on the vidicon target increases, the target current also increases. This current flows through large resistor R1 and causes the voltage at the vidicon target to decrease, thereby decreasing the signal current. The large resistor, therefore, exerts a stabilizing effect on the signal current. This circuit may be used as an adjunct to an amplifier-rectifier type of automatic-target-voltage circuit. The combination (shown in Figure 3-321) increases the range of control beyond that of either individual circuit.

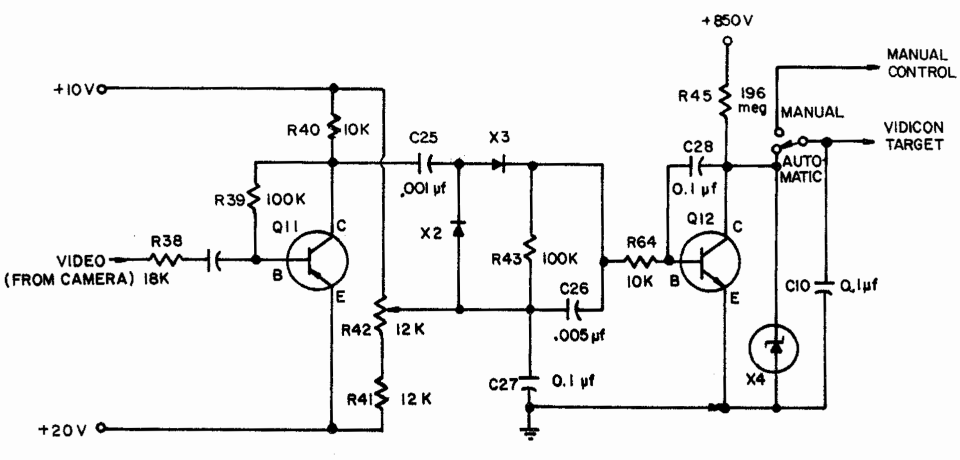


Figure 3-348. Automatic Target Voltage Circuit

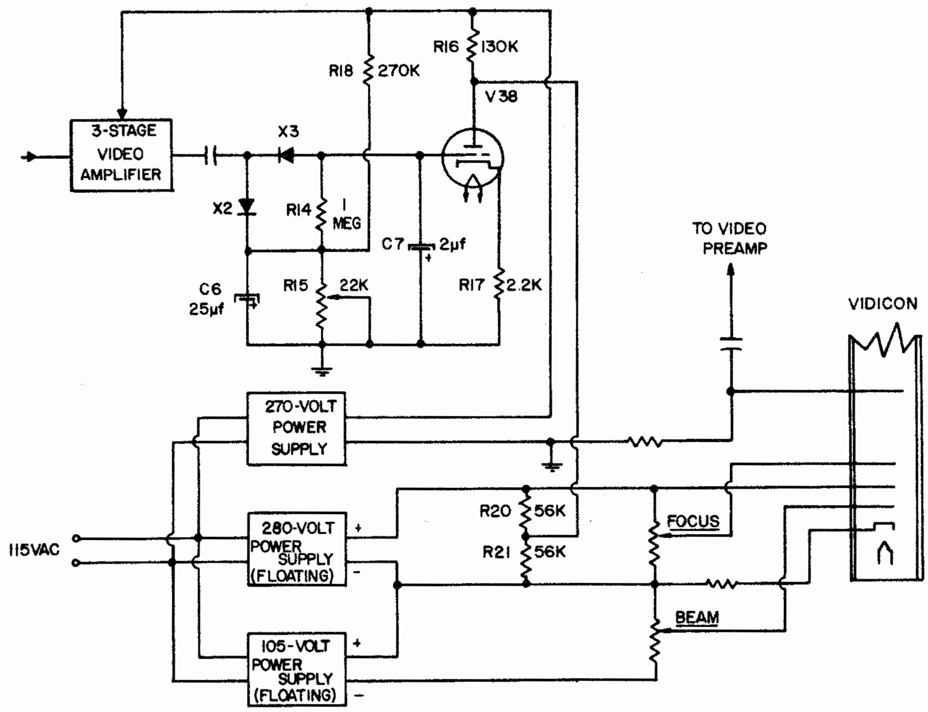


Figure 3-349. Automatic Sensitivity Control (Simplified)

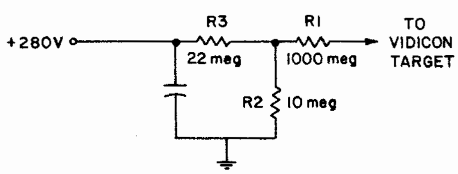


Figure 3-350. Automatic Sensitivity Circuit

3-10.16.13 Camera Blanking

As the vidicon beam completes each scan of the target and returns to scan the next line, it needs to be cut off to prevent premature discharge of the vidicon target. The same is true as it returns to the top of the raster at the end of each field. Vidicon (camera) blanking performs this function. Many cameras utilize the horizontal- and vertical-drive signals for camera blanking. This is especially true of those using EIA synchronization signals. Cameras using industrial types of sync and those designed to use more than one type must have a different means of vidicon blanking, since an EIA type of drive signal may not be present. The signal that is added to the video blanking the kinescope is not suitable for blanking the camera tube, because the camera-blanking interval should be narrower than the system-blanking interval. Horizontal- and vertical-blanking signals for the camera tube are therefore often derived from the deflection signals, usually with some reshaping. Figure 3-351 illustrates this condition

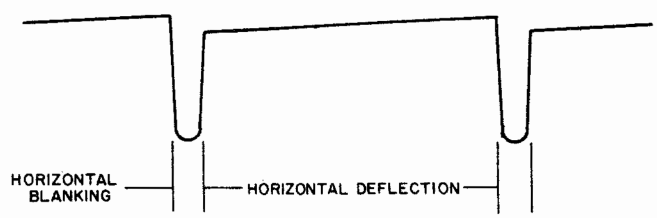


Figure 3-351. Vidicon Blanking and Deflection Signal

for a horizontal-deflection pulse. When flyback pulses are used for vidicon blanking, horizontal signals must be kept isolated from the vertical to prevent mixing of the deflection signals. This may be done by using separate isolating amplifiers before mixing the signals, by using steering diodes, or by applying one signal to the cathode and the other to the control grid of the vidicon. (Either positive pulses at the cathode or negative pulses at the control grid may be used to blank the vidicon.) In the circuit of Figure 3-352 standard EIA horizontal- and vertical-drive pulses are used to blank the vidicon. These are combined through diodes X1 and X2, amplified and inverted by Q3, and then applied to the cathode of the vidicon through C9. The pulses at the cathode are positive and have an amplitude of about 11 volts. The circuit shown in Figure 3-353 uses the flyback pulses of the horizontal-and vertical-deflection signals to blank the vidicon. Sweep-failure-protection circuitry is combined with camera blanking. The negative portion of the horizontal-deflection signal is clipped by X14, amplified by buffer stage Q14, and then fed to the base of Q15.

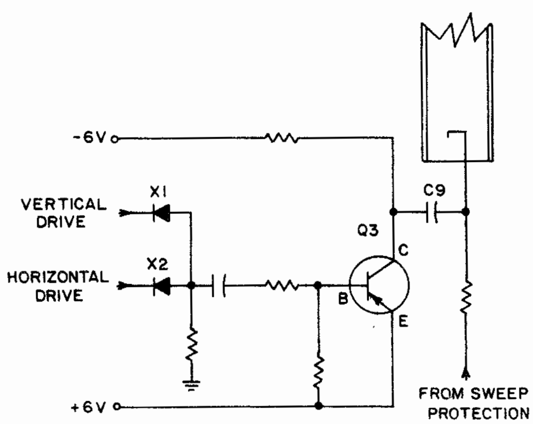


Figure 3-352. Vidicon Blanking Circuit

The positive portion of the vertical-deflection signal is clipped by Q18 and fed to the base of Q16. In the absence of signals, Q15 and Q16 are biased by R47 and R49 to conduct heavily. In this stage the collectors (and the vidicon cathode) are held at +20 volts, which is sufficient to cut the tube conduction off. When strong positive signals are received at the bases of Q15 and Q16, the transistors are cut off, causing the collectors (and the vidicon cathode) to go to -20 volts, thereby turning the beam on. Should either input fail, Q15 and Q16 (as appropriate) will conduct heavily, cutting off the vidicon. This circuit therefore serves as both a blanking amplifier and as a vidicon-protection circuit. In the circuit of Figure 3-354, the horizontal- and vertical-blanking pulses are not mixed, but are applied to separate electrodes of the vidicon tube. Negative vertical-blanking pulses are fed to the cathode. The vertical sawtooth triggers the blanking multivibrator so that it produces positive pulses of the desired width. The vertical-blanking-amplifier tube (V3A) provides the low-impedance output necessary to transmit the signal over the camera cable. The DC voltage from the beam control is superimposed on the pulses, and the resultant is fed to the vidicon control grid. The flyback pulse from the horizontal-deflection amplifier is utilized directly for vidicon horizontal blanking.

3-10.16.14 Deflection Protection

If a television camera should be energized before the sync generator, the beam of the pickup tube would in most cases not be deflected. It would be concentrated in a small spot of the target and would rapidly destroy the phosphor in this sopt. The resultant permanent burn would render the tube useless for many

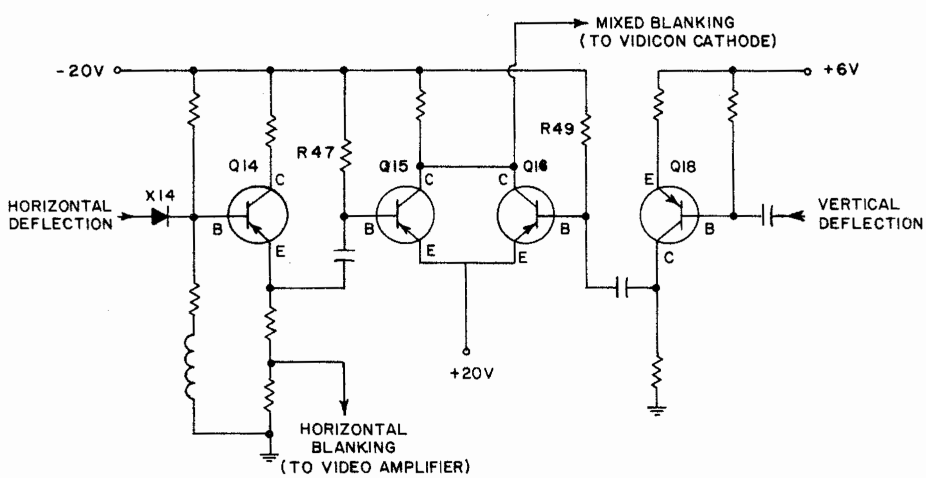


Figure 3-353. Vidicon Blanking and Sweep Failure Protection Circuits

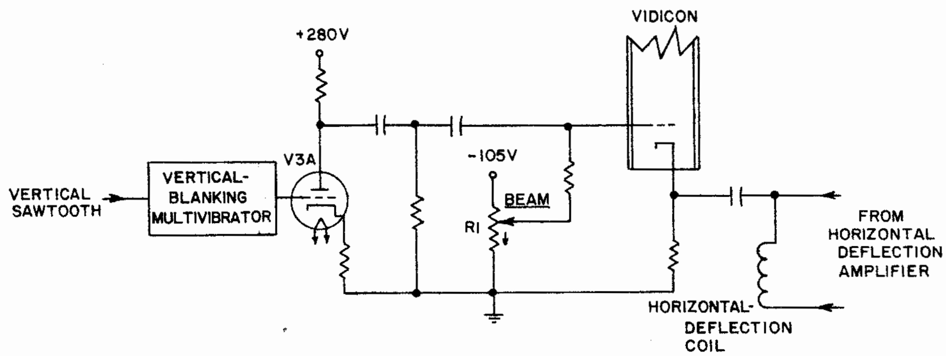


Figure 3-354. Camera Blanking Circuit

purposes; in any event, the deterioration would be undesirable. Similarly, the pickup tube would be damaged should the sync generator fail or be turned off during operation of the camera. If horizontal scan of the camera should fail or the horizontal-drive pulses be removed, only vertical scan would be present. The result would be a vertical line burned into the target surface of the tube. Similarly, removal of the vertical-drive pulse or failure of the vertical-scan circuitry would cause a horizontal line to be burned into the target surface. Since vidicon tubes are expensive and difficult to install correctly, the use of a deflection-protection circuit to prevent these burn-out effects is desirable. This is accomplished through cutting off the beam of the camera tube by removing target voltage, removing grid-2 voltage, biasing grid 1 highly negative, or biasing the cathode highly positive. Many modern cameras incorporate deflection protection with the camera-blanking circuit. Figure 3-353 shows a circuit that is representative of these provisions. Figure 3-355 shows a deflection-protection circuit which is separate from the camera-blanking circuitry. Resistor R19 is the vidicon cathode resistor. Rectifier X4 and filter R14-C14 produce a forward bias on Q5 from a sample of the horizontal-deflection waveform. Likewise, the combination of X5, X6, C13, and R15 produces a forward bias on Q6 from a sample of the vertical-deflection waveform. When both deflection waveforms are present, the transistors conduct, the collector of Q5 goes to near ground potential, and the vidicon can conduct. Should either deflection waveform be lost, one of the transistors is cut off, and neither can then conduct. The collector voltage of Q5 therefore becomes the value established by voltage divider R18-R20 - about 25 volts. This cuts off the vidicon and prevents screen damage. In this circuit, blanking is also fed to the vidicon cathode

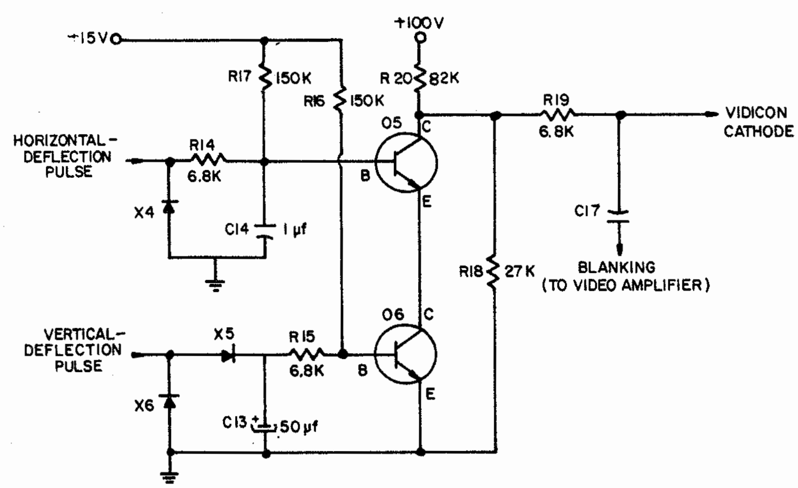


Figure 3-355. Vidicon Protection Circuit

through C17, thus has no effect on the operation of the protection circuit.

3-10.16.15 Cable-Delay Compensation

A video or drive signal does not pass instantly through a coaxial cable - the transit time varies with cable construction. For a polyethylene cable such as RG-59/U, it is about 1 microsecond for 700 feet of cable. When a camera-control unit is separated from a camera by a cable of significant length, the video signal arriving at the camera-control unit is delayed by the transit time of the drive signals to the camera plus the time for the video signal to return. This means that the timing relationship between the video signal in the processing amplifier and the pulses used in processing the signal has changed. For example, if there are 700 feet of camera cable, the video signal reaching the control unit is delayed by the one microsecond required for the drive signals to travel from the control unit to the camera plus the one microsecond required for the video signal to travel from the camera to the control unit. However, the clamp, blanking, and sync signals used to process the video signal in the control unit do not travel this round trip. They are therefore two microseconds ahead of the video signal when it arrives at the control unit. Figure 3-356 demonstrates that with a vidicon-blanking signal duration of 8 microseconds and with a system-blanking duration of 12 microseconds, up to four microseconds delay in the signal from the camera can be completely hidden by system blanking. A delay greater than four

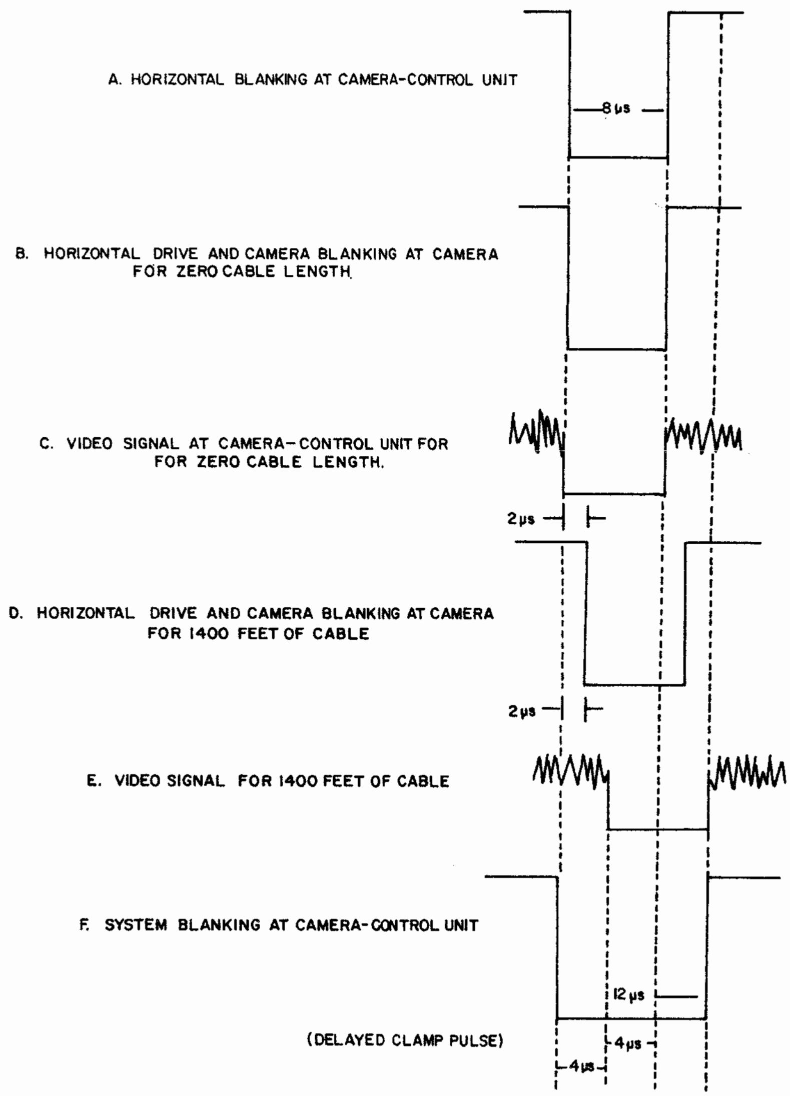


Figure 3-356. Waveforms Resulting From Transit Delays in Cables

microseconds will cause the video signal to be shifted to the right in the screen of the monitor since the camera-blanking signal will no longer be completely covered by the system-blanking signal. A black bar will appear at the left side of the picture. In order for the clamp circuit to function, the clamp pulse must coincide with the camera-blanking signal. Under the conditions of Figure 3-356D and E it is apparent that the clamp pulse must be delayed by four microseconds. If the width of the clamp pulse is reduced to four microseconds or less, it will fall within the camera-blanking interval for any camera cable up to 1400 feet (which produces the four-microsecond delay). By using a fixed delay, satisfactory operation can be achieved with no adjustment required for any length of cable up to 1400 feet. For greater lengths of cable the delay must be increased while the clamp and camera-blanking pulses are decreased. However, when the camera-blanking pulse is reduced to six microseconds for a delay of six microseconds (corresponding to a camera-cable length of about 2000 feet), the area of overlap between the undelayed signal and the six-microsecond delayed signal decreases to zero. No interval is left during which the clamp pulse may be applied. The maximum length of cable can be extended beyond that of the foregoing example by using a variable delay in the clamp-pulse circuit and decreasing the camera-blanking interval. The practical minimum is about four microseconds, which allows for eight microseconds of delay with a 12-microsecond system-blanking interval. This method of delay compensation was favored in cameras manufactured before about 1962. Most cameras of more recent design allow blanking, clamp, and sync pulses to go undelayed. Instead, horizontal-drive pulses feeding the camera deflection circuit are delayed by a full line, less the amount of delay required. In effect, the horizontal-drive pulse to the camera leads the other pulses by the amount of time required for the round trip of the signals through the camera cable. This form of delay is ordinarily accomplished by an adjustable multivibrator, and can compensate for a practically unlimited length of camera cable. An additional advantage of this circuit is that all pulses maintain their proper phase relationships, and no marginal timings of pulses need be involved, regardless of camera-cable length. Compensation of vertical-drive pulse delay is not required ordinarily because of the relative insignificance of a few microseconds delay as compared to the approximately 17-millisecond field time. Figure 3-357 shows the circuit of a camera-control unit in which only the clamp pulses are delayed. Here a delay of about two microseconds, corresponding to about 700 feet of camera cable, is accomplished in an artificial delay line. Performance is satisfactory with as much as 1400 feet of cable, without adjustment. Transistor Q1 is an emitter follower for isolation from other circuitry; it provides impedance match for the delay line consisting of C1, L1, C2, L2, C3, L3, and C4. The pulses leaving the delay line are sharpened by the differentiating network consisting of C5 and R7. Monostable multivibrator Q2-Q3 generates a clamp pulse of about four microseconds. Figure 3-358 shows the circuit of a camera-control unit that delays the horizontal-drive pulse by nearly a full line. Transistor Q1 isolates the delay circuitry from other sections of the camera-control unit that use the horizontal-drive pulses. Monostable multivibrator Q2-Q3 has an unstable state with a period that corresponds to the amount of delay required. The period of its stable state is the length of time by which the horizontal-drive pulses will lead the other pulses. Differentiation circuit C3 and R4 sharpens the pulses from Q1, and diode X1 clips negative-going overshoots from the differentiated driving waveform. Both Q2 and Q3 are normally conducting. The positive-going pulse from X1 cuts off Q2, causing the collector to swing negative. This negative-going output pulse is coupled to the base of Q3, cutting it off and

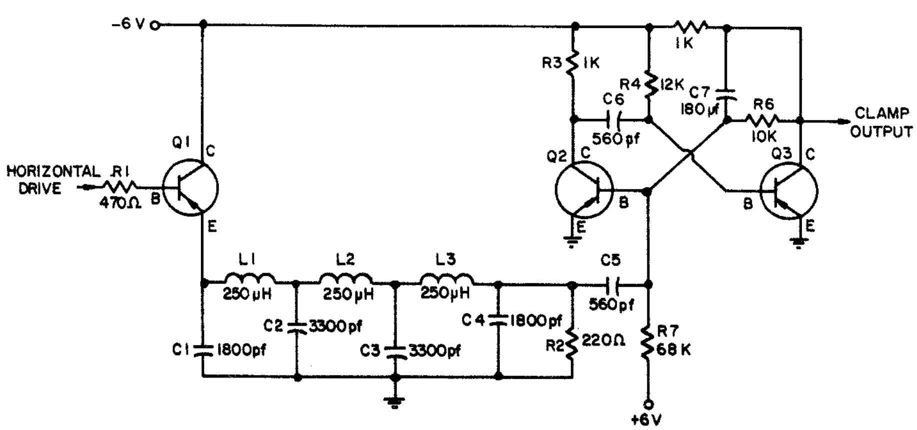


Figure 3-357. Clamp Pulse Delay Circuit

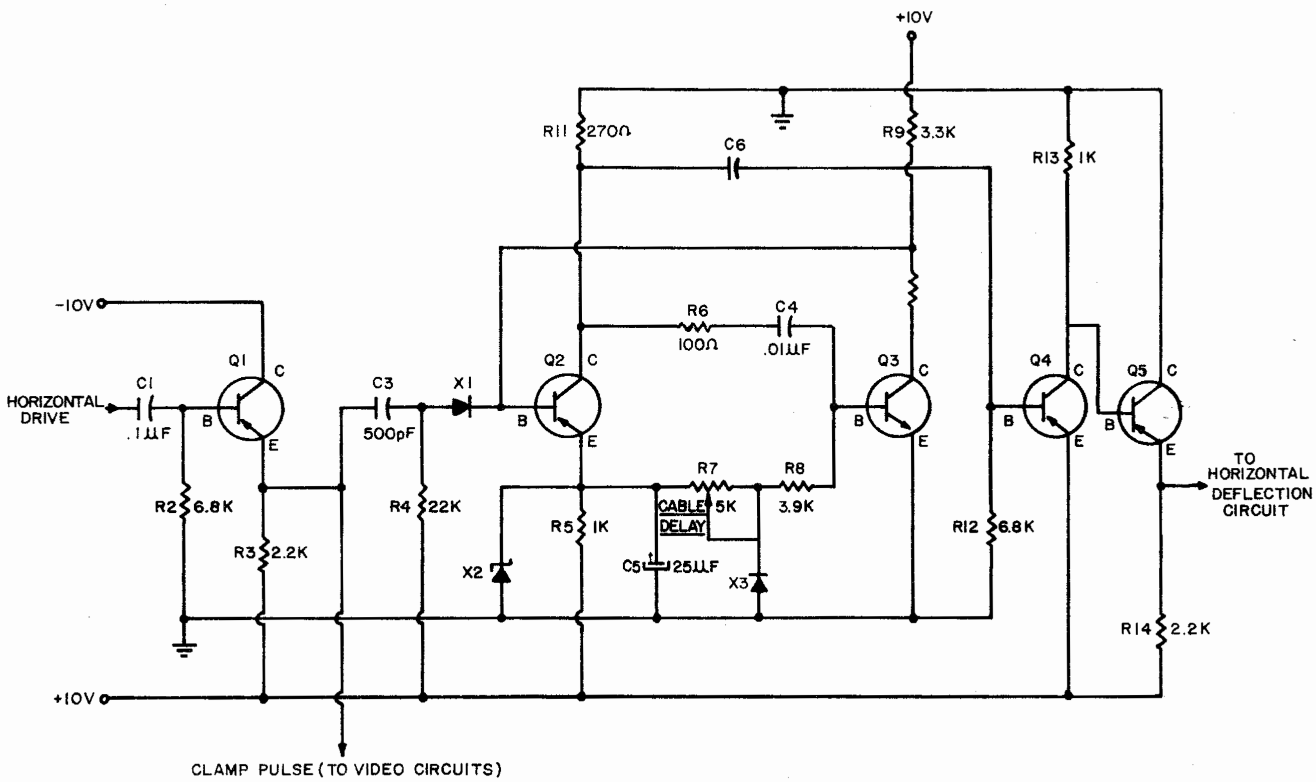


Figure 3-358. Cable Delay Compensation Circuit

causing its collector to swing positive. The positive pulse at the collector of Q3 is coupled to the base of Q2, completing the multivibrator loop. The circuit will maintain this unstable state until the added charge on C4 drains sufficiently to bring Q3 back into conduction. This time is adjustable by R7. Diode X3 allows C4 to charge very rapidly, but it forces the capacitor to discharge through R5 and R7. Zener diode X2 maintains a constant voltage at the emitter of Q2 and stabilizes the bias current of Q3. The rectangular waveshape at the collector of Q2 is adjusted to the amount of delay required. This signal is coupled by the differentiating circuit, consisting of C6 and R12, to the base of Q4, which is normally saturated. The negative-going pulse, corresponding to the leading edge of the waveform at the collector of Q2, has no effect on Q4. The positive-going pulse, corresponding to the trailing edge of the waveform at the collector of Q2, cuts off Q4, producing negative pulses at its collector. This waveform is further amplified by emitter-follower Q5 and is fed to the horizontal-deflection circuit of the camera.

3-10.16.16 Vidicon Filament Compensation

Where the camera is separated from its control unit by a relatively long length of cable, the I^2R loss in the camera cable becomes significant with regard to the vidicon filament. The lack of proper voltage at the filament pins of the vidicon tube may be compensated for by a variety of methods. A very simple method is to feed a high voltage (say 115 volts) through the camera cable, thus lowering the current and therefore the I^2R loss. A filament transformer is located within the camera, and the filament voltage does not vary significantly over a wide range of cable lengths. Where economics or space limitations do not allow for a filament transformer to be installed in the camera, other methods are used. The filament transformer may be located in the camera cable to produce the required 6.3 volts at the vidicon filament. In this case, a variable resistor is inserted in series with the cable to provide for adjustment to suit cable length. A disadvantage of this system is that the variable resistor must be readjusted whenever the length of camera cable is changed. This method is illustrated in Figure 3-359. By Ohm's law, if the voltage across the filament is correct, the current through the filament will be the same regardless of cable length. Therefore, when the vidicon filament is correct, the voltage drop across R2 in the camera-control unit will always be the same. If the vidicon current is 600 mA, the voltage across R2 will always be 0.6 volt. Terminals TP-1 and TP-2 and resistor R1 may be located in the camera-control unit to facilitate adjustments. Other tube filaments may be in parallel with the vidicon filaments, in which case the proper voltage drop across R2 will need to be recalculated. A more advanced

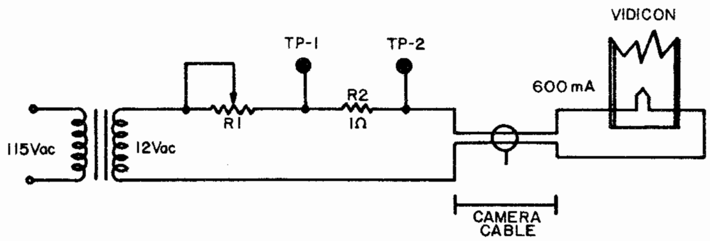


Figure 3-359. Camera Cable Compensation for Vidicon Filament

method of compensation provides a current regulator in place of R1. Compensation will then be automatic. In this case, the filaments will use direct current. Figure 3-360 is the schematic of such a current regulator.

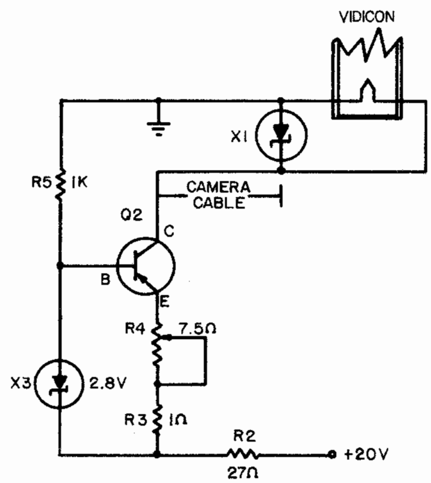


Figure 3-360. Filament Current Regulator Circuit

As resistance of the camera cable increases due to length, the current through Q2 tends to decrease and the voltage drop across R3 — R4 decreases. Zener diode X3 maintains a constant voltage between R2 and the base of Q2. The increased voltage drop across R3 and R4 increases the forward-bias voltage of Q2 and increases the current through Q2 and the camera cable. Potentiometer R4 sets the initial bias level of Q2 and provides the calibration adjustment for filament voltage and current. Since the circuit is a current regulator, care must be taken that the value of current selected does not exceed the rating of the vidicon tube in the camera. Vidicon tubes are available with recommended current ratings ranging from 95 mA up to 600 mA. If a 95-mA tube is placed in a circuit adjusted for operation at 600 mA. If a 95-mA tube is placed in a circuit adjusted for operation at 600 mA, the full 20 volts of the power supply will appear across the vidicon filament and destroy it. As a precautionary measure a Zener diode, X1, having a breakdown voltage slightly higher than the rated voltage of the vidicon filament is often placed in parallel with it.

3-10.16.17 Dynamic Focus

The vidicon beam must travel a slightly greater distance to reach the corners of the scanned area than to reach the center. For this reason the optimum focus setting for best corner resolution will be slightly different than for best center resolution. Optimum focus in both center and corners is attainable through application of appropriate parabolic waveforms is synchronism with the scan to the vidicon focus electrode. Separate horizontal and vertical waveforms are necessary and are generated separately. This type of compensation is complex and is found only in the most expensive, high-resolution cameras. Figure 3-361 shows the dynamic-focus circuit of such a camera. The horizontal parabolic waveform is formed by Q16 and Q17. The vertical waveform is formed by Q18 and Q19. These are mixed at C32 and amplified by Q20 and Q1. Q20 and Q1 are conventional linear-amplifier circuits. Neon bulb M1 fires at about 70 volts, protecting Q1 during the charge time of C9 at turn-on. After warm-up the bulb has no function. The combined parabolic waveforms are added to the DC focus voltage at grid 3 of the vidicon tube.

3-10.17 LIGHTING

A television camera cannot function without illumination on the televised scene - the quality of the picture produced depends to a large extent on how the scene is illuminated. There is a minimum amount of light that must be present for optimum picture quality. The amount of light reaching the sensitized surface of the camera depends on the amount of light falling upon the scene, the lens setting, and the amount of light that the scene reflects. In many closed-circuit installations the camera is required to operate in sunlight. There is little or no control over the amount of light striking the scene in this case. Fortunately, reasonably good pictures can be obtained in sunlight. In another case pictures may be required around the clock in an outdoor area. When night falls, artificial

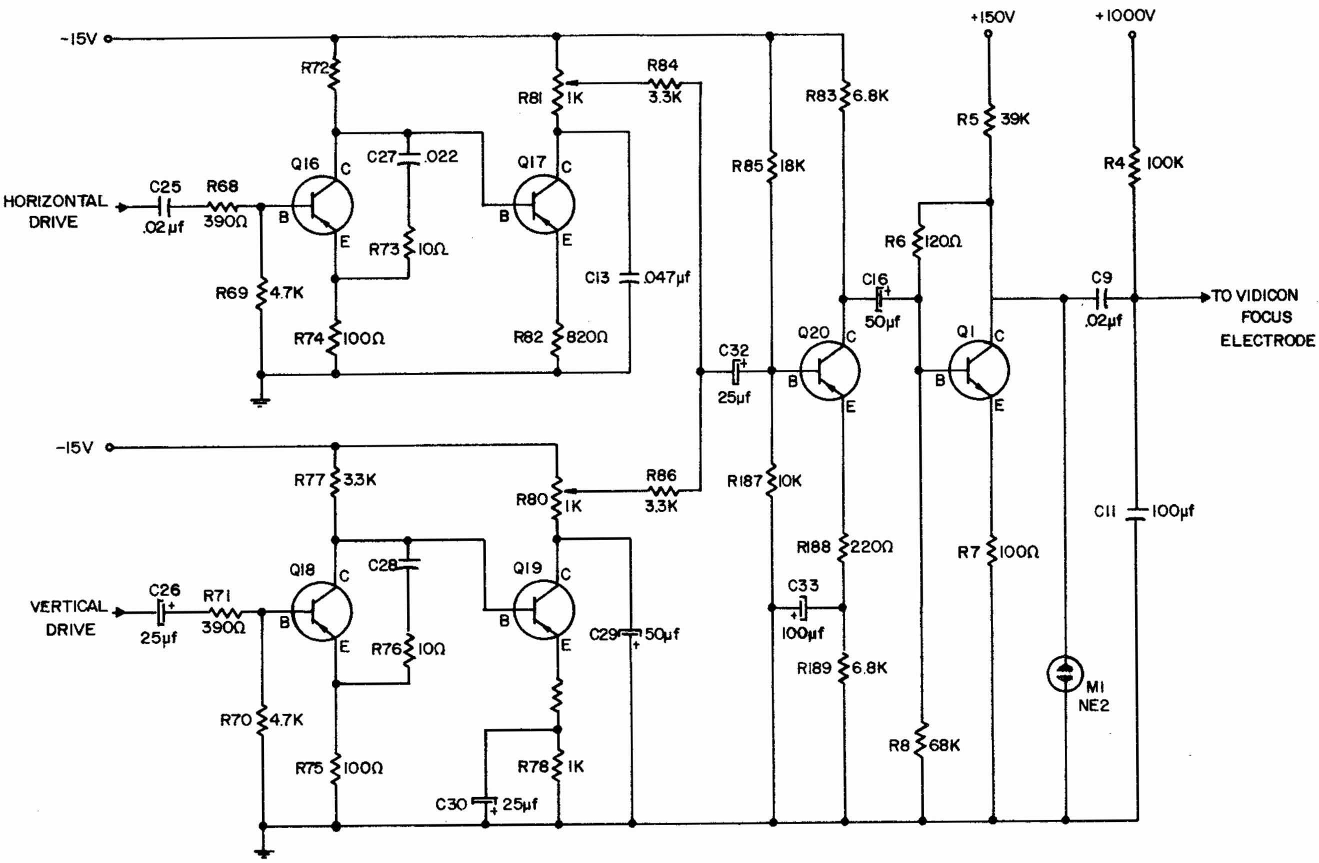


Figure 3-361. Dynamic Focus Circuit

lighting will be needed. In rare cases there may be sufficient light installed for other purposes, but generally special illumination must be present. Where there is a choice, consistently good pictures can be obtained with the vidicon cameras if lighting is flat or diffuse, as with a lightly overcast sky. But in many cases the scene cannot be lighted to best advantage. For example, a camera on a pan-and-tilt unit may be called upon to view several areas, each with different lighting.

3-10.17.1 Principles of Light

The Illumination Engineering Society defines light as visually-evaluated radiant energy. The average human eye is sensitive to radiant energy which lies between wavelengths of 380 and 760 nanometer (25,400,000 nanometer = 1 inch). Figure 3-362 shows the visible spectrum in relation to visible portions. This definition excludes the infrared and ultraviolet portions of the spectrum which are also loosely called

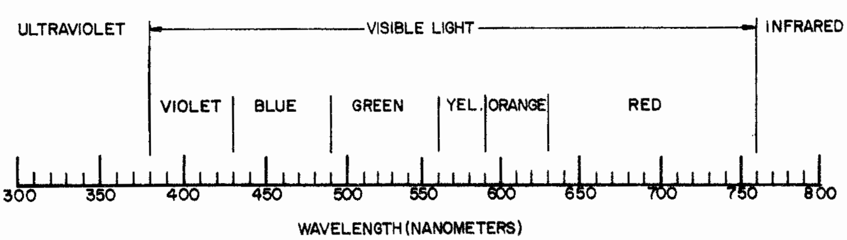


Figure 3-362. Visible Portion of Electromagnetic Spectrum

light. In many instances these wavelengths are of interest in television work. The terms which express light quantities are based on their relationship to a one candlepower source, or a source having a luminous intensity of one candela. The source is imagined to be centered in a hollow sphere which has a radius of one foot (Figure 3-363) to radiate equally in all directions,

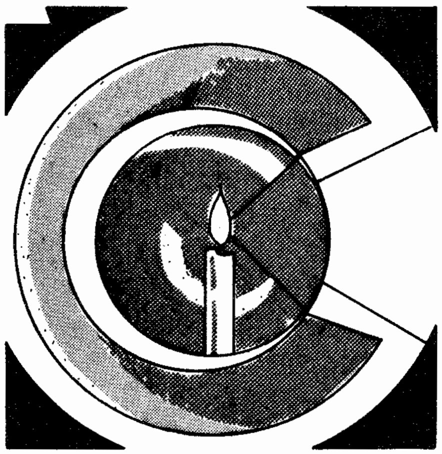


Figure 3-363. Concept of Light Quantities

and to be small enough that it can be considered a "point source." The surface of the sphere has an area of $4\pi r^2$. Since the radius of the sphere is one foot it may be termed a unit sphere. The surface area of the sphere is 12.57 square feet. By definition, the one-candlepower source radiates one lumen to each square foot of the sphere's surface. The total luminous flux from a source which radiates one candlepower in all directions is therefore 12.57 lumens. The solid section formed by the one-foot radii and the surface arcs is called a unit solid angle, or steradian. A lumen is the amount of light which passes through a steradian from this one-candlepower source. If the radius of the sphere were doubled, the area of the surface section subtended by the unit solid angle would quadruple, and the amount of light would be spread thinner. The total amount of light in this larger area would still be the same as in the original smaller area - one lumen. Illumination striking a surface is measured according to the number of lumens per square foot. One lumen per square foot is called one footcandle (fc). For example, the sphere shown in Figure 3-363 has one lumen per square foot and is therefore lighted to one foot-candle. If the radius of the sphere is doubled while the intensity of the light source remains the same, the sphere has one lumen to four square feet or .25 footcandle. Because we are measuring the amount of light striking the sphere, we refer to the footcandle as the unit of incident light. Another important term in light measurement is the unit of brightness, called the footlambert (fL). If the sphere in Figure 3-363 has a theoretical surface that perfectly diffuses and transmits the light falling upon it; it emits light on its surface at the rate of one lumen per square foot. A surface which emits light at this rate has a brightness of one footlambert. The emission may be due to: 1) self luminosity, as in the case of the sun, a light bulb, or the phosphor of a kinescope; 2) reflection, as in the case of the moon, the wall of a lighted room, and most objects we ordinarily view; or 3) light transmitted through an object, as in the case of clouds in an overcast sky, or a white glass globe enclosing a light bulb. Table 3-26 lists the brightness of some familiar light sources.

3-10.17.2 Light Sources

Light sources are divided into two types: incandescent, and luminescent. The former include the filament of the ordinary incandescent lamp, the flame of a candle, and the carbon arc. Luminescent sources include the fluorescent lamp, the neon lamp, the xenon arc, and mercury lamps.

3-10.17.2.1 Incandescent Lamps

When a material is heated above approximately 873°K (600°C), it radiates visible energy. Practical incandescent-lamp filaments operate at 2700° to 3450°K. Lamps having a relatively long life operate at the lower end of this range. As temperature increases, lamp-life decreases, but lamps become more efficient. Photoflood lamps, which have an average life of only six hours, operate with their filaments at approximately 3400°K, only 225 degrees below the melting point of tungsten, and produce more light than do longer-life bulbs of the same wattage. Similarly, it is possible to operate ordinary bulbs with increased efficiency and output by increasing filament voltage, but with a decrease in life. Theoretical life expectancy of a lamp approximately doubles for every four-percent increase. At the same time light output approximately doubles or halves with each increase or decrease of 13 percent. As the temperature of a tungsten filament increases, radiation at the shorter wavelengths increases more rapidly than at the longer wavelengths. When a lamp is operated at a relatively low temperature, it radiates a reddish-colored light. As the filament temperature increases the light becomes whiter since the radiation in the blue region becomes proportionately greater. This is illustrated in Figure 3-364. The color-temperature

Table 3-26. Brightness of Some Familiar Light Sources

North sky, clear day	1000 ft.
North sky, very dark day	10 ft.
Moon's surface	1500 ft.
Candle flame	1500 ft.
Snow in sunlight	9000 ft.
40-watt incandescent lamp, inside frosted, brightest spot	15,000 ft.
100-watt incandescent lamp, inside frosted, brightest spot	95,000 ft.
Sun	450,000,000 ft.
40-watt fluorescent lamp, T17, cool white	1150 ft.

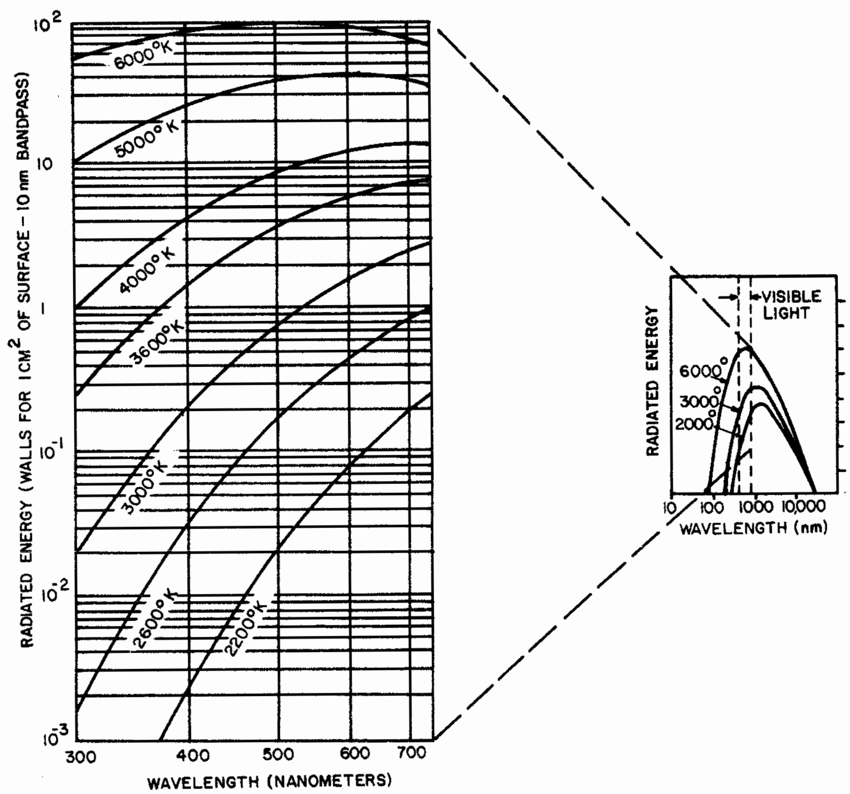


Figure 3-364. Spectral Distribution From Black-Body Radiator at Different Temperatures

rating of a lamp, measured in degrees Kelvin (K) is an indication of its spectral balance. A particular rating means that the lamp has the same distribution of energy in the visible spectrum as does a theoretically perfect incandescent radiator (black body radiator) at that temperature. Many lamps sold specifically for television and photographic use have a color-temperature specified. This is generally not critical however for black-and-white television or photography.

3-10.17.2.2 Tungsten-Halogen Lamps

The tungsten-halogen lamp is used extensively for television lighting because of its relatively high efficiency coupled with smaller size and lighter weight than regular incandescent lamps. The name arises because of the tungsten filament (similar to other incandescent lamps) and the inclusion of a small quantity of an element of the halogen family within the bulb - normally iodine or bromine. The first lamps of this type used iodine and were made with quartz-glass bulbs. Hence they were often called "quartz-iodine" or simply "iodine" lamps. Later designs sometimes used other types of glass and bromine rather than iodine. The broader term "tungsten-halogen" describes this entire family of lamps. The tungsten-halogen regenerative cycle is used to obtain an increased lamp life at higher operating temperatures than is possible with standard tungsten lamps. In addition, a more uniform light output throughout the life of the lamp is possible because of reduced blackening of the glass envelope as the lamp ages. To understand the process of the tungsten-halogen regenerative cycle refer to Figure 3-365. When the lamp filament is heated, tungsten atoms evaporate and are

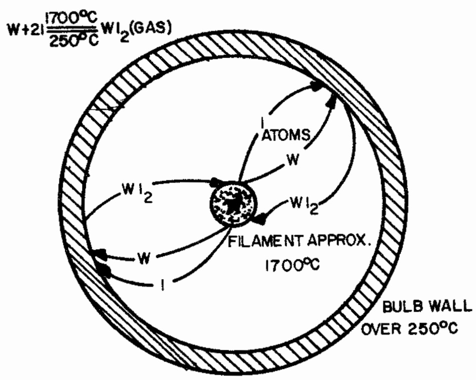
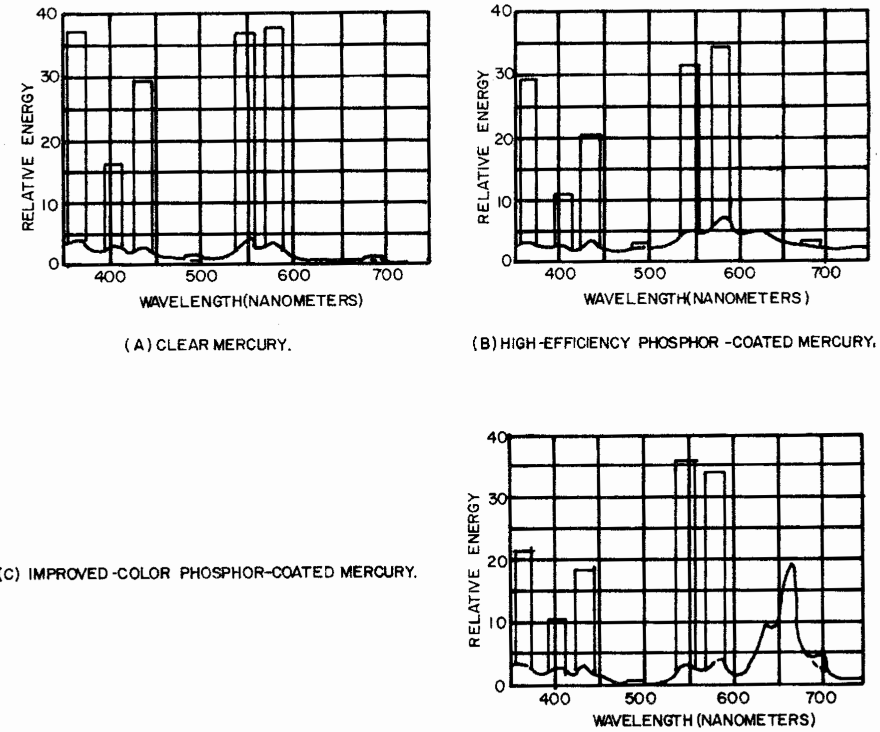


Figure 3-365. Tungsten-Iodine Regenerative Cycle

deposited on the bulb wall. These combine with iodine to form gaseous tungsten iodide. Tungsten-iodide atoms which strike the filament are decomposed by heat; tungsten is redeposited on the filament; and the iodine atoms are free to repeat the cycle. Thus, tungsten atoms that evaporate from the filament are returned by the halogen regenerative cycle. Without such regeneration the filament would be weakened by the loss of metal and the light output would be reduced because of the blackening of the bulb wall. The bromine cycle is similar to the iodine cycle. For the regenerative cycle to be effective the temperature of the glass envelope should be above 250 °C. For this reason, tungsten-halogen lamps have smaller envelopes than conventional lamps, and glasses having relatively high melting points, such as quartz, are used. Spectral distribution of a tungsten-halogen lamp is substantially the same as that of conventional incandescent sources operating at the same temperature.

3-10.17.2.3 Other Light Sources

Other types of light sources may be used on occasion where television is involved. Among these are fluorescent, mercury, sodium, and xenon lamps, and carbon arcs. Due to their awkward shape, fragility, and difficulty in providing a concentrated beam, fluorescent lamps are rarely, if ever, used in portable and environmental television lighting systems. They are most likely to be used when television is installed in laboratory or office areas. No special difficulty is encountered in their use, and their broad radiating surfaces give flat illumination with light shadows. Vidicon cameras produce excellent pictures under these conditions. Mercury lamps produce light by the passing of an electric current through vaporized mercury. The bulbs may or may not be coated with phosphors to convert the ultraviolet portion of their output into visible light. Mercury lamps are somewhat more efficient than incandescent lamps and have a long life, but they are relatively expensive and usually require an external ballast. Figure 3-366 shows the spectral response of several types of mercury lamps. Note that all are weighted in the purple region. This is quite noticeable visually, and does not coincide with the sensitivity curves of standard vidicons. Thus, the increased efficiency is largely lost by the vidicon transfer characteristic. Mercury lamps are frequently used for industrial lighting and in floodlighting. A drawback is that, after a brief power interruption, the lamp will not restart until it cools. Sodium discharge lamps are similar in principle to mercury lamps except that vaporized sodium is used in place of mercury. Their light is concentrated in the yellow region, but, as light souces go, they are exceptionally efficient. They are used principally in outdoor floodlighting and in street lighting. Ordinary vidicons have good sensitivity to their output. They can restart immediatly after a power interruption. A xenon lamp consists of two electrodes in a xenon-filled bulb. The light source is an arc passing through the gas. The equivalent color temperature of a xenon lamp is about 6000 °K, which approximates sunlight. Figure 3-367 shows the spectral distribution. Note the infrared radiation. Xenon lamps are available in sizes ranging from 30 watts up to 30 kilowatts. Lamps of 10 kw and higher require liquid cooling. The xenon gas in these lamps is under a pressure of approximately 10 atmospheres during operation, therefore a protective enclosure is mandatory. Protection against their ultraviolet radiation is also required. The larger sizes of xenon lamps are used extensively and have largely replaced the smaller carbon arcs. The short arc of a xenon lamp facilitates concentration of a high-intensity beam. For television purposes they generally are used outdoors where the light source cannot be located near the subject of interest. Nighttime viewing of missile launch pads or test facilities is an example of this application. Most xenon lamps are designed to operate from direct current. All require auxiliary equipment for starting. Initially,



(A) CLEAR MERCURY.

(B) HIGH-EFFICIENCY PHOSPHOR-COATED MERCURY.

(C) IMPROVED-COLOR PHOSPHOR-COATED MERCURY.

Figure 3-366. Spectral Distribution Typical of Most Mercury Lamps

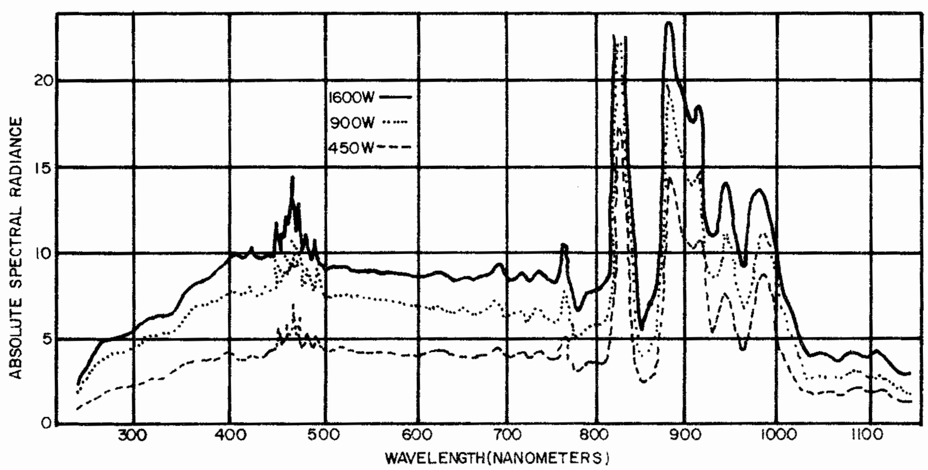


Figure 3-367. Spectral Distribution of Xenon Lamps (Three Types)

a brief high-voltage pulse, up to 50kV, ionizes the gas between the electrodes. After the arc is initiated the lamp operates at 16 to 75 volts. There are three categories of carbon arcs: the low-intensity arc; the flame arc; and the high-intensity arc. The high-intensity arc, is used in searchlights and motion-picture projectors. Television uses include missile illumination and sun simulators for imitating outer-space conditions. The high-intensity carbon arc has a color temperature that can be varied from 2900° to 6500°K. Units have been built having ratings as high as 600 kilowatts.

3-10.17.3 Invisible Light

On occasion it may be necessary to view materials or people without the use of visible light.

Applications include examination of materials that could be damaged by visible light, and medical study of persons or animals in what they believe to be darkness. For example, a medical researcher may use television to study sleeping patients. Special camera tubes are required for this application. Infrared light is more suitable for illuminating live subjects than ultraviolet light since prolonged exposure to ultraviolet light may cause sunburn, and eye protection is required. Ordinary incandescent lamps are excellent sources of infrared energy. Specially-designed infrared lamps are basically incandescent lamps operated at low color-temperature, say 2500°K. Lamp-life is practically indefinite under these conditions. A number of filters are available that absorb the visible light from an incandescent-light source. Wratten types 87, 87C, 88A, and 89B are among these filter types. Special glasses for the same purpose include Corning 7-56, 7-57, and 7-69.

3-10.17.4 Lighting Requirements

High-sensitivity vidicons, such as the 7735B, 7262A, and 8134, can produce usable pictures in modern cameras with highlight illumination as low as 0.1 footcandle striking the target. The relationship between scene brightness and target illumination is expressed as:

$$E = 0.116 \frac{B}{F^2}$$

where

B is the scene luminance, in foot-lamberts,

E is the illumination of the vidicon target,

F is the f number of the lens.

The formula assumes that the lens transmission is 70 percent, the object is 40 times the focal distance away, and the object is not appreciably off the center axis of the lens. It allows for a small amount of vignetting. Thus, if a high-sensitivity vidicon of one of the foregoing types (7735B, 7262A, or 8134) is used with an f/4 lens setting, the following expression obtains:

$$B = \frac{EF^2}{.116} = \frac{(0.1)\ (4^2)}{.116} = 14 \text{ fL}.$$

If an f/2 lens is used, the figure drops to about 3.5 fL. If the average highlight areas of the scene reflect 50 percent of the light striking them, the amounts of light striking the scenes are 2 and 7 footcandles for the foregoing examples. These figures are minimums for high-sensitivity vidicons. Scenes lighted at these levels will produce pictures which are of lower than optimum quality. Best signal-to-noise ratio and lag characteristics will be achieved with light levels ten times greater than these, or about 1.0 footcandle striking the vidicon. Therefore, for an f/4 lens, about 280 footcandles is desirable for a high-quality picture, while at f/2 about 70 is necessary. Figure 3-368 is a nomograph for calculating incident light where the luminous intensity of the source is known.

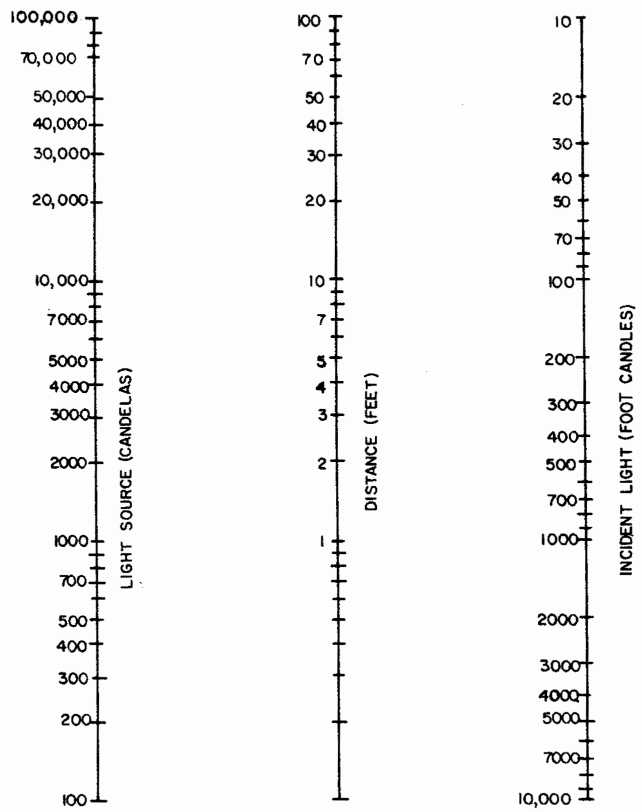


Figure 3-368. Nomograph for Calculating Incident Light From Source Intensity

3-10.18 TELEVISION MONITORS

The television monitor displays the desired picture in a convenient viewing location. Its circuits resemble the corresponding circuits of the ordinary television receiver, but without the RF, I-F, or audio circuitry. Home receivers are sometimes used for monitoring in installations which use RF distribution systems; but where best picture quality is desired, the television monitor will be incorporated into an all-video distribution system. The monitor offers wider bandwidth, and it eliminates various types of distortion that are virtually unavoidable with RF distribution systems. They are usually designed to have a uniform frequency response to 10 megahertz, matching that of most cameras and distribution equipment. Both transistor and vacuum-tube versions of the television monitor are available. While the transistor types are more complex and expensive, this is often offset by reductions in maintenance costs, reduced power consumption, and, in many cases, elmination of rack-cooling equipment. In applications requiring continuous service coupled with high reliability, transistor monitors are very frequently more economical than their vacuum-tube counterparts. Figure 3-369 shows the block diagram of a representative vacuum-tube monitor. The three-stage video amplifier has two outputs: one feeds the kinescope; the other feeds the sync circuitry. The DC restorer reestablishes the DC component of the video signal, since this is normally lost in the capacitive coupling circuits of the video amplifiers. As shown in Figure 3-369, when external sync is used it is fed to amplifier V4 through S1. When internal sync is used, V4 is fed by a signal from the last video-amplifier stage. Sync separator V5 removes the video portion of the signal when present, and its associated circuitry separates the sync signal into horizontal and vertical components. The vertical component of the sync signal triggers sawtooth generator V13 which feeds the vertical-output stage. This supplies the driving signal for the vertical-deflection coils. Phase-inverter V6 produces horizontal-sync pulses of both negative and positive polarity. These are compared in AFC circuit V7 with a feedback pulse from the horizontal-output circuit. The AFC is a phase-detector that supplies a DC correction signal to otherwise free-running multivibrator V8 to keep it in step with the incoming signal. The output of the horizontal multivibrator has a nearly sawtooth shape. However, the waveform required by the horizontal-deflection coils is rectangular. This is generated by horizontal output V9 in conjunction with horizontal damper V10. A portion of the signal from the horizontal-output circuit is rectified by V11 to supply the 18 kilovolts required at the ultor of the kinescope. Shunt regulator V12 stabilizes this voltage. Figure 3-370 presents the block diagram of a transistor monitor. The general arrangement of the circuits is similar to that of the vacuum-tube monitor except that, as is frequently the case, the high-voltage supply is independent of the horizontal-output circuitry. Notice also that the horizontal-output stage is capacitively coupled to the horizontal-deflection coils. Separation of the high-voltage section from the horizontal-deflection section permits improved high-voltage regulation by keeping the high voltage independent of raster width.

3-10.18.1 Monitor Circuits

Many circuits in the television monitor correspond in function to camera circuits. Some are practically identical. The video-amplifier circuits are so similar that there is no need for a separate discussion. (Refer to paragraph 3-10.16.3.) Others have to meet special requirements of the monitor that are not present in the camera. These circuits are detailed in the following sections.

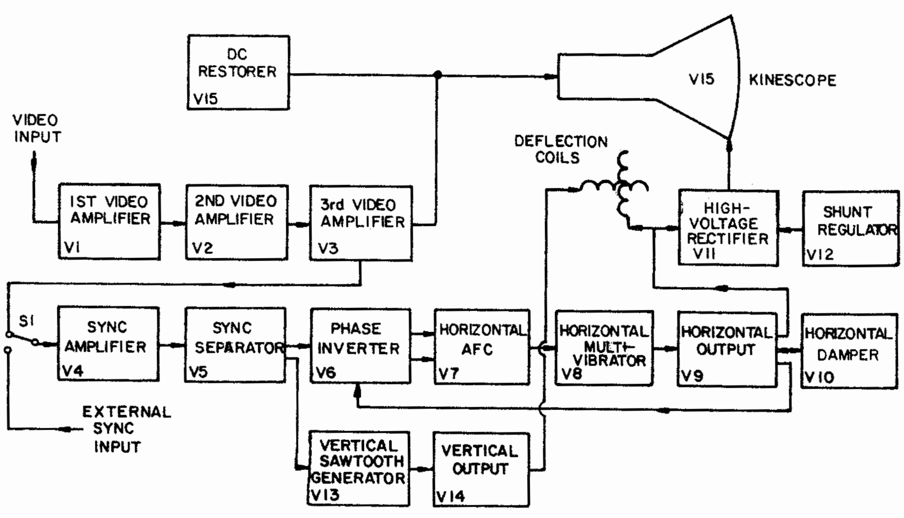


Figure 3-369. Vacuum Tube Television Monitor

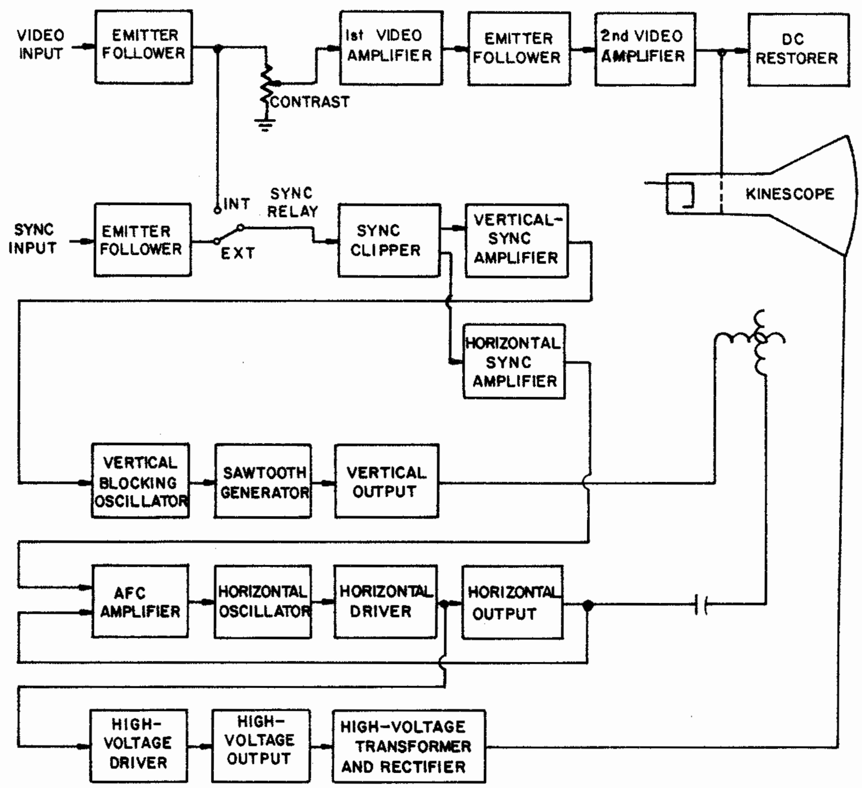


Figure 3-370. Transistorized Television Monitor

3-10.18.2 Video-Output Stages

The video-output stages of television monitors are similar schematically to other video amplifiers. However, they must meet special requirements which do not apply to interstage amplifiers. This stage must be capable of delivering a video signal of 70 volts or more to either the cathode or the control grid of the kinescope. The control grid requires a slightly stronger signal than does the cathode. Figure 3-371 shows the schematic of the video-output circuit of a monitor that uses conventional circuitry. A cathode-peaking circuit (C14, C39, and R53) compensates in part for losses due to the distributed capacitances of the circuit. A shunt-series peaking circuit, consisting of L5 and L6, provides additional correction. The output signal passes through C15 to the kinescope grid; the DC restorer maintains a fixed reference level for the signal. Figure 3-372 shows the schematic diagram of the output stage of a transistor monitor. This circuit has no voltage gain but has considerable power gain. Its output impedance is relatively low, therefore the circuit is relatively undisturbed by stray circuit capacitance. Transistor monitors are also manufactured which use single-ended video-output circuits. These are quite similar to conventional video amplifiers, the principal difference being the selection of a transistor able to handle the relatively high (for transistors) circuit voltage necessary to produce a video signal of approximately 70 volts.

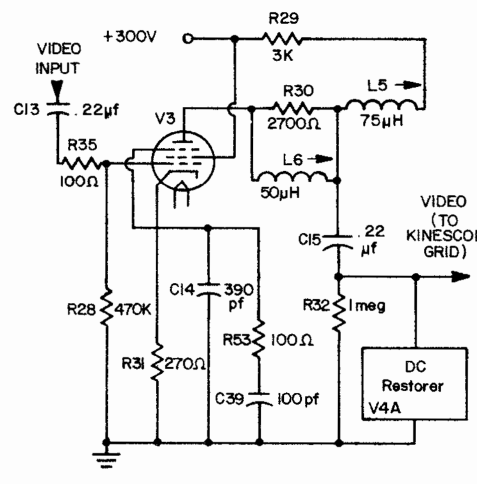


Figure 3-371. Vacuum Tube Video Output Circuit

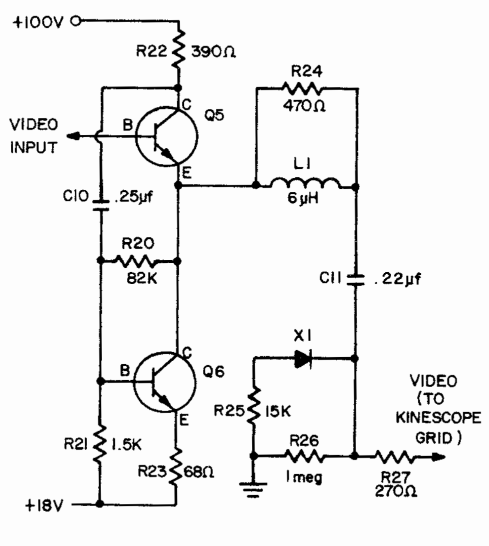


Figure 3-372. Transistorized Video Output Circuit

3-10.18.3 DC Restorers

Video signals are usually fed from one stage to another through coupling capacitors. Since a capacitor cannot pass direct current, the DC component of the signal is lost in the coupling, possibly with some low-frequency components of the signal. A capacitive-coupling network causes a video signal to average itself around the bias level of the input circuitry of the stage being fed. To maintain predominantly bright scenes in a proper relationship to predominantly dark scenes, the monitor must recover this DC component. Otherwise, all scenes will have the same degree of average brightness. Thus, if a monitor is adjusted for best picture when a dark scene is displayed, the details in the dark areas will be lost when a bright scene is viewed. The same is true going to the opposite direction - from a bright to a dark scene. The monitor must be readjusted if these details are to be seen correctly. The DC restorer automatically maintains the correct brightness level regardless of scene content. Figure 3-373 illustrates the function of the DC restorer. Video signals from three different scenes are shown as they leave the target of a vidicon tube in Figure 3-373A. Note that the lightest and darkest portions of all scenes are at the same brightness level. After the signal has passed through a capacitive coupling network (Figure 3-373B), the average voltage of each scene is the same, but the voltage levels of the peaks are shifted. The DC restorer returns these signals to their original relationship. This is usually, but not always, the last step in signal processing before the signal enters the kinescope tube. In any event there will be no coupling-capacitor between the DC restorer and the kinescope. This circuit functions by holding either the positive or the negative extremes of the signal to a given reference level. Ordinarily the sync tips are used in a composite signal, and the blanking tips in a noncomposite signal. The polarity depends on whether the signal is injected at the grid or at the cathode of the kinescope. Figure 3-374A and B show the two most common DC-restorer circuits. The principal difference is that in (A) the signal feeds the cathode of the kinescope, while in (B) it feeds the grid, therefore the signals have opposite polarities.

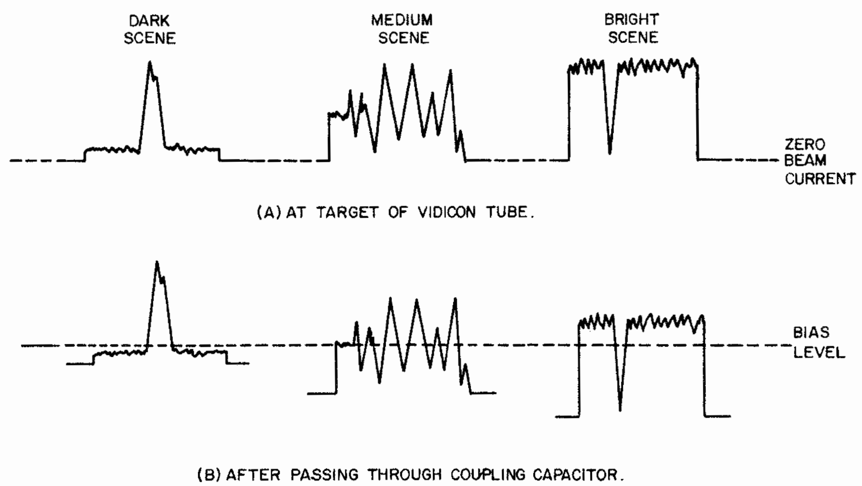


Figure 3-373. Video Signals

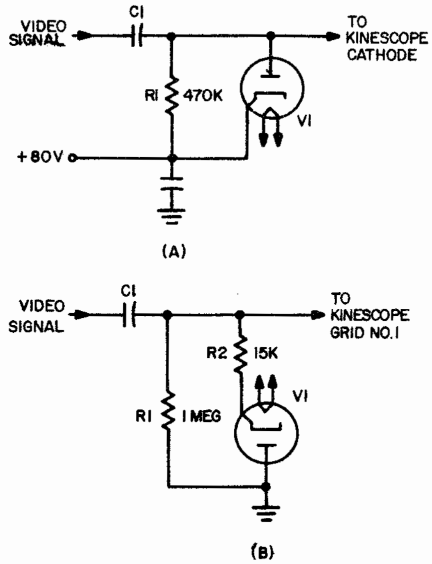


Figure 3-374. DC Restorer Circuits

This requires that the diodes be reversed. In both circuits the kinescopes are cathode-biased and the grids are at ground potential. For this reason the circuit feeding the grid of the kinescope is referenced to ground, while the circuit feeding the cathode is referenced to the positive-bias voltage of the kinescope cathode. In the circuit of Figure 3-374A, R1 conducts until the charge on C1 (and therefore the charge at the kinescope cathode) is +80 volts. The most positive portions of the video signal (the sync tips) cause an increase in the voltage at the plate of V1. When this occurs, V1 conducts and the voltage returns to +80 volts. When the signal moves in a negative direction, V1 cannot conduct because its plate is then more negative than its cathode. All portions of the signal more positive than +80 volts are quickly reduced to +80 volts. Consequently V1 holds the most positive portions of the signal to a fixed level. The action of the diode in Figure 3-374B is similar except for the reversal in polarity. The grid of the kinescope starts at ground potential. A negative-going signal causes V1 to conduct, returning the diode and kinescope grid to ground potential. The diode is cut off during the positive swings of the signal, therefore the video signal is always positive with respect to ground. In both circuits the value of R1 is large enough to prevent a rapid discharge of C1 because this event would cause the circuit to introduce tilt into the signal. The keyed-clamper circuit performs a function similar to that of the DC restorer, except that in the clamper the action is practically instantaneous, and the circuit functions only during blanking time. The DC-storer circuit is ordinarily adequate for maintaining a black level at the kinescope. Where the requirement is more exacting (e.g., the blanking mixing circuits of a camera or the modulator stage of a transmitter), the clamper will usually be used.

3-10.18.4 Differential Amplifiers

Differences in DC potential between the video source and the monitor will cause current to be present between the two locations in the shield of the interconnecting coaxial cable. This difference in potential also appears at the center conductor of the cable. For this reason it is sometimes called a "common-mode" signal. The current in the center conductor will cause a hum signal in the picture. Since the video signal in the center conductor is measured relative to the shield, and the hum signals in the two conductors are in phase and have approximately equal amplitudes, it is possible to separate the two signals in a differential amplifier. Figure 3-375 is a schematic of one type of differential

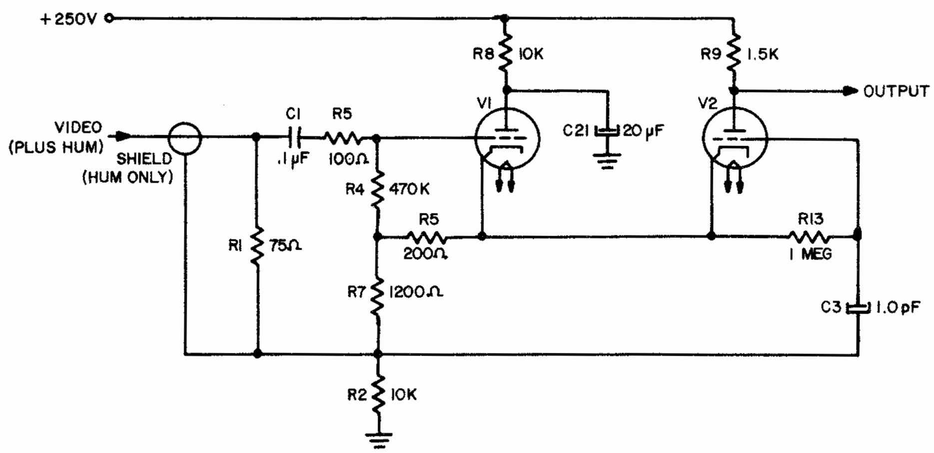


Figure 3-375. Differential Amplifier

amplifier. The circuit has two inputs: one from the center conductor; and one from the shield of the coaxial cable. The shield, which is grounded at the transmitting end, feeds the junction of R1 and R2. As far as the amplifier is concerned, the hum signal is generated across R2. The video-plus-hum signal feeds the grid of the cathode-follower, while the in-phase hum signal feeds only the cathode. The video signal from V1 is cathode-coupled to V2. The hum-only signal from the shield of the coaxial cable is fed to the V2 grid. Under these conditions the hum signal subtracts from the hum-plus-video signal in V2, leaving only the video signal at the plate of V2. Figure 3-376 shows the circuit of another type of differential amplifier. As in the preceding version, the signal at the center conductor of the coaxial cable feeds the grid of the first tube, while the signal on the shield becomes the AC reference for the stage. This time the signal feeding the grid is inverted in the plate circuit and fed to the grid of V2. The signal from the shield of the coaxial cable also feeds V2, where it becomes the AC reference for the stage. It is out of phase with the video-plus-hum signal at the grid. Hum cancellation therefore takes place across R10 due to the out-of-phase relationship between the two hum signals. Potentiometer R8 is adjusted for exact cancellation of the two signals.

3-10.18.5 Sync Separators

When a television monitor is fed a composite video signal, the sync portion of the signal must be separated from the incoming signal and channeled to the sync circuitry. In addition, the horizontal and vertical portions of the signal must be channeled to the appropriate deflection circuits of the monitor. An overdriven amplifier usually performs the sync-separation function. A combination of high-pass and low-pass filters then separates the horizontal-sync and vertical-sync signals. Figure 3-377 shows a typical sync-separation circuit. Sync amplifier V401 is unbiased, so partial limiting takes place in this stage, eliminating part of the video signal. Sync clipper V402 cuts off both positive and negative excursions of the sync signal, thereby delivering an output that is clean and of uniform amplitude. It is positively biased at grids one and two. Grid one is unbypassed and is therefore partially degenerative. A grid-leak bias circuit on grid three consists of R408, R406, and C402. The otherwise heavily forward-biased stage is thus easily saturated by a positive-going signal because of the grid-leak bias. Therefore the driving signal can switch the stage between saturation and cutoff. There are two outputs from V402. The first passes through C405 to the horizontal-AFC circuit, which is sensitive to the horizontal-sync pulses. The second passes through the integrator circuit which, for practical purposes, eliminates the horizontal-sync pulses and vertical serrations from the composite-sync signal. The output of the integrator is a single pulse, corresponding to the vertical-sync interval of the composite-sync signal. The circuit will function using either EIA or industrial (unserrated) sync signals. Figure 3-378 shows the sync-separation circuit of a transistor monitor. The isolation provided by Q2 is necessary when an external sync signal is fed

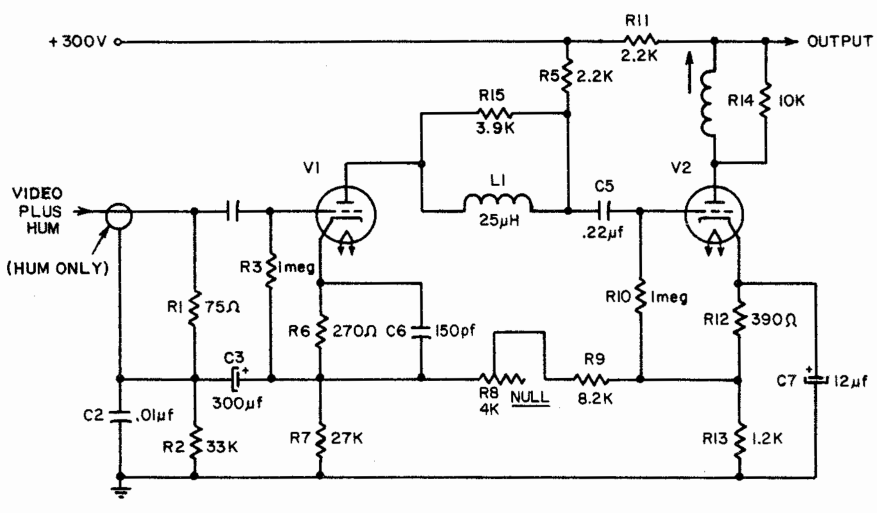


Figure 3-376. Differential Amplifier

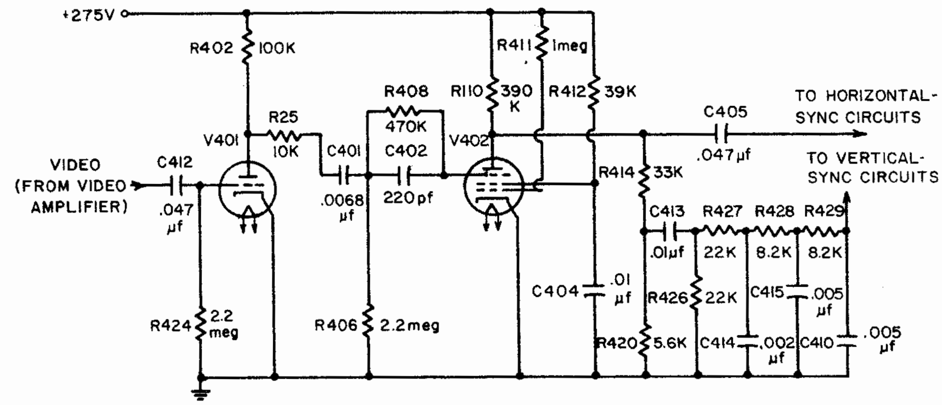


Figure 3-377. Typical Sync Separation Circuits

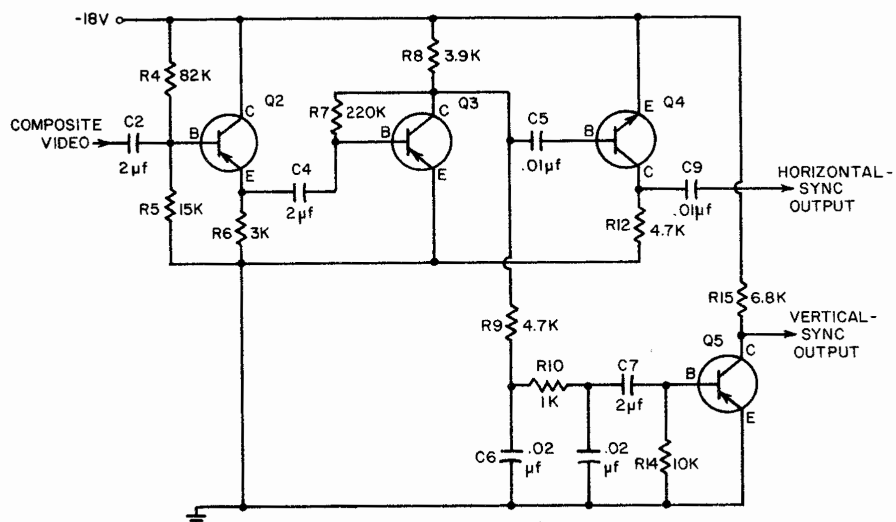


Figure 3-378. Sync Separation Circuit of a Typical Monitor

to the monitor. Transistor Q3 is a forward-biased stage that goes into saturation easily and removes the video signal by limiting action, leaving only the sync signal in the output. Two outputs are taken from the collector of Q3. The first drives horizontal-sync amplifier Q4, in which further limiting action takes place. The second feeds an integrator circuit consisting of C6, R9, R10, and C8, which removes the horizontal-sync pulses and the serrations in the vertical pulse. This leaves a signal corresponding to the vertical-sync interval of the composite signal. From the integrator, it goes to vertical-sync amplifier Q5, which is also a limiter stage. In turn, this stage drives the vertical-deflection circuitry. Figure 3-379 shows the sync-separator circuit of a receiver designed especially for use as a master monitor. The video signal enters at Q3, which is an isolation stage, and proceeds to Q4, which clips when fed signals of normal amplitude. An overdriven stage (Q5) clips both positive and negative peaks of the driving signals, leaving only sync at the collector. Two outputs are taken from this circuit. One feeds the horizontal-sync processor (7Q3), the other feeds the vertical-sync processor (7Q4). The signal that feeds the horizontal-sync circuit is differentiated by 7C8 and 7R35 and is then fed to the base of 7Q11. Differentiation, clipping in 7Q11, and

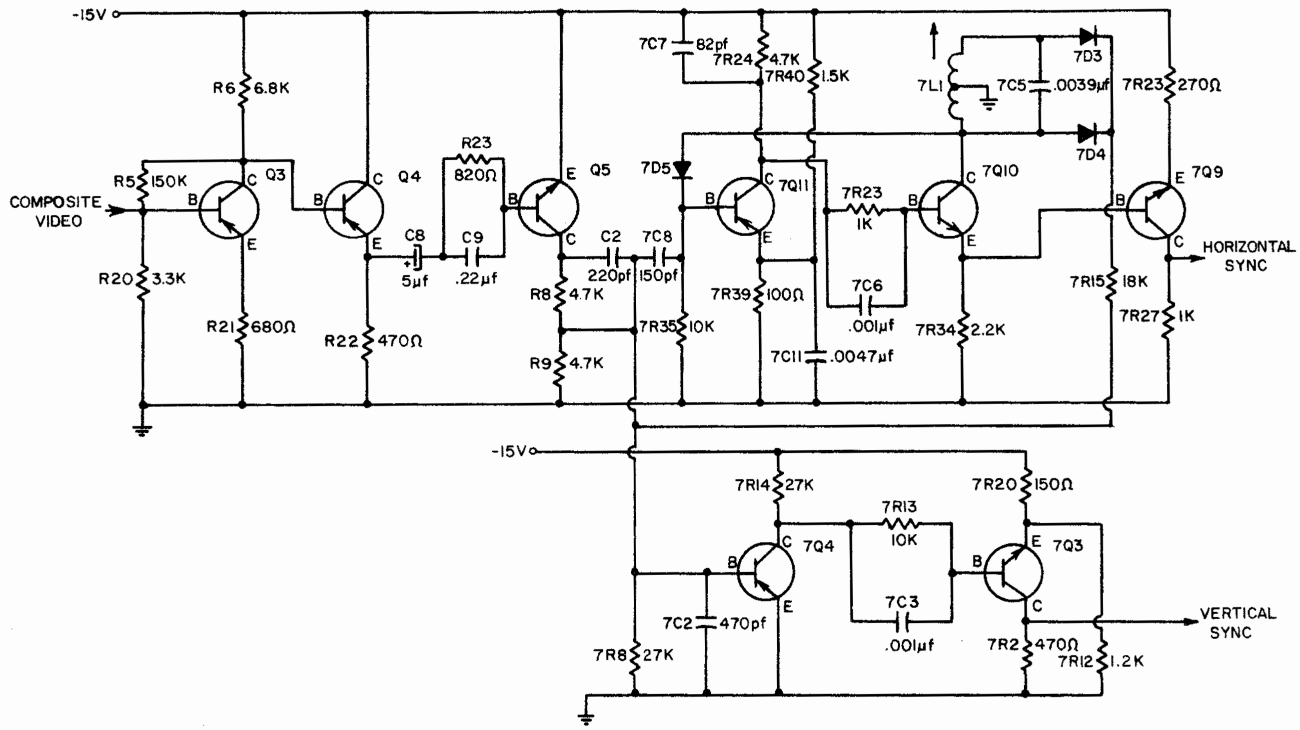


Figure 3-379. Sync Separation Circuit of a Master Monitor

stretching of the pulses in RC network 7C7-7R24 give the equalizing pulses (and pulses corresponding to leading edges of the vertical-sync pulses) the same duration as the horizontal-sync pulses when they arrive at the base of 7Q10. Transistors 7Q10 and 7Q9 amplify and limit the signal before feeding it to the horizontal-deflection circuits. The signal at the collector of 7Q10 is applied to ringing-coil 7L1, which is tuned to 15.75 kHz. The voltage at the collector of 7Q10 is therefore a large-amplitude sine wave. Diode 7D5 allows the positive half of this sine wave to be fed back to the base of 7Q11. This signal coincides with the unwanted equalizing pulses (those that occur at the half-line interval) and cuts off 7Q11 during this interval. This gates-out the unwanted equalizing pulses. A second output from the ringing coil feeds the vertical-sync circuitry to improve the timing of the vertical-sync pulses. Full-wave rectifier 7D3-7D4 converts the 15.75 kHz sine-wave signal into 31.5-kHz negative pulses. These coincide with the six vertical-sync pulses. The sync signal from Q5 is added to this signal. Transistor 7Q4 is biased to conduct only when the two signals coincide. The output of 7Q4 is six pulses during each vertical-sync period. This assures accurate interlace in the picture since the vertical-deflection circuit is fed only the actual sync signals. Limiting and integrating action takes place in 7Q3 and the intervening circuitry, resulting in a signal that corresponds to the vertical-sync interval.

3-10.18.6 Vertical-Deflection Generators

The purpose of the vertical-deflection generator is to cause the electron beam to travel linearly from the top of the kinescope to the bottom in step with the scan of the camera tube. Its output should provide a sawtooth current in the vertical-deflection coil. Figure 3-380 shows the schematic of a vertical-deflection circuit typical of many vacuum-tube monitors. Vertical-hold control P502 sets the natural frequency of cathode-coupled multivibrator V501. Locking occurs when the natural frequency of the multivibrator is slightly lower than the frequency of the driving pulses. The amplitude of the multivibrator output is controlled by height-control P501. The charging and discharging of capacitors C504 and C505 by the square-wave signal from the multivibrator results in a sawtooth waveform that feeds the output

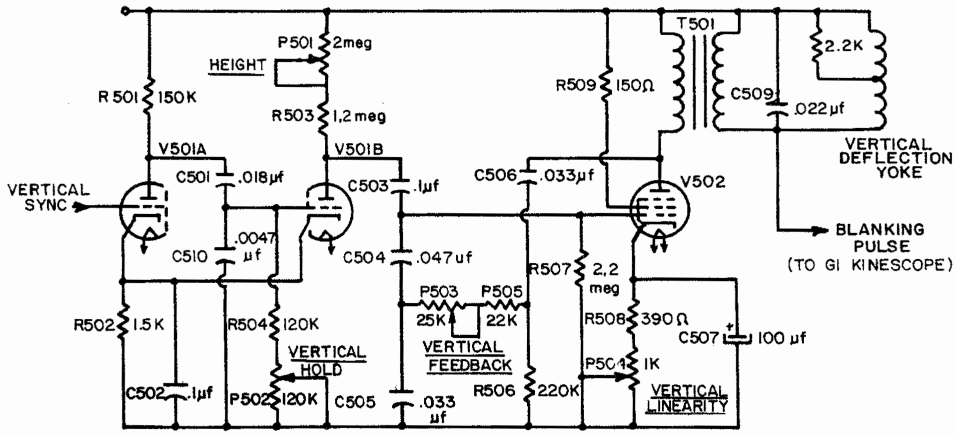


Figure 3-380. Vertical Deflection Circuit for a Vacuum Tube Monitor

stage (V502). Linearity control P504 sets the bias of V502. Ideally, this sets operation over a nonlinear portion of the characteristic curve that just cancels the nonlinearity of the charge curve of C504 and C505. Vertical-feedback control P503 determines the amplitude of the small step at the start of each vertical scan, corresponding to the top of the picture, through adjustment of the amplitude of a feedback signal around V502. The step originates at retrace due to the collapse of the magnetic energy stored in the vertical-deflection coil. The control is adjusted such that the step is of proper amplitude for best linearity at the top of the picture. The output of V502 is transformer-coupled to the deflection coil. A damping network formed by C509 and the 2.2K resistor minimizes oscillation of the yoke at its natural frequency. A secondary output of the circuit is utilized as a blanking pulse, which adds to the effectiveness of the blanking signal. In subsequent circuitry the sawtooth component is filtered out, leaving only the vertical-flyback pulse. This is essential, since the sawtooth waveform would cause nonuniform shading in the kinescope presentation. Figure 3-381 shows the vertical-deflection circuitry of a professional-quality transistorized monitor. Here Q4 is a blocking oscillator synchronized by the vertical-sync pulse. Vertical-hold control R15 determines the natural frequency of the blocking oscillator through adjustment of the bias current. Damper-diode X1 prevents sustained oscillation of the blocking oscillator. The output of Q4 is a positive pulse of about 400-microseconds duration, recurring at the vertical rate. This is coupled through emitter-follower Q2 to power amplifier Q3. The pulse cuts off Q3, causing the magnetic field about L301 to collapse. Following this event, the transistor returns to conductance, causing L301 to charge at a linear (sawtooth) rate. The waveform is impedance coupled to the deflection yoke. Diode X3 rectifies the flyback pulse and feeds a DC signal proportional to the deflection voltage to the base of Q1. Here it is amplified and then fed to the base of Q2 as a negative feedback voltage. This improves height stability. The NE2 lamp is a protective device for preventing excessive drive to Q3 during the time immediately after the set is turned on i.e., while C4 is charging. Height-control R17 also functions through Q1, adjusting the bias of Q2 and thereby controlling the gain of the circuit. The circuit has two additional feedback paths: one through R20; the other through R12. The latter provides the regular vertical-linearity adjustment and determines overall linearity by varying the amplitude of a predominantly low-frequency feedback signal. Resistor R20 feeds back a high-frequency signal, so it controls principally the linearity near the top of the picture. The Q5 is a regulator stage. Ripple which may be present on the - 18 volt buss is fed to the base of Q5 through C17 and inverted. This is then applied as a cancelling signal to the collector of Q2. A secondary output of the vertical-deflection circuit feeds a blanking pulse to the kinescope. Capacitor C14 isolates the output transistor from the yoke to allow an adjustable centering-current to pass through the yoke without affecting the output stage. Figure 3-382 shows yet another approach to vertical-scan generation in a transistor monitor. Multivibrator Q1 is a conventional unijunction-transistor relaxation oscillator. As is normal with television monitors, it is free-running, synchronized to a drive pulse. The sawtooth voltage results from the charging of C2 and C3 through R11 and R25. Unijunction-transistor (UJT)

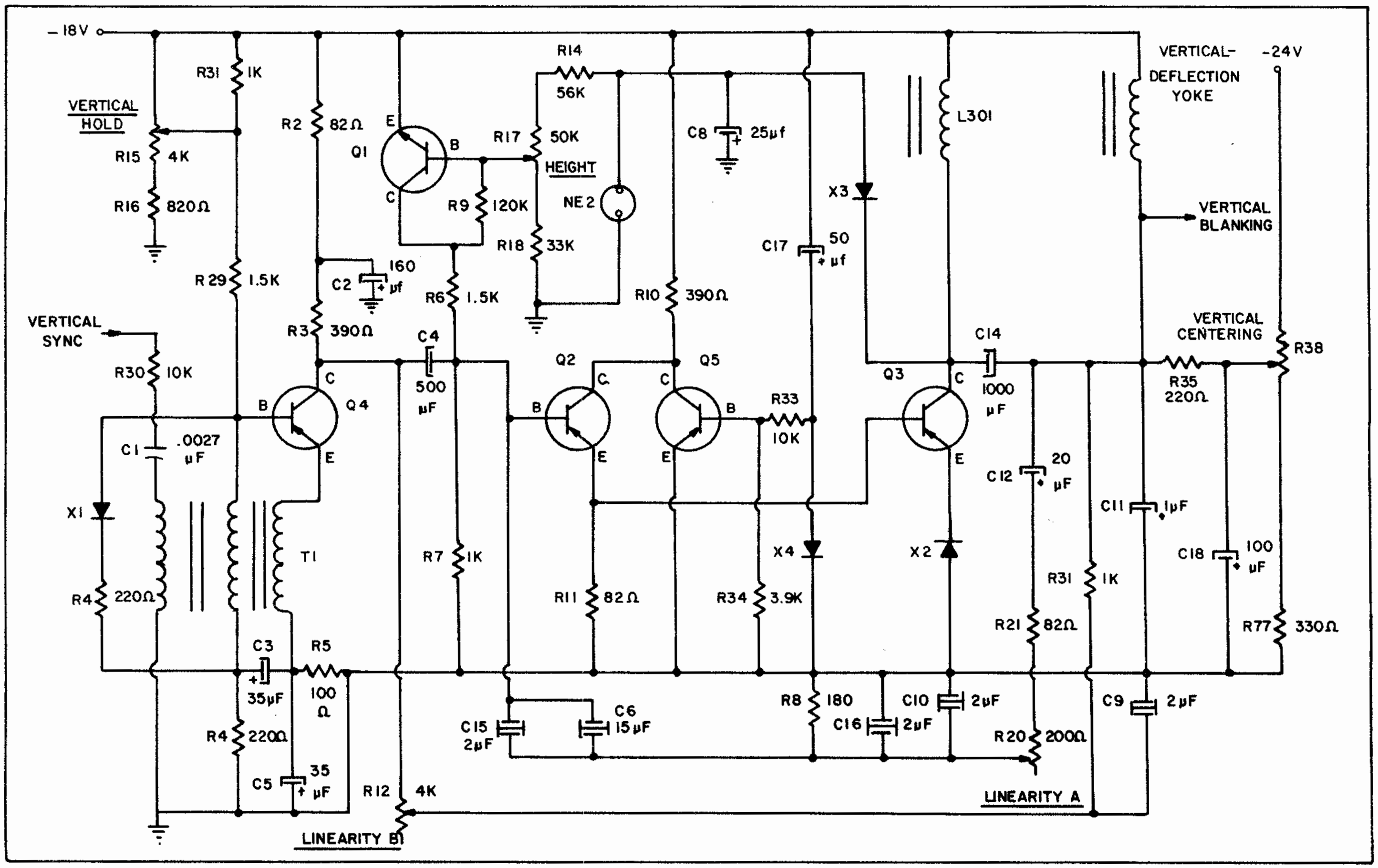


Figure 3-381. Vertical Deflection Circuit for a Transistorized Monitor

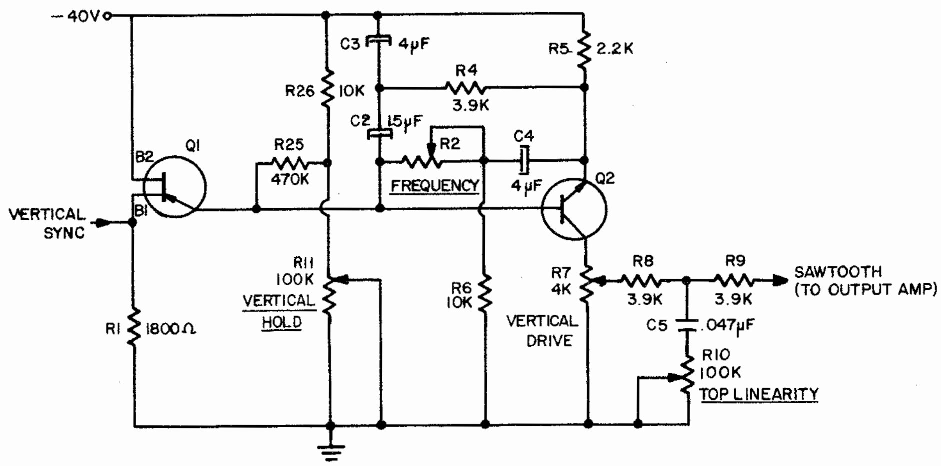


Figure 3-382. Vertical Scan Generator

Q1 is cut off until the drive pulse at base one causes it to conduct. Capacitors C2 and Ce then discharge through Q1. Resistors R11 and R2 set the free-running frequency of Q1 by adjusting the charging of C2 and C3 through R11 and R25. Resistor R4 provides a feedback path to improve linearity. The original nonlinearity in the charging circuit is applied to C3. It adds to the original sawtooth, leaving a signal across C2 which consists only of the difference between the two signals. The result is a linear sawtooth, since the nonlinearity exists at both ends of the capacitor. Height-control R7 adjusts the amplitude of the output signal. In combination, R10 and C5 form an additional charging circuit that affects linearity principally at the top of the picture. The potentiometer is employed to adjust the amplitude of the "step" occuring at the end of retrace.

3-10.18.7 Horizontal Deflection

The conventional television monitor uses a free-running oscillator synchronized to the horizontal-sync pulses for control of horizontal deflection. Synchronization is almost universally obtained through the use of an AFC circuit that compares the timing of the incoming sync signals to a feedback signal from the horizontal-output circuit. The output of the AFC circuit is a DC correction voltage that is applied to the oscillator. Figure 3-383 shows the horizontal-oscillator circuitry used in most vacuum-tube monitors. The frequency of oscillation of a conventional cathode-coupled multivibrator (V10) is determined by C38 and R80. A parallel-resonant circuit (17 and C37) stabilizes the multivibrator frequency at 15.75 kHz for standard scan rates. The square-wave output of the multivibrator is modified by C40 and R81 to form a sawtooth-like waveform that feeds the horizontal-output stage. An automatic frequency-control circuit (AFC), consisting in part of phase-detector V7, establishes exact control of the phase and frequency of the multivibrator. Phase-inverter V6 feeds sync pulses of positive polarity to the cathode of V7A, and of negative polarity to the plate of V7B. Horizontal-flyback pulses from a secondary winding on the horizontal-output transformer are reshaped by R54 and C27 to form sawtooth pulses. The sawtooth pulses are applied to the remaining plate and cathode of V7A and V7B. When the oscillator is in step with the sync pulses, the correction voltage at the grid of V10A is approximately that established by horizontal-hold control R90. Should the oscillator tend to drift, one diode will increase conductance while the other decreases. This causes a current to flow in R50 and R51. A correction voltage results that is applied to the grid of V10A. This change in voltage causes a change in frequency in the proper direction to correct the drift sensed by the AFC diodes. The time-constant of the correction voltage is determined by C36. Certain types of industrial sync may cause a bending of vertical lines near the top of the picture due to loss of horizontal-sync pulses during the vertical-sync interval. The time required for correction may be increased by decreasing the value of C36. Too small a value, however, will cause instability when no signal is present, and will increase the sensitivity of the circuit to noise. The phase-detector (AFC) circuit is generally preferred over directly synchronized circuits, such as are used in vertical-scanning circuits, because of its relative immunity to noise. Noise pulses tend to average out in the AFC circuit. A popular variation of this circuit places the horizontal-hold network at the grid of V10A, giving control directly at this point, rather than through the diodes. The change has no effect on the operation of the circuit. Figure 3-384 illustrates a different approach to maintaining synchronization

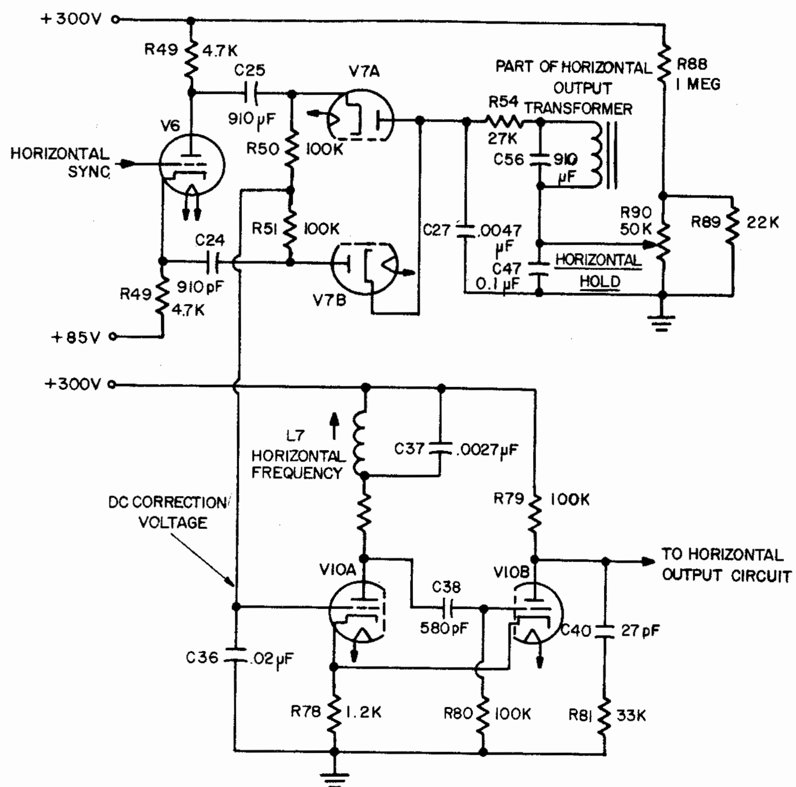


Figure 3-383. Typical Horizontal Oscillator Circuitry

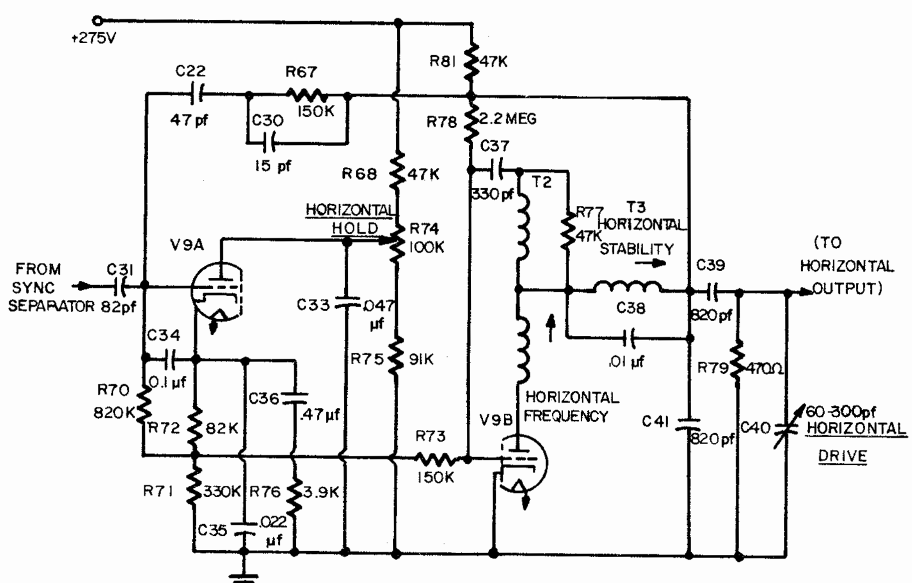


Figure 3-384. Horizontal Oscillator Circuitry

with relative noise immunity. The frequency of blocking-oscillator V9B is determined by C37, R78, R73. The slug adjustment of blocking-oscillator transformer T2 determines the width of the pulse from the blocking oscillator, and, to a lesser extent, its natural frequency. Coil T3 resonates with C38 near 15.75 kHz, thus stabilizing the oscillator at this frequency. In addition, this combination serves to filter the output of the blocking oscillator. The pulse across R81 is converted to a sawtooth signal by C41, R79, and C40. The control tube (V9A) keeps the blocking oscillator in phase with the incoming horizontal-sync pulses. The triode, having grid-leak bias, conducts only on the positive peaks of the driving signal. The driving signal consists of the incoming sync pulses and the sawtooth signal fed back through C22, R67, and C30. The integrator network in the cathode circuit of V9A filters its pulsating cathode current to form a control signal for V9B. This is then fed to the grid as a bias signal. In normal locked-in operation, the sync pulse is superimposed on the steep leading edge of the sawtooth signal. Should the phase of the oscillator tend to shift, the position of the sync pulse superimposed on the sawtooth will also shift, and V9A will conduct for a longer or shorter period of time. This changes the voltage appearing on the integrator network in the cathode circuit, and, in turn, the bias voltage at the grid of V9B. The change is of the proper polarity to correct the frequency of the blocking oscillator. Manual adjustment of the blocking-oscillator frequency is accomplished by changing the plate voltage of V9A at R74.

3-10.18.8 Horizontal-Output Circuits

The horizontal-deflection system of a television monitor is usually associated with the power supply for the kinescope ultor. Figue 3-385 shows a typical circuit. For linear deflection of the scanning beam a sawtooth current is required in the horizontal-deflection coil. At standard scanning rates it is highly inductive, requiring a nearly rectangular voltage waveform across it to produce a sawtooth current in it. The grid voltage driving V1 has a nearly sawtooth waveform, Figure 3-386. During the interval from *a* to *b*, V1 is cut off. At time *b* the grid of V1 goes sufficiently positive for it to conduct. The plate-current path of V1 is through a portion of T1, V3, and L1 to the 275-volt supply. Autotransformer T1 couples the signal to horizontal-deflection coil L3, causing the beam to move to the right. At times the driving waveform at the grid of V1 suddenly goes negative, cutting off V1. If all the circuitry on the plate side of V1 were purely resistive, current in the deflection coil would stop, and the beam would go to the center of the screen. However, all of these components are reactive and contain energy when V1 is cut off. The portion of greatest interest is the magnetic energy in the deflection coil. Current here at time *c* is at its maximum value, Figure 3-386. When V1 is cut off, the magnetic field surrounding the coil collapses and resonates with the distributed capacitance in the circuit. This causes the current in the coil to reverse polarity. At the time the current is at zero, the voltage across the combination of L3 and C_S (Figure 3-385) is at its maximum.

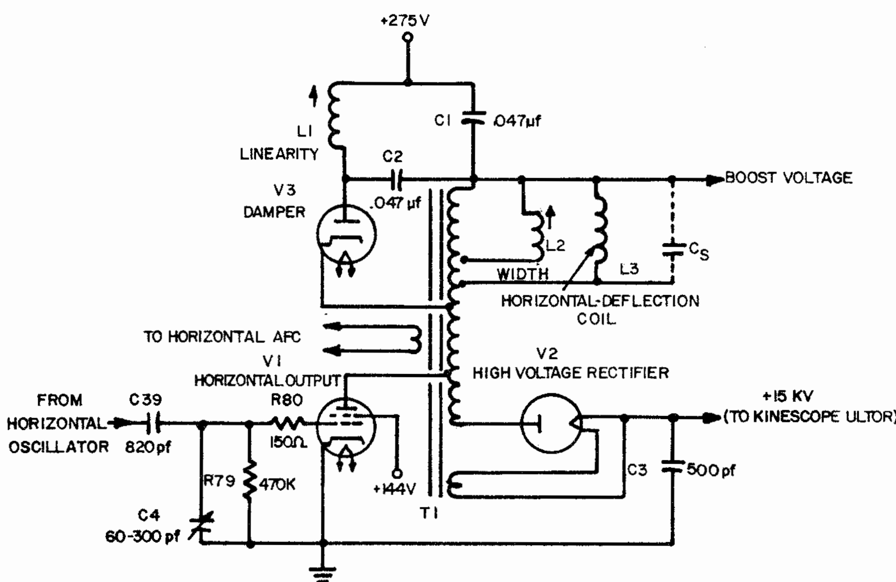


Figure 3-385. Horizontal Deflection and High Voltage Power Supply Circuit

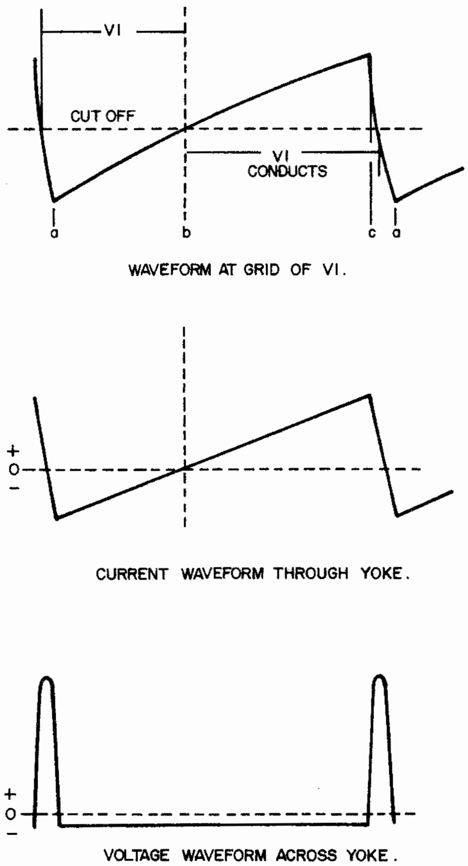


Figure 3-386. Waveforms in Typical Horizontal Deflection Circuit

The beam is now midway through retrace. Capacitor C_S now discharges through the deflection coil, causing the beam to move still father to the left. When C_S has fully discharged, the beam has completed the retrace. The energy in the circuit now is again magnetic but has reversed polarity since the start of retrace. This is point a, again in Figure 3-386. The current from the coil is now of the proper polarity to flow through damper tube V3 as the field around L3 collapses. Capacitor C2 is placed in parallel with C_S, lowering the resonant frequency of the circuit and giving the current waveform a more gentle slope. This is the interval from a to b in Figure 3-386. The beam is moving from left to right. At time b the current in the yoke is again zero. Most of the energy in the circuit is in the form of a charge on C2, and the beam is approximately at the center of the screen. Capacitor C2 cannot discharge now because of the reversed polarity of V3. At this time the signal from the horizontal oscillator causes V1 to conduct. The current through V1 thus is provided with an additional path. Since the charge on C2 adds to the power-supply voltage, the plate current of V1 now passes directly through the horizontal-deflection coils to C2. The charge on C2, after several cycles of operation, will probably be equal to or greater than the power-supply voltage. The voltage at the end of C2 connected to the deflection coil (yoke) will therefore be at least double the power-supply voltage with respect to ground. A small amount of current may be drained from C2 for circuits requiring a higher voltage than that supplied by the regular power supply. Thus, C2 is commonly referred to as the "boost" capacitor, and the voltage at the positive side is called the "boost" voltage. At time c, V1 is again cut off, and the cycle repeats. Width is controlled by placing a variable inductor, L2, in parallel with a portion of the autotransformer. The shunting effect of L2 reduces the total inductance of the circuit, the amount of such reduction determining the width of the picture. Where the high-voltage supply is separate from the horizontal-deflection circuitry, width may be controlled by adjustment of the B+ to the circuit. The somewhat nonlinear discharge of the yoke causes expansion of the picture at its left. Linearity is improved by the insertion of L1 in series with the B+ lead for the entire circuit. The coil resonates at the horizontal-deflection frequency, causing the picture to be stretched through the center. Best linearity is achieved near the point of resonance. This is also the point where the picture has maximum width. Capacitor C1 tunes L1 to resonance. Adjustment is ordinarily made slightly off-resonance to compensate for stretch normally occuring at the left edge of the picture. Adjustable capacitor C4 controls the amount of signal delivered to the grid of V1. Excessive signal at this point will cause V1 to conduct before the damper tube is cut off. Nonlinear travel of the scanning beam across the picture produces a vertical white line (drive line) through the center of the picture. A weak driving signal will not create sufficient grid-leak bias for V1, and excessive plate current may flow. To prevent this, C4 should be set by first increasing it until the white line appears, and then decreasing it until the line disappears. Figure 3-385 shows that a large-amplitude voltage pulse (flyback pulse) appears across L3 during retrace. This voltage is several times the power-supply voltage. It is stepped up even further by T1 and then applied to the plate of V2, which rectifies it to form a pulsating DC voltage. The flyback then passes through a filter network to become the kinescope ultor voltage - usually 12,000 volts or more. Another winding on T1 supplies the filament voltage for V2. Normally, the current drain

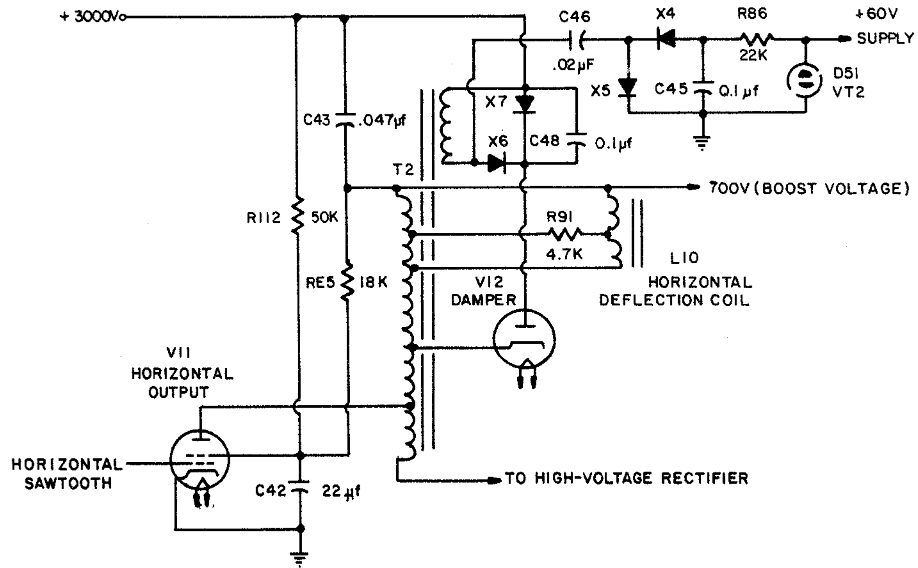


Figure 3-387. Horizontal Output Circuit

from this supply is rather minute, and a single capacitor of small value (C3) is sufficient for filtering. Another winding sometimes found on T1 couples a feedback pulse to the horizontal-AFC circuits to stabilize the scan frequency. The circuit of Figure 3-387 employs a different approach to linearity correction. During scan through the right side of the raster, V11 conducts through autotransformer T2, V12, and X7. During retrace time the added winding of T2 couples the flyback pulse through X6 to C48, causing it to charge. The charge on C48 adds to the +300 volt supply, increasing the voltage at the plate of the damper tube. During scan through the left side of the picture, the current induced by the collapsing field on L10 passes through V12 and dissipates the voltage on C48. The voltage on the plate of V12 is therefore higher when the beam is at the left side of the picture than when it is at the center. This slows the beam travel through the left side of the picture, which corresponds to compression, and therefore cancels the inherent expansion. The optimum amount of cancellation can be achieved by adjusting the value of C48. An added bonus of the circuit is the 60 volt supply driven by the linearity winding. This is a conventional voltage doubler. The neon regulator stabilizes its output.

3-10.18.9 Transistor Horizontal Output Circuits

Figure 3-388 depicts the horizontal-deflection circuit of a popular transistorized monitor. Notice that there are many differences between this design and its vacuum-tube counterpart. Rather than substituting the transistor equivalents of the older vacuum-tube circuits, new types of circuits are employed. For example, the horizontal oscillator is a Hartley-type oscillator. The AFC circuit controls the frequency of the oscillator through a reactance amplifier. This more closely resembles the AFC system used in RF tuners than the horizontal-AFC system commonly employed in vacuum-tube monitors. Most significant of all is the horizontal-output circuit, which is transformer-coupled from its driver but is direct-coupled to the deflection yoke. Also, two transistors are employed in a series amplifier arrangement. Two damper diodes are used - one across each output transistor. Transistor Q2 is a conventional Hartley-type oscillator. As in vacuum-tube monitors, this is a free-running oscillator held in phase with the horizontal-sync pulses through an AFC circuit. The latter consists of phase-detector X1-X2 and reactance amplifier Q1. The horizontal-flyback pulse is shaped into a sawtooth waveform by R5 and C2 and is applied to the diodes as the AFC feedback signal. Both diodes are cut off except during the sync-pulse interval. When the oscillator is in proper phase with the incoming sync pulse, the latter coincides with the steep negative-going slope of the sawtooth, and both diodes conduct equally. Should the oscillator tend to drift downward in frequency, current in X2 will increase and in X1 will decrease during the sync-pulse interval, causing the DC voltage across C3 to move in the positive direction. Components R6, R1, and C1 form a voltage-dropping and decoupling network that reduces

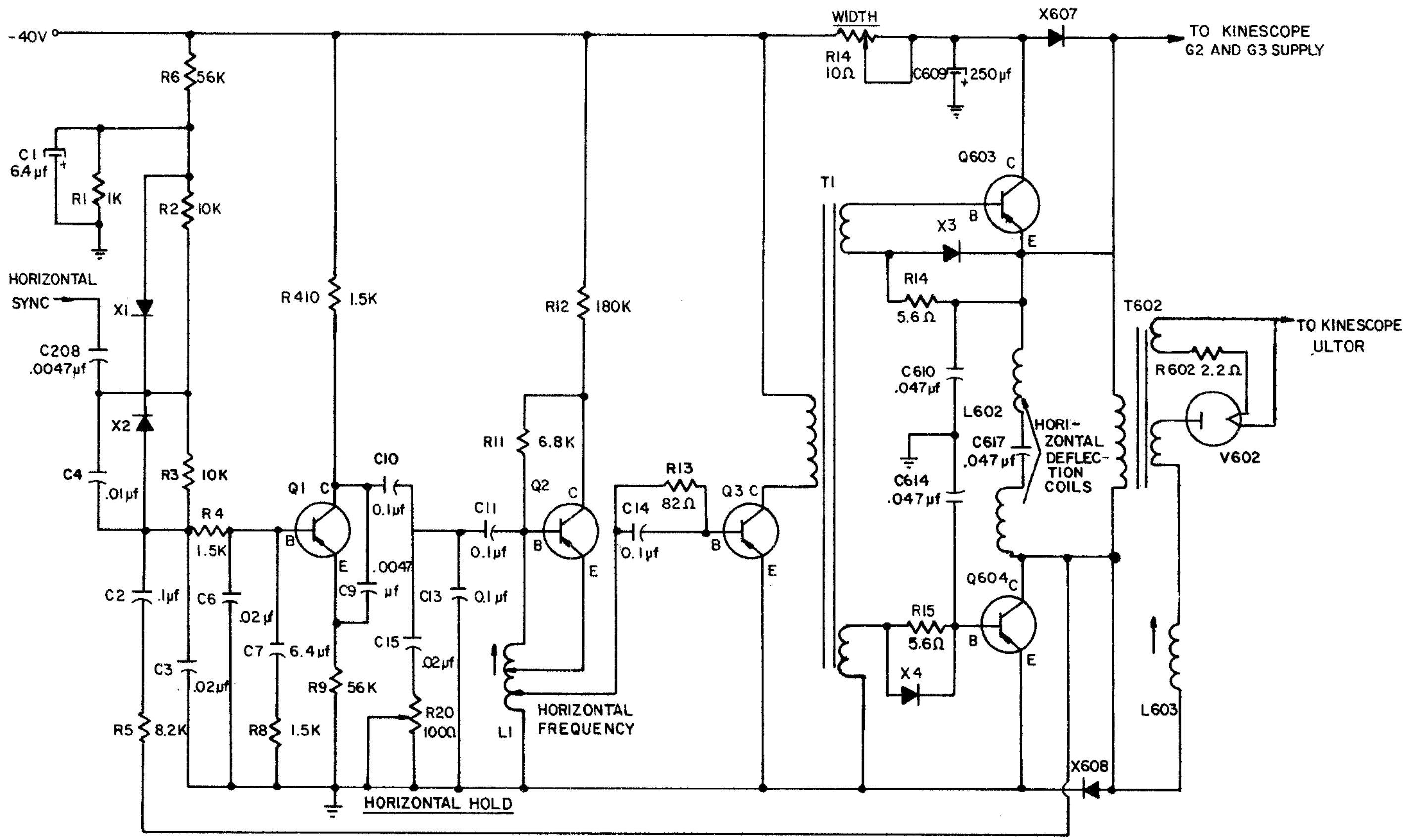


Figure 3-388. Horizontal Deflection Circuit

the -40 volts of the power supply to the operating range of the AFC circuit. The values of components R4, C6, R8, and C7 in the integrator network determine the AFC time constant. If a different time-constant should be required, the value of R8 may be changed. The output of the AFC discriminator is a DC voltage that determines the effective capacitance of reactance amplifier Q1. Transistor Q1 maintains control of the frequency of oscillation by acting as a capacitor in parallel with oscillator coil L1. The stage has two inputs: the DC voltage at the base, and the oscillator waveform at the collector. Capacitor C9 couples the oscillator waveform to the emitter of Q1. Since the current in a capacitor leads the voltage by 90 degrees, the signal voltage at the emitter of Q1 lags the voltage of the oscillator by 90 degrees. This signal, amplified by the transistor amplifier, appears at the collector and therefore across L1. The inductor sees the signal as a capacitance since its characteristics are capacitive. Consequently, the DC signal level at the base of Q1 - controlling its gain - controls the capacitive effect and therefore the frequency of oscillator Q2. Horizontal-hold controls the capacitive effect and therefore the frequency of oscillator Q2. Horizontal-hold control R20 and the slug adjustment of L1 set the nominal frequency of the oscillator. potentiometer R20 serves as the front-panel control. The output of the oscillator is coupled from a tap on coil L1. Transistor Q3 is a conventional overdriven amplifier that forms square waves from a sine-wave input. This is transformer-coupled to output amplifiers Q603 and Q604. The negative-going square-wave signal causes the transistors to conduct heavily, and the beam spot moves to the right. When the square-wave signal swings positive, the transistors are cut off, and the energy field of the yoke windings collapses, discharging rapidly into capacitors C610 and C614. These then discharge rapidly into the yoke. The polarity has reversed, and the beam is at the left side of the picture. The yoke now has a discharge path through X607 and X608 to C609. This decreases the resonant frequency of the circuit, and the inductance discharges more slowly, causing the beam to move to the right, but at a much slower rate than it moved to the left. This is the forward-scan rate. When the beam approaches the center of the picture, the transistors again conduct, causing the beam to continue to move to the right. The cycle then repeats. Capacitor C617 aids linearity by having a charging rate that complements the nonlinearity of the scan rate due to losses caused by unavoidable resistance in the circuit. The capacitor also prevents the direct current of the transistors from flowing through the yoke and thus causing decentering. The DC path for the transistors is through the primary of transformer T602. In this particular circuit, the high voltage for the kinescope ultor is derived from the horizontal-flyback pulse. It is stepped up in the conventional way (through a transformer) and rectified by V602. Inductance L603 tunes the secondary to a harmonic of the hroizontal frequency, which provides for more effective operation. An additional output is taken from the emitter of Q603. This is rectified and employed as the DC voltages on grids 2 and 3 of the picture tube.

3-10.18.10 Dynamic Focus

Since the beam of the kinescope must travel farther to reach the edge of the picture than to reach the center, a higher focus-voltage is required for optimum focus at the edge of the raster than at the center. In order to have optimum focus throughout the entire raster, the focus voltage should change as the beam passes across the screen from left to right, and from top to bottom. Dynamic (parabolic) focus waveforms are required; one at the horizontal rate, and one at the vertical rate. These are ordinarily mixed together and fed to the electostatic-focus electrode of the kinescope. Figure 3-389 shows the dynamic-focus circuit of a popular high-resolution television monitor. The AFC pulse from the horizontal-output transformer drives the horizontal parabolic-shaping network which consists of L1 and C2. The pulse excites the parallel-resonant circuit, causing a sine wave to appear at the grid of V1. The frequency of the sine wave is adjusted to be the same as the horizontal-scan rate. This is phased so that the most negative portion of the waveform coincides with the edges of the raster. The waveform is then inverted by amplifier V1, causing the voltage to be most positive at the edges of the picture. The waveform is not the ideal parabola, but is sufficiently close to one for the purpose. A waveform of near-parabolic shape and of the proper phase, occuring at the vertical rate, appears at the cathode of the vertical-output tube. This is because the cathode bypass capacitor has a significant amount of reactance at the 60-Hertz vertical-scan frequency. This waveform feeds the cathode of V1 which amplifies it without inversion. The two waveforms are combined at the plate of V1 and superimposed on the DC electrostatic-focus voltage of the kinescope. The amplitude of the combined waveforms is over 200 volts. Resistor R6 isolates the parabolic-focus circuit from the focus potentiometer. This allows the amplitude of the parabolic waveform to remain constant regardless of the DC focus setting. Resistance R16 protects circuit components in case of arc-over within the picture tube.

3-10.18.11 High-Voltage Supplies

In the preceding discussions the high-voltage supplies were described as being integrated with the horizontal-deflection circuits. This is the usual method of high-voltage generation and is found in most commercial television monitors. However, there are cases where the high-voltage supply operates independently of the deflection circuitry(except, perhaps, for a drive pulse for synchronization). In high-resolution monitors this may be necessary because kinescope voltages are quite high. In this case, a single tube may not be able to deliver both an unusually high amount of deflection power and the high-voltage power. Where a wide range of raster width control is required, integration is not feasible, since the high voltage would be dependent on raster width. Figure 3-390 is a schematic of the high-voltage circuits of a high-resolution monitor having a supply independent of the horizontal-deflection circuitry. High-voltage transformer T1 is similar in construction and appearance to the conventional horizontal-output transformer. Inductance L2 tunes to T1 to resonate at the horizontal-scan rate. Amplifier V1 is driven by a flyback pulse from the horizontal-output transformer. This is considered preferable to a free-running oscillator because it minimizes the effects of cross-talk. Resonance causes a sine wave to appear at the rectifiers. Tubes V2 and V3 form a conventional voltage-doubler circuit. Additional boost is accomplished by coupling the horizontal-flyback pulses from the

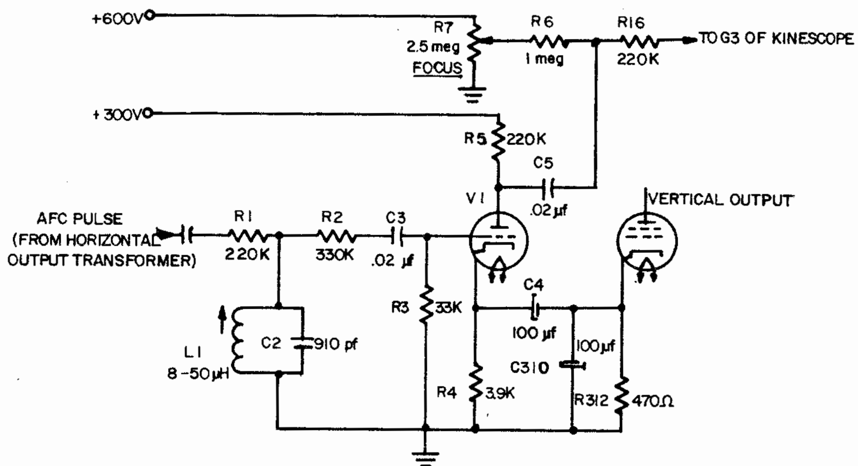


Figure 3-389. Dynamic Focus Circuit

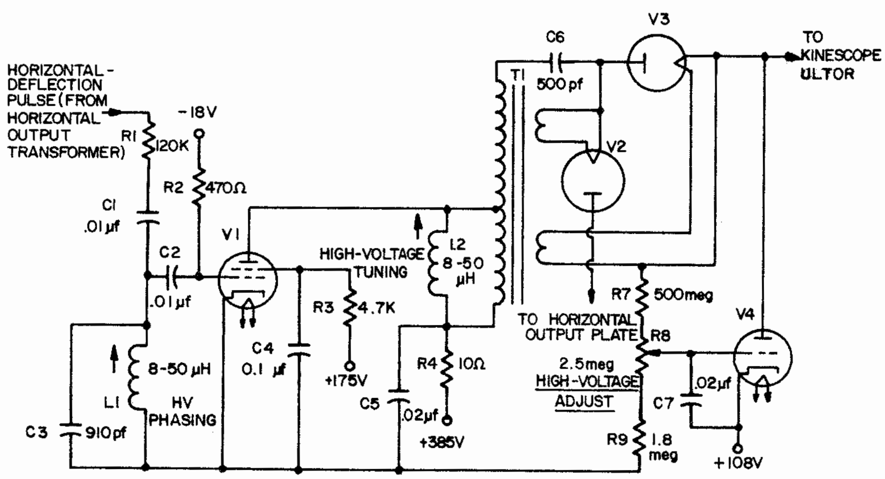


Figure 3-390. High Voltage Circuit

plate of the horizontal-output tube to the plate of V2. Inductor L1 allows phasing of the sine-wave output of V1 to add to the pulse at the plate of V2. L1 and L2 are normally adjusted for maximum high-voltage output. Shunt regulator V4 has its cathode referenced to the +108-volt supply. A voltage divider (R7, R8, and R9) across the output of the rectifiers provides a value suitable for the grid of the regulator. If the high-voltage rises, the grid voltage of V4 also rises, increasing the current in V4 and loading the supply. The increased loading causes a reduction in high voltage. The output voltage can be adjusted to the exact value desired by manipulating R8.

3-10.19 SUPPORTING EQUIPMENT

Complex television systems require auxiliary equipment in addition to the sync generators, cameras, and monitors described eariler. These bind the cameras and monitors into a single system, increasing flexibility and extending the usefulness of the system. Examples are:

1. Video switchers, so one of several cameras can feed a particular monitor.
2. Video distribution amplifiers that allow one camera to feed several cameras.
3. Pulse distribution amplifiers so a single sync generator can feed several cameras.
4. Video-insert amplifiers that add countdown or other identification to the picture.
5. AGC amplifiers to supplement gain controls, especially on signals originating outside the system.
6. Equalizing amplifiers for transmission of signals over long lengths of coaxial cable.
7. Stabilizing amplifiers to correct signals originating remotely. (These reshape distorted pulses and remove hum components.)
8. A pan-and-tilt unit that adds versatility to systems where remote control is necessary.

3-10.19.1 Video Distribution Amplifiers

Video distribution amplifiers allow the video signals to branch out in many directions so a single camera with one output can feed many picture monitors. One output of a camera may loop through five modern distribution amplifiers with a minimum of distortion and loading. Each of these has three or more outputs that may, in turn, feed five more from each output. By this means a single signal can feed many viewing points, or it can feed a number of inputs where loop-through connections are not feasible (switcher inputs and remote feeds). Distribution amplifiers also provide circuit isolation in event of a short circuit, cross talk, etc. Modern video distribution amplifiers are usually made up as plug-in modules so that several can be mounted on a single shelf or frame. Figure 3-391 shows the block diagram of a video distribution amplifier. The circuits are similar to those found in other video equipment. The input is high impedance - normally a loop-through connection. The model shown has a Darlington input stage, followed by two amplifier stages and a totem-pole output circuit. Sync mixing is optional and may be controlled remotely by using the sync-drop feature. The sync signal is fed into circuit similar to the video channel and is then mixed with the video. Gain is considerably less in the sync channel than the video channel because of the higher input level (4 volts) of the sync signal compared with the 0.7 to 1.0-volt video signal. Gain of the video channel is normally unity, but commercial units have adjustable controls to allow some gain or loss. Schematics of the individual circuits are described in Chapter 4.

3-10.19.2 Pulse Distribution Amplifiers

The pulse distribution amplifier allows sync, blanking, and drive signals from a single sync generator to branch out in many directions. A single horizontal-drive output, for example, may feed a hundred cameras and a hundred insert amplifiers without appreciable distortion when pulse distribution

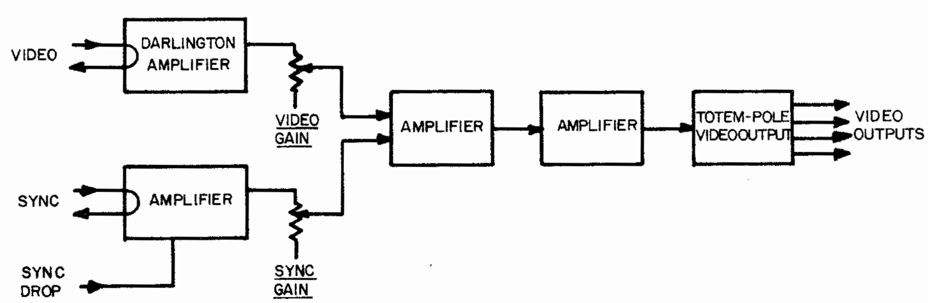


Figure 3-391. Video Distribution Amplifier

amplifiers are used. These differ from video distribution amplifiers in the levels of signals they carry, the use of a pulse-squaring method for reshaping distorted pulses, and the lack of the sync-mixing function. The pulse-squaring feature maintains a 4-volt output signal regardless of broad variations of the input signal. Figure 3-392 shows the block diagram of a typical distribution amplifier. It consists of a high-impedance input stage, a gain stage, a Schmitt trigger for pulse squaring, and an output stage feeding several jacks. Since a high-impedance input stage is used, a single source can feed several amplifiers, using loop-through connections. The Schmitt trigger reshapes pulses so they are rectangular. The output stage provides power amplification sufficient to feed several monitors or other equipment. This latter stage usually has an output impedance near zero, therefore, the 75-ohm impedance. The resistors also isolate the individual outputs in event of a short circuit or other difficulty in the distribution system. The input circuit of a pulse distribution amplifier is normally an emitter follower or a Darlington pair.

3-10.19.2.1 Schmitt Trigger

The Schmitt trigger, Figure 3-393 used almost universally for pulse squaring, resembles the emitter-coupled multivibrator, but has direct coupling throughout. When there is no input, Q1 is cut off and its collector voltage approaches the power-supply voltage. This is coupled to the base of transistor Q2 through R3. The base voltage of Q2 is approximately that established by voltage divider R2-R3-R5, and Q2 is conducting heavily through R1 and R4. The voltage drop across R1 maintains the emitter of Q1 at a negative potential, so that it is held at cutoff. A negative signal of sufficient amplitude, applied to the base of Q1, overcomes the reverse bias and causes Q1 to conduct. The decrease in potential at its collector is coupled to the base of Q2, therefore, current through Q2 decreases and lowers the potential across R1. This, in turn, decreases the reverse bias across Q1 and causes it to conduct more heavily. The regenerative action continues until Q1 is conducting at saturation and Q2 is cut off. The output voltage at the collector of Q2 is now approximately that of the power supply. This condition is stable until the end of the input pulse. The positive-going trailing edge of the pulse decreases the base potential of Q1 and increases the reverse bias. This causes the collector voltage to swing more negative and the emitter current to decrease. Simultaneously, the negative-going voltage at the collector of Q1 is coupled to the base of Q2 and drives it more negative. The decreasing voltage across R1 reduces the reverse bias of Q2, and it starts to conduct again, cutting off Q1 and returning the circuit to its original state. A small capacitor, C1, is often added to shorten the rise time of the output signal.

3-10.19.2.2 Output Stages

Output stages of pulse distribution amplifiers are often similar to those of camera-control units and video distribution amplifiers. Because of the multiple outputs required, a series amplifier such as the totem pole is usually employed. The complementary circuit shown in Figure 3-394, may also be used, but has not been discussed previously. The circuit Q1-Q2 is two emitter followers, each having the other transistor for an emitter load. Functionally, the circuit Q3-Q4 is the same as that of Q1-Q2. Transistor Q1 is npn, Q2 is pnp. A negative-going signal forward biases Q2 causing it to increase conduction while causing Q1 to decrease. Conversely, a positive-going signal forward-biases Q1, causing it to increase conduction while tending to cut off Q2. The result is a powerful voltage swing at the emitters of both transistors. This is direct coupled to a similar circuit, Q3-Q4, which supplies additional power gain. Capacitor C1 isolates the direct current at the emitters from external circuitry. Resistors R5 through R8 increase the impedance of the circuit to about 75

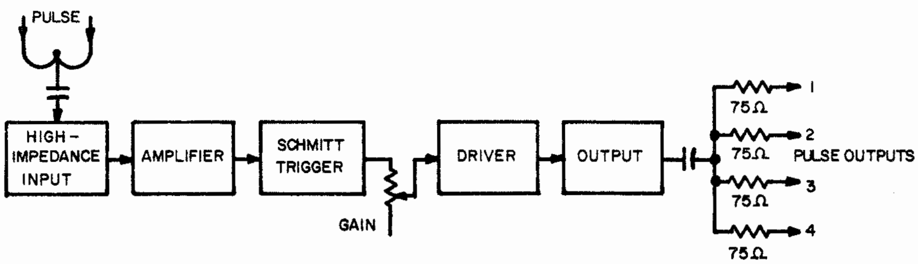


Figure 3-392. Pulse Distribution Amplifier

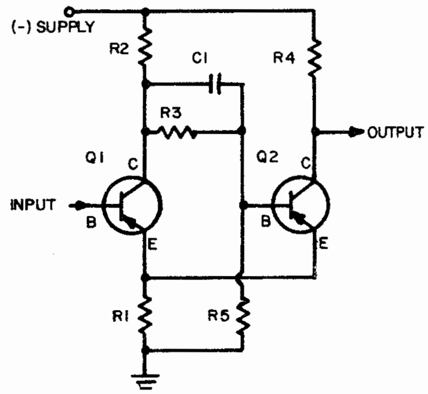


Figure 3-393. Schmitt Trigger

ohms; without these it would be nearly zero. These resistors also isolate the individual outputs from one another. In a variation of this circuit the collectors of Q1 and Q3 are connected to a +8.5-volt supply, with the collectors of Q2 and Q4 connected to a -8.5-volt supply. This sets the emitters of Q3 and Q4 at ground potential in their quiescent state; coupling capacitor C1 can thus be eliminated.

3-10.19.3 Video-Insert Amplifiers

The video-insert amplifier, or switching amplifier, is a version of the special effects generator used by broadcasters for title insertion, split screens, and montages. Industrial and scientific users find this a versatile device for insertion of countdown information, picture identification, and timing information. During tests of equipment a split-screen effect allows recording of two signals simultaneously on the same video tape to give positive timing correlation between two separate pictures. All signals to the insert amplifier must be timed to a common sync source, either directly or through locking circuitry. If they are not thus timed, inserted material will drift through the picture. The video-insert amplifier is a high speed electronic switch. Figure 3-395 shows its operation in highly simplified form. Only one signal - either video-A or video B - feeds the video output at a given instant. This is video-A when no signal is present at the key-signal input. When a positive-going signal of a predetermined amplitude is present at this input, video-B feeds the output. This switching is done electronically in a practical unit. Figure 3-396 shows how the insert amplifier may be used to provide a split-screen effect. Here a separate video signal feeds each input. A square-wave signal at the horizontal-scan frequency feeds the key-signal input. The output switches from video-A to video-B each time the square wave swings above a preset level midway across the picture in the example shown. Therefore, switching takes place in the center of each scanning line. The dividing line may be moved left or right by adjusting the symmetry of the square wave. Alternately, a sawtooth signal may feed the key-signal input, so the point at which switching takes place would be adjustable by using the sensitivity (key level) control. Figure 3-397 shows how timing or identification signals may be inserted into a picture. The signal of interest feeds the A channel. The video-B input is an all-white signal: normally a blanking signal of reduced amplitude. The key-signal input is the output of a camera viewing only a clock display. This signal operates the switching but does not appear in the output. Waveforms for a single scan (the line through the letters) is shown below the picture. Each time the key signal swings above a predetermined level, the output switches to video-B. This occurs several times during the line indicated in this example. An alternate hookup would be to omit the

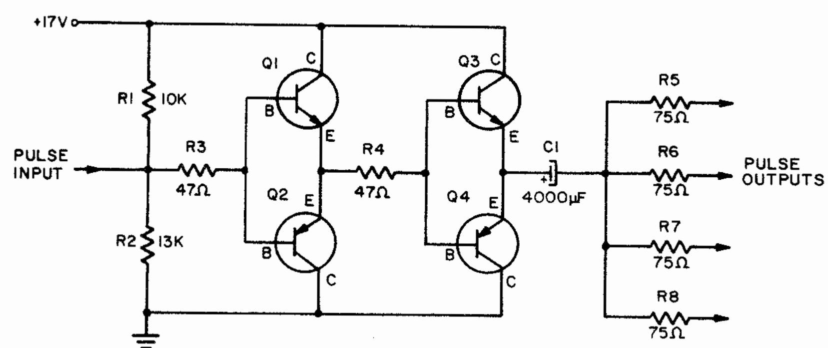


Figure 3-394. Output Circuit of Pulse Distribution Amplifier

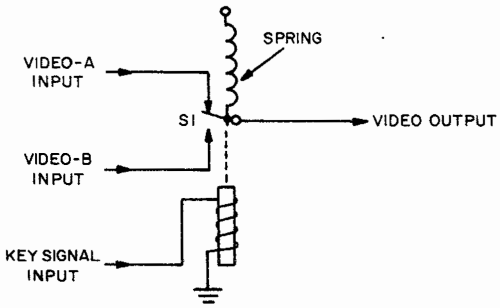


Figure 3-395. Insert Amplifier (Simplified)

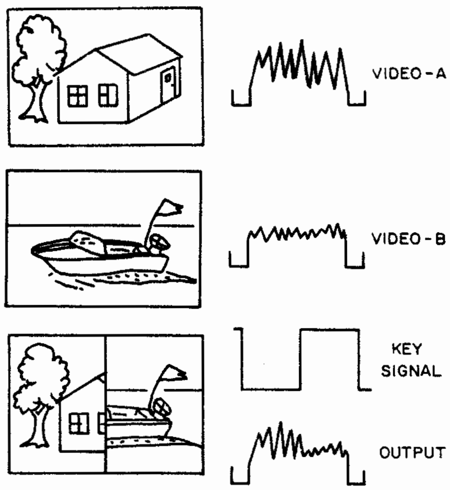


Figure 3-396. Split-screen Effect

white signal and feed the timing signal to both the key-signal and video-B channels. The method chosen is usually more satisfactory, since any shading in the signal from the camera viewing the clock display will not appear in the output. Figure 3-398 shows the block diagram of a practical video-insert amplifier. This unit has five inputs: video-A, video-B, key signal, horizontal drive, and blanking. There are usually three or four identical video outputs. The video-A input serves as the main video channel; usually the video-B signal is inserted into the picture. The key-signal channel supplies the switching information. The blanking signal may be either added to or subtracted from the blanking in the incoming video. The horizontal-drive signal is used for clamp keying. The video-A and video-B channels are identical. The clamp circuits in the video channels maintain equal black levels prior to switching. This keeps black portions of both channels alike regardless of picture content and prevents a shift in the brightness level of one channel in comparison to the other with changes in picture content. The signals are direct coupled from the clamp stages to the emitter follower which serves as a mixer. Each video channel has an electronic switch, but only one - either the video-A or the video-B switch - is turned on at a given time. This in turn, allows only one of the two input signals to reach the emitter-follower mixer at any given instant. The input section of the key-signal channel is similar to that of the video channels. The clamp stage maintains a constant keying level, regardless of changes in the duty cycle of the key signal. This signal is direct-coupled from the clamp stage to the Schmitt-trigger stage. White signals above a level preset by the key-level control activate the Schmitt trigger. Outputs of opposite polarities are used to drive the two video switches so that only one can be on at a time. The video signals pass from the emitter follower through a conventional blanking mixer to the output stages. A special-effects generator (Figure 3-399) includes a pattern generator

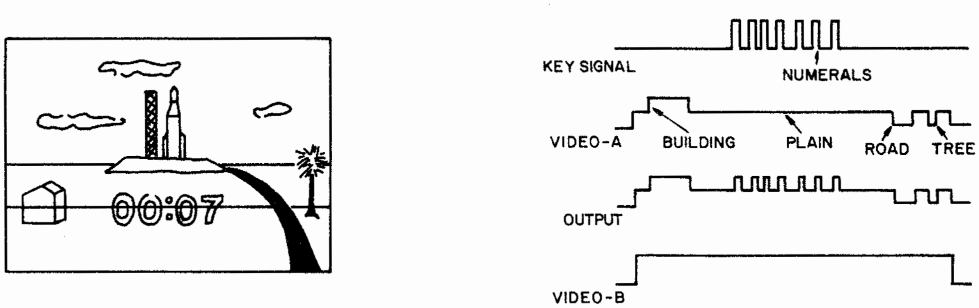


Figure 3-397. Insertion of Timing Information into a Picture (left)

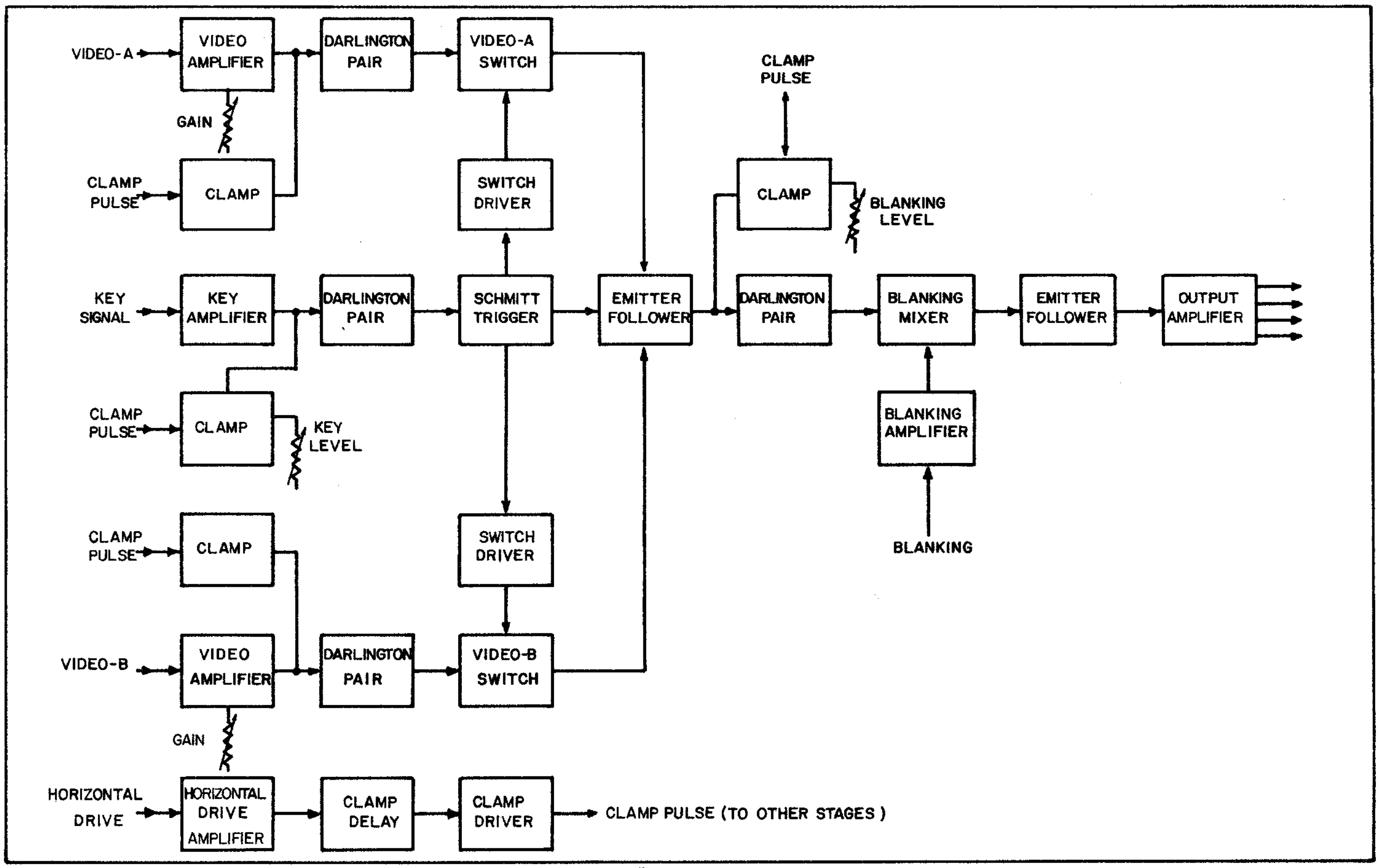


Figure 3-398. Video Insert Amplifier

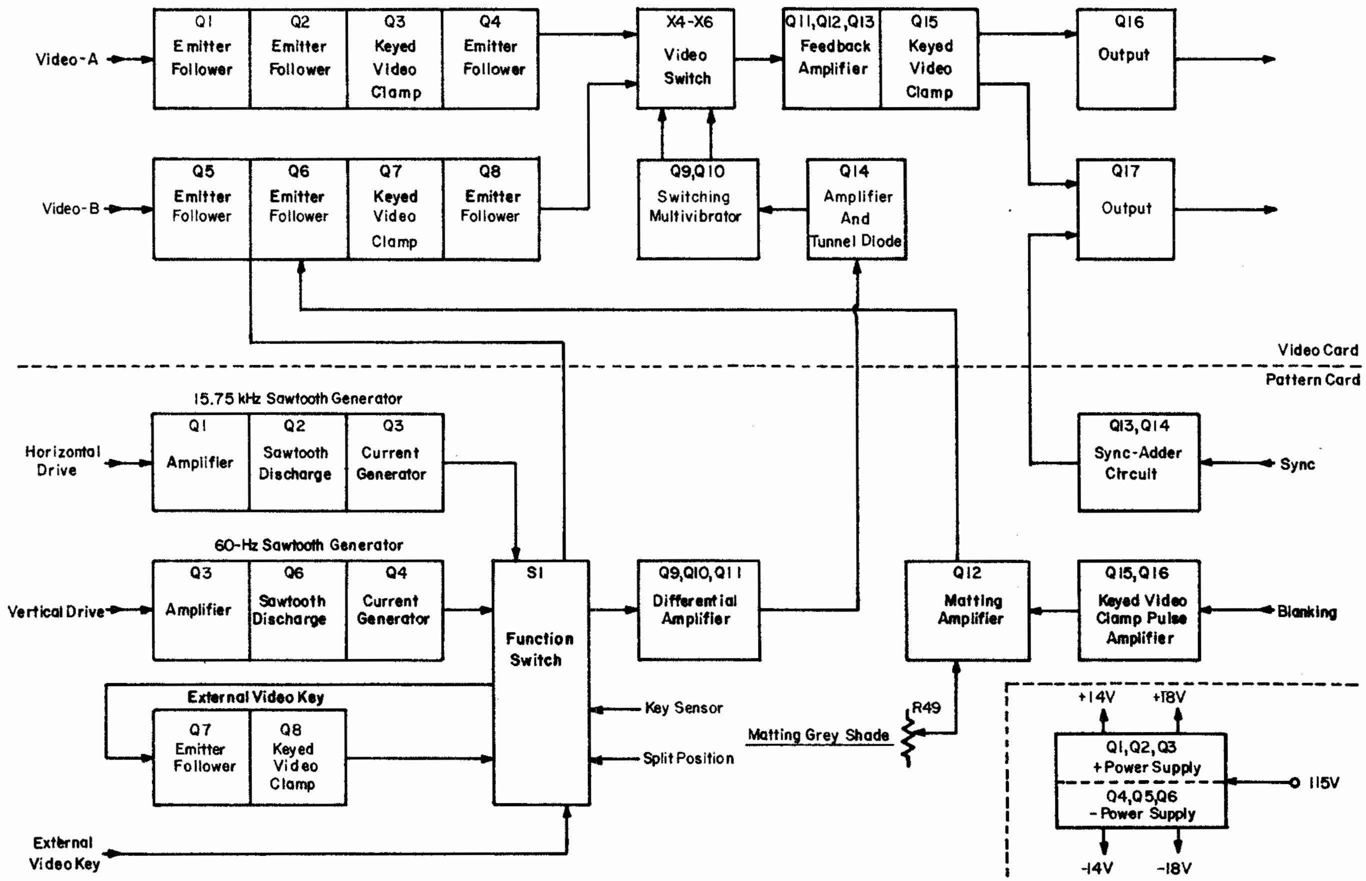


Figure 3-399. Special Effects Generator

for creating split-screen effects. The video card is functionally similar to the unit shown in Figure 3-398 but without the key channel. The pattern card contains the additional circuitry for split-screen effects, the key channel, and addition of sync. The 60-hertz and 15.75-kHz sawtooth channels are used in generating the waveforms necessary for the split-screen effect. The 60-hertz waveform is used to split the screen vertically; the 15.75-kHz sawtooth is used to split the screen horizontally. A combination of the two is used for a corner wipe. Switching is controlled by a DC voltage from the Split-Position control connected with function switch S1. When the sawtooth exceeds the DC level, Q9 conducts and triggers the video switch on the video card. The result is the output switches from video-A to video-B. Video may also be switched by a third video signal fed into the external video-key channel. This is useful in generating more complex patterns than those possible with the sawtooth generators. The "matting" function is used for insertion of letters or other information. When this mode of operation is selected at S1, the output of Q5 in the video-B channel is fed into the external-video-key channel which, in turn, operates the video switch. A blanking signal is fed during keying through Q15, Q16, and Q12 to Q6 in the video-B channel. The amplitude of this blanking signal, as set by potentiometer R49, determines the shade of the lettering from black to white. Video amplifiers, clamping and sawtooth generators, sync addition, and blanking insertion in insert amplifiers are similar to those of camera chains. Figure 3-400 shows the switching circuit of a popular model. Transistor Q307 is a conventional emitter follower. Transistor Q308 is a switching stage that is cut off by a negative signal at its base and is turned on by a positive signal. When Q308 is turned on, the video signal at the emitter of Q307 has a path to the video amplifier; when the transistor is cut off, the path is closed. Transistors Q316 and Q317 operate identically to Q307 and Q308 except that the keying signal is inverted; therefore, when Q317 is turned on, Q308 is turned off and vice versa. The result is that either video-A or the video-B channel is coupled to the video amplifier at a given instant, but never both at once. Figure 3-401 shows the video-switching circuit of another insert amplifier. The two video signals feed X4 and X16, respectively. Square waves of opposite polarity from the two sides of a multivibrator also feed the switching circuitry through the gate inputs. When the gate signal to the upper section swings positive, both X4 and X5 are forward-biased, and video passes from the video-A input to the video amplifier. Simultaneously, a negative gate pulse reverse-

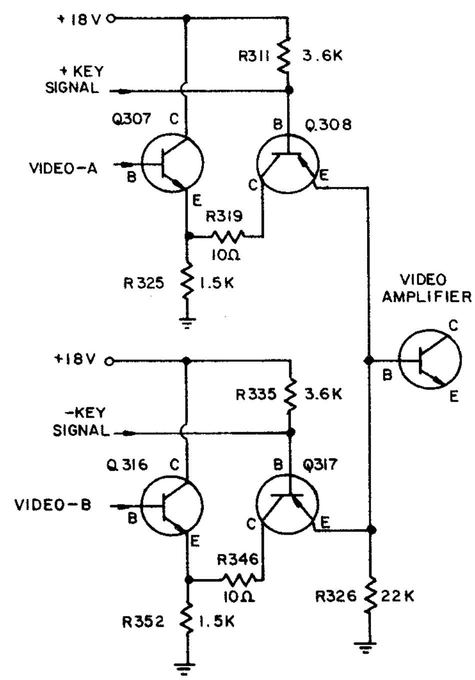


Figure 3-400. Switching Circuitry of Insert Amplifier (Simplified).

biases X16 and X7, and the signal from the video-B input is blocked. However, X8 is forward biased, therefore it shorts out any video which may leak through X16, since C10 offers a low impedance to ground for the video signal. When switching waveforms reverse polarity, X4 and X5 become reverse-biased, which blocks video-A, and X16 and X7 become forward-biased, which allows video-B to pass. Diode X6 then shorts video-A through C10 to ground.

3-10.19.4 Equalizing Amplifiers

Video signals transmitted over appreciable lengths of coaxial cable suffer from frequency-versus-phase distortion. Figure 3-402 shows losses in signal strength at various frequencies for some common types of 75-ohm cable. The equalizing amplifier has frequency and phase characteristics complementary to those of coaxial cable. Practical units are adjustable.

3-10.19.5 Stabilizing Amplifiers

A rather broad variety of units are called "stabilizing amplifiers." If a video signal becomes degraded by transmission, the basic function of the stabilizing amplifier is to restore the impaired signal to its original condition as nearly as possible. Stabilizing amplifiers are able to reshape distorted

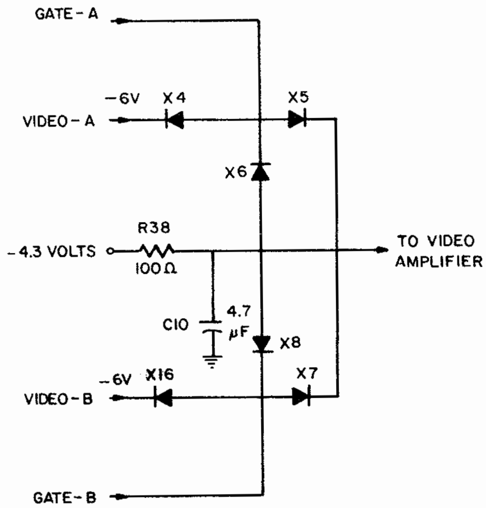


Figure 3-401. Switching Circuitry of Insert Amplifier (Simplified)

sync pulses by removing them from the video signal amplifying them in a separate channel, and then recombining them with the video signal. Stabilizing amplifiers may also include equalizing, sync-stripping, sync-adding, and automatic-gain-control functions. A simplified block diagram of a basic stabilizing amplifier appears in Figure 3-403. This unit amplifies and processes the sync signal separately from the video signal. The sync circuitry may be used in a variety of ways, depending on whether the video input is composite or noncomposite and whether sync is to be added or removed. A composite incoming signal may have its sync signal stripped, or it may have new sync substituted. A noncomposite signal may or may not have sync added. Finally, a separate sync output is available when the input signal is composite. Tracing the signal through this unit the input circuitry splits it in two directions, sync one way, and video the other. Next, the video signal is clamped to remove low-frequency disturbances. The video amplifier may contain correction circuitry consisting of AGE, equalization, chroma control, white clipping, blanking insertion, or gamma correction. Sync - either from the signal itself as it entered the unit or from an external source - is then added to the video signal. Figure 3-404 shows the block diagram of a commercial stabilizing amplifier that includes adjustable video-cable compensation and AGE. Video signals from the compensated loop-through input go from emitter follower to an L-pad attenuator consisting of two Raysistors. These are used in both the manual and automatic gain-control functions. Low-frequency disturbance are removed by the keyed clamp stage. The signal then passes through the equalizer stages, which may be adjusted to compensate for the high-frequency losses in as much as 1500 feet of RG-11/U cable. The following stage provides chroma control. Here a band of frequencies around 3.58 MHz are amplified separately from the rest of the video signal to allow for separate gain control (using Raysistors) of the chroma portions of color signals. This circuitry is normally bypassed when the

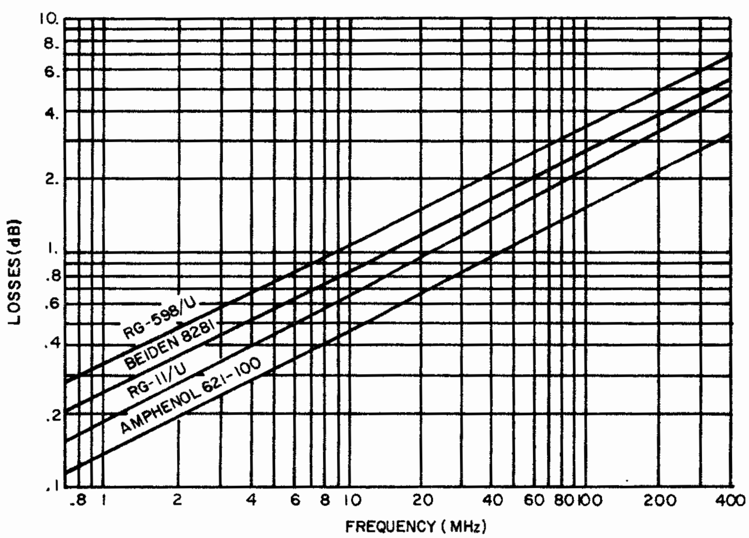


Figure 3-402. Frequency-versus-Amplitude Characteristics of Some Coaxial Cables

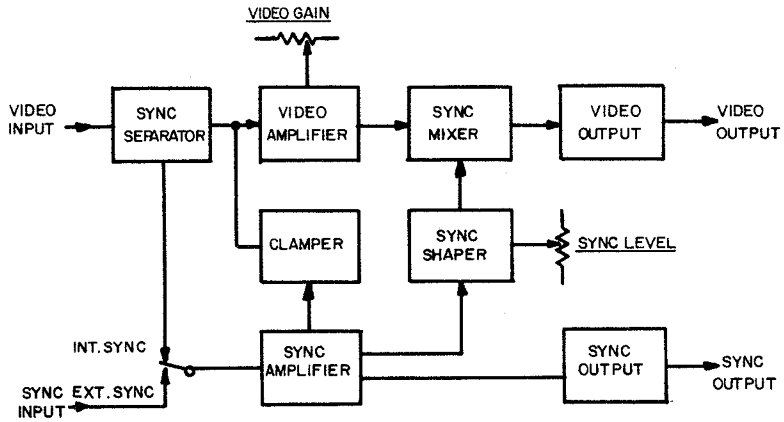


Figure 3-403. Stablizing Amplifier (Simplified)

stabilizing amplifier is used solely with black-and-white signals. Next, the signal is clamped in preparation for the white-clip, black-clip, AGE, and white-stretch functions. These stages provide corrections for distortions that may occur later in a television transmitter or, in some cases, may correct for distortions that may have occurred in preceding equipment. A feedback signal taken following the second clamp is used in the AGE and blanking-correction functions. Next, the signal passes through white-clip and sync-mixing circuitry. Sync is added at a level in excess of that normally required and is set to the desired level in the sync-clip stage. Finally, the signal passes through the white-stretch section to the output amplifier. This has four isolated outputs, much like a video distribution amplifier.

3-10.19.5.1 Circuits

Figure 3-405 shows the AGC section of the stabilizing amplifier described previously. Video is fed back from the second clamp through emitter follower Q1 to differential amplifier Q2-Q3. This stage provides gain and level discrimination such that only the white peaks of the video signal appear at the collector of Q3. The base voltage of Q3 is obtained from Q4, a current amplifier. The base of Q4 is fed from the AGC level control. The setting of this control determines the operating point of the AGC circuit and, therefore, the level at which video peaks are maintained by this circuit. The white peaks appearing at the collector of Q3 are direct-coupled through emitter follower Q14 to the base of peak detector Q5. The voltage at the emitter of Q5 is approximately proportional to the AGC error. This feeds Q6 through a low-pass filter controlling the reaction time of the AGC system. Transistor Q6 is a common-emitter DC amplifier that also inverts the AGC error-signal voltage. The gain of the stage, and therefore the gain of the AGC loop, is adjusted by the AGC damping control. This is set to accurately maintain the AGC level without hunting or overshoot. Buffer transistor Q7 feeds the AGC error voltage to a front-panel switch that selects either manual or automatic video gain. Transistors Q9 and Q8 are current amplifiers used to drive the lamps of Raysistors RY1 and RY2. The resistance of the resistor elements of the Raysistors depends on the brightness of their associated lamps. The signals feeding the two Raysistors are out of phase: when the resistance of RY1 increases, the resistance of RY2 decreases. These form an L-pad video attenuator. A variation of this circuit is also possible with a single Raysistor and a fixed resistor. The circuit with two Raysistors obviously has a wider range of control. Transistors Q10 through Q13 form an AGC-inhibit channel that senses whether the AGC peaks are below or above a preset threshold level. When video peaks are below this level, Q13 is switched into conduction. This cuts off Q5 and disables the AGC system. The gain of the amplifier is then established by the inhibit-unity-gain control, normally one. The inhibit-threshold control is usually set so Q13 switches when video peaks drop below 25 to 30 percent of the normal level. The inhibit channel serves a "squelch" function by preventing the AGC from operating at an excessive level when very low-amplitude signals, or no signals, are present. This prevents an annoying noise signal from appearing in the picture during fades to black in motion-picture pickups.

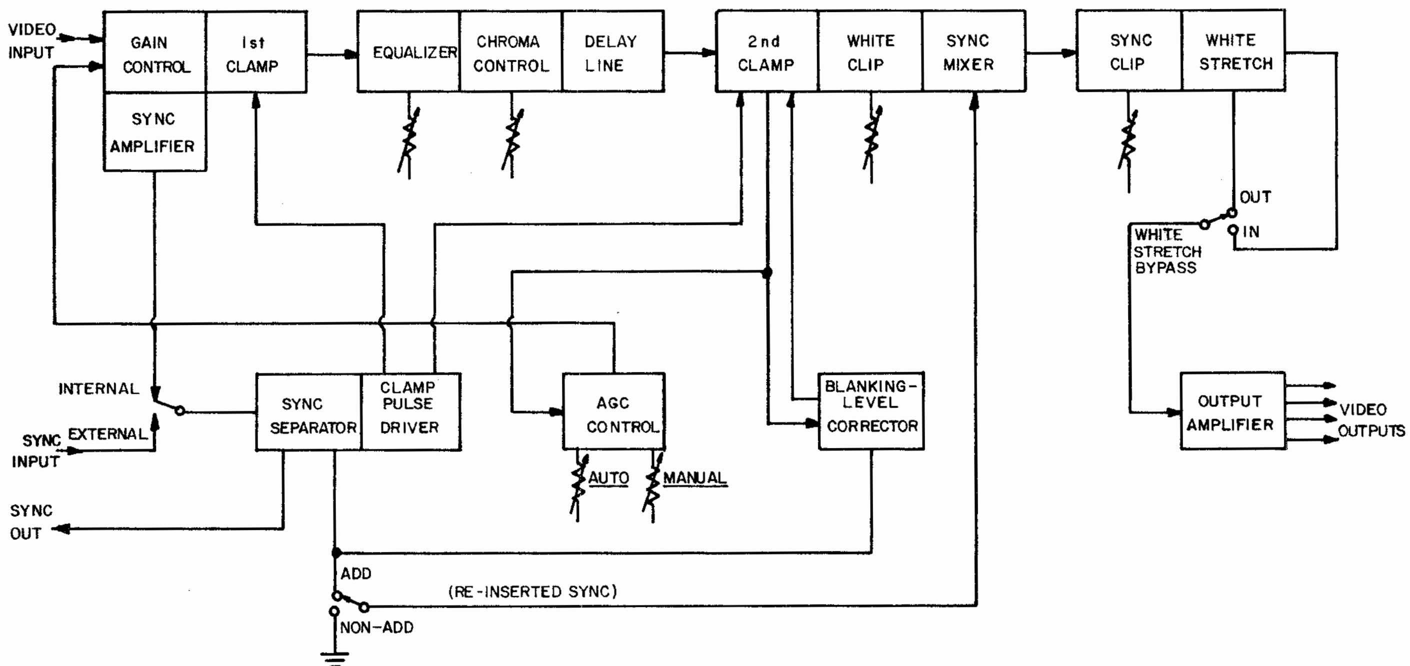


Figure 3-404. Stabilizing Amplifier

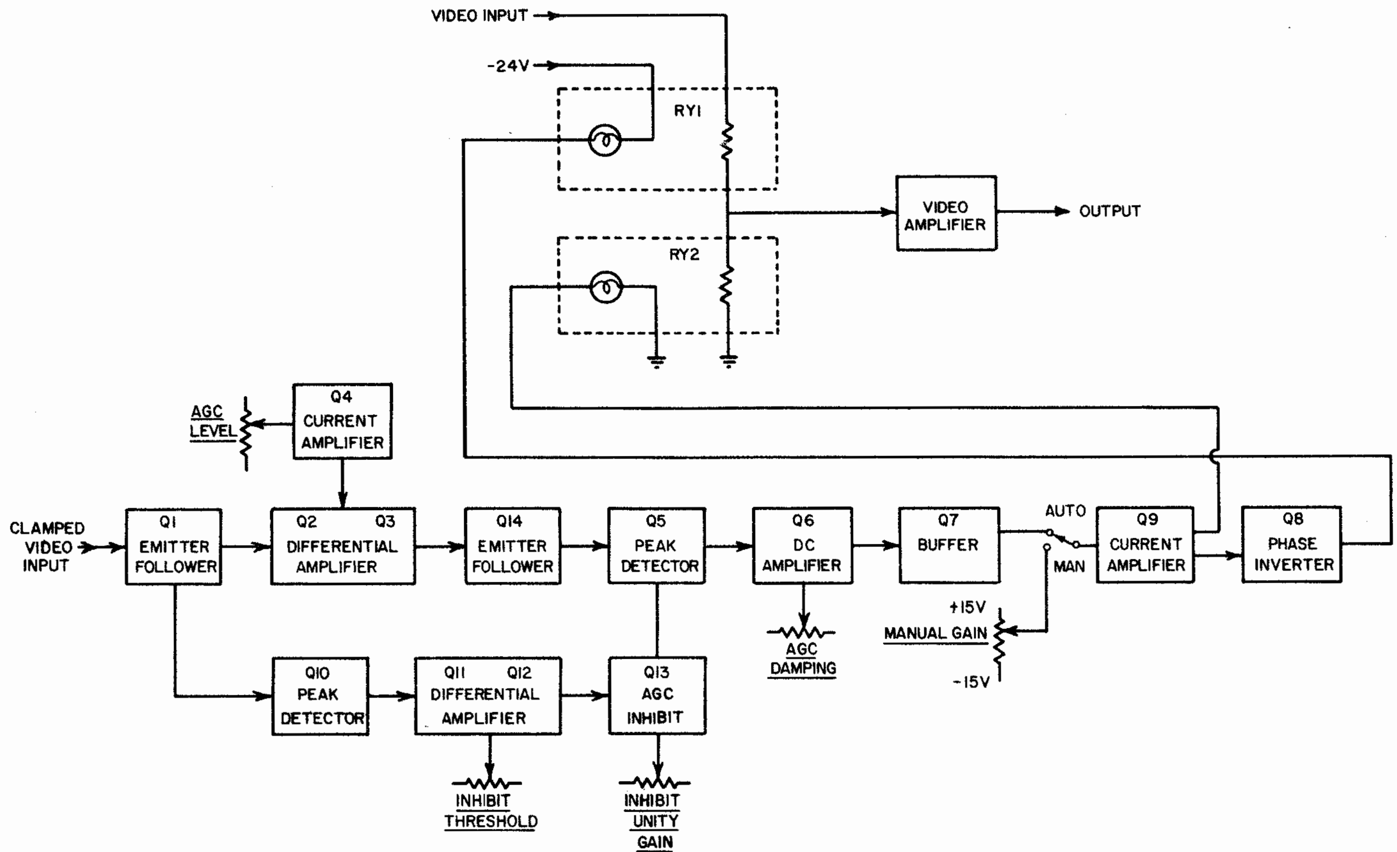


Figure 3-405. AGC Circuits of Stabilizing Amplifier

3-10.19.5.2 Equalization Circuits

Figure 3-406 shows the video-equalization circuits of the stabilizing amplifier. The incoming video signal is split into two paths: one through R16, the other through Q1 and Q2. The signal through R16 is not corrected; that through Q1 and Q2 has a rising frequency-response characteristic. When R17 is set to its maximum counterclockwise position, the signal output is principally that passing through R16 and is uncorrected. When R17 is in its maximum clockwise position, the signal is principally that passing through Q1 and Q2, and maximum correction results. The RC networks in the emitter circuit of Q1 give this stage its rising frequency-response characteristic. The feedback network C6-R11 is used principally for fine adjustment of phase response.

3-10.19.6 Video Switchers

Video switchers range in complexity from the simplest component in the television system to by far the most complex. It may be a simple push-button-operated mechanical switcher, or it may perform the switching function by means of conventional relays, crossbars, solid-state circuits, or reed relays. The simple mechanical switcher does not ordinarily have any electronic or control circuitry. This type is practical when there are a number of inputs but only a single output. Complex mechanical switchers require processing amplifiers in the input circuitry, output circuitry, or both. Solid-state switchers employ no moving parts at the points of actual switching. They are high speed units, produce almost no transients, and require no mechanical adjustments. Solid-state electronic switching is adaptable to both simple and complex switching systems. Compactness and low power consumption are other desirable features. Reed-relay switching has features of both mechanical and electronic systems. Signals pass through no electronic circuitry except input or output amplifiers. Switching is done mechanically, but the relays do not require adjustments. Because the relays usually have a latching feature, they draw current only during the acutal switching operation. Reed relays have exceptional reliability and lend themselves to either simple or complex switching systems. They are ordinarily mounted on printed-circuit cards since the overall physical size of a reed switcher is comparable to that of a solid-state switcher.

3-10.19.6.1 Single-Output Mechanical Switchers

Single-output mechanical switchers do not require amplifiers or other electronic circuitry, as shown in Figure 3-407. The switcher is simply a number of mechanically interlocked switches. The selected input passes through the switcher to its destination, where it is terminated in 75 ohms. All other inputs are connected to individual 75-ohm resistors in the switches. Therefore, loop-through inputs are not practical at the switcher inputs. This mechanical switcher is useful where only a single output is needed. A common application is the switching of a master monitor or waveform monitor to any of several key points in a complex television system. The auxiliary section (Figure 3-407) may be used when simultaneous switching of tally lights or audio signals is required.

3-10.19.6.2 Complex Mechanical Switchers

Complex mechanical-switching systems are usually represented by a group of grid lines (matrix),

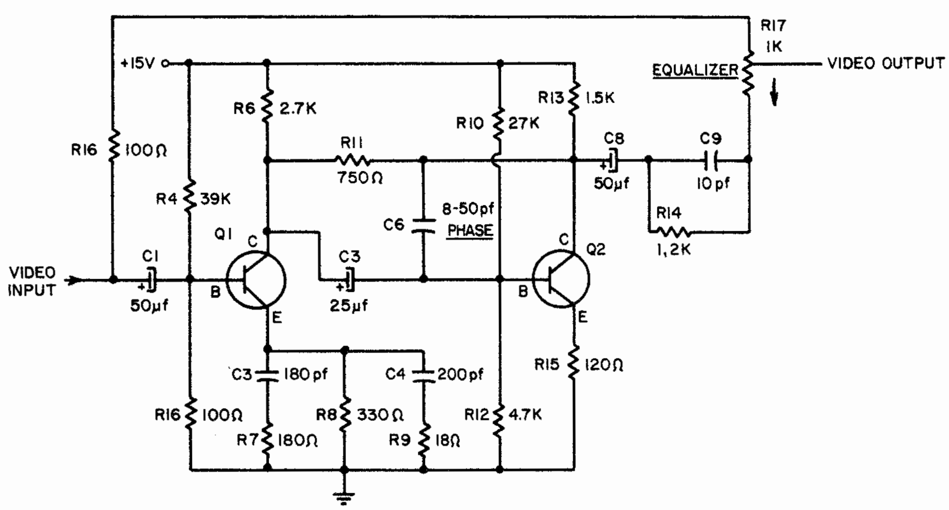


Figure 3-406. Video Equalization Circuits

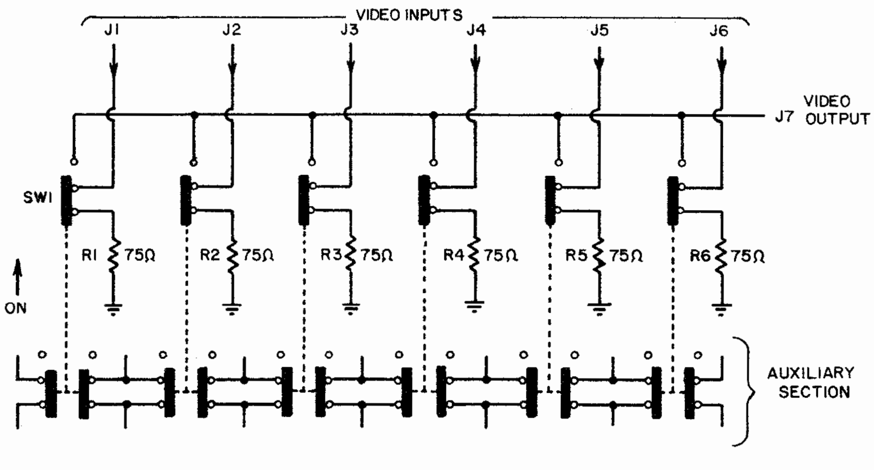


Figure 3-407. Video Switcher Schematic

as shown in Figure 3-408. Switching takes place at the junctions of the matrix (crosspoints). These may actually be coaxial relays, reed relays, or the crosspoints of a mechanical-crossbar module. A switcher like this can function without electronics as long as no onput is connected to more than one output at a time, since the demand for such a switcher is quite modest, and is not often encountered. Complete versatility in the switching matrix is possible through the addition of video amplifiers at the inputs, at the outputs, or both. Where the crosspoints are mechancial and the amplifiers are in the outputs, the inputs terminate in 75-ohm load resistors, and the output amplifiers have high-impedance inputs. The effect of adding one or all across the 75-ohm load resistor is negligible. When the amplifiers are in the input circuitry, they have multiple outputs and supply a separate isolated 75-ohm signal to each of the crosspoints. A single-pole double-throw switch at the crosspoint can switch the signal to the output or to the 75-ohm load resistor. Mechanical switchers sometimes include amplifiers in both the input and output channels. When this is done, the input amplifiers usually produce a low-impedance balanced signal, thereby minimizing crosstalk. The output amplifiers reconvert the signals to the regular 75-ohm unbalanced format.

3-10.19.6.3 Electronic Switchers

The advantages claimed for all-electronic switchers over mechanical switchers are:

1. Space savings.
2. Lower power consumption (no relays).

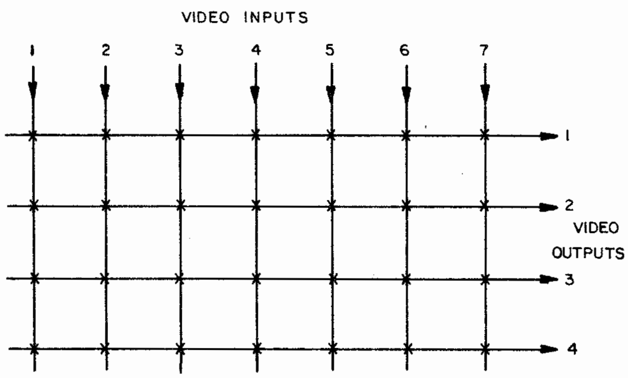


Figure 3-408. A Typical Switching Matrix

3. Simplicity (input and output amplifiers may be combined with switching circuitry.
4. Reliability (no relay wear and no adjustment required).

The foregoing advantages are realized to the fullest in the more complex switching systems. Switchers tailored to the particular installation are the rule. They can be built to accommodate any number of inputs and outputs. Often they are remote controlled, usually from a control panel located in an operator's console. Remote control signals are transmitted to where the actual switching takes place. Figure 3-409 shows the block diagram of a 20 by 5 switcher. Signals enter at the top and exit at the right. Any input may feed any or all outputs. The control circuitry, entering from the bottom, feeds DC signals to the

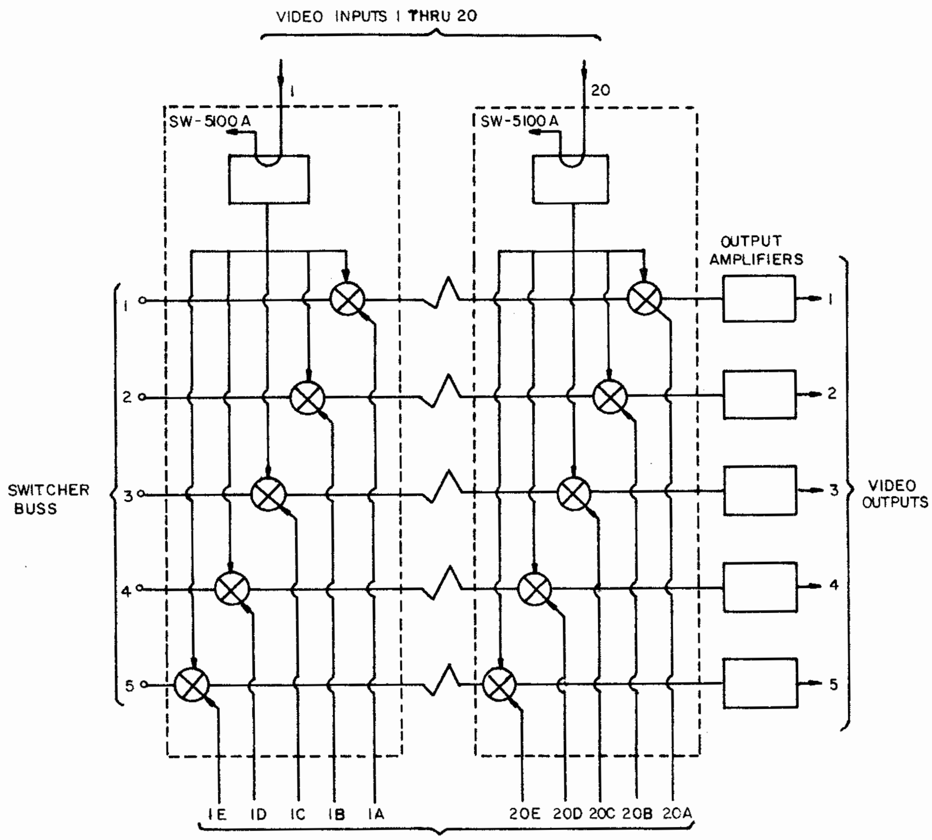


Figure 3-409. Switch Control

switching junctions from the remote-control panel. Two types of modules comprise the switcher shown: the switch module, and the ouput module. The circuitry of the latter is similar to that of a conventional video distribution amplifier. Figure 3-410 shows the block diagram of the switch module; this particular unit contains a total of five switching junctions. The schematic of one of the switching junctions is shown in Figure 3-411. Switching takes place in X1, Q4, and X2. When no control voltage is applied, X1 is reverse-biased, and no signal passes through it. Likewise, Q4 has reverse bias through R11 and is cut off. A +15-volt signal applied to the control input forward-biases X1, allowing the signal to pass. The positive voltage also biases Q4 into conductance. This produces a positive signal at its emitter, causing X2 to conduct. At this time the switching crosspoint functions as an ordinary video amplifier. Diode X2 also functions as a steering diode, isolating the many switch modules that are usually connected together at this point. Thus, the presence or absence of a control voltage determines whether the signal will be passed through.

3-10.20 OPTICS

An acquaintance with the principles of optics is necessary if television equipment is to be used to its best advantage. There are many terms that need to be defined and explained.

3-10.20.1 Optical Terms

The purpose of a lens in a television camera is to focus an image of a scene on the photosensitive surface of a pickup table. This is accomplished through refraction - the bending of a light ray when it passes obliquely from one medium to another in which its velocity differs. An example of refraction familiar to everyone is the apparent bend in a soda straw partially immersed in a glass of water. This is caused by a decrease in the velocity of light waves when they pass from air to water, and an increase to the original velocity when they emerge. Figure 3-412 shows a wave front as it passes from air, through a piece of glass, and back into air. Divisions a through z indicate equal increments of time. From a through g all parts of the wavefront are travelling at the same velocity. At point h part of the wavefront is in air and the rest of it is in glass. The

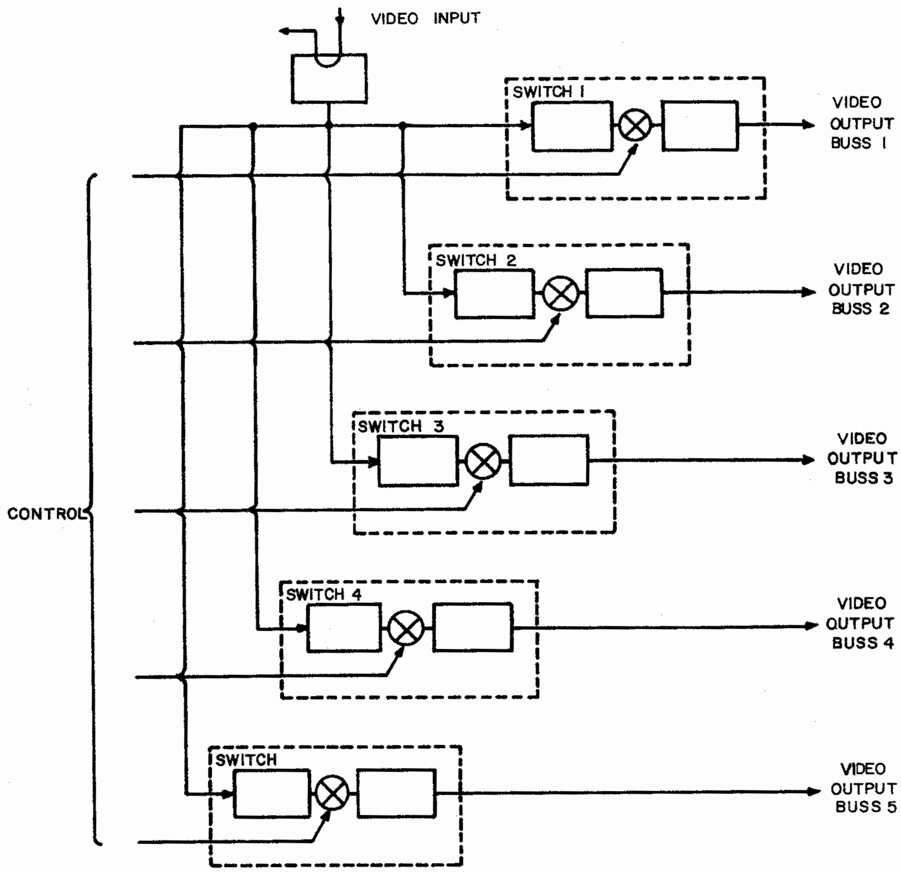


Figure 3-410. Electronic Video-switch Module

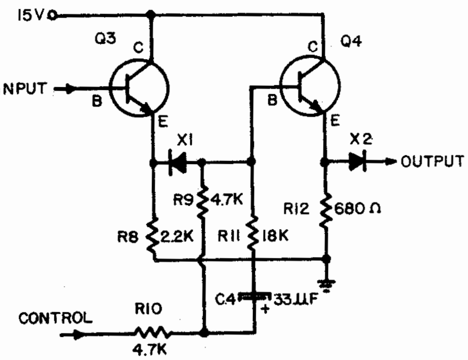


Figure 3-411. Schematic of Electronic Switching Junction

lower part now travels at a lower velocity (2/3 that in air) than does the upper part. The wavefront continues to enter the glass at points h, i, and j with the upper portion bending as did the lower. At points k through u the entire wavefront is in the glass and no bending occurs. Note, however, that it

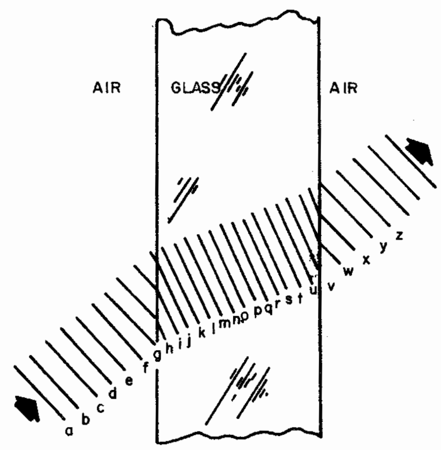


Figure 3-412. Refraction of Wavefront as it Passes Through Glass

is now travelling in a different direction. At point v the lower portion of the wavefront begins to emerge. Since it is travelling at a higher velocity than the upper portion, it is bent in the opposite direction. Bending continues as the wavefront passes points w and x. When it emerges at point y it is travelling in the original direction. Refraction gives prisms and lenses their optical properties. The section of glass in Figure 3-413 does not have parallel sides, so the light emerges

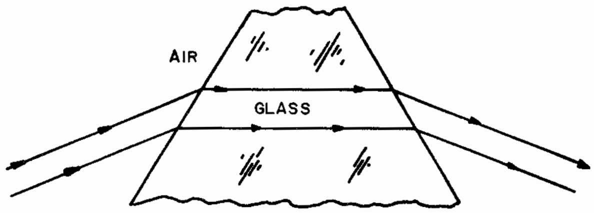


Figure 3-413. Refraction of Light by Glass

travelling in a different direction. The amount of refraction depends on the angle at which the light enters the medium. For example, a light ray arriving at the refracting medium exactly perpendicular to it is not refracted, but is only slowed as it enters. As the angle of incidence deviates from the perpendicular, the amount of refraction increases. The amount of bending may be calculated if the index of refraction is known. This is the ratio of the velocity of light in air to the velocity in the medium.

$$n = \frac{\sin i}{\sin r}$$

where

i is the angle of incidence,
r is the angle of refraction,
n is the index of refraction of the refracting medium.

A lens has one or both of its sides curved. When parallel light rays enter Figure 3-414 they are caused to converge at point p by the process of refraction. Here it is readily seen that the light rays entering the upper part of the lens are refracted downward, while those entering the lower part of the lens are refracted upward. Toward the center of the lens the angle of incidence is greater than near the edge - and the amount of refraction is correspondingly less. Figure 3-415 shows how a lens forms an image out of the reflected light rays from an object. All parts of the arrow are in the same plane. Rays from the center of the arrow (point b) at the left diverge and are converged by the lens to point b at the right. From point a the rays are converged to point d. Likewise, the rays from all other points are converged to corresponding points. Note that the image is reversed from the object. Figure 3-416 shows the basic types of simple lenses. Converging or positive lenses are suitable for forming images on the photosensitive surfaces of camera tubes. Diverging or negative lenses are used with converging lenses to aid in correcting aberrations. Aberrations are the convergence to more than one point of light rays originating at a common point, (see Figures 3-417 and 3-418) or the deviation of such rays from a single focus. Common types encountered are:

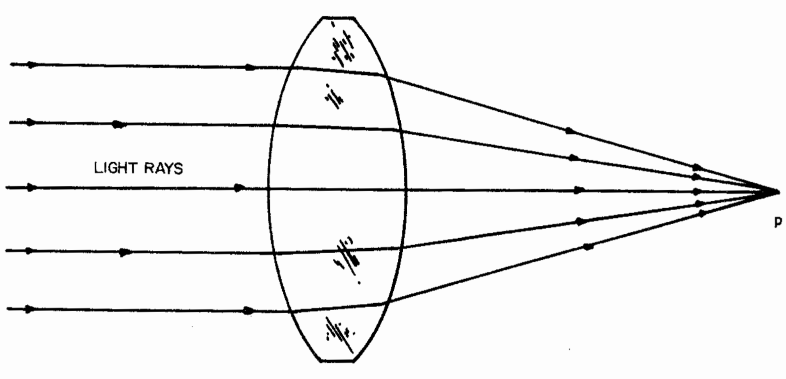


Figure 3-414. Refraction by a Lens.

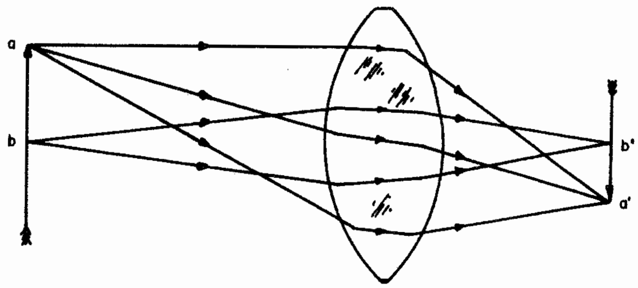


Figure 3-415. Formation of an Image by a Lens

1. Spherical aberration - the convergence of rays passing through the center of the lens to a point farther from the lens than rays passing near the edge (Figure 3-417). Spherical aberration can be minimized by:

a. Making the diameter of the lens very small.

b. Stopping the lens down by means of an iris.

c. Dividing the total bending of the rays equally between the two surfaces of the lens. The side of the lens toward the object must have a sharper curvature than the side toward the image.

d. Using a combination of positive and negative elements. Ideally, the distortion contributed by each of the two elements is equal and opposite.

2. Coma (Figure 3-418) is similar to spherical aberration in that rays passing through the center of the lens focus in a different plane from rays passing near the edge, but it affects rays which originate from points away from the lens axis. The image from a point source off the lens axis forms a comet-like image that gives the aberration its name. It can be reduced by the same methods which reduce spherical aberration.

3. Astigmatism (Figure 3-419) also affects rays which pass through the lens from a point away from the center axis of the lens.

a. Here, rays passing through a vertical diameter come to focus at a different distance than do those entering along a horizontal diameter. Rays first converge to form a line image.

b. farther back the rays form a circle, and still farther back the rays form another line image which is perpendicular to the first.

c. The point of best possible focus is where the circle is formed (between b and c).

4. Image distortion occurs when the shape of the image is different from that of the object. The two types are illustrated in Figure 3-420. Pincushion distortion results when points in the image which are off-center are magnified more than points which are on the center axis. Barrel distortion is the opposite of pincushion distortion.

5. Chromatism represents a difference in refraction of light rays having different colors. Each color has a different focus, as shown in Figure 3-421. In all simple lenses the shorter wavelengths are refracted more than are the longer wavelengths. Chromatism is corrected by using two or more elements of different shapes and different indexes of refraction. This is shown in Figure 3-422. Flint glass has a higher index of refraction than does crown glass.

Reduction of chromatism is called color correction. It is possible to achieve this for two or three well-spaced points of the visible spectrum. In usual terminology, a lens corrected at two wavelengths is said to be achromatic. A lens corrected for three wavelengths, as well as for spherical aberration and coma, is

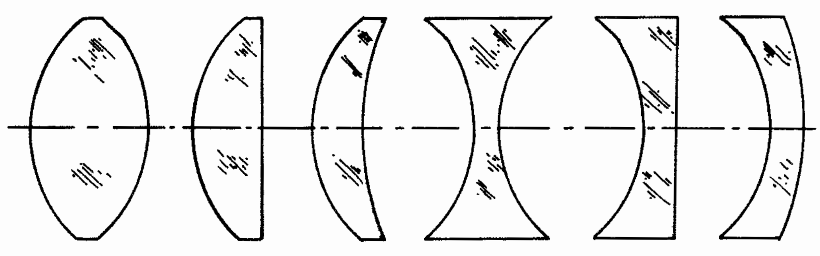


Figure 3-416. Lens Shapes

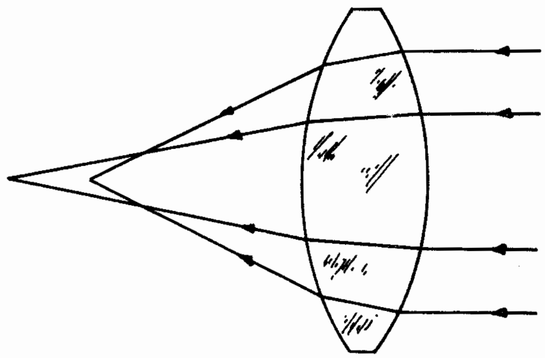


Figure 3-417. Spherical Aberration

said to be apochromatic. Most TV lenses are coated with one or more layers of transparent materials having indexes of refraction between those of air and glass. The purpose of the coatings is to reduce the amount of light reflected from the surfaces of the lens elements and thereby increase the efficiency of the lens. Each element surface which is not cemented to another element is coated. A surface of crown glass, if uncoated, will reflect about 5 percent of the light which strikes it. If this seems trifling, remember that a modern lens is comprised of a number of elements, and that the 5-percent figure applies to each surface which contacts air. Thus, if there are eight elements in a lens, the amount of light which it transmits is reduced almost 35 percent. In addition, the light that reflects from the inner and rear elements will reflect back toward the camera tube in the same proportion from the surfaces it strikes on its return path. In a four-element lens there are 28 beams of doubly reflected light toward the camera tube from a single point source of light. These form ghost images and reduce overall contrast of the image. The antireflection coating works very much like a quarter-wave matching section between two RF lines having different impedances. The coating is one-quarter wavelength thick at the center of the visible spectrum. Light which is reflected toward the source cancels, while that which reflects toward the image plane adds to incident light. Aberrations and reflections in a lens decrease the contrast of the image. Scattering of light rays, either by the lens elements or by the internal walls of the housing, will cause light to spill over to the darker areas. This makes light areas darker than they should be and dark areas lighter than they should be. When a good lens is compared to a poor one, the difference may be equivalent to an f-stop. Two basic characteristics of a lens are the focal length and the lens speed. The focal length (f) is the distance from the optical center of the lens to the point behind it where an object at an infinite distance away forms an in-focus image. For example, if an object an infinite distance away forms an in-focus image two inches behind the lens, focal length is two inches. The focal length fixes the image size for a given object and the angle of field covered. Longer focal-length lenses bring distant objects close up, while shorter focal-length lenses give a wider angle of view. Thus, when installed on a camera, a lens which has a high magnification actually gives a small image size for a given object, but covers a wide angle of view. The lens-speed designation is the ratio of focal length (f) to lens diameter (D), and is called the f-number (n_f). The ratio is given by:

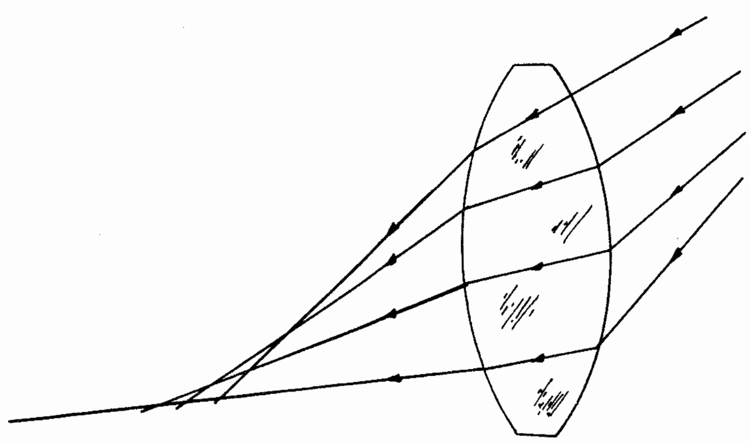


Figure 3-418. Coma

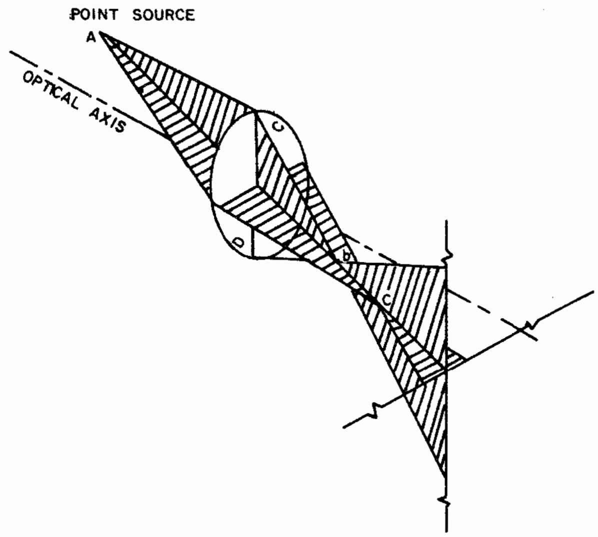
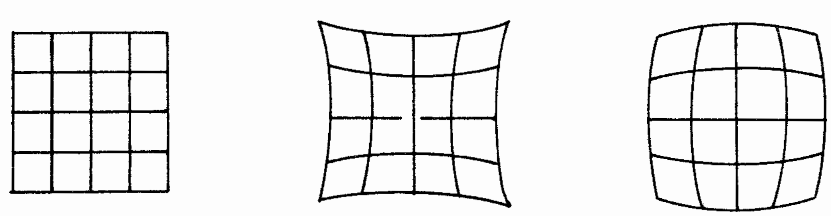


Figure 3-419. Astigmatism

$$n_f = \frac{f}{D} \text{ or } D = \frac{f}{n_f}$$

As the f-number increases, the diameter of the lens and the light it will pass decreases. Since the f-number is inversely proportional to the diameter of the lens, but the amount of light that a lens will pass is proportional to its area, then the latter is inversely proportional to the square of the f-number. Thus, an f/2 lens will pass four times as much light as will an f/4 lens, and 16 times as much light as an f/8 lens. The f/2 lens is said to be faster than the f/8 lens. Practical lenses are equipped with an iris so the effective diameter of the lens can be mechanically adjusted. Depth of field is the area that is plainly in focus ahead of and behind the subject. As the f-number increases, depth of field decreases. As focal length increases, depth of field decreases for a given f-number. The closer the object on which the lens is focused, the shorter the depth of field. The shorter the depth of field, the more critical the focusing. Thus, focusing is most critical for a zoom lens when it is set to its narrowest angle, at its widest aperture, and focused to its closest point. The hyperfocal distance for a given lens at a given f-number is the nearest point at which objects are in sharp focus when the lens is focused for infinity. If the same lens is focused instead to its hyperfocal distance, all points from one-half the hyperfocal distance to infinity are in sharp focus. Many fixed lenses have an engraved scale to indicate depth of field. High-quality photographic and television lenses are made of several elements since this is the most practical method of reducing aberrations. The various elements have different shapes and are made of different types of glass with indexes of refraction that vary with the type of glass. A lens suitable for a televison camera is equipped with an iris to change the effective diameter of the lens and compensate for varying light conditions. This is calibrated in terms of f-numbers. The f rating of a lens is the largest possible opening of the lens. The iris is ordinarily located at a nodal point and does not affect the field of view. The f numbers calibrated on most lenses are in terms of full stops. Each stop admits twice as much light as the one having the next higher number. For example an f/1.4 lens would be calibrated f/1.4, f/2, f/2.8, f/4, f/5.6, etc. Many lenses are also calibrated in terms of T-stops. This method takes into account the absorption and reflection of light by the lens. If one-half of the light is absorbed and reflected at an f-stop of f/2.0, the lens is rated T-2.8. The format of a lens describes the field of view for a lens of a given focal length. For example, a lens of a given focal length need supply an image of only .50 inch by .38 inch to a one-inch vidicon, but to a 1.5-inch vidicon it must supply an image of 0.8 inch by 0.6 inch. Thus, for a one-inch vidicon, whatever portion of the image falls outside the useful area is



(A) OBJECT. (B) PINCUSHION DISTORTION (C) BARREL DISTORTION

Figure 3-420. Image Distortion

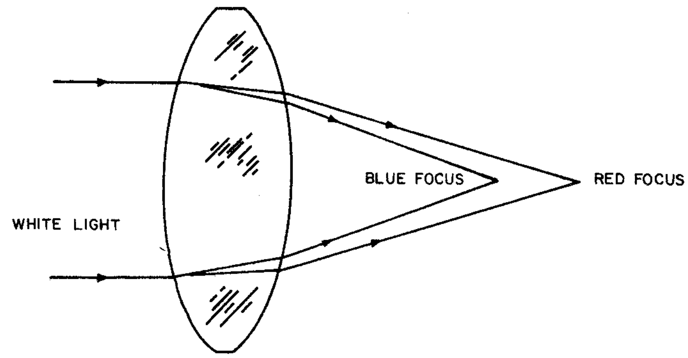


Figure 3-421. Chromatism

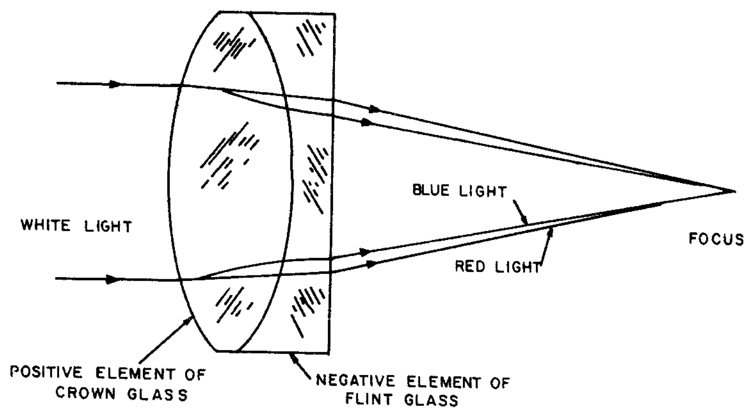


Figure 3-422. Correction of Chromatism

superfluous, and a lens designed for use with a one-inch vidicon is not likely to supply an image appreciably outside this area. This field of view is determined by the diameters of the rear elements of the lens. For a lens having a given focal length, the field of view, therefore, is not the same for camera tubes having different sizes of scanned areas. The field of view of a lens is a simple geometric proportion that is calculated by the formula shown and is illustrated in

$$\frac{f}{w} = \frac{D}{W}$$

where

- f is the focal length of the lens (inches or millimeters),
- w is the width of the image (same units of f),
- D is the distance from the lens to the subject, measured from the iris if it is critical (inches or millimeters),
- W is the width of view (same units as D).

The same proportion applies for calculation of the height of view. Table 3-27 gives the scanned areas of the commonest types of camera tubes for a 3:4 aspect ratio Figure 3-424 through 3-426 allow quick calculation of the fields of view of the three popular sizes of vidicons. One-inch vidicon cameras, like most 16-mm motion-picture cameras, are usually equipped with type-C lens mounts. In many cases, 16-mm

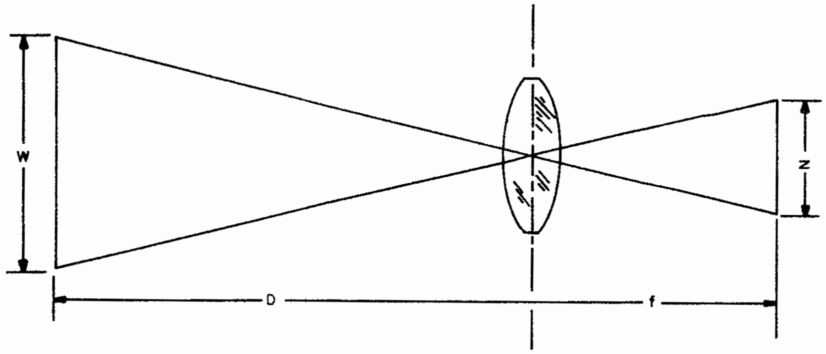


Figure 3-423. Calculation of Field of View

Table 3-27. Picture areas for Camera Tubes and Films

TUBE OR FILM TYPE	PICTURE SIZE			
	HEIGHT		WIDTH	
	IN.	MM	IN.	MM
1/2" vidicon	.18	4.57	.24	6.1
1" vidicon	.375	9.52	.5	12.7
1 1/4" plumbicon	.47	12.0	.63	16.0
1 1/2" vidicon	.6	15.2	.8	20.3
2" image orthicon	.72	18.2	.96	24.4
3" image orthicon	.96	24.4	1.28	32.5
4 1/2" image orthicon	.96	24.4	1.28	32.5
4 1/2" vidicon	1.5	38.1	2.0	50.8
Std. 8-mm motion picture	.145	3.68	.192	4.88
Super-8 motion picture	.166	4.22	.245	6.24
16-mm motion picture	.295	7.49	.404	10.26
35-mm motion picture	.631	16.03	.868	22.05
35-mm slide	.964	24.49	1.496	38.0

motion-picture lenses are satisfactory for use with television cameras. However, since the scanned area of a one-inch vidicon (.375" x .500") is larger than a 16-mm motion-picture frame (.295" x .404"), lenses designed for the latter cannot be relied upon. Use of these may result in a loss of image at the sides of the raster and in the corners. This is most common with wide-angle lenses. It is safest to specify a lens made for the type of camera to be used.

3-10.20.2 Fixed Lenses

Though fixed lenses lack the versatility of zoom lenses, they are often used for any of the following reasons:

1. Cost. Several fixed lenses may be purchased for the price of one zoom lens of equal performance.

2. Speed. Many fixed lenses are available with stops to f/.95. Zoom lenses are limited to about f/2.0 for one-inch vidicons.

3. Focal length. Zoom lenses are made to cover the normally-used range, say 15-150 mm for a 10:1 zoom range. Where focal length either longer or shorter than the foregoing is required, a fixed lens is usually necessary.

The variety of fixed lenses available is virtually limitless. Standard types have focal lengths ranging from 5.7 mm to over 300 mm, corresponding to horizontal viewing angles of from 96 to 2.2 degrees for a one-inch vidicon. Lenses with longer focal lengths are usually slower than others because of the large physical size and the amount of glass necessary. For example, a 300-mm lens with a speed of f/2 would be about a

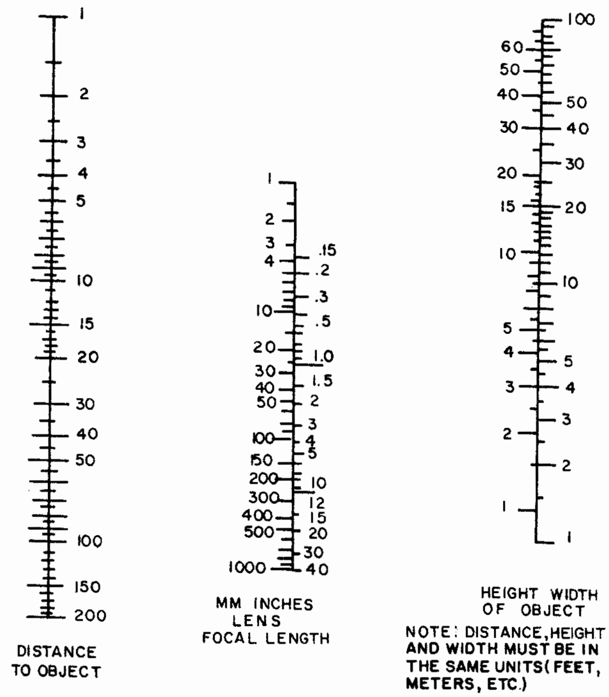


Figure 3-424. Field of View Nomograph for 1/2 inch Vidicons

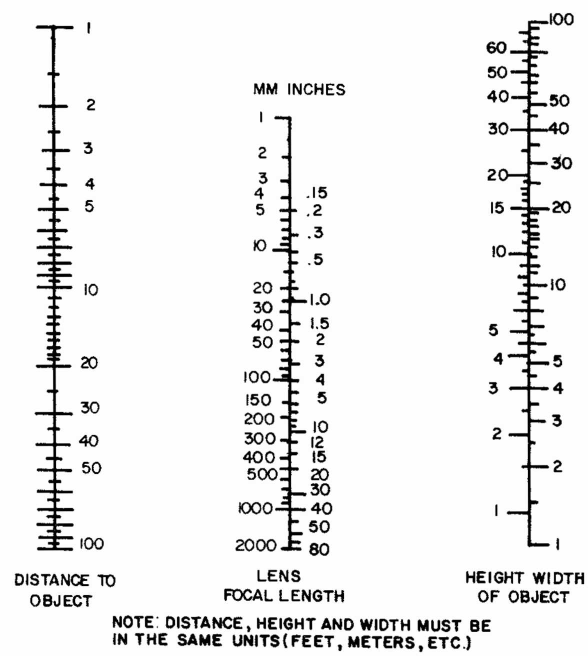


Figure 3-426. Field of View Nomograph for 1-1/2 inch Vidicons

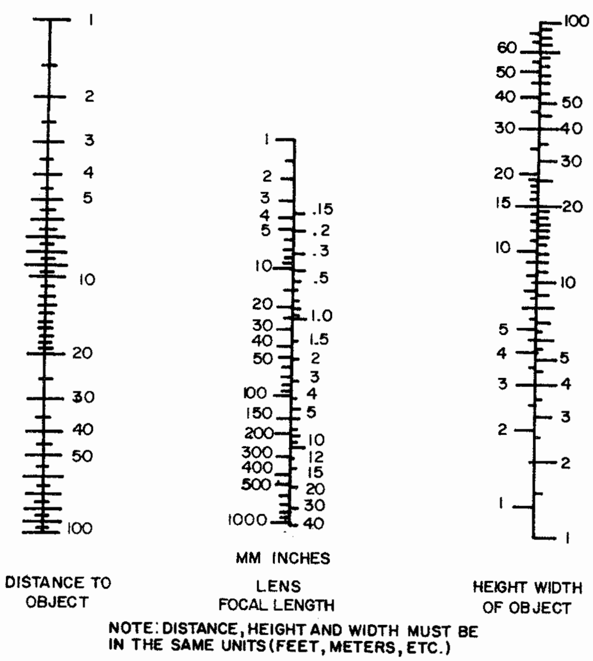


Figure 3-425. Field of View Nomograph for 1 inch Vidicons

foot long and about six inches in diameter, and certainly would require special support on a modern transistor camera. Fixed lenses ordinarily have two adjustments: focus and iris. The closest distance to which a lens will focus is limited only by the maximum distance to which the optical elements can be adjusted from the photosensitive surface. When an object to be viewed is closer than the normal range of focus of the lens, it is only necessary to insert an extender tube between the lens and its mount. This is simply a section of metal tubing with the appropriate thread on both ends. These are available in various lengths. Of course the extender must be removed if distant objects are to be viewed. The focus calibration of the lens is not valid with the extender tube installed.

3-10.20.3 Zoom Lenses

A zoom lens is one in which the focal length is made continuously variable by moving one or more of the elements along the axis of the lens. The image remains in focus in all positions. Zoom lenses are especially useful in remote-controlled installations with permanently mounted cameras. For example, a camera may be used for surveillance: when unusual action occurs in a particular area of the picture the zoom lens may be operated to allow close-up viewing. The higher price may well outweigh the lower price of a fixed lens since a single zoom lens may take the place of a number of fixed lenses. It also has complete flexibility in industrial explosion-proof, heat-resistant, or underwater housings. Optically, a zoom lens is a fixed lens with a variable-power telescope mounted in front of it. Many variations are possible and in use. Figure 3-427 shows the arrangement of the elements in a simplified zoom lens. Elements A, B, and C form the variable-power telescope; D is the regular fixed lens; B is the zoom element. In practical lenses, the functions of elements C and D are usually combined in a single group of elements, but are shown separately here to illustrate the principle. When the zoom element is at mid-range (Figure 3-427B) the overall image size is the same as if there were no zoom section; the light rays exit from element C on the same paths as they entered element A. Therefore, refraction in the zoom element just cancels that in the front and rear elements, and the overall focal length of the entire lens assembly remains the same as that of the fixed-lens portion by itself (element D). When the zoom element is moved forward (Figure 3-427) it more than cancels the refraction of the front element. Only the rays which pass through the center of the front element strike the rear element, and the effect is that of a short focal length. Note that the focal plane of the lens shifts as the lens is varied from the narrow-angle to the wide-angle setting. Therefore, this lens requires a different focus setting in the midrange position. Adjustment of elements A, C, D, or of the entire assembly will refocus the image. In practical lenses this is accomplished automatically as the zoom setting changes. Two types of compensation are used: mechanical, and optical. Figure 3-428 shows the arrangement of the elements in a modern mechanically-compensated zoom lens. Elements C and D, minus the front section of C, form an ordinary fixed lens. This contains the iris - located just as in regular fixed lenses. Elements A, B, and the front of C form the telescope section. The arrangement here is essentially symmetrical except for the smaller physical size of the rear elements. As B (the zoom element) is operated through its range, a cam automatically adjusts C through the range show lens; A is the regular focusing element. In a variation of this type of lens, A can be the mechanically compensated element, rather than C. Figure 3-429 shows an optically-compensated zoom lens. It is similar to Figure 3-427 except that the negative zoom element is split in two. One is located as in the mechanically-compensated lens; the other is located in front of the entire assembly. These move as a unit - with the spacing between them always the same. This arrangement keeps the focal plane relatively uniform. This combination is quite flexible. The negative elements may be fixed, with the positive elements performing the zoom function; or the whole zoom assembly

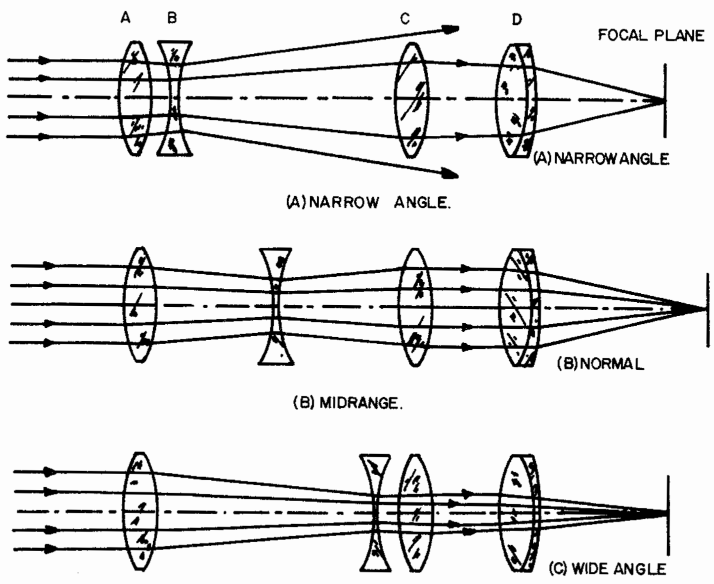


Figure 3-427. Optically Compensated Zoom Lens

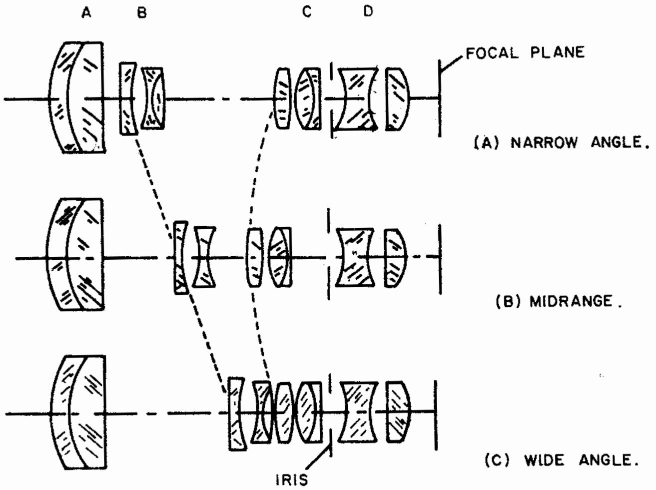


Figure 3-428. Mechanically Compensated Zoom Lens

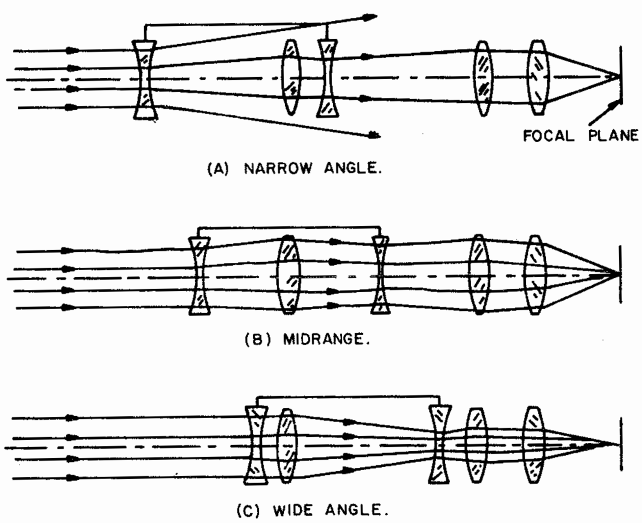


Figure 3-429. Optically Compensated Zoom Lens

may be reversed, with either the positive or negative elements moveable. Two types of accessory lenses are available for zoom lenses: the range extender and the close-up lens. The range extender changes the effective focal length of the lens. For example, a 2X extender installed on a 15-150-mm zoom lens changes its range to 30-300 mm. A disadvantage in the use of this extender is that the f-stop must be multiplied by two. If the lens normally has a maximum aperture of f/2.8, a 2X extender reduces it to f/5.6. It also sharply reduces the resolution of the lens. The range extender mounts between the lens and its regular mount. The close-up lens allows a zoom lens to focus on objects nearer than its regular range while maintaining zoom tracking. It also reduces, quite sharply, the maximum distance to which the lens will focus, but it has no appreciable effect on the amount of light reaching the camera tube. The close-up lens is a single-element attachment and fits on the front of the lens.

3-10.20.4 Lens Care

Lenses are built with incredible precision, yet they require only a minimum of care in order to last practically forever. Here are some everyday precautions:

1. Avoid contact with the glass surfaces. Oils from finger marks attack lens coatings. Metallic and other hard objects will scratch the coating and the glass itself.
2. Cover the glass surfaces with lens caps when the lens is not in use.

Field maintenance should be restricted to two categories:

1. Adjustment and lubrication of the electric motor assemblies used for remote controls.
2. External cleaning. Lens surfaces are highly polished and usually coated to improve optical performance. The very thin coating can be damaged by improper cleaning methods, reducing the performance of the lens.

Remove dust with a camel-hair brush and blower. Afterward, wipe gently with a clean cotton cloth. To remove grease and finger marks, first use a blower. Next, moisten lens tissue with alcohol or a lens-cleaning fluid. Remove marks with a gentle action, changing tissues frequently. Remove greasy deposits with benzene and then clean with alcohol or lens cleaner. Avoid an excessive amount of any solvent since it may flow into the lens and attack the black finish. Dust again and wipe with a clean cotton cloth. Persons without special training in the assembly of lenses, should not disassemble one unless the manufacturer specifically calls for it in an operation and maintenance manual. If so, then those directions must be followed carefully.

3-10.21 TELEVISION RECORDING

Through television recording it is possible to preserve the information displayed on the monitor for future viewing. For many installations a prerequisite of the installation is that an event be viewed again and again, often frame by frame, perhaps months later and miles away from the original televising.

3-10.21.1 Magnetic-tape Recording

The comparatively recent development of video tape recording offers many attractive advantages over film recording. Principal among these are:

1. Video tape requires external processing. The recording may be played as soon as the tape can be rewound.
2. Video tape may be reused when the recorded material is no longer of value.
3. The same machine can be used for both recording and playback.
4. Monitoring of recorded levels is very simple compared to kinescope recording.
5. Excellent quality sound may be recorded on the same tape as the video.
6. No optics are required in the recording process.

The foregoing advantages are often very attractive to industrial and scientific users, for whom delays encountered in processing can be expensive. In many cases a tape need be replayed only a few times or even not at all, where no significant data has been recorded; this is an ideal use of tape. Some information on the theory of magnetic-tape recording is in order. When a material that can be magnetized is placed in a magnetic field, the domains of the material become magnetized in accordance with the direction and density of the field. On a magnetic tape, the recording material is comprised of microscopic iron-oxide particles loosely bonded to a plastic surface. A number of turns of wire wound on a core of a highly permeable material forms the magnetizing device (the head). During recording and reproduction a gap in the core of the head contacts the tape coated with iron oxide so the magnetic circuit of the core is completely through the tape surface. If the current through the core winding varies at a given rate as the tape moves past the gap, a magnetic pattern, directly proportional to the coil current, is laid down on the tape as it leaves the gap (Figure 3-430). When the process is reversed and the magnetized tape passes the gap with no current flowing (Figure 3-431), the magnetic circuit is completed again. A voltage proportional to the change rate of the magnetic flux at the gap is induced into the coil. If both recording and reproducing processes are

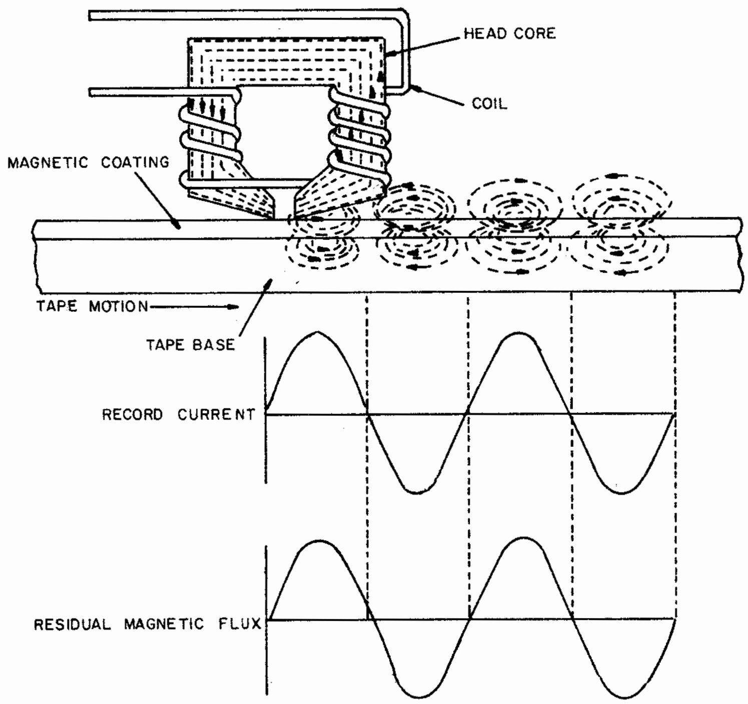


Figure 3-430. Magnetic Pattern During Record Mode

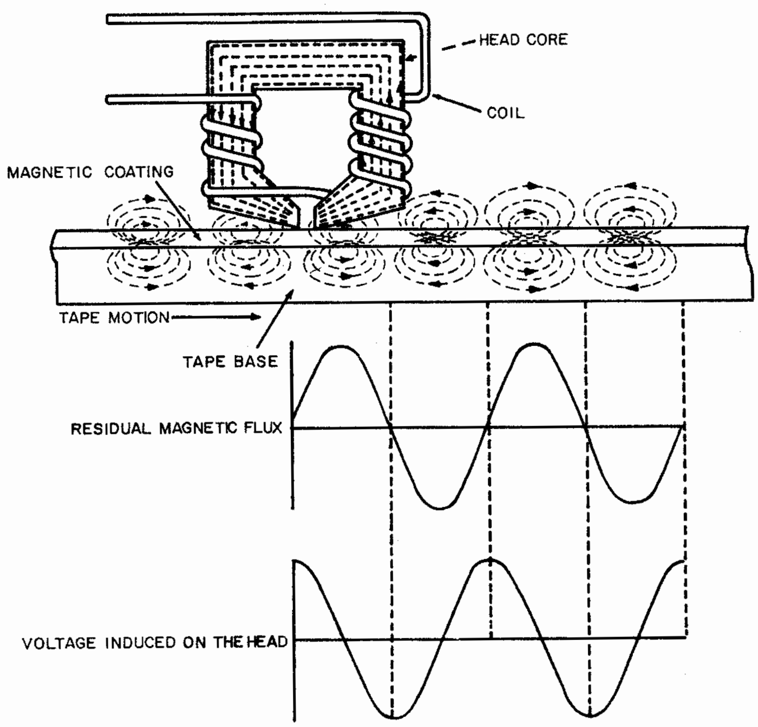


Figure 3-431. Magnetic Pattern During Reproduce Mode

performed at the same constant speed, the voltage induced during the reproduce-mode follows faithfully the current used during the record mode. The wavelength of the signal recorded on the tape is a function of tape velocity and signal frequency, as shown in the formula:

$$V = \lambda f$$

where

- λ is the signal wavelength on tape (in inches),
- V is the tape velocity (in inches per second),
- f is the signal frequency (in hertz).

This formula shows that, for a given tape velocity, the wavelength becomes progressively shorter as the frequency is increased. Conversely, as the tape speed is increased, the wavelength for any given frequency increases proportionately. The upper frequency limit in the record mode is determined mainly by the inductive reactance of the head coil and its distributed capacitance. As a practical matter, frequency limitations of magnetic-tape recordings are attributed almost entirely to the reproduce head. The voltage induced in the core coil during the reproduce operation is proportional to the rate of change of the magnetic flux at the gap. When the magnetic wavelength of the tape equals the width of the gap, the rate becomes zero, so there is no induced voltage (gap effect). Frequencies above this cutoff point cannot be reproduced dependably. The rate of change of the magnetic flux for a given tape speed varies with frequency. As the frequency is halved, the change rate is halved, and the induced voltage decreases by 6dB. If a recording is made, at constant current, covering all frequencies up to cutoff (for a given tape speed and a given reproduce head), the reproduce-head output will show a 6 dB per octave rolloff as the frequency drops. This rolloff is characteristic of all magnetic-recording systems and limits to about 10 the number of octaves that can be reproduced. In a 10-octave recording the signal-to-noise figure for the first octave will be 60 dB poorer than that for the tenth octave. Losses caused by tape thickness, head gap loss, etc. add progressively in the order shown in Figure 3-432 to produce a decrease in the high-frequency region from the ideal indicated by the dashed line to the resultant indictated by the lowest solid line.

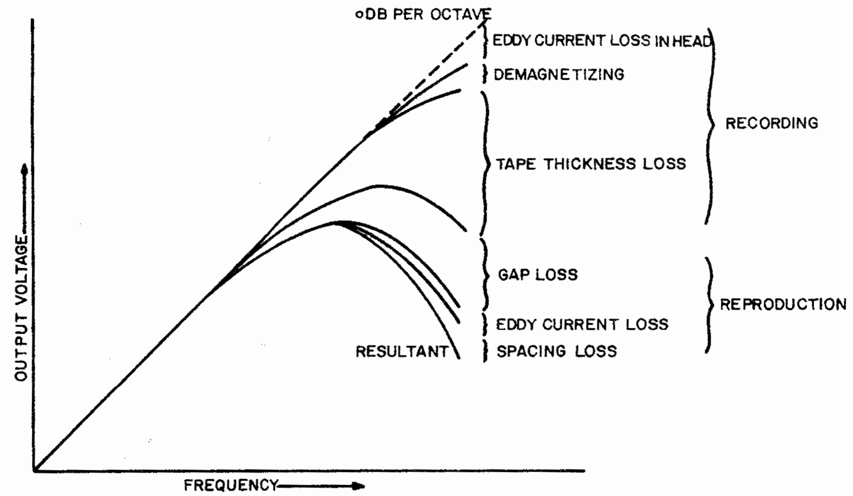


Figure 3-432. Induced Voltage versus Signal Frequency

3-10.21.2 Vertical Scan

The first commercial video tape recorders employed essentially vertical scanning. The excellent quality and other obvious advantages of these recorders over kinescope recording created ready acceptance in the broadcast field and in numerous industrial, scientific, and military applications where the high initial cost of the machines could be absorbed in either film savings or a reduction in processing time. Mechanically these machines consist of four video heads located on a headwheel assembly that rotates at high speed (14,400 rpm) perpendicularly to the motion of the tape (Figure 3-433). Each revolution of the headwheel records 4 tracks that represent a total of 64 television lines; an entire frame is recorded in 33 tracks. A control track recorded along the edge of the tape maintains synchronization between the recorded tracks, the rotating video heads, and the linear motion of the tape. A vacuum arrangement holds the tape in contact with the rotating heads. The video tracks are recorded essentially transversely on the tape. These machines normally have two audio tracks: the first is regular audio, and the second may be used for a stereo or an additional conventional audio track. For industrial users this is useful for recording timing, countdown, or narration. Electronically, the video signal is frequency modulated prior to recording. This eliminates the need for the bias signal used in conventional audio recording. The signal-to-noise ratio is built up through limiting. A complex servomechanism maintains correct phasing of the drive motor, the headwheel motor, and the incoming sync signal.

3-10.21.3 Helical Scan

The introduction of helical-scan video tape recorders in 1962 made the video-tape process economically feasible. In addition, the relative portability of the machines was very attractive to industrial users who needed to install the recorder in various locations on an "as required" basis. The light weight of the machines appealed to commercial airlines; they installed the machines along with television monitors for the entertainment of their passengers. The simplicity of operation and maintenance appealed to all. However, there are limitations to the quality obtainable with helical scan: vertical-scan instruments

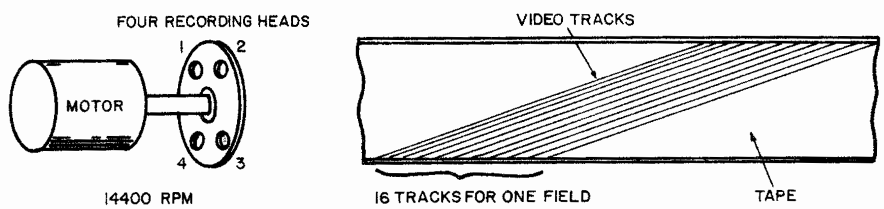


Figure 3-433. Four Head Video Tape Recording System

remain standard for critical commercial broadcast operation. Two basic systems of helical recording are in use. One has either a single head or records most of the signal with a single head and fills in the gap (caused by this head passing from one edge of the tape to the other) with a supplementary head. The second system uses two heads placed on opposite sides of the headwheel assembly. Figure 3-434 illustrates the magnetic patterns on tapes recorded by the two methods. The distinguishing feature of the helical recorder is that an entire video field is recorded during each pass of the head across the tape. This makes possible still-framing, slow motion, reverse motion, and highspeed viewing of the recorded material. Both one-head and two-head methods are in general use, and both are available in models using one-inch and two-inch tape. Figures 3-435 and 3-436 show how the tape runs in a helix around the head assembly, causing the video materials to be recorded in a long diagonal on the tape. As in vertical-scan machines, the video signal is frequency modulated prior to recording. There are also the control head, a two-channel audio head, a fullwidth erase head, and an audio erase head. Figure 3-437 shows their positions. The tape wraps around the head drum at a slant for slightly less than 360 degrees. While the main video head passes the gap, the sync head comes into use, preventing a void from occuring in the video signal. In the video-record mode, all previously recorded tracks are erased by the full-width erase head. The tape then passes the rotating head drum where the video and sync tracks are recorded. Next the stationary head assembly adds the control and audio tracks. Since the audio-erase head acts only on track one, it is possible to record a new sound track while retaining a previously-recorded video track. The audio tracks are recorded and played back as in a conventional audio tape recorder. Figure 3-438 shows the positions of the tracks on the tape. During a given revolution of the head drum the video head records track 1-2 followed by the sync head recording track 3-4. Next the video head records track 5-6, followed by the sync head recording track 7-8. To simplify equalization of the complex frequency-amplitude characteristic of a magnetic reproducing head, video tape recorders use an FM recording system. In the machine shown, the 3.0-MHz carrier deviates 2.0 MHz for 100-percent modulation - always in an upward direction. The sync tips of the video signal produce a carrier of 3.0 MHz and peak whites produce a carrier of 5.0 MHz. Both sidebands are used. The result is an FM signal with a comparatively low modulation index.

3-10.21.4 Electronics Assembly

Figure 3-439 is a simplified block diagram of the electronics of the recorder. The servo system controls the rotation of the head-drum motor. In the record mode the head drum is synchronized to the 60-Hz vertical-sync signal that is derived from the composite video-input signal. During reproduction the head-drum motor is locked to the 60-Hz signal recorded on the control track. Figure 3-440 shows the recorder in the record mode. From the video-input connector, the signal passes to the video-level control (located on the front panel). The signal then goes to the modulator where it is amplified, clamped, and is then used to modulate the 3-MHz carrier. The modulator also samples the vertical-drive signal from the incoming video and feeds this to the servo system, where it becomes the reference signal for the head-drum motor. The modulator has two FM output

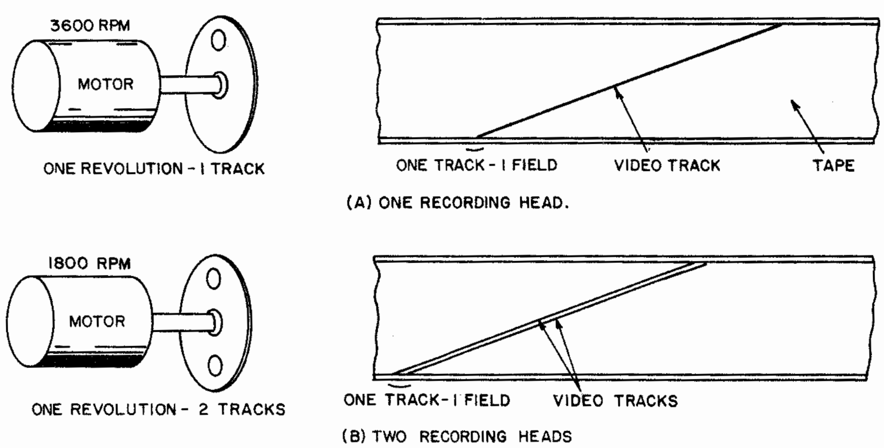


Figure 3-434. Magnetic Patterns of Helical Scan Video Tape Recorders

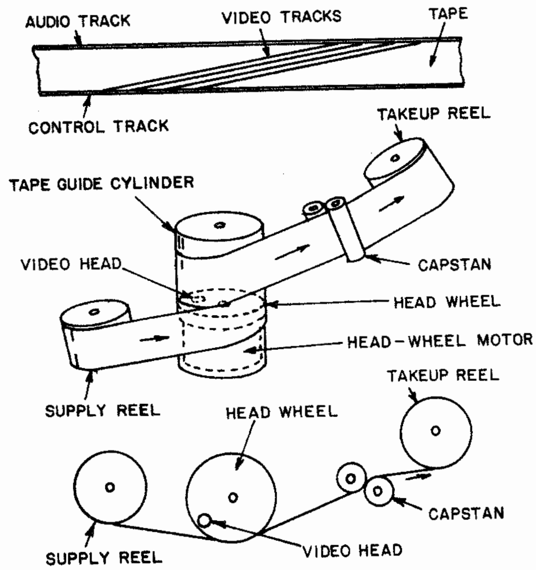


Figure 3-435. Recording Mechanism of Single Head System

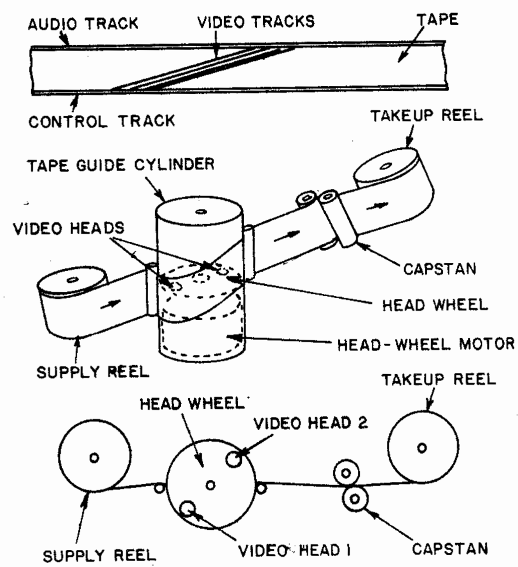


Figure 3-436. Recording Mechanism of Double Head System

signals, the first of which feeds the demodulator for monitoring pruposes. The second feeds the record amplifier, which supplies signal to the video and sync heads through a rotary transformer assembly (roto-coil assembly). Although identical signals are fed to both the video and sync heads, the signal is recorded only by the head that is actually in contact with the tape. The rotary transformer provides electrical

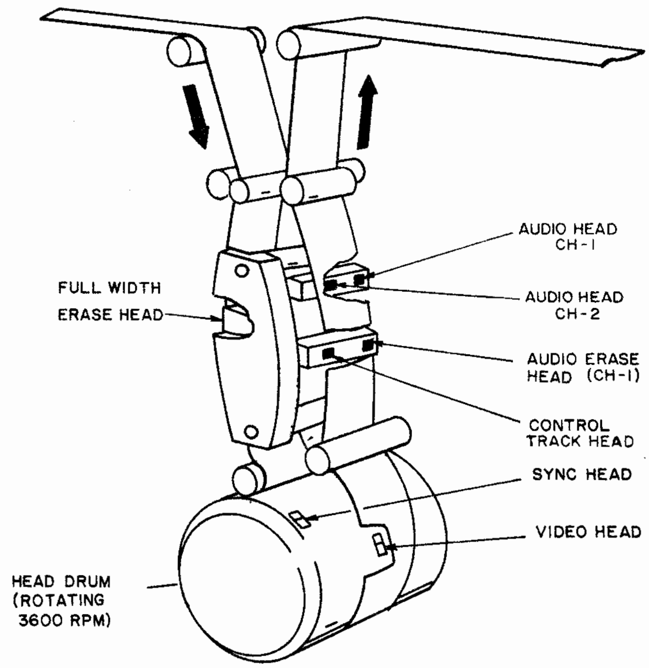


Figure 3-437. Typical Head Positions

coupling between the record amplifier and the heads without the necessity of brushes or slip rings. The two audio channels are completely independent. When an input from a microphone is used, the signal passes through a preamplifier and then to the level control. A signal fed to the line-input jack is routed directly to the control. The audio signal next passes to the record amplifier, where it is amplified, equalized, and mixed with a bias signal, as is conventional for audio recording. The signal from the audio-level control is also fed to the line amplifier as a monitor signal. Channel 1 contains an audio-erase head that is not included in Channel 2. In the record mode, the servo system maintains the rotation of the head-drum motor in such phase and speed that the sync head contacts the tape during the vertical-blanking period, and the video head contacts the tape during the video period. A vertical-drive signal from the modulator triggers a multivibrator in the servo system. The output of the multivibrator is a 60-Hz square wave that feeds the rotating phase comparator attached to the head-drum motor shaft. The phase comparator varies the symmetry of the 60-Hz square wave to correct for headphase displacement. The output of the phase comparator is integrated, and the resulting error signal

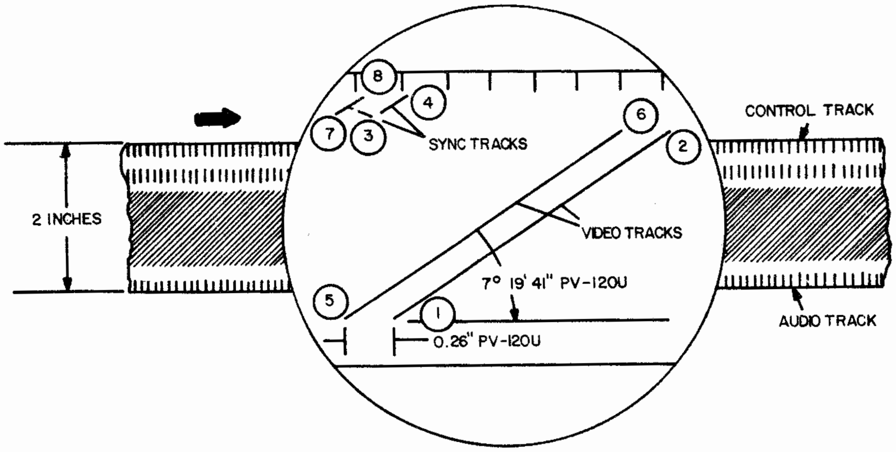


Figure 3-438. Typical Track Positions on Video Tape

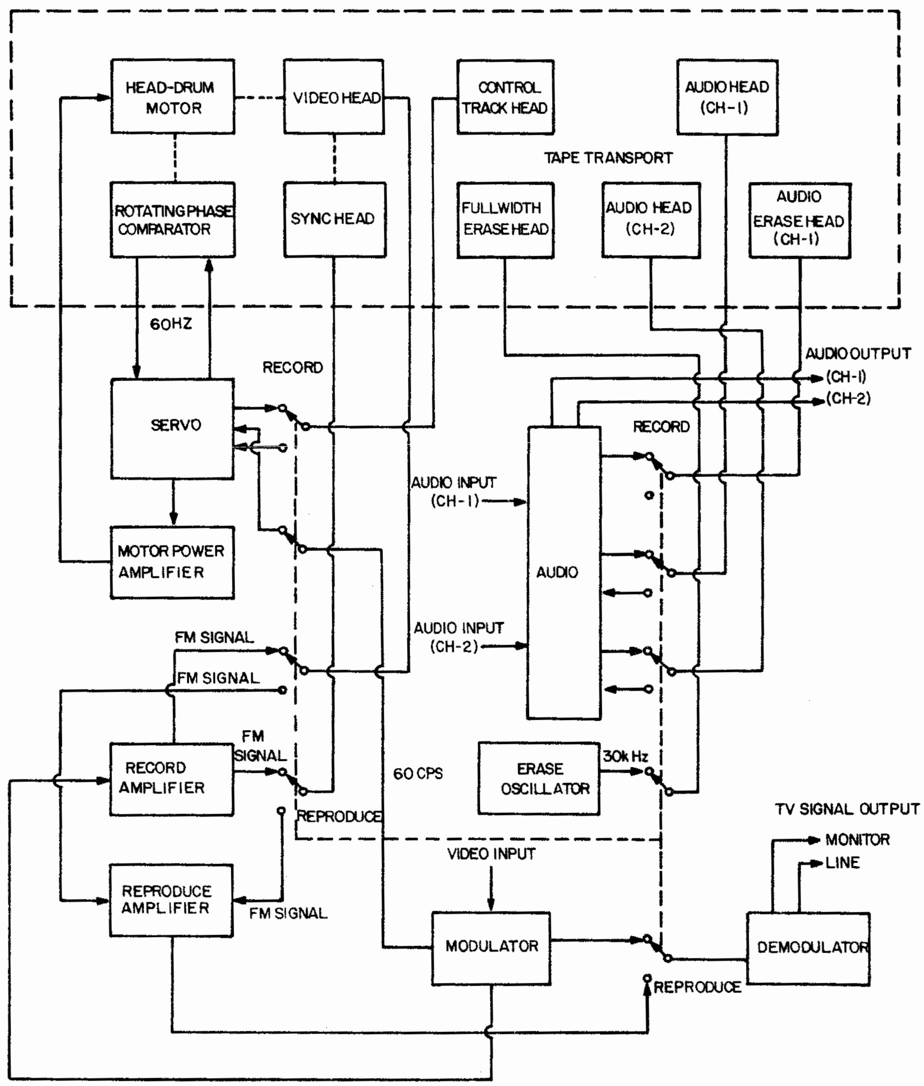


Figure 3-439. Electrical System of Video Tape Recorder (Simplified)

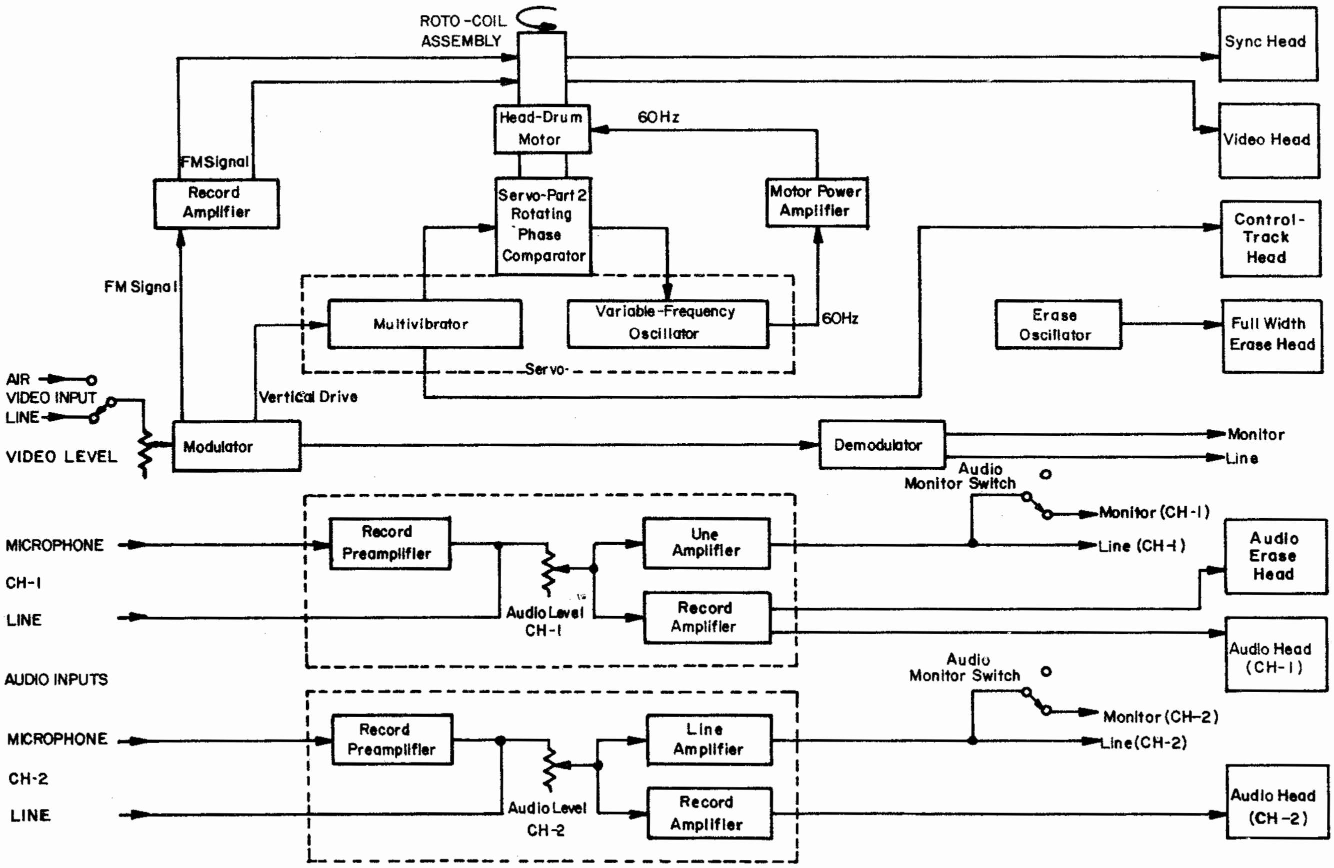


Figure 3-440. Typical Recorder in Record Mode

is fed as a control signal to a 60 Hz sine-wave oscillator. This, in turn, feeds the motor power amplifier that drives the head-drum motor. A second output of the multivibrator is fed to the control-track head where it is recorded in the control-track position on the tape. This signal maintains the proper phase relationship between the head drum and the tape during playback. In the reproduce mode (Figure 441) the frequency-modulated signal from the video and sync heads passes through the rotating transformer to the reproduce-head amplifier. Following amplification, the signal passes through several stages of limiting before demodulation. Monitor and line outputs are fed directly from the demodulator. The audio reproduce amplifier processes the audio signal very much like the conventional audio tape recorder. Servo control of the head-drum motor is similar to the record mode except that the prerecorded control-track pulses trigger the multivibrator in the servo system.

3-10.21.5 Tape-Transport System

Three subassemblies make up the tape transport - the rotary head assembly, the stationary head assembly, and the tape-transport system. The latter drives the tape by means of a capstan and a pinch roller at a nominal speed of 4.25 inches per second. The capstan is driven by a hysteresis synchronous motor rotating at a constant speed. In the fast-forward and rewind modes a reversible induction motor drives the tape. The heart of the video tape recorder is the rotary head assembly. This is the most critical assembly of a video recorder and deserves the best of care. The rotation of the heads, in conjunction with a very small head gap, makes possible the recording of the 4-MHz FM signal on magnetic tape. The speed of the tape relative to the heads is 740 inches per second. The rotary head assembly consists of the head motor, the head-drum disc housing, the video and sync heads, the rotating phase-comparator, and a set of inclined tape guides. The latter maintain the proper slant angle of the tape in relation to the heads. The stationary head assembly houses the fullwidth erase head, the control-track head, the audio-erase head, and the audio record-playback heads. Figure 3-437 shows

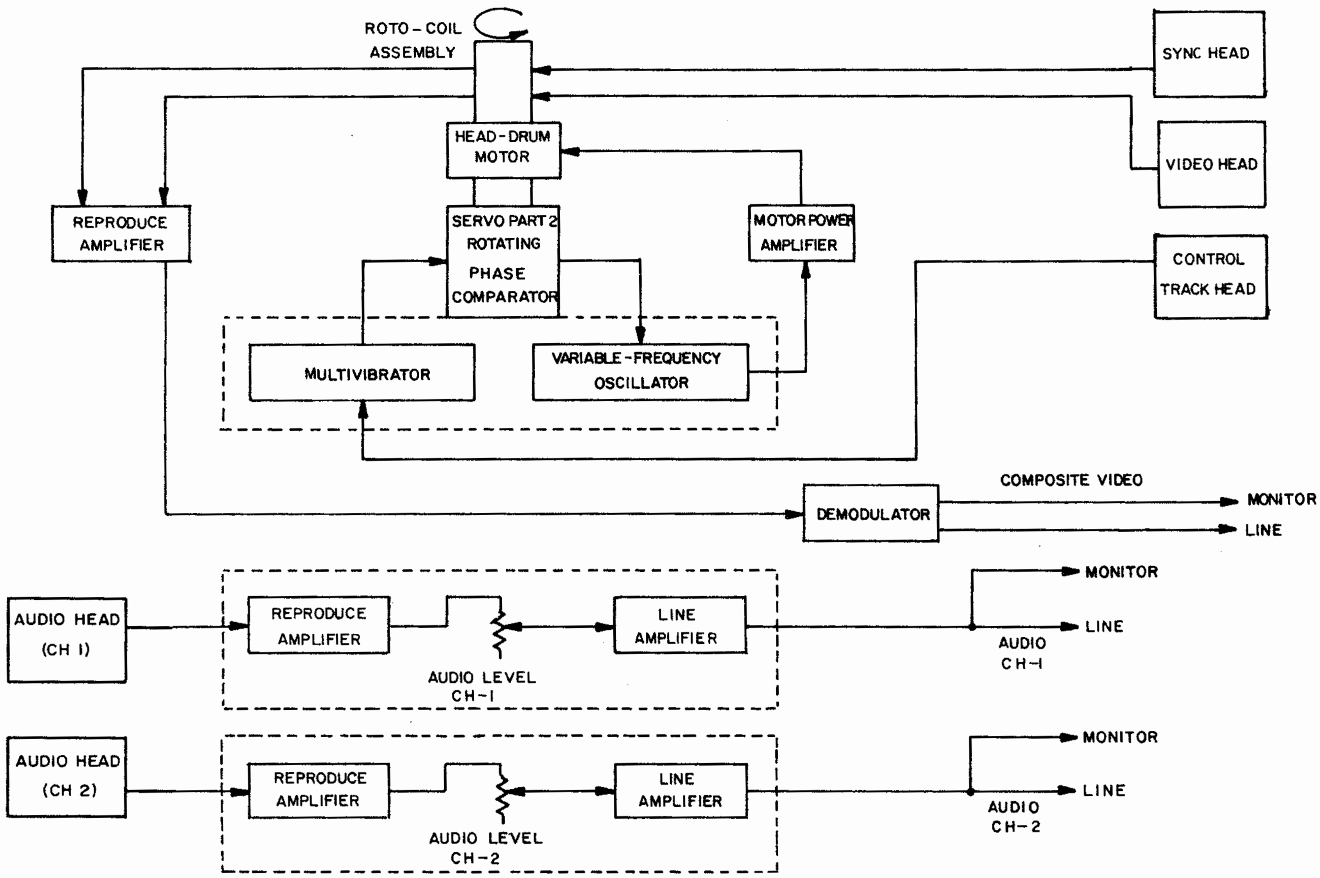


Figure 3-441. Typical Recorder in Reproduce Mode

the location of these. The assembly also houses a photocell sensing device that provides an automatic-stop feature.

3-10.21.6 Circuit Descriptions

This section describes some of the circuits which are unique to the video tape recorder. First to be considered is the modulator, which is divided into two parts - preparation of the video signal, and modulation of the carrier. Figure 3-442 shows the modulator block diagram of Part 1. The incoming signal is first fed to the gain control on the system-control panel. The peak-to-peak signal at test point 101 is about 0.3 volt in normal operation. The signal next passes to a conventional high-frequency pre-emphasis network. Amplifiers Q101 and Q102 are conventional common-emitter video amplifiers. The low-pass filter inserted between them cuts of at 3.5 megahertz. Clamp X101 is a conventional DC restorer. Control R120 sets the bias on the diode, thereby establishing the clamp level. This circuit causes the sync tips to correspond to a frequency of 3.0 megahertz in Part 2 of the modulator. The signal is direct-coupled from the clamping point to the point of actual modulation, thereby preserving the clamping reference. Diode X102 is a conventional whitepeak clipper whose purpose is to prevent carrier deviation above 5.0 megahertz. Voltage regulators Q106 and Q107 ensure stable operation of the clamping and clipping circuitry. The clamping level is about +4.0 volts DC. The input signal is also routed to a conventional vertical-sync separator. The vertical-drive pulse is separated by the low-pass filter, shaped by Q111 and Q112, and fed to the servo system. In Part 2 of the modulator Figure 3-443 the video signal from Part 1 frequency modulates two similar oscillators operating at 54 and 57 megahertz in their quiescent states. The signals from the two oscillators are mixed together to form beat frequencies of 3 megahertz on peak white signals. A video signal that causes the frequnecy of Q311 to increase causes the frequency of Q321 to decrease. In this way a small

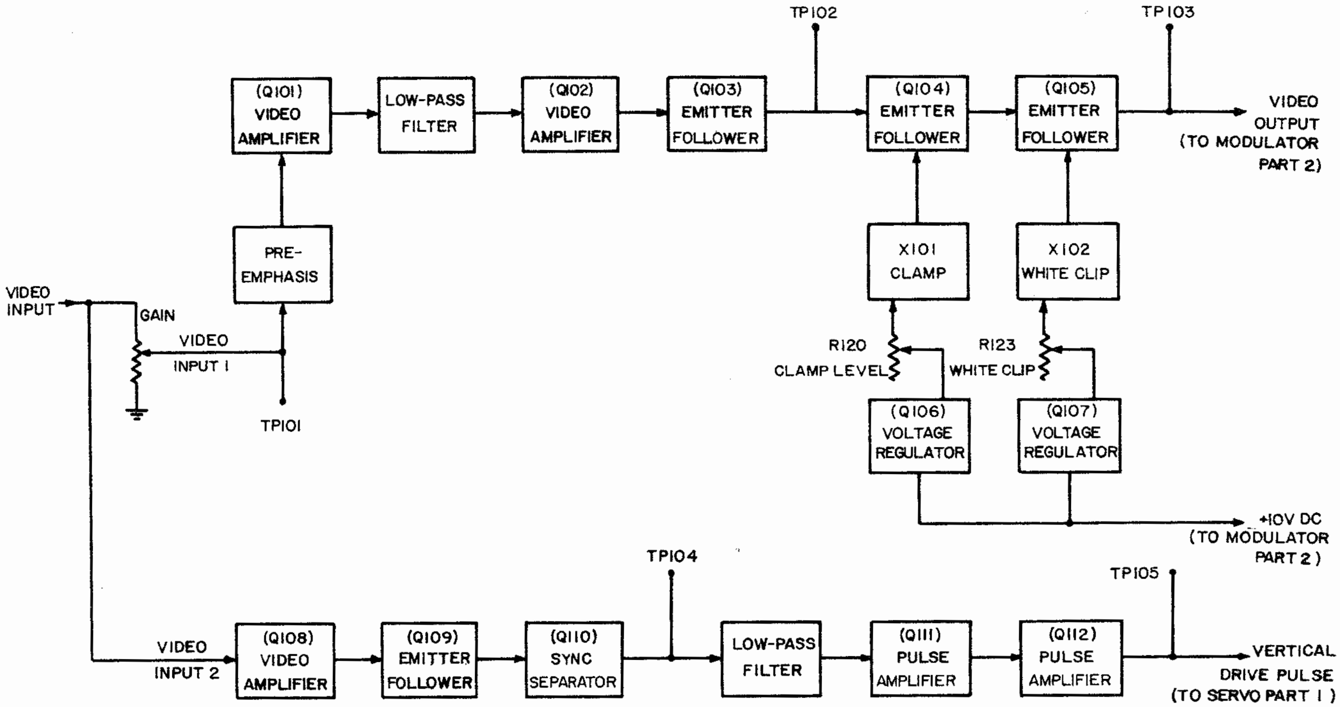


Figure 3-442. Modulator-Part 1

amount of frequency modulation at approximately 55 megahertz is translated into an exceptionally wide deviation of the FM carrier. The lowpass filter removes the 55-megahertz components from the output of the mixer. The signal is then amplified and fed through a limiter to form a square-wave signal. The square-wave shape allows the highest possible average signal to be recorded on the tape. Modulator Part 2 has two outputs - one feeds the record amplifier, the other feeds the demodulator for monitoring purposes. Figure 3-443 shows the circuit of oscillator Q311, which is a version of the Clapp oscillator. Coil L311 forms the tank circuit. Varactor X201, acting in parallel with L311, varies the frequency of the tank in accordance with the modulating video signal. Variable capacitor C312 sets the quiescent frequency of the oscillator. Capacitor C313 resonates with T311 to form a symmetrical sine wave. Diodes X311 and X312 clip at their characteristic 0.7-volt level to form square waves from the sinusoidal output of T311. The forward resistance of the diodes is sufficiently high that the signal is not shorted. RF chokes L201, L202, L312, and L313 serve to prevent interaction between the oscillators. Oscillator Q321 (Figure 3-444) is similar schematically except that the polarity of the varactor is reversed, and it is biased positively instead of negatively in relation to the incoming signal. Figure 3-445 is the block diagram of demodulator Part 1. Its input is either from modulator Part 2 (in the stop or record modes), or from the reproduce amplifier (in the playback mode). The circuit consists of six consecutive 3-stage limiters that elminate amplitude variations of the FM carrier, thereby obtaining the best possible signal-to-noise ratio. Symmetry controls R10 and R66 balance positive and negative clipping levels of the signal. If those are not balanced, the 3-megahertz carrier will appear in the picture. The circuitry of the final 3-stage limiter, which is representative of all the others, is shown in Figure 3-446. Common-emitter amplifier Q16 provides voltage gain. Limiter-driver Q17 is a low-impedance source for limiter diodes X11 and X12. Symmetry-balance adjustment R66 compensates for minor variations in diodes. Transistor Q18 is a current amplifier feeding the next stage. Six of these 3-stage limiters make up demodulator Part 1. Each has 10 dB of gain and 10 dB of limiting. Figure 3-447 is a block diagram of demodulator Part 2. The actual process of demodulation comprises four steps - the delay line, the mixer, the phase splitter, and the detector. This circuit first converts the FM signal to AM before it is detected. The low-pass filter removes remaining RF components from the signal. All amplifier

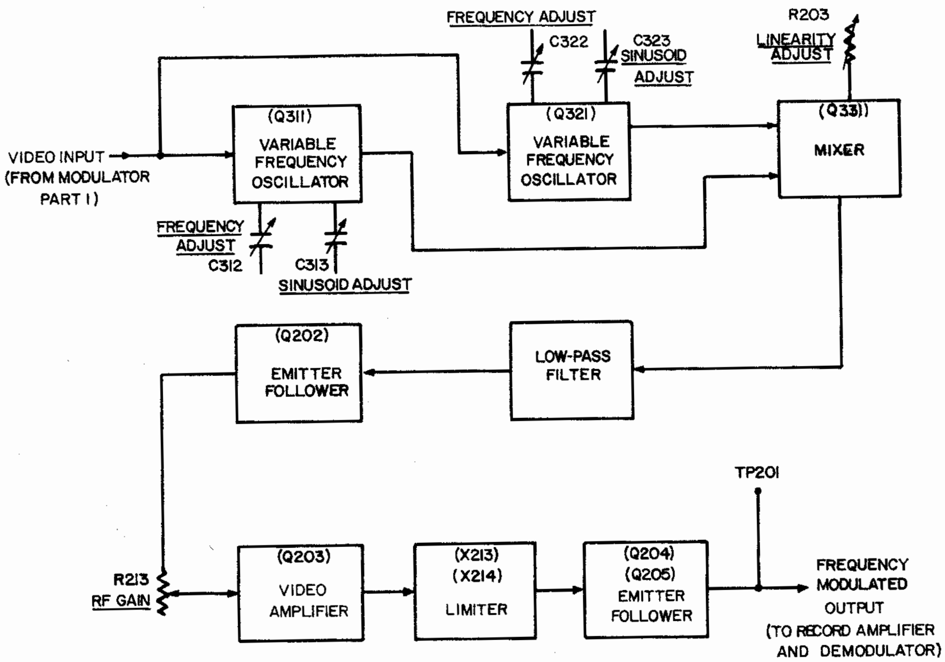


Figure 3-443. Modulator-Part 2

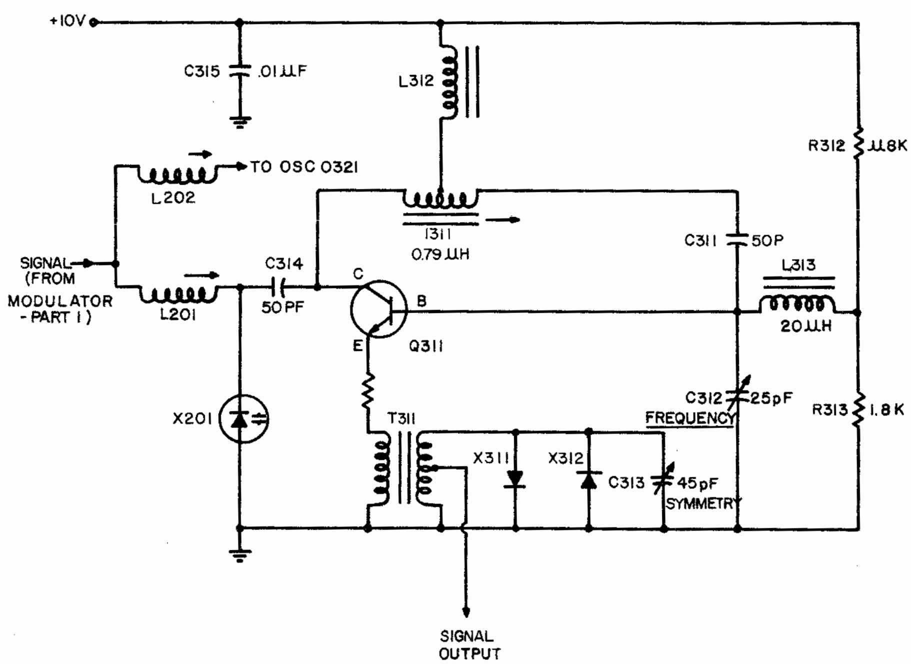


Figure 3-444. FM Oscillator

stages are conventional. The three output stages, Q11, Q12, and Q13 are similar and conventional. The latter feeds a video signal to a panel meter for level monitoring. Potentiometer R48 provides the calibration adjustment for the meter. Figure 3-448 shows a portion of the schematic of demodulator Part 2. Amplifier stage Q1 provides driving power for the delay line and for Q2. Transistors Q2 and Q3 form a mixing circuit. The common collector resistor of Q2 and Q3 (R11) mixes the signals that are applied to the respective bases. The

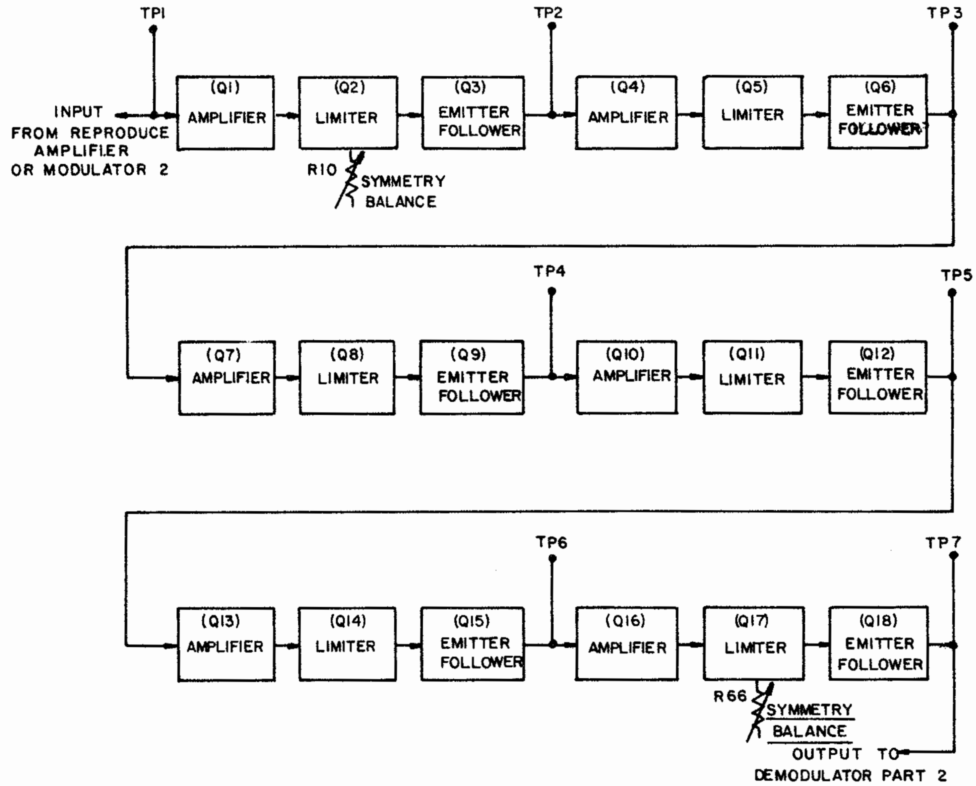


Figure 3-445. Demodulator-Part 1

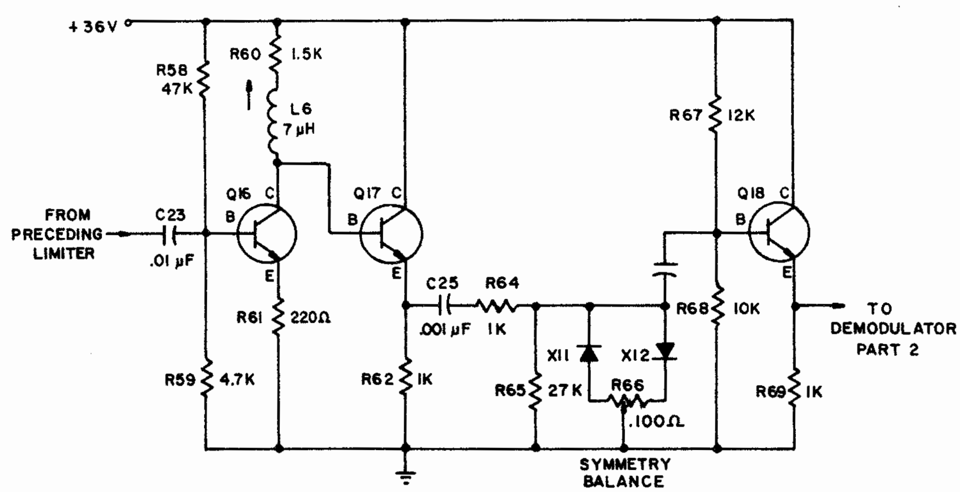


Figure 3-446. Three-stage Limiter

two signals are adjusted to equal amplitude by R5, balancing the loss through the .0625-microsecond delay line. The effect of the delay line on the mixed signal is shown in Figure 3-449. The average voltage in C is far greater than in F, although the amount of delay is .0625 microsecond in both cases. Thus, the average voltage decreases as the width of the pulses decreases due to a rise in the FM carrier frequency. Complete cancellation of the two signals occurs at 8 megahertz. From the mixer the signal passes to phase-splitter Q4

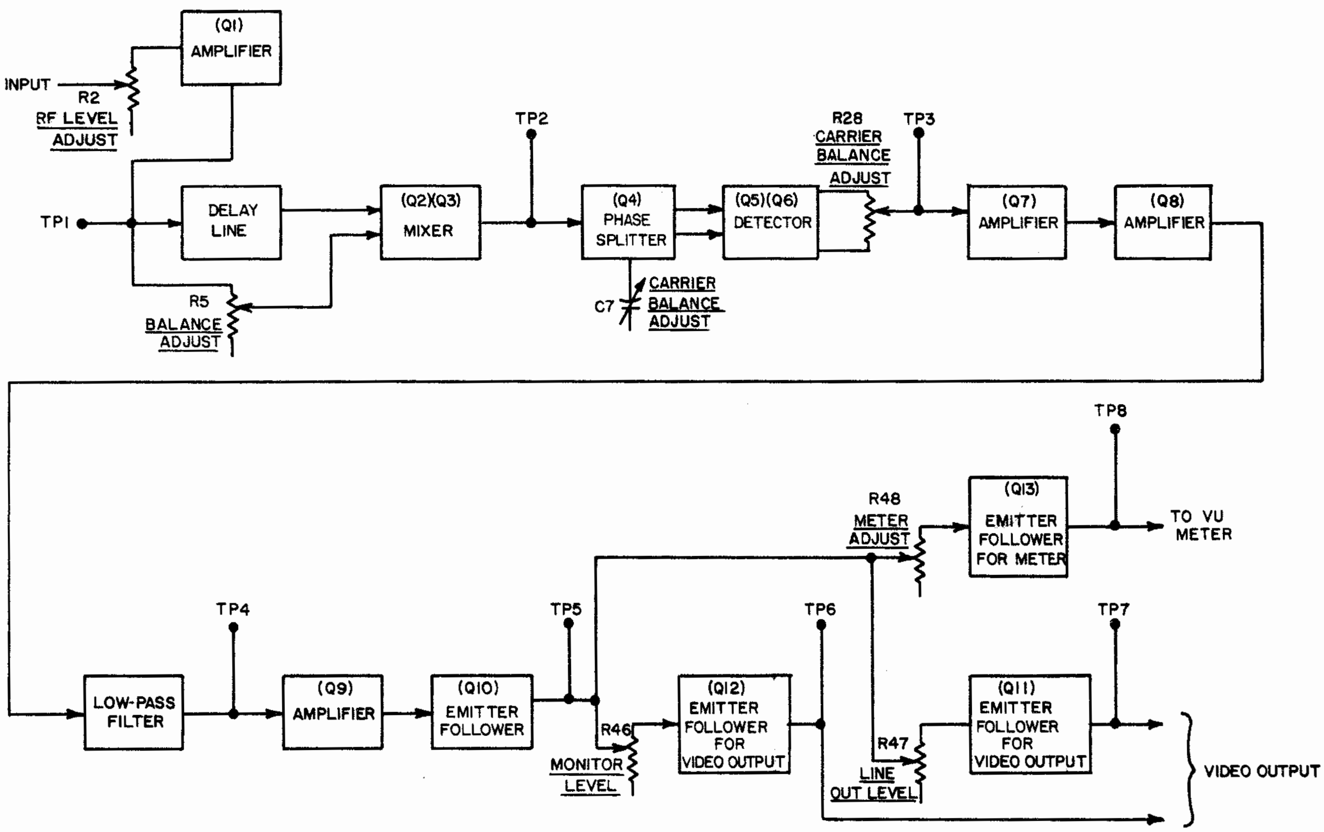


Figure 3-447. Demodulator-Part 2

(Figure 3-448). The signals of opposite phase are fed to the balanced detector consisting of Q5 and Q6. The carrier component of the signal cancels across C13 and C14. However, the modulation component causes a complementary variation of the charges across C13 and C14 and a proportional current through R28. Potentiometer R28 is adjustable for best cancellation of the carrier component. In addition, C7 is adjustable to compensate for minor phase variations in the signals through Q5 and Q6. Since C12 adds phase shift, C7 can compensate for all variations. The record amplifier (Figure 3-450) consists of two separate and identical channels. Both are linear RF amplifiers having a common input. The separate outputs feed the video head and the sync head respectively. Independent record-current adjustments allow optimum adjustment for each head. Meter calibration is provided by controls R132 and R232. These are normally set for a zero reading on the VU meter after record-current adjustments have been optimized. All circuitry of the record amplifier is conventional. Figure 3-451 shows a block diagram of the reproduce amplifier. The signals from the video and sync heads are amplified separately in two essentially similar RF channels of conventional circuitry. After being mixed together they are fed to emitter follower Q10. The gain of the sync channel is adjusted to one-third that of the main video channel to avoid the possibility of complete cancellation if the overlapping portions of the signal should happen to arrive out of phase. The difference in amplitude is eliminated through limiting in the succeeding stages. The output of the reproduce amplifier is a single, frequency-modulated RF signal which is fed to the demodulator. Circuitry that switches the heads between the record and playback amplifier is also located in this assembly. The requirements for speed stability in a video tape recorder are far in excess of those for an audio tape recorder. Variations in tape speed or motor phasing that would go unnoticed in a conventional audio recorder would be totally unacceptable in a video recorder since a uniformly stable picture could not be maintained. Even

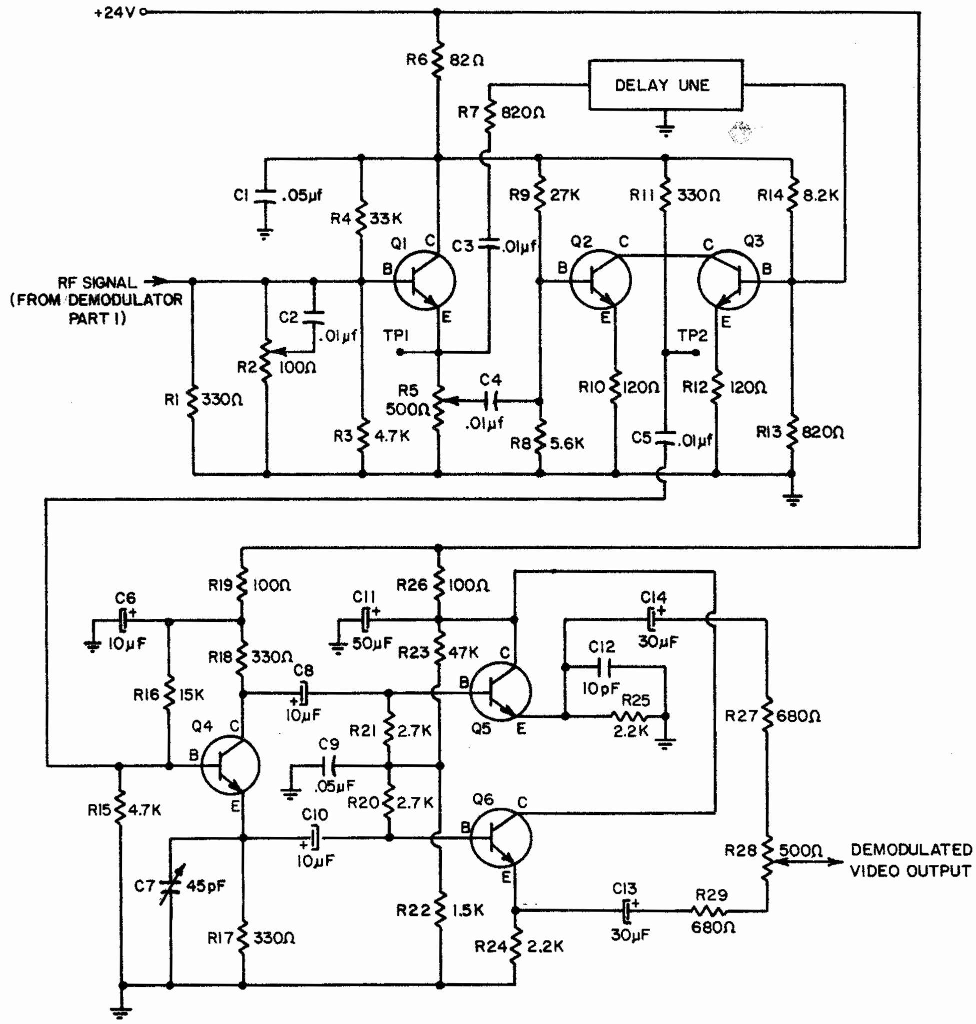


Figure 3-448. Schematic of Demodulation Circuits

slight phase variations of the head drum or the main drive motor will affect the duration of the video lines and might cause the monitor to lose synchronism. In addition, the rotating heads must pass directly over the recorded tracks during reproduce for optimum picture quality and they must maintain a fixed relationship with the forward motion of the tape. During the recording of video tape the speed and phasing of the head drum must place the sync head in contact with the tape during the remaining time. When the tape is played back, the heads must pass directly over the recorded tracks if they are to supply a signal of optimum quality. The servo system maintains this delicate control over the motion of the tape and the rotation of the head drum. Figure 3-452 shows a block diagram of the servo system. In both record and playback modes the capstan motor is driven directly from the 60-hertz line. In the record mode monostable multivibrator Q1-Q2 is triggered by the vertical-drive pulse that was stripped from the video signal in the modulator section. The output of the multivibrator is a symmetrical 60-hertz square wave. This signal is fed through emitter follower Q3 to the control-track head and is recorded on the tape for use as a reference signal during playback. A second output from Q3 is fed to oscillator Q301, which is turned on and off by a shield mounted on the rotating head drum, passing in and out of its coils. The circuit will oscillate only when the square-wave signal from Q3 swings positive and the shield passes between the coils. In normal

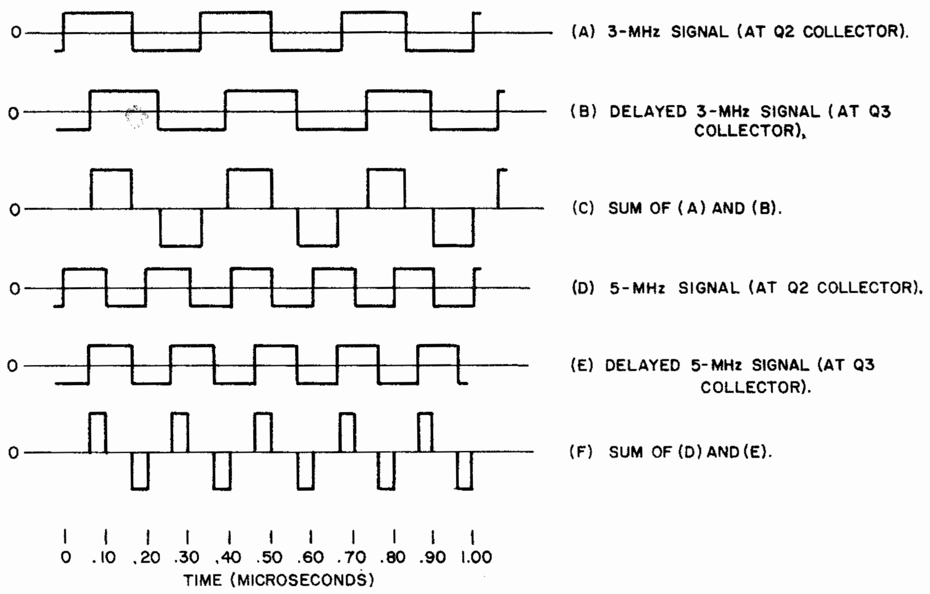


Figure 3-449. Signals in Mixer Circuit Shown in Figure 3-448

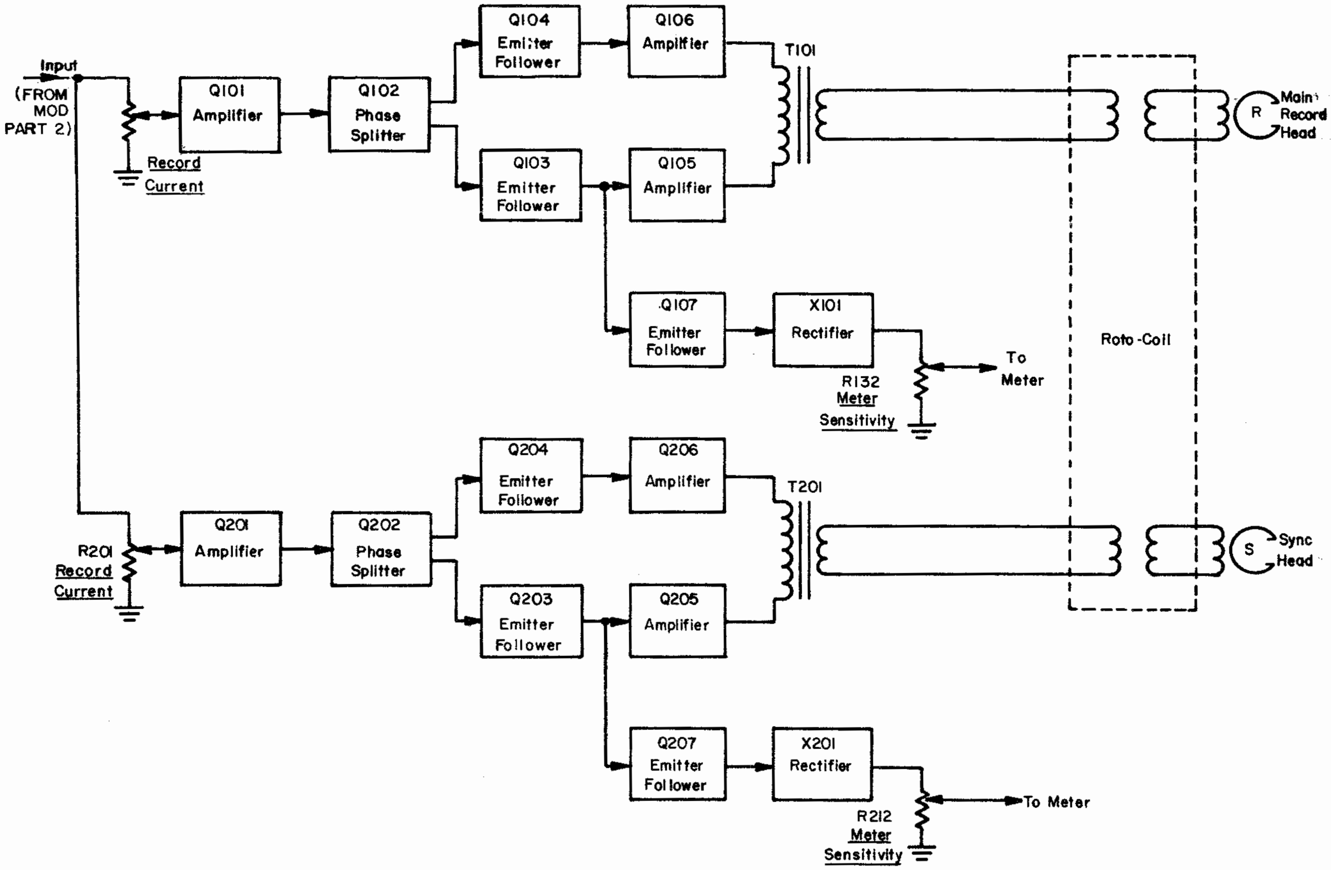


Figure 3-450. Record Amplifier

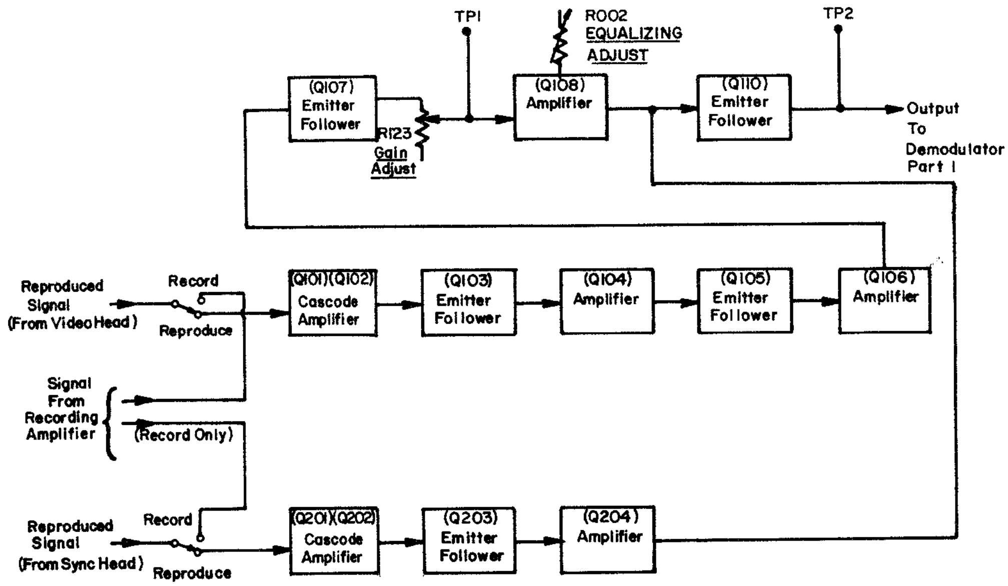


Figure 3-451. Reproduce Amplifier

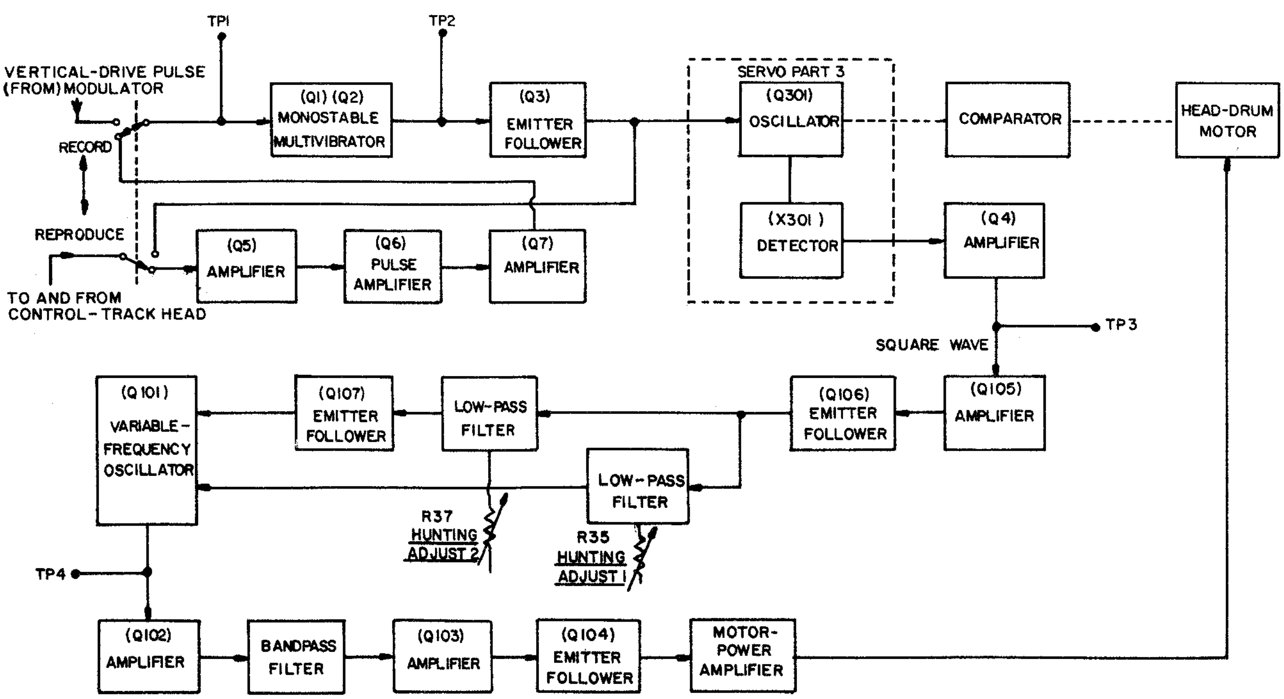


Figure 3-452. Servo System

operation the phasing is such that the oscillator is turned off three-fourths of a revolution and is on for one-fourth of a revolution. The resultant signal is rectified by X301 to form a rectangular wave. This is amplified and fed through an integrator to form a DC signal that controls Q101. The speed of rotation of the head-drum motor is controlled by the frequency of oscillator Q101. The output of this oscillator has a rectangular shape, and its duty cycle is controlled by a DC signal whose voltage is dependent on the pulse width of the signal from Q4. The output of Q101, when amplified by the motor power amplifier, supplies the power for the head-drum motor. To summarize, the head-drum motor is controlled by a closed-loop servo system. The speed of rotation corresponds to the frequency of the vertical-drive signal. Proper phasing of the motor is maintained by means of a comparator having two inputs: one corresponding to the vertical-drive signal, and the other corresponding to the angular position of the motor. The output of the comparator controls the frequency of the oscillator whose output drives the head-drum motor. This circuitry may seem unnecessarily elaborate for the job, but, in the playback mode, the exact phasing necessary for proper tracking of the video heads can be maintained only by a closed-loop servo system. In the playback mode the operation is identical to that of the record mode except that the prerecorded signal from the control-track head is now the reference signal. The capstan motor, being driven from the 60-hertz power line, pulls the tape past the control-track head, which produces an output having the same phase-relationship with the angular position of the capstan as did its input when the signal was recorded. In this manner the tape remembers what the angular position of the head-drum motor was at any particular instant when the tape was recorded. This information is processed by the servo system to assure that the corresponding position is maintained when the video tape is played back. Figure 3-453 shows the circuit of the comparator oscillator - a conventional oscillator with a separate feedback winding. The 60-hertz square wave from Q3 turns Q301 on and off at a 60-hertz rate. (This square wave also supplies the oscillator with all of its power.) The oscillator is also turned off and on at a 60-hertz rate by a shield, having a 50-percent duty cycle, located on the motor shaft. The shield passes between the two windings of the oscillator coil. The pulsating oscillations are rectified by X1 and filtered by C304 to form a square wave. Another noteworthy circuit is the 60-hertz control oscillator and its associated circuitry, shown in Figure 3-454. Transistors Q5 and Q6 amplify the signal from the comparator circuit. (Zener diode X6 limits the level of signal fed to Q6.) The output of Q6 feeds two separate integrator circuits: the one containing R37 controls high-frequency hunting, and the other (containing R35) controls the low-frequency hunting. Hunt adjustments control the amount of output signal. Additional gain for the higher frequency network is provided by Q7. Diodes X1, X2, and X3 limit the extent of control which the circuit ejects over the speed of the head-drum. The frequency of oscillator Q1, controlled by the signal from Q5, Q6, and Q7, establishes the speed of the head-drum motor. Potentiometer R3 is a phase adjustment for the oscillator and sets the phase of the head-drum motor.

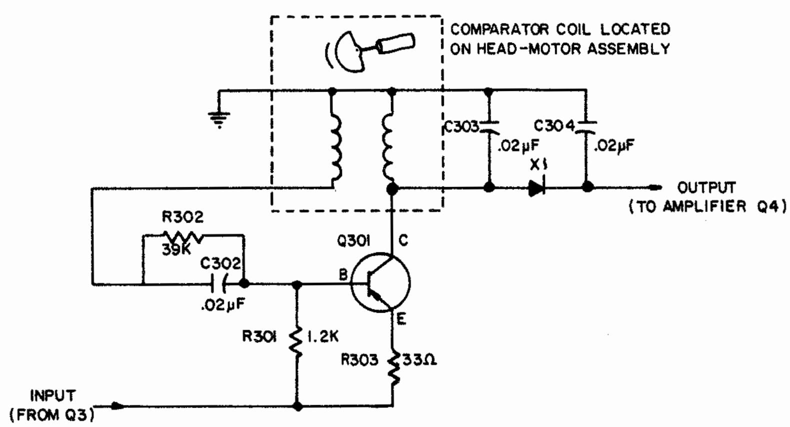


Figure 3-453. Comparator Oscillator

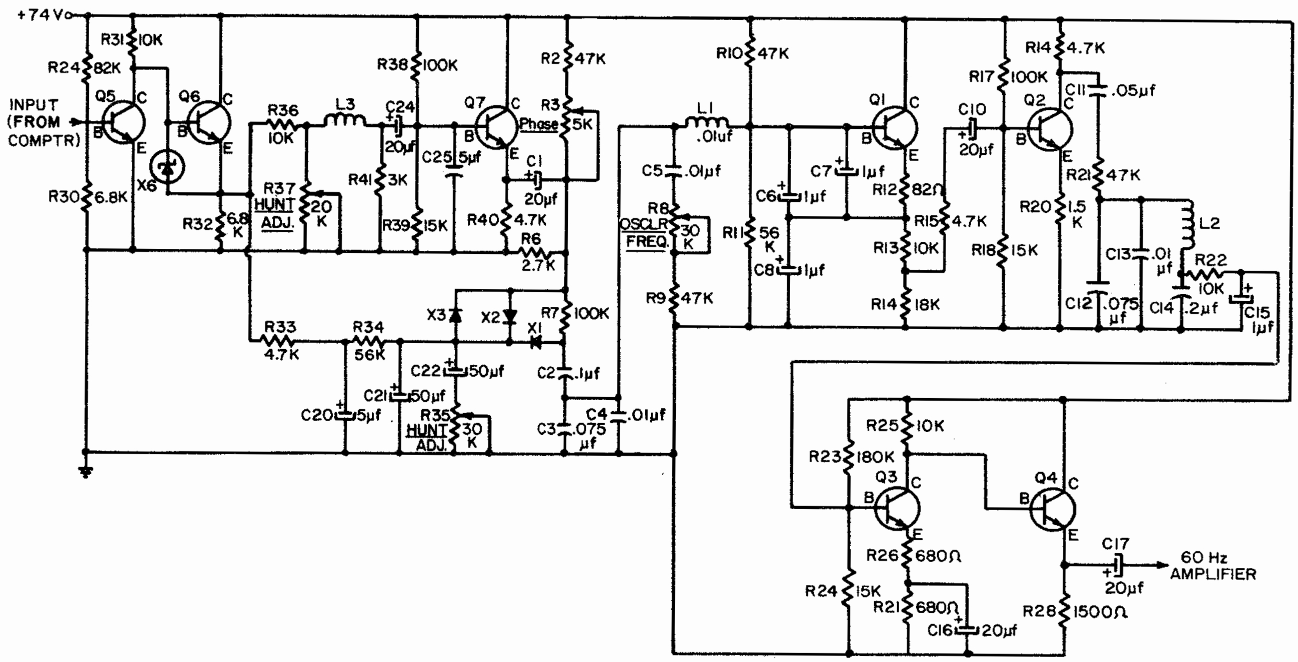


Figure 3-454. Control Oscillator and Related Circuits

The rectangular wave at Q5 should (ideally) have a 1:3 ratio. Oscillator Q1 outputs a low-frequency (60-hertz). The fundamental frequency is controllable by the network C6, C7, C8, L1, C3, C4 and C5, and is adjustable through R8. Fine frequency and phase control is maintained by the signal feed through C2. The output signal is taken from the emitter of Q1 and amplified by Q2. The signal is then fed through a low-pass filter (to form a clean sine wave) and passes through two or more gain stages before feeding the motor-power amplifier. Figure 3-455 is a block diagram of the motor-power amplifier. The signal is divided into two channels that are similar except the lower one contains a 90-degree phase-shift network. This supplies the quadrature voltage for the second winding of the two-phase head-drum motor. Control R24 is used for setting the phasing of the two signals. The motor-power amplifiers feed an assembly that consists of eight power transistors; these furnish the driving power for the head-drum motor. A blower mounted on the capstan motor cools the power transistor assembly. Figure 3-456 is a block diagram of the audio system which is conventional. Inputs may be either via a low-impedance microphone or a 600-ohm balanced line. The output to the audio monitor feeds a standard 600-ohm balanced line. Two similar channels make up the audio system. The schematic of the full-width erase oscillator is shown in detail in Figure 3-457. This is a conventional Hartley oscillator and operates at 30 kHz. When the instrument is not recording, the circuit is turned off by the grounding of the base of Q1 through X1 and R6. Potentiometer R4 allows adjustment for a clean sine wave varying the amount of feedback.

3-10.22 TELEVISION TESTING AND MAINTENANCE

Deterioration of picture quality will inevitably occur in a television system with the aging of components. Personnel responsible for servicing should have a periodic testing schedule, a familiarity with television test methods, a knowledge of maintenance procedures, and an acquaintance with the necessary test equipment.

3-10.22.1 Test Equipment

The first and perhaps most useful item of test equipment is the volt-ohm-milliameter, familiarly known as the multimeter. The vacuum-tube version is preferable for some measurements, but a meter of 20,000 ohms-per-volt sensitivity will generally suffice for nearly any measurement, and the portability is often advantageous. This item is a must for both troubleshooting and preventive maintenance measurements. The second most important item of test

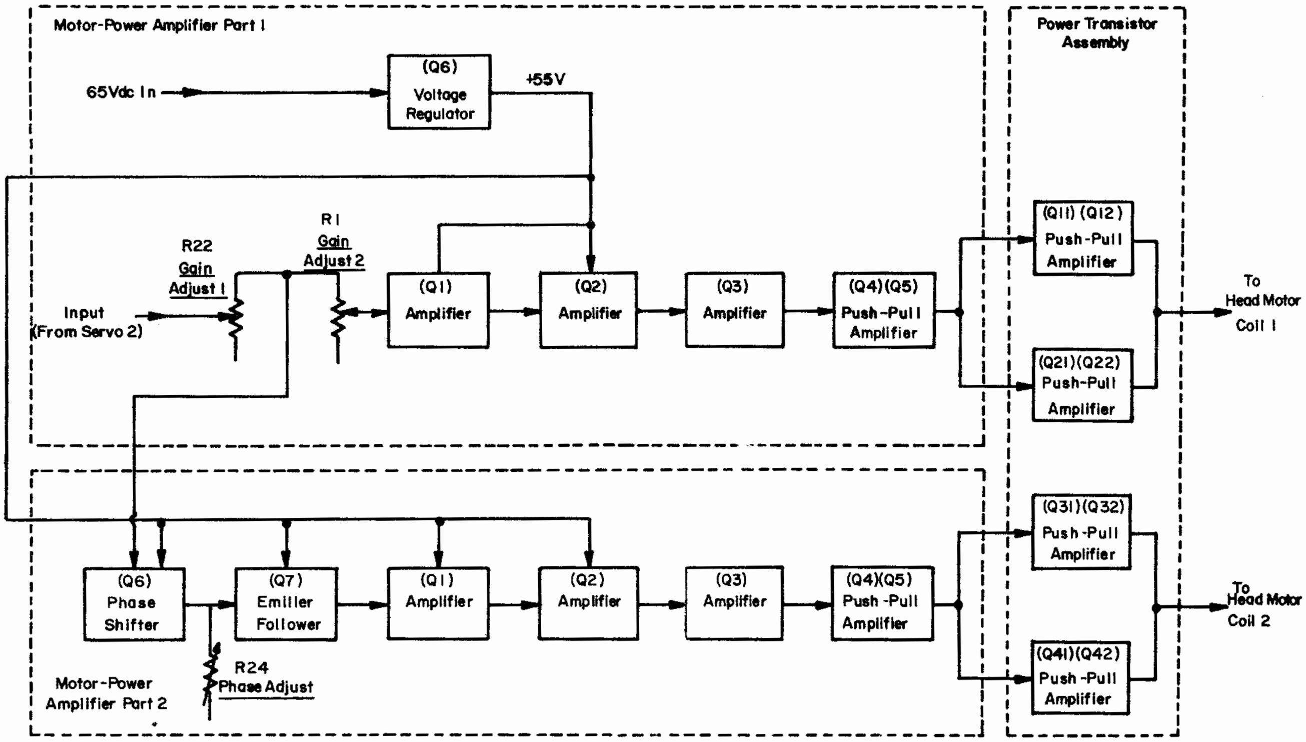


Figure 3-455. Motor Power Amplifier

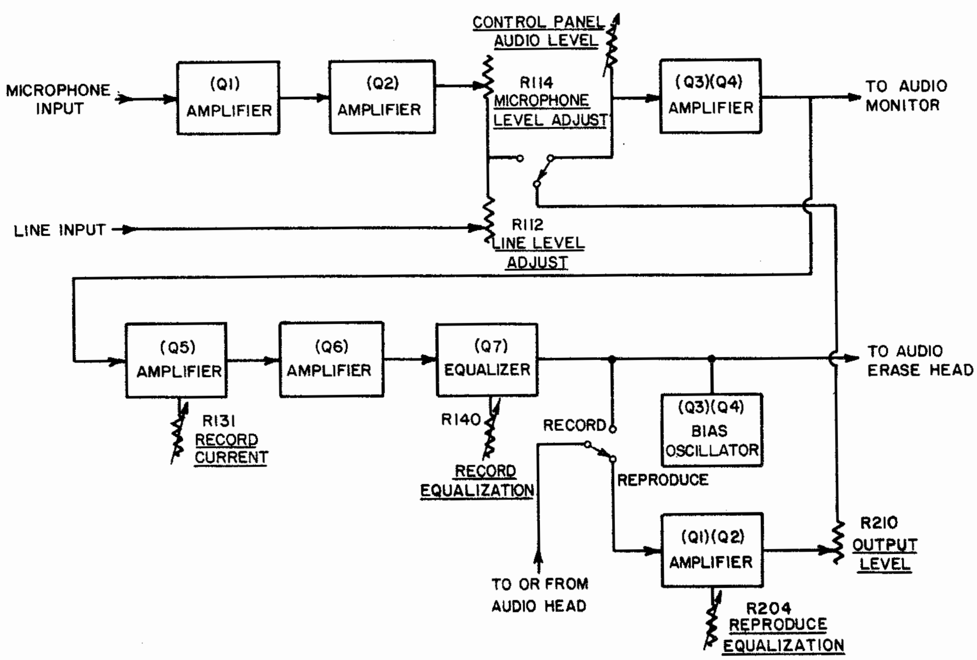


Figure 3-456. Audio System

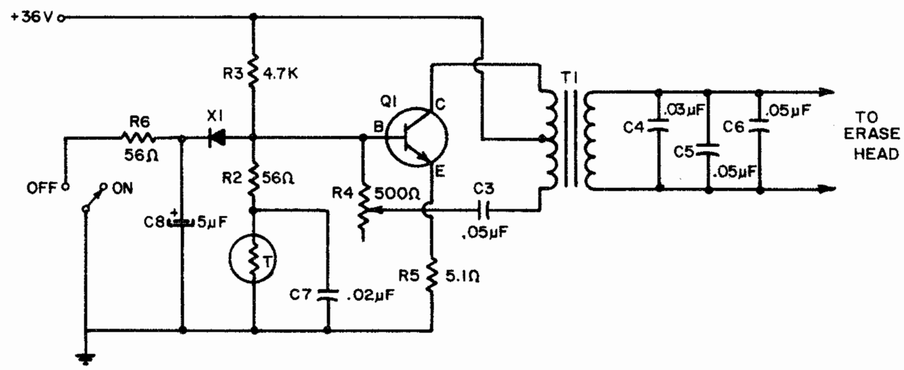


Figure 3-457. Erase Oscillator

equipment is the oscillosope - necessary for both troubleshooting and preventive maintenance. The variety of oscilloscopes that are available is practically limitless. Those having relatively narrow bandwidths (0 to 100 kHz) have limited value for television servicing. However, when used with a detector probe and a sweep generator they are practical for making frequency-response measurements. This is about the extent of their value. Pulse measurements, many frequency-response measurements and troubleshooting require an oscilloscope having moderate gain on a bandwidth at least equal to that of the channel to be measured. For waveform monitoring and for most pulse measurements an oscilloscope with sufficient sensitivity to display the 0.7- to 1.4-volt video signal is needed. Waveform monitors are designed to operate from television signals without needing lock-in adjustments. They may accept composite signals, or non-composite signals and an external sync signal. In the more complex systems there may be several waveform monitors. The number may be reduced in some cases by using it in conjunction with a video switcher, to allow waveform monitoring of video signals from several sources. Three categories of probes are used with oscilloscopes, and all are needed occasionally. These are the attenuator probe (available in 10:1 and 100:1 versions), the cathode-follower probe, and the detector probe. The probes allow measurements to be made in sensitive circuits without the capacitance in the coaxial cable feeding the oscilloscope exerting an appreciable effect on the circuit. The 10:1 probe increases the input impedance of the oscilloscope by about 10 times, and isolates the capacitance of the coaxial cable feeding the oscilloscope from the point being measured. This is most important for frequency-response measurements. Typical input capacitance for a 10:1 probe provides even more isolation, but it is of limited value except where high-level signals are to be measured; a typical input capacitance is 2.5 pF. The cathode-follower probe overcomes the heavy attenuation of the votlage-divider probe. It contains a vacuum-tube cathode-follower circuit within the probe housing, taking advantage of the high input impedance and low output impedance of that circuit. Since the voltage gain of the cathode-follower circuit is always less than unity, a small amount of attenuation takes place. This is far less than the attenuator probe, however. A typical cathode-follower probe has about 40 megohms shunted by 2.0 pF. Also useful is the detector probe. It is needed for alignment of video-amplifier circuits and is also valuable for frequency-response measurements in distribution systems. With the detector probe, sweep-frequency measurements can be made that are unaffected by the rolloff characteristics of the oscilloscope. (Even very expensive oscilloscopes often have significant rolloff at ten megahertz.) The detector probe rectifies the sweep signal and feeds the oscilloscope a signal corresponding to the envelope of the input signal. This contains only relatively low frequencies. The detector probe is quite simple in construction, and many technicians build their own. Figure 3-458 shows the schematic of a voltage-doubler type. The frequency response of this unit is very good to 300 megahertz. Frequency response of cameras, monitors, distribution amplifiers, and related equipment can readily be measured using the video sweep generator. This unit has an output that scans across a band of frequencies at a linear rate, usually 60 hertz.

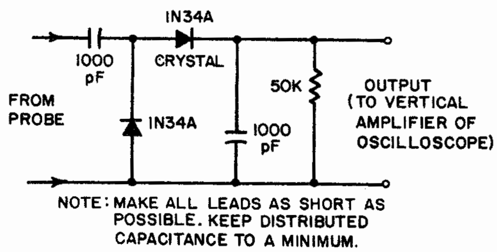


Figure 3-458. Detector Probe Suitable for Video Frequencies

The output is normally displayed on an oscilloscope with a detector probe. The pattern shows the frequency response of the circuit under test (Figure 3-459). Some generators have built-in detector circuits. These are useful when individual components are tested, but when the system is spread over a wide area and a complete check is required, a detector is needed that is separate from the sweep generator. If an RF distribution system is used, an RF sweep generator will be needed. These are similar to the video sweep generator except for the frequency range, and

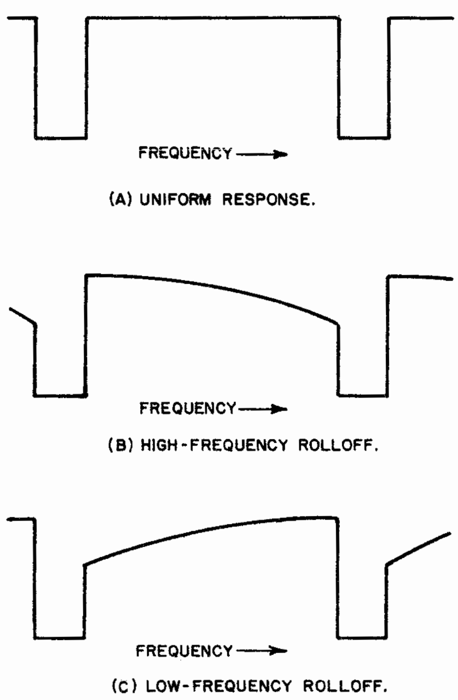


Figure 3-459. Video Sweep Signal Displayed on Oscilloscope

commercial units are available that cover both ranges. The sweep generator is also useful for checking the frequency response of transmission systems. However, its usefulness is limited to devices in which clamping does not take place, unless these circuits are disabled or bypassed. Since stabilizing amplifiers and insert amplifiers employ these circuits, the multiburst signal is more convenient here. The grating generator is essential for scan-linearity adjustments of both television cameras and monitors. Its output is a carefully timed series of electronically generated pulses that appear in the television picture as a grating or crosshatch pattern. The grating generator is used in conjunction with the EIA linearity chart. The television test-signal generator is the most versatile piece of equipment available for servicing a television distribution system. In a well-maintained system it will be used frequently. Its three principal outputs are:

1. The multiburst signal, used for making frequency-response measurement;
2. The stairstep signal, used for making video-amplifier linearity, differential-phase, and differential-gain measurement;
3. The sine-squared and square-wave signal, useful for measuring phase distortion.

The outputs of the television test-signal generator are useful for distribution-system testing since they contain video, blanking, and (if desired) sync pulses, as in normal video signal. Clamper and sync-adding circuits need not be disabled or bypassed when this unit is used. A special generator may be required for systems having nonstandard scan rates. The high-low cross filter is used with a television test-signal generator and an oscilloscope for making differential gain measurements. The television-waveform monitor is normally equipped with this filter as part of its circuitry. The television test-signal receiver is used with a test-signal generator for making differential-phase and differential-gain measurements. The diascope is useful for setting scan sizes and making preliminary resolution measurements and other performance tests of a camera channel. It consists of a small projector that can be installed in the camera-lens mount in place of the usual lens. The device projects a test pattern on the vidicon target that this is of proper size (Figure 3-460) and in the proper location for correct adjustment of the camera scan. The advantage of the diascope over a test pattern viewed through a lens is that minor variations between lenses are avoided, and this test pattern is known to be properly oriented and centered upon the vidicon. The diascope is also

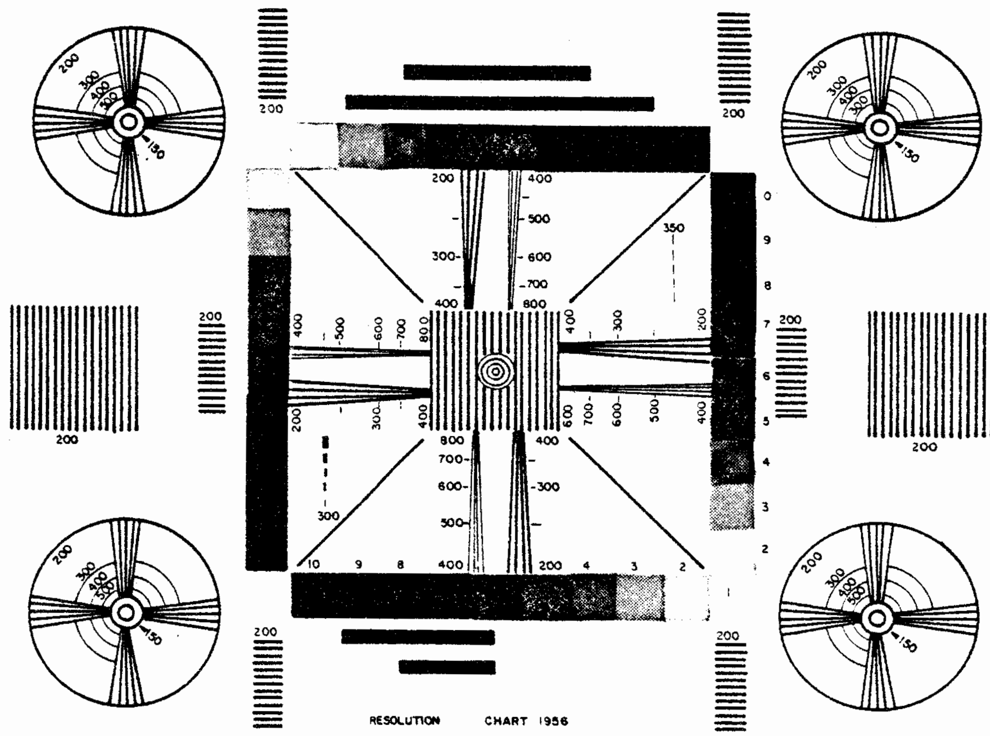


Figure 3-460. EIA Resolution Chart

useful for resolution measurements when a lens of known characteristics is not available or when the lens is to be elminated from the measurement. All popular test patterns can be obtained. Test patterns are needed for checking and aligning cameras; these are also useful for testing on a system-wide basis. The four most often used which may be purchased directly from EIA, are:

1. The EIA resolution chart.
2. The EIA linearity chart.
3. The EIA logarithmic gray scale.
4. The EIA linear gray scale.

These are 18- x 24-inch opaque versions. Gray scales for the resolution chart are supplied separately because of the different printing processes required for best reproduction of each portion of the chart. Transparencies of reduced size are also commercially available for use in specially built rear-illuminated light boxes. These boxes have the advantage of providing uniform lighting so no external illumination is needed, but there is some sacrifice in gray-scale accuracy. However, the convenience of having a light-source available in a single unit with the test pattern may often outweigh minor inaccuracies. The EIA resolution chart (Figure 3-460) is useful for many measurements besides resolution, and it must be used frequently in a properly maintained system. Its principal purpose is testing cameras, but it is also useful in checking monitors and other system components. Among the measurements possible with this test pattern are vertical resolution, horizontal resolution, gamma, interlace, phase distortion, ringing and linearity. The horizontal wedges in the central area are used for measurement of vertical resolution. Limiting resolution is read by observing the point on the wedge beyond which each individual line cannot be recognized with certainty. This measurement is highly subjective. It is therefore essential, when checking a camera (for example) that the monitor, along with any intervening equipment, be in good condition. Vertical resolution in the corners is measured by using the wedges within the small circles. Horizontal resolution is checked in the same way as vertical resolution, but by using the vertical wedges. The numbers along the left side of the pattern indicate the fundamental frequency (in megahertz) created in the video signal by the adjacent portion of the wedge. This calibration is valid only for a 525-line system. The horizontal-resolution calibration is with reference to picture height. This allows a ready comparison of horizontal and vertical resolution, and is universal television practice. The 9-step gray scales allow measurement of the transfer characteristic, or gamma, of the television camera. The white background of the central area forms a tenth step. The gray scales supplied for use with the pattern are logarithmic; in a channel having a gamma of one they will form a logarithmic curve on the oscilloscope. Vidicon tubes have a gamma of about 7, therefore they require some white stretch to form a correct curve. In practice this is rarely done, due to roughly complementary characteristics of the phosphors in the picture monitor. For general applications, all nine steps of the gray scale should be distinguishable. Since this is also a subjective measurement, it is useful only for cursory checking. The four diagonal lines within the square formed by the four gray scales may be used to check interlace quality. A jagged line indicates there is pairing of scan lines. This test is ineffective, however, if interlace is 100 percent. A check for nonuniform phase delay with respect to frequency employs the horizontal black bars near the top and bottom of the large white circle. Phase distortion will appear as smear, or streaking, at the right of the bars. Depending on the direction of misalignment, the smear may appear either black or white. Vary the "high-peaker" or "phase-shift" adjustment to correct this. To make the most accurate adjustment, an exaggerated setting of the contrast control of the monitor may be necessary. The resolution chart is very effective in that case. Ringing, if present, will be most noticeable at the right of the vertical wedges in the small upper right and lower left circles. Ringing will be most pronounced at the right of the line corresponding to the frequency of ringing. It may be caused by improperly installed coaxial cables, incorrect types of coaxial cables, improper video alignment, or incorrect termination. A rough check or horizontal linearity is possible by judging the relative sizes of three groups of vertical bars at the left, center, and right of the pattern. All bars should be of equal size, and all groups should be of equal width. The narrow groups of horizontal bars at either side of the circle may likewise be used for estimating vertical linearity. These measurements are rather crude, however, and a more accurate check can be made using the EIA linearity chart. (Figure 3-461). The white triangles along the edges of the chart indicate the edge of the raster. Resolution measurements are accurate only when the points of the triangles coincide with the edges of the raster. The pattern should be positioned by using an underscanned monitor. The EIA linearity chart (Figure 3-461) is necessary for accurate adjustment of scan linearity of both monitors and cameras. This is done by comparision of the

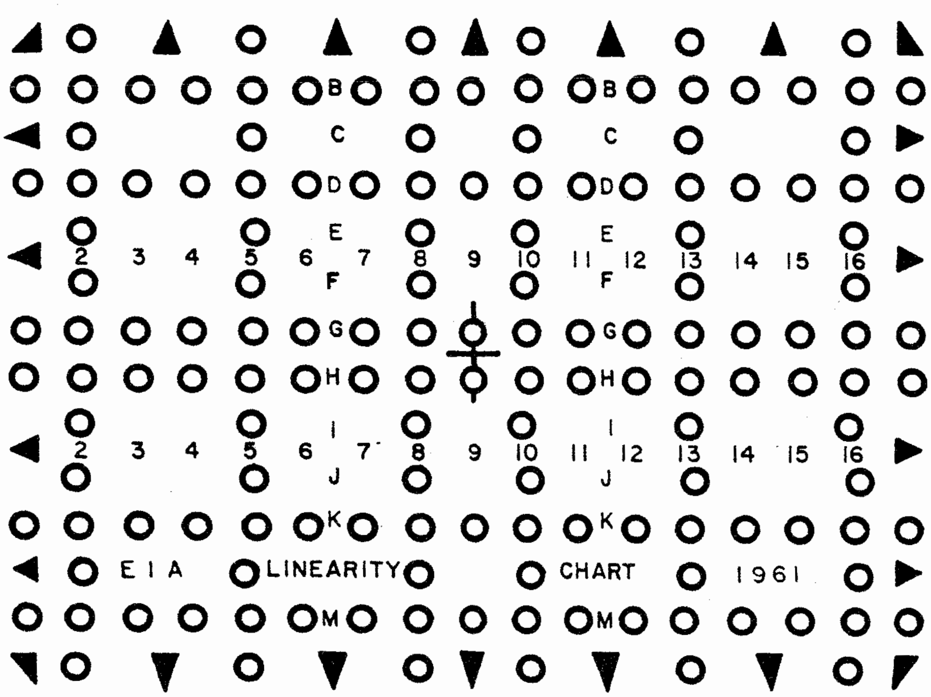


Figure 3-461. EIA Linearity Chart

uniformly spaced circles with the electronically-generated pattern of a grating generator. For camera adjustments, the camera is focused on the pattern, which is viewed on a monitor with a superimposed grating pattern. The camera scan-linearity adjustments are then set for best coincidence of the two patterns. For monitor adjustment, either of two methods may be used. Both require that the grating pattern be displayed on the kinescope. In one method, a transparency of the linearity chart is placed in front of the monitor. Linearity adjustments are then made while viewing the grating pattern through the transparency, adjusting for best coincidence of the grating with the linearity chart. In the second method, the grating pattern is displayed on the kinescope while a 35-mm slide of the linearity pattern is projected on the kinescope from the front. This method has the advantages of allowing a single slide to be used for all monitors, regardless of picture size. It also aids in the elimination of possible parallax. The EIA logarithmic reflectance chart, or gray scale, is used in a manner similar to the gray scales of the EIA resolution chart. On this pattern, the gray scales are much larger and less obscure in an oscilloscope display. The relative brightness of the patches of the EIA linear reflectance chart (Figure 3-462) have a linear rather than a logarithmic relationship. This chart is therefore useful for estimating the transfer characteristic (gamma) of a camera or other equipment. If the equipment between the chart and the oscilloscope presentation has a gamma of unity, the oscilloscope presentation will be two crossed linear stairsteps, when viewed at the horizontal rate. The human eye has a built-in "AGC" characteristic and is therefore a poor judge of absolute light intensity. The light meter is used for measuring light quantities in a televised scene. The most useful type reads in footcandles.

3-10.22.2 Camera Measurements

Since the video signal originates in the camera, every effort should be made to ensure that this signal is as near to perfect as possible. If it is of poor quality, there is probably very little that can be done to correct it in succeeding equipment. Seven major factors affecting picture quality are vertical resolution, horizontal resolution, geometric distortion, phase distortion, gamma, noise, and shading.

3-10.22.2.1 Vertical Resolution

Camera resolution is measured by using the EIA resolution chart or a similar test pattern. This pattern is described in the previous section. Vertical resolution is limited by the number of scanning lines making up the raster, by the beam-spot size of the pickup tube, by optical and electronic focus, and by the quality of the lens. The practical maximum that can be realized is usually taken as the number of active scanning lines in the picture times 0.7. This is called a Kell factor. Thus, a 525-line system would resolve about 340 lines (525 lines, less 40 for blanking, times 0.7 equals 339.5 lines). This may often be exceeded when a test pattern is viewed, but the foregoing figure is realistic for more ordinary scenes. The limitation imposed by the beam size is somewhat fixed by the design of the camera and by the size of the aperture within the camera within the camera tube.

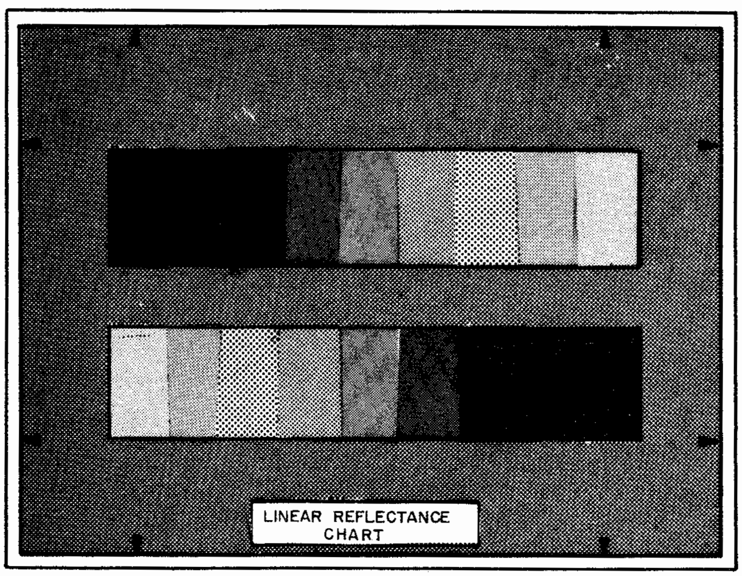


Figure 3-462. EIA Linear Reflectance Chart

In electromagnetically-focused tubes a decrease in focus current, accompanied by a compensating decrease in electrostatic focus voltage, will cause the spot size to increase. Another factor affecting picture resolution is vidicon-beam alignment magnets, or improper adjustment of the current through the electromagnetic alignment coils, will cause poor resolution and poor shading in certain areas of the picture. This adjustment varies with individual vidicon tubes and must be made whenever the tube is changed. Beam alignment may be checked by rotating the electrostatic-focus adjustment while viewing an EIA resolution chart. If the adjustment is incorrect, various points of the picture will come into focus at different settings of the control, and shading will change as the control is rotated. With vidicons having grids 3 and 4 connected together, the point of optimum focus for the center will probably be different from that of the corners but if alignment is properly set, all four corners will come into focus together. Shading will be symmetrical at the point of best focus.

3-10.22.2.2 Horizontal Resolution

Horizontal resolution, like vertical resolution, is affected by the electrical- and optical-focus settings and by the beam alignment of the pickup tube. An additional factor is the alignment of the video amplifier. The stages having the most pronounced effect are the high-peaker and aperture-correction circuits. These are also the stages most likely to be misaligned. The high-peaker can be adjusted while the horizontal bars of an EIA resolution chart are viewed; adjust for minimum smear to the right of the bars. The aperture-correction circuit may be ajdusted while the vertical wedges of the resolution chart are viewed; adjust for the best resolution below the point where noise becomes intolerable. The alignment of the remaining video amplifiers will also affect horizontal resolution. Many popular transistor cameras rely on heavy negative feedback for uniform frequency response, so there are no adjustable peaking coils. If a loss of response is traced to a video-amplifier circuit, it will probably be necessary to replace a component to restore resolution. Many cameras have adjustable peaking coils. While it is rare for these adjustments to drift, it may be necessary to align a video amplifier. For this procedure the manufacturer's manual should be consulted. The usual test equipment in such tests is a video sweep generator, a detector probe, an oscilloscope, and circuit components for disabling the high-speaker, clamper, and aperture-correction circuits. An item often overlooked when good resolution is not attainable is the quality of the lens of the camera. It is a common misconception that even relatively low quality lenses are capable for providing resolution in excess of that of a 525-line television system. This is not so. Comparative tests with expensive lenses will reveal differences in the resolution attainable in a high-quality television system. Nearly all lenses will show deficiencies in corner resolution, overall resolution, and shading. This is particularly true of many zoom lenses. Deficiencies become most noticeable when horizontal resolution is measured since it is frequently twice the value of vertical resolution. In any standard 525-line television system, this is limited to about 340 lines.

3-10.22.2.3 Geometric Distortion

Geometric distortion is defined as an aberration that causes the reproduced picture to be geometrically dissimiliar to the perspective plane projection of the original scene. The most usual cause is nonlinearity of scanning in either the vertical or horizontal plane. This is best corrected through using the EIA linearity chart (Figure 3-461) and observation of the grating pattern. First, adjust the camera to provide an otherwise good picture. Set blanking to EIA standards. The monitor should be adjusted for best linearity and should be underscanned - all edges visible. The accuracy to which the monitor linearity is adjusted does not limit the accuracy to which the camera can be adjusted, but it is desirable that the monitor be well adjusted. If the height and width of the camera picture have not been previously adjusted, do so with the diascope or deflection gauge. If the foregoing tools are not available, position the camera in front of the chart at such a distance as to cause the corners of the chart to just meet the edges of the sensitized area. Be sure the center axis of the lens coincides exactly with the center axis of the linearity chart. The horizontal and vertical-size controls may now be adjusted so that the edges of the linearity chart coincide with the edges of the raster. If the image is not exactly upright at the point of best focus, adjust the deflection yoke to bring it upright. Mix the signal from the camera with the signal from the grating generator set for a vertical-bar pattern of 315 kHz and a horizontal-bar pattern of 900 hertz (for a 525-line system). The picture should contain 17 vertical lines and 14 horizontal lines. The sync input of a video-distribution amplifier is sometimes convenient for mixing the grating signal with the video signal. An insert amplifier or the mixing amplifier of a switcher-fader is also suitable. If no other method is available, the two signals may be fed into the looping input jacks of the picture monitor with usable results, since the resultant mismatch will affect only resolution and not linearity. The phasing adjustments of the grating generator may now be adjusted for best coincidence of the grating intersections with the centers of their corresponding circles. The horizontal-and vertical-linearity adjustments may now be set for best coincidence within the circles. This will usually affect the size of the scanned area, due to interaction of the controls. The height and width controls should therefore be reset for correct raster size and aspect ratio. If a line intersection falls within its corresponding circle, the position error at that point is one percent or less of the picture height. If it falls on the dark portion of the corresponding circle, the position error at that point is between one and two percent of the picture height. If it falls outside its corresponding circle, the position error at that point is greater than two percent. Many cameras omit either, or both, of the linearity controls due to inherently linear circuits. Failure of one of these to pass the linearity test may require a component replacement. Other factors which may cause geometric distortion are deflection-yoke characteristics and lens characteristics. Wide-angle lenses should be avoided when making geometric-distortion measurements.

3-10.22.2.4 Gamma and Video Linearity

Except for a possible gamma-correction circuit, which usually may be set for linear operation, the video-amplifier portion of a camera should be linear. The gamma of a vidicon tube is normally about 0.7. This is noticeable as a slight white compression when a linear reflectance chart is viewed. This presentation on the oscilloscope is useful in recognizing a distorted signal. Severe white compression may occur because of low emission of the cathode of the vidicon tube. This is most prevalent in low-current types. The emission may often be restored by operating the tube at an increased heater voltage for a short period of time. Defective tubes still within their warranty period should be returned to their manufacturer. The following procedure is suggested for tubes beyond their warranty period, but which are unusable due to low cathode emission. Operate the doubtful tubes for 30 seconds with the heater at 10 volts AC and with G1 and G2 floating. Following this, operate the tube for 40 minutes with the heater at 8 volts AC and with G1 at +3 volts ac and G2 at +300 volts DC. Since there is some danger of filament burn-out while performing the foregoing procedure, it is recommended for use only with tubes that would be otherwise unusable. Poor video linearity may also be caused by aging of vacuum tubes in video amplifiers. A more accurate measurement of video-amplifier linearity can be made by coupling the stairstep signal to the video amplifier through a suitable coupling network (Figure 3-463). The interpretation of the stairstep signal is discussed later.

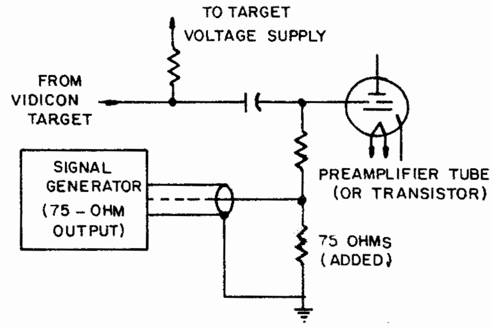


Figure 3-463. Signal Injection into Camera Video Preamplifier

3-10.22.2.5 Noise

Currently, there is no universally accepted method of noise measurement for video signals. Camera manufacturers differ in their methods, and their figures are meaningless unless the method of arriving at a noise figure is given. One popular way to measure noise requires the use of an oscilloscope having a bandwidth equal to that of the video amplifier. The signal-to-noise ratio is then determined by dividing the peak-to-peak video level by the rms noise level. The latter may be taken as one-sixth of the peak-to-peak noise level with reasonable accuracy. The noise level of a camera is somewhat dependent on the setting of the aperture-correction circuit. Even with the foregoing method the figure is meaningless unless this setting is known. It should be adjusted to give flat frequency response or to give a certain resolution figure. The principal source of noise in a vidicon camera is the first video amplifier. If a significant amount of noise is found in the output signal, and it cannot be traced to this stage, it is probably due to a defective component. The amount of noise appearing in the output signal depends on the relative settings of the gain, iris, and target controls. An abnormally high setting of the gain control, accompanied by a low setting of either the target or iris, will increase the relative amount of noise in the signal.

3-10.22.2.6 Shading

Shading is a special kind of noise signal originating in the vidicon tube. This appears as nonuniform intensity in an area which should have uniform intensity. A check for shading can be made by focusing the camera on an evenly illuminated all-white piece of paper (or other material) or on a light box without a slide. The optimum appearance of the signal on an oscilloscope is a level line at both the horizontal and vertical rates. Poor shading not traceable to the optical system may be due to:

1. Improper beam alignment of the vidicon tube.
2. Poor scan linearity; or
3. Beam-landing error.

Improper beam alignment will cause beam-landing error. This should be corrected by adjusting the beam-alignment magnets or the current through the alignment coils. Poor scan linearity results in concentration of the beam in certain areas of the picture. There is a greater amount of signal current in one area of the picture than in another, even though illumination is even. Beam-landing error may also be caused by deficiencies of the deflection yoke or focus coil. This may result in the beam not arriving at the target at a 90-degree angle throughout the picture, and some electrons will glance off the target. Beam-landing error will always be present to a certain degree; it is most prevalent in cameras having a high focus current. The yoke designer may have sacrificed a certain amount of shading in order to improve another quality of the picture; say, corner focus. Because of the many variables involved where such tubes are used, vidicon manufacturers do not ordinarily guarantee their tubes to have shading better than 67 percent of the maximum white signal in a given tube. However, when good grades of yoke assemblies are used, this figure is easily exceeded with most tubes.

3-10.22.2.7 Moire

The decelerator electrode of the vidicon, grid No. 4, is a fine-mesh screen that will cause a moire pattern to appear in the picture as the electrostatic-focus control is adjusted through its range. This is a beat pattern caused by interaction of the mesh wires with the line structure of the picture. It is most pronounced when the vidicon is rotated to an angle where the mesh wires are parallel to the horizontal scan of the tube. It is least noticeable when the mesh wires are at a 45-degree angle to the direction of horizontal scan. Normally, vidicon tubes are manufactured with the mesh wires positioned at a 45-degree angle to the indexing pin. Therefore, the moire pattern should be at a minimum when the tube is installed according to the manufacturer's instructions. Where the moire pattern appears in the picture at the point of best focus, a slight rotation of the tube will often reduce or eliminate the pattern. Where the pattern appears in a portion of the picture at the point of best general focus, it can often be eliminated by resetting the beam alignment. (See previous section.)

3-10.22.3 Monitor Testing

Since one camera may feed several monitors, it is necessary that they be accurately aligned if all are to display similar pictures. In scientific applications where measurements are being made, this is absolutely essntial. An often-heard objection to the use of television for many scientific purposes is that variation exists in pictures. This problem is most often due to poorly aligned monitors. Accurate alignment is also desirable for aesthetic reasons. It is very discomfiting to view several monitors supposedly displaying an identical picture, but each having varying degrees of over- or under-scan, nonlinearity, and improperly positioned deflection yokes. Monitor testing will be considered from the standpoints of resolution, scan linearity, and gamma.

3-10.22.3.1 Resolution

The three prinicpal factors affecting the resolution of a monitor are: the condition of the video amplifier, the focus of the kinescope, and the kinescope itself. A quick check of the resolution of a monitor can be made by displaying a resolution chart that originates in a properly adjusted camera. (The camera must have a resolution capability equal to, or greater than, that of the monitor.) The pattern displayed should present resolution equal to, or greater than, the manufacturer's specifications on all wedges. Failure to achieve this indicates that the monitor requires adjustment or maintenance. First, check to see that the video input of the monitor is correctly terminated. If the signal is looped-through to another device, the termination switch should be placed in the "HI-Z" position. If not, it should be placed in "75-ohm" position. In the absence of a switch, a 75-ohm resistor should be placed on a unused input jack. This is very important, since most video sources will not perform properly unless terminated in the correct impedance; in fact, many will not function at all. If a coaxial cable of any significant length is used, no source will perform acceptably unless it is terminated in the correct impedance. If the proper resolution does not appear on a monitor, check that the focus control is correctly adjusted. A usually reliable indication is to see that the line structure of the raster is sharply defined. If no setting of the focus control will produce this effect, the high-voltage setting may be too low. The latter is accompanied by an excessively large raster in most cases. Where the source of the problem is in a horizontal-deflection circuit, the horizontal size may not be appreciably increased, or may even be decreased. In monitors having unregulated high-voltage supplies, a poorly functioning high-voltage section can be isolated by rotating the brightness control through its range. The raster size should not change through the range of normal operation. If it does, and the monitor has a regulated high-voltage supply, the power supply is out of regulation or below the value at which the regulator is set to function. The value of high voltage may be measured by using a high-voltage probe and a multimeter. (The probe must have been designed for use with the particular meter.) If the high voltage is below the proper value, the monitor requires maintenance. Blooming of the picture - that is, a different focus setting is required for a brightly illuminated area than for a dimly illuminated one - indicates poor regulation of the high-voltage supply, or excessive current in the picture tube. In the latter case, the tube must be replaced. The beam-spot size of the kinescope tube also limits the amount of resolution possible from a monitor. This, however, is quite uniform within tubes, and does not change appreciably with age. The third factor affecting monitor resolution is the bandwidth of the video amplifier. Poor frequency response will appear as poor horizontal resolution of the monitor. If this seems to be the problem, the condition of all tubes should first be checked. If the difficulty cannot be traced to tubes, proceed with the alignment, a video sweep generator and an oscilloscope with a detector probe. (A wideband oscilloscope having known characteristics may be used with an attenuator or cathode-follower probe if a detector probe is not available.) In any case, the capacitance of the probe should be as low as possible, since measurements will be made on high-impedance circuits. This is especially important when aligning wideband monitors. For a cursory check, the multiburst signal may be used in place of the video sweep generator. However, this is not completely satisfactory since response is checked only at spot frequencies, and peaks or dips occuring between these spots will go undiscovered. In addition, the highest frequency of the standard multiburst generator is 4.2 megahertz, while commercial 525-line monitors are built to have video bandwidths of 8 to 10 megahertz, and certain high-resolution monitors have bandwidths above 30 megahertz. The manufacturer's procedure for video alignment should be strictly followed. In the absence of a procedure, the video sweep signal should be fed to the video input jack and properly terminated. The low-capacitance probe may then be connected to the pin feeding video to the cathode or grid of the kinescope tube. The intervening peaking coils, and any other adjustments affecting frequency response, should then be adjusted for the most uniform frequency response of the amplifier. It is best to repeat the adjustments several times. If frequency response cannot be adjusted to normal, the signal should be traced through the circuit until the faulty stage is established. Normal troubleshooting procedures may then be followed. The principal factor affecting vertical resolution is scan interlace. If interlace is not correct, it can usually be achieved by slight adjustment of the vertical-hold control. If this is not effective, maintenance is required. Poor interlace may be caused by cross coupling of the horizontal-drive signal into the vertical-deflection circuitry.

3-10.22.3.2 Geometric Distortion

Monitor linearity should be checked at periodic intervals and adjusted as required. A simple check can be made by feeding the monitor a test pattern from a camera of known good linearity. All circles should appear round. A more exacting adjustment can be made through use of the EIA linearity chart and a grating generator. For a 525-line system, a standard generator produces vertical lines at 900-hertz rate and horizontal lines at a 315-kHz rate. For other scanning systems, different frequencies will be required. Setting scan linearity requires either a transparency of the EIA linearity chart (of the exact size of the raster) or a slide projector and a slide of the chart. With the former, the transparency is laid over the monitor face for adjustment. When the slide-projector method is used, the pattern is projected directly on the kinescope phosphor from the front. The latter is more satisfactory as a shop procedure, since the test pattern appears in the same plane as the grating pattern, and no parallax exists. Care should be taken, however, to ensure that the slide projector is mounted on the optical center of the picture tube. For setting scans, the vertical- and horizontal-blanking signals should be set to the values indicated by the solid lines at the edges of the test pattern. The grating generator should be connected to the monitor input. The grating signal should be set at a convenient level with respect to the blanking signal. Height and width controls should be set to nominal values and the centering adjusted to obtain the proper raster position. The chart is then placed over the monitor screen or is projected on it. The raster size should now be adjusted to exactly fill the area defined by the boundary of the chart. Set phasing controls of the grating generator for best coincidence of the grating intersections with the circle centers. The linearity controls of the monitor should now be adjusted for best coincidence of the intersections with the circle centers. Height and width controls will probably require resetting following linearity procedures in order to maintain the proper aspect ratio. A secondary adjustment found on many monitors is via the pincushion magnets. These may be clipped to the deflection yoke or located near the yoke and supported by other means. They compensate for minor yoke irregularities and for the curvature of the kinescope face. Move them to give the best geometry in the corners and around the edges of the picture. In some instances it may be necessary to replace one or more with magnets of different strength. If a line intersection falls entirely within its corresponding circle, position error at this point is less than one percent of picture height. If it falls within the dark portion of the corresponding circle, position error is between one and two percent of picture height. Most monitors are guaranteed to have two percent or better linearity in both directions. This seems to be the practical limit because kinescope faces are usually not flat, and measurement beyond this value is very difficult.

3-10.22.3.3 Gamma

The gamma of the monitor picture is normally about 2.1 - a value fixed by the characteristics of the phosphors used in the tube. The video-amplifier stages ordinarily will have a gamma of unity. Any deviation from this figure is due to nonlinear operation of the video amplifier and is usually caused by defective tubes. Gamma may be checked using the stairstep pattern of the television test-signal generator. A low-capacitance probe connected to the output of the video amplifier may be used to feed the signal to the oscilloscope. All steps viewed on the oscilloscope should be of equal height when measured at the normal video level. The measurement should be made at 10, 50, and 90 percent of the average picture level.

3-10.22.4 Sync Generators

The sync generator is the central device that controls the timing and scan rates of all equipment in a system. Therefore, the sync generator and related equipment deserve careful attention.

3-10.22.4.1 Pulse Widths and Amplitudes

Table 3-28 presents the EIA standards for pulse widths. These standards should be carefully maintained to ensure uniformity between various systems. The table follows EIA practice of giving pulse widths in fractions of the duration of one horizontal scanning line (1 line = 1 H = 1/15,750 = 63.5 microseconds). However, the use of microseconds in measuring pulse widths is often more convenient, since oscilloscopes calibrated in H may not always be available. Table 3-28 therefore also gives pulse widths in microseconds. These are measured at the 10-percent amplitude point, as shown in Figure 3-464. Also shown is the method of measuring rise time. The values of

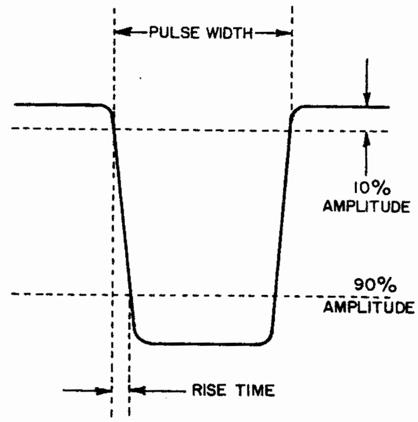


Figure 3-464. Measuring Pulse Width and Rise Time

Table 3-28 are nominal, but should be followed as closely as possible. In cameras using industrial types of sync, the manufacturer's recommendations may differ from the table values, therefore adjustments should be made to his specifications. This also applies to pulse equipment in systems having nonstandard scan rates. Pulses feeding television equipment are normally set at four volts. This value should be checked periodically. The sync amplitude appearing in a composite signal should be set at 0.4 volt in a 1.4-volt system or 0.3 volt in a 1.0-volt system. Outputs of pulse-distribution amplifiers are normally set at the four-volt level in either case. Most modern pulse-distribution amplifiers cannot change pulse widths in normal operation. Certain older equipment, employing clipper circuitry rather than Schmitt triggers for pulse regeneration, can cause stretching if the clipper adjustment is improperly set. The pulse outputs of these amplifiers should be measured frequently to ensure that pulse stretching does not occur.

3-10.22.4.2 Pulse-Cross Display

The pulse-cross display affords a quick operational check of pulse widths and timing. It is not as accurate as the oscilloscope method, but the relative ease of measurement allows frequent and rapid observation of sync and blanking signals. The only instrument required is a special picture monitor that may be used at other times as an ordinary picture monitor. The pulse-cross switch should be flipped occasionally so the technician will know whether the pulse widths are approximately correct, the correct number are present, and the proper timing exists. In a

Table 3-28. EIA Standard Pulse Widths

PULSE	H	MICROSECONDS
Horizontal Line	1.0	63.5
Horizontal Sync Pulse	.075	4.76
Equalizing Pulse	.0375	2.38
Horizontal Blanking Pulse	.175	11.1
Front Porch	.022	1.4
Vertical Serration	.07	4.44
Vertical Blanking	19.7	1250
Horizontal Drive	.1	6.35
Vertical Drive	10.5	666.6

normal monitor, sync and blanking pulses generally occur during the retrace intervals, therefore they are not visible in the picture display. In the pulse-cross display, the vertical scan of the monitor is shifted in phase to place the vertical-blanking interval near the center of the picture. This area is normally expanded for ease of viewing. Horizontal scanning is likewise shifted in phase to allow the horizontal-sync and -blanking pulses to appear in the viewing area. This shift is normally about one-fourth the width of the screen, since placing the horizontal-blanking interval near the center of the picture would cause the equalizing pulses and vertical serrations that normally occur near the center of the picture to fall during retrace of the pulse-cross display. Figure 3-465 shows the presentation of one type of pulse-cross display. In other types the polarity of the video signal is reversed for ease of viewing, and in still other types the vertical-deflection rate may be changed from 60 to 30 hertz. In this case, the vertical-blanking interval following each field may be viewed separately, rather than interlaced as shown in the illustration.

3-10.22.5 Measurements in Video-Distribution Equipment

Since signal quality over a television channel is limited by the poorest component in the channel, it is important that all equipment between the camera and monitor be capable of transmitting the signal with a minimum of distortion. The equipment, therefore, requires periodic maintenance to ensure optimum performance. Tests to be made on video-distribution equipment include frequency response, differential gain, differential phase, noise, and phase distortion. For purposes of measurement, video switchers and other equipment carrying video signals may be processed in a manner similar to that of distribution amplifiers. Frequency response is best checked by using the video sweep generator in conjunction with the video-detector probe and the oscilloscope. Cursory measurements may be made by using the multiburst signal. If alignment is required, the method recommended by the manufacturer should be used. The most practical method of measuring

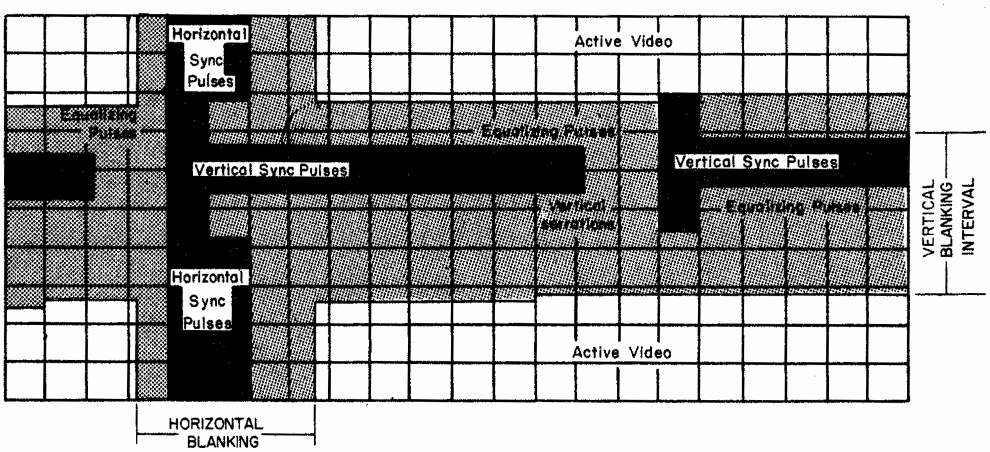


Figure 3-465. Interlaced Pulse Cross Display

phase distortion employs the sine-squared and window signal. The method of interpretation is described in the following sections. Phase shift is usually present when frequency response is unsatisfactory. Differential gain and differential phase are measured by using the stairstep signal with 3.58-MHz subcarrier superimposed. The procedure, as described in the following section, involves the use of the television test-signal generator, the television test-signal receiver, the high-pass filter, and the oscilloscope. Noise in distribution equipment should be far below the video-signal level. No noticeable amount should be added to the video signal, even in extensive distribution systems. If any is added, this is an indication that maintenance is required. The most useful item of test equipment for making measurements on video distribution equipment is the television test-signal generator. Outputs thus provided include multiburst, stairstep, sine-squared, and window signals.

3-10.22.5.1 The Multiburst Signal

The multiburst signal consists of a square wave followed by six individual sine-wave bursts at discrete frequencies. Usually, generators supplied for broadcast purposes employ 0.5 MHz, 1.5 MHz, 2.0 MHz, 3.0 MHz, 3.6 MHz, and 4.2 MHz. If a distribution system has a broader bandwidth than the above, a generator having a wider spread of frequencies should be used. These are available on special order. Figure 3-466 shows the appearance of a normal multiburst signal. Various types of distortion observed when using this signal appear in Figure 3-467. The tilt in the square-wave signal and in the sync pulses is more indicative of phase distortion than of amplitude-versus-frequency distortion, but the two normally go hand in hand. If a distribution system is equalized for best

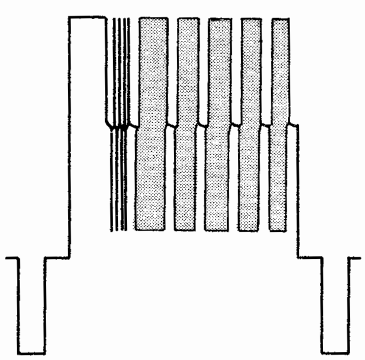


Figure 3-466. Multiburst Signal

amplitude-versus-frequency response, the phase response will be very accurate unless many successive stages of amplification and equalization are present. Any of the distorted signals of Figure 3-467 may be caused by incorrect equalizer settings. The waveform of 3-467A is typical also of an unterminated video coaxial cable. In this case, the overall amplitude will be increased beyond normal. Figure 3-467B might be caused by an excessively low value of termination resistance, in which case the signal of Figure 3-467C could be caused by a defective coaxial cable if the cable were long enough for a standing wave to develop. The curve of Figure 3-467D shows axis shift. The unsymmetrical waveform was caused by excessive pre-equalization that overloaded the amplifier on the negative clipping and the resulting unsymmetrical waveform. The higher frequencies were then re-equalized to near normal amplitude. If the clipping were not present, the waveform would show excessive high-frequency response. The square-wave signal shows evidence of this. Note also the phase shift evidenced by the unsymmetrical shape of the square wave. A pitfall to avoid when making quantitative comparisons is the measurement of the positive peaks only with reference to the white-flag signal. (In a properly equalized system, the peaks of the bursts will be equal to the white-flag signal.) To be accurate, measure the peak-to-peak signal and compare it to what a perfect peak-to-peak signal would be. If the positive peaks only are measured, with reference to blanking, a burst that is completely lost will show a 50-percent (6-dB) loss when the proper measurement would indicate infinity.

3-10.22.5.2 The Stairstep Signal

The television test-signal generator also supplies a stairstep signal, usually consisting of ten steps in linearly ascending amplitude Figure 3-468 shows the normal waveform. Note that each step is of uniform amplitude and width. Modern generators also produce 3-step and sawtooth patterns that may be used for making the same tests as the ten-step pattern. Figure 3-469 shows a stairstep waveform exhibiting white compression. In a more severe case, clipping would be evident, and one or more of the upper steps would be missing. These conditions are called incremental-gain distortions. The usual cause is operation of an amplifier stage outside the linear portion of its transfer curve, and may result from excessive input-signal level or component failure. Compression may take place at either the positive or negative peaks. The latter would be evidenced principally by loss of sync amplitude. Incremental gain always results in differential gain, which is defined as the difference in gain (in decibels or percentage) experienced by a

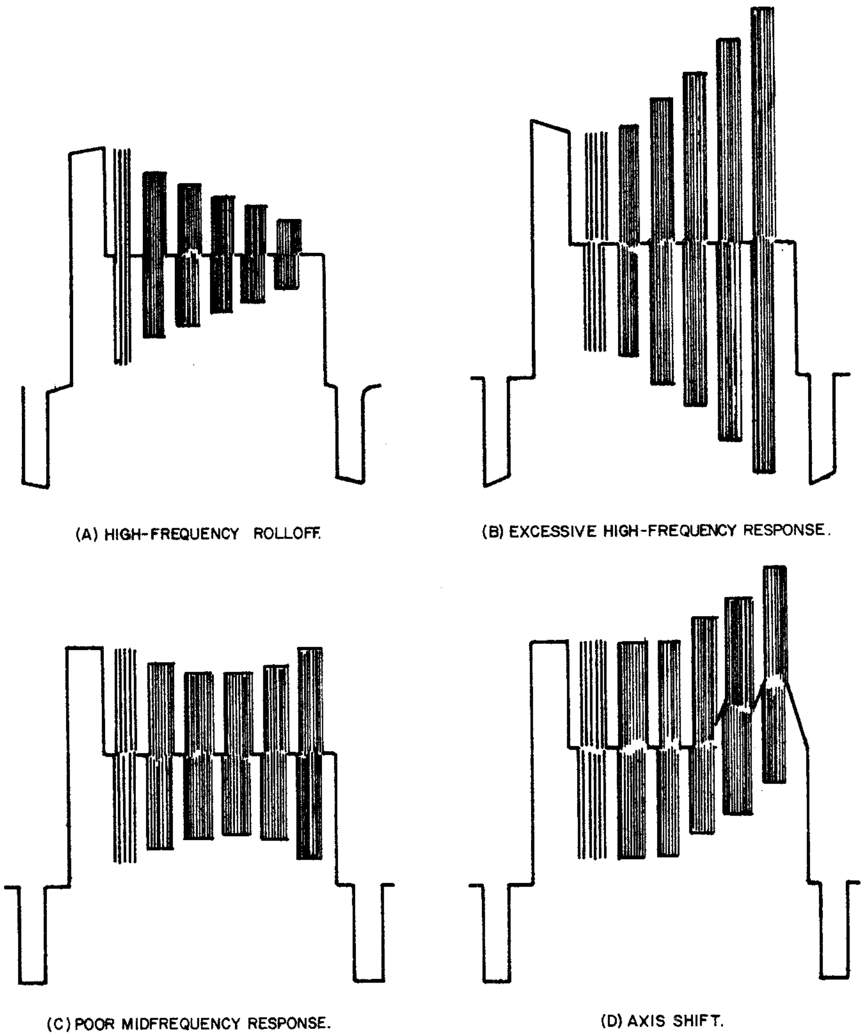


Figure 3-467. Multiburst Signal Showing System Deficiencies

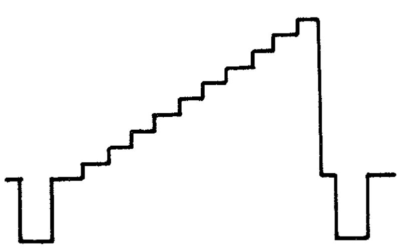


Figure 3-468. Normal Stairstep Waveform

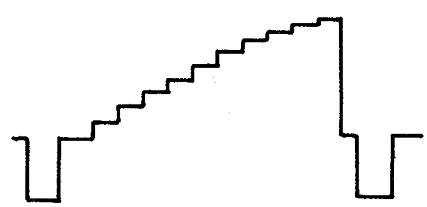
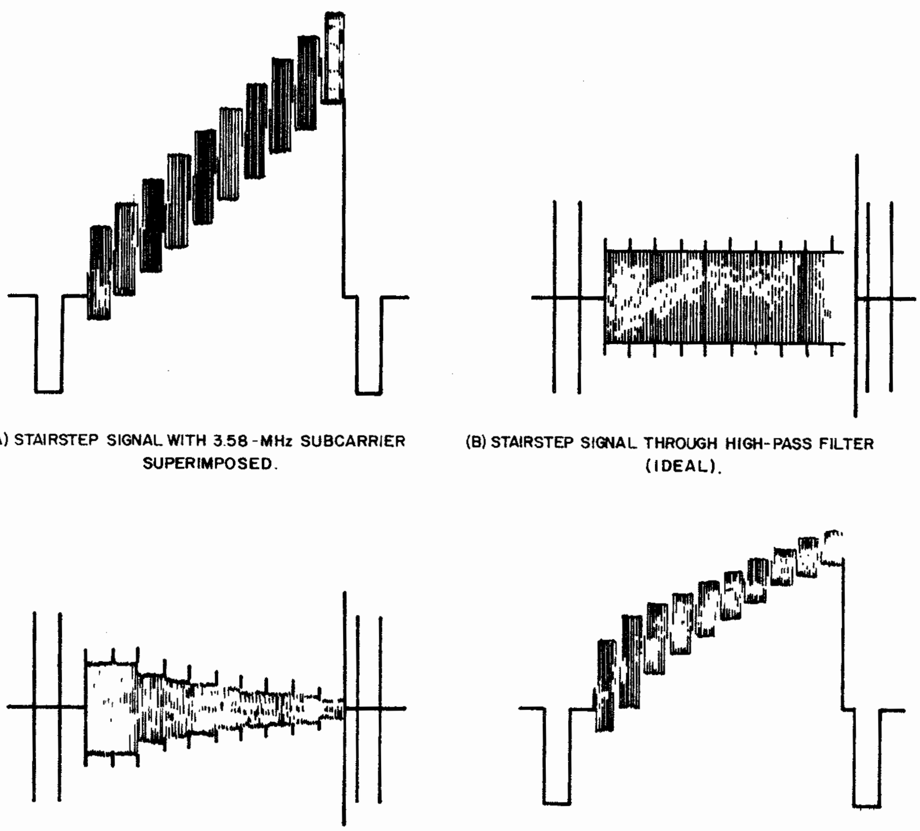


Figure 3-469. Stairstep Waveform, Showing White Compression

small high-frequency signal at two stated levels of a low-frequency signal on which it is superimposed. As a practical matter, the differential gain should be a minimum at all signal levels, therefore it should be measured between the extremes. Differential gain is measured when distortion due to nonlinearity of amplifiers is suspected. The usual method is to feed a stairstep signal to the unit under test with a 3.58-MHz subcarrier superimposed. Commerical generators provide this type of signal. The output of the unit under test is fed through a high-pass filter to an oscilloscope where only the 3.58-MHz signal is displayed. If the amplitudes of the sections of the filtered waveform, corresponding to the steps of the staircase waveform, are equal, then the unit under test has no differential gain. Figure 3-470 shows distorted and undistorted waveforms that may be encountered when differential-gain measurements are made. The gain of the oscilloscope may be increased to allow the most accurate measurement possible after the high-pass filter is switched in. The first measurement of differential gain is usually made at 50 percent average picture level (APL). This represents an average signal produced by a picture of normal blacks and whites. Often signals are not average; they may be nearly all white or nearly all black. This is especially true of industrial and scientific installations where lighting conditions cannot always be controlled. Therefore, to establish a true picture of system performance, these conditions must also be simulated. Nearly all white is simulated by transmitting the stairstep signal every fifth line with a 100-percent white signal during the intervening lines. This is designated as 90-percent APL on the test-signal generator. Conversely, the nearly all black signal is simulated by transmitting the stairstep every fifth line with a black signal in the intervening lines. Figure 3-471 shows an oscilloscope presentation of these waveforms with the high-pass filter out. The 10-percent tests are more



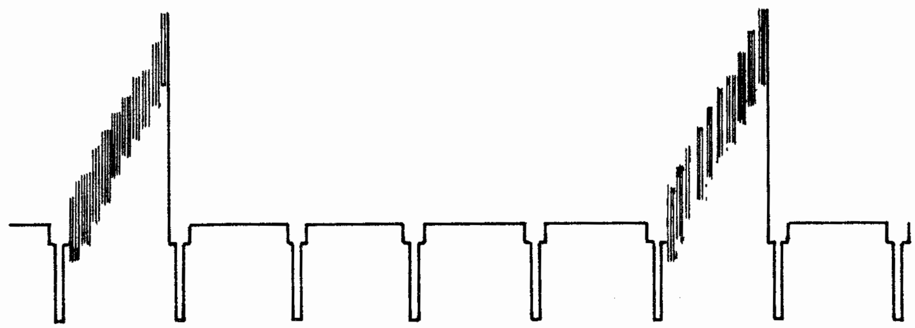
(A) STAIRSTEP SIGNAL WITH 3.58-MHz SUBCARRIER SUPERIMPOSED.

(B) STAIRSTEP SIGNAL THROUGH HIGH-PASS FILTER (IDEAL).

(C) DIFFERENTIAL GAIN IN WHITE REGION.

(D) WAVEFORM OF (C) WITHOUT HIGH-PASS FILTER.

Figure 3-470. Waveforms Encountered in Differential-gain Measurements



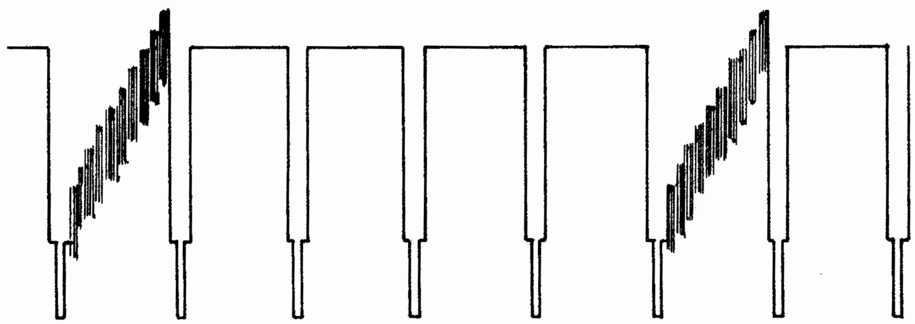


Figure 3-471. Stairstep Signal Simulating Various Scenes

severe than the 50-percent APL, and therefore will generally indicate higher levels of distortion. Differential phase is defined as the difference in phase shift experienced by a small high-frequency sine-wave signal at two stated levels of a low-frequency signal on which it is superimposed. The test waveform is also the stairstep signal with the 3.58-MHz subcarrier superimposed. The output signal from the system under test is then fed through the high-pass filter to the test-signal receiver, which in turn feeds an oscilloscope. In the test-signal receiver the 3-58-MHz component is matched against an internally generated 3.58-MHz signal. The output of the receiver is a signal proportional to the phase difference between the two signals. Another device that is useful for measuring differential-phase distortion is the vectorscope; an oscilloscope having a circular scan. Commercial units are also suited to monitoring waveforms and measuring differential gain.

3-10.22.6 Noise Measurements

Noise is defined as any spurious signal that is undesired. It is always present to some degree in any television channel but is annoying when excessive. Various types include thermal noise (generated within electrical components), transistor shot noise, 60-hertz ripple from a power supply, crosstalk from another camera, and unwanted signals induced by RF radiation into the channel of interest. These will be annoying in varying degrees, depending on their nature and frequency. Perhaps the most troublesome signal is crosstalk from another television system operating from a different sync generator. Television systems are also very sensitive to 60- and 120-hertz ripple which has the appearance of drifting through the picture. This can be made stationary by locking the sync generator to the 60-hertz power line, in which case the system is very tolerant of this type of interference. In general, stationary patterns are less objectionable than moving patterns, and low-frequency noise is more objectionable than high-frequency noise. Various methods have been proposed for noise "weighting" when making noise measurements; these involve the insertion of a filter ahead of the measuring device. The filter has increasing attenuation with increasing frequency, with rolloff beginning about 100 kHz. Attenuation at 3 MHz may vary from 10 to 20 dB. The purpose of the filter is to reference all noise signals to a level of supposed equal annoyance at a single frequency where it is measured. Thus, if a noise signal of 1 megahertz is estimated to be only one-half as annoying as one of the same amplitude at 50 kHz, the original signal is reduced to one-half its original amplitude for the measurement. A convenient approximation of one of these curves is the 1958 IRE rolloff curve, available on many oscilloscopes. In the absence of any standard curve, however, the value of the weighting filter is dubious. In any event, such a filter must be used for all noise measurements if comparisons are to be meaningful. A curve which may be valid for a 525-line television system will not be valid for a different scan rate. Signal-to-noise measurements in television systems are ordinarily expressed as the ratio (in dB) of peak-to-peak video to rms noise. Thus signal-to-noise ratio =

$$\frac{20 \log_{10} \text{peak-to-peak video}}{\text{RMS noise}}$$

By convention, rms noise is taken as 1/6 the peak-to-peak noise as measured on an oscilloscope. This is accurate enough for most field measurements. Measurement is ordinarily accomplished by first setting all system levels to their normal operating levels, using a window or other convenient test signal. The test signal is then removed, and the amplitude of the remaining signal (noise) is compared to the original signal level. In circuits such as camera-control units, which have black clippers, care must be taken to ensure that noise is not being elminated in the clipper circuitry.

3-11 RADIAC EQUIPMENT TESTING

3-11.1 GENERAL

Radiac equipment is designed for the detection and measurement of radioactivity. The short designation RADIAC is derived from RAdioactivity Detection Identification And Computation. Because radiac equipment is basically electronic in nature, it falls into the Navy System for design, maintenance and use of electronic equipment. In general, the testing of radiac equipment can be accomplished with the aid of test equipment available to the electronics technician. Some components, such as subminiature tubes, Geiger-Mueller tubes, and resistors of very high values cannot be thoroughly checked with ordinary testing equipment, and therefore require special consideration. A brief outline of the basic types of radiac equipment is presented first. The test equipment necessary to check the special components encountered in radiac equipment and to calibrate some radiacmeters is then discussed.

3-11.2 RADIAC FUNDAMENTALS

Radioactivity is the disintegration or breaking up of the atoms of an unstable element. Many chemical elements, such as uranium, radium, radon, etc, have natural radioactive properties. These elements emit (radiate) specific kinds of particles and waves in various quantities and intensities, depending on the nature of the element from which the emission originates. These particles and waves are emitted without the addition of any external energy to the element. Small amounts of different elements are continually created by the disintegration of these radioactive elements. In turn, these different elements disintegrate into still other elements. This decay process is not affected by such physical factors as pressure and temperature. The rate at which this decay occurs varies with each element. Thus, the half-life of radium (the length of time elapsed before radium loses one-half its original activity) is approximately 1600 years; the half-life of radon (a new element created by the decay of radium) is 3.82 days. Natural radioactive properties are shown usually by elements whose atomic weight is greater than that of lead (lead is the final stable element in the uranium series of disintegration). Under certain conditions, normally stable elements can be made artificially radioactive. These elements then have their own characteristic half-life of decay. The principal emanations from radioactive materials are four types of rays (particles or waves): alpha (α), beta particles (β), gamma waves (γ) and neutrons (η). Their characteristics and properties are important aids to their detection and measurement.

3-11.2.1 Alpha Particles

Alpha particles are helium nuclei (helium ions with a double positive charge). They have velocities up to about 7 percent of the speed of light (3×10^{10} cm/sec). These particles have short range, poor penetrating power, and very strong ionizing power. The two protons which make up the helium nuclei give the particle a positive charge equal to twice the negative charge of an electron.

3-11.2.2 Beta Particles

Beta particles are simply high-speed electrons. The charge of a beta particle is therefore negative. These particles can move with a speed almost equal to the speed of light (about 95 percent). Beta particles have strong ionizing power (about 1/100 that of the alpha particle), and are able to produce measurable effects after passing through shields 100 times the thickness required to stop alpha particles.

3-11.2.3 Gamma Waves

Gamma waves, or rays, constitute a type of electromagnetic radiation, similar to X rays, but in general have a much higher frequency of vibration and are far more penetrating. However, for the same wavelength of radiation the properties of the two types of rays are the same. Gamma rays are not particles and carry no charge. They are considered photons or bundles of electromagnetic energy similar to light waves. The wavelength of gamma rays is much shorter than that of light waves. In fact, this difference in wavelength is the main distinction between different types of electromagnetic radiation, including radio waves, radiant heat, infrared, visible light, ultraviolet, X-rays, gamma rays, and cosmic rays. The sequence given is in order of decreasing wavelength and increasing penetrating power. Gamma rays have mild ionization power (about 1/10,000 that of the alpha particle) and are intensely penetrating (about 10,000 times that of the alpha particle). These gamma rays can be detected after passing through 12 inches of steel.

3-11.2.4 Neutron Particles

The neutron is the neutral constituent of all stable nuclei, but is emitted during the fission cycle used in nuclear reactors or fusion used in many types of nuclear warheads. They have no electrical charge, hence, their name neutron (neutral).

3-11.3 RADIOACTIVE MATERIALS

The activity of a given quantity of radioactive materials is defined in terms of the number of alpha or beta disintegrations taking place over a predetermined unit of time. Thus, to be "curie" is defined as 3.7×10^{10} disintegrations/per second.

3-11.4 UNIT OF RADIATION MEASUREMENT

The unit of measurement of radiation is called the "roentgen", or "r"; and is defined for X rays or gamma rays as an exposure that releases 2.58×10^{-4} coulomb/kilogram of dry air. The "rad" is a unit of absorbed radiation dose, and is equal to 1/100 joule/kilogram of ionization energy. The term is used as a measure of radiation damage.

3-11.5 QUALITY FACTORS

The "Quality Factor" (QF) is a measure of biological cell damage resulting from an exact absorbed dose caused by a particular type of radiation, in relation to damage caused by the same amount of impinging electrons. Typical quality factors of various types of radiation include: alpha particle (10), fast neutron (10), slow neutron (5); X rays, gamma rays and Beta particles (1); etc. The "rem" (roentgen-equivalent-man) is a measure of dose equivalents received by man, and is equal to the product of rad x QF or rem = rad x QF.

3-11.6 RADIATION DETECTORS

3-11.6.1 Ionization Chamber

The ability of alpha, beta, and gamma radiation to ionize gases is the characteristic most frequently used to detect the presence of radiation. The ionization chamber is used for collecting the ions formed by the ionization of air by radiation. A cross-sectional view of the chamber and the associated schematic are shown in Figure 3-472. The chamber consists of two electrodes. The outer electrode (anode) is formed by an aquadag coating on the inner surface of the polystyrene walls. The inner electrode (cathode) is a rectangular loop of wire placed so that it is approximately 1-1/4 inches from the aquadag coating. The wire loop is insulated from the wall coating. The wall coating has an atomic number equal to about 7 or 8 (approximately the atomic number of air). The chamber is filled with air to a pressure of 760 millimeters of mercury (equal to the standard atmospheric pressure at sea level). The ionization chamber is sensitive to the effects of radiation. (Specially

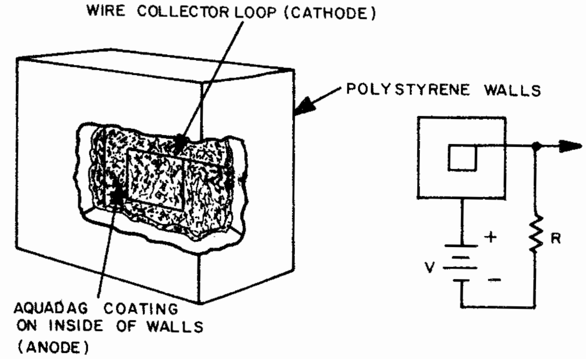


Figure 3-472. Ionization Chamber and Associated Circuit

constructed chambers are used for alpha and beta or gamma radiation measurement.) When ionizing rays enter the chamber, they collide with the gas atoms in the chamber. These collisions release electrons from the gas atoms and the gas becomes ionized. Under the influence of the electric field maintained between the two electrodes, the positive and negative ions move to the cathode and anode, respectively. The movement of the ions in the chamber results in a minute current flow. (With no radiation present, the chamber acts as an open circuit, and no current can flow.) Potential V is applied across the chamber through resistor R (Figure 3-472). Resistor R is a specially treated, high megohm resistance. The current flow through the ionization chamber is extremely small (a fraction of a microampere). This current flows through the extremely high resistance R and produces a potential difference that is fed to an amplifier. In an ionization chamber, the atoms of air are normally in a neutral state and are not affected by any potential difference between the electrodes of the chamber. When ionized, they are influenced by this potential. Ionization consists of removing one or more electrons from an atom. This electron is now termed a negative ion. The electron is emitted with enough energy to force additional electrons from other atoms by collision. However, each time the electron collides with an atom it loses energy and is thus slowed down and left with a reduced ion creating capacity. If no potential is applied to the ionization chamber, electrons released by the original ionizing event would eventually be slowed down to a point where they would be captured by the positive ion. The charge of the positive ion would be neutralized, and would thus become a netural atom. If a small voltage (for example 100 volts) is applied to the chamber, an electric field exists in the space between the positive aquadag coating on the walls and the negative collector loop. Positive ions formed by the ionization of the gas in the chamber tend to drift toward the negatively-charged collector loop; negative ions (electrons) tend to drift toward the positively-charged aquadag coating, since there is a tendency for some of the ions to recombine and form neutral atoms. The electrons lost by this recombination do not contribute to the final current flow in the chamber. The longer the time that the electrons require to reach the anode, the greater the possibility of recombination. If the applied voltage is increased to 160 volts, the electric field is sufficient to prevent recombination of the positive and negatively-charged collector loop of the ionization chamber, while all the negatively-charged electrons move toward the positively-charged aquadag coating. Thus, ionization within the chamber results in a current flow through the chamber. When the current in the chamber reaches the saturation value, further increases in applied voltage (within the limits of the ionization chamber region) do not increase the current flow (the gas amplification factor equals one). Thus, the chamber is operated over fairly wide limits of applied voltage without change in the current flow within the chamber. Variations in the ionization current result solely from changes in the intensity of the radiation.

3-11.6.2 Geiger-Mueller Tube

A simple device for radiation detection is a Geiger-Mueller (G-M) tube (Figure 3-473). The tube is filled with a gas mixture at low pressure. A thin wire, which constitutes the anode of the tube, is oriented axially to a cylinder and insulated from it. A voltage is impressed across the tube such that the wire is positive with respect to the cylinder. The magnitude of the impressed voltage is just below that necessary to ionize the gas molecules and cause conduction. In this dormant condition, no current flows. When radiation is present in the vicinity of the tube, an incoming radiation usually ionized some molecules of the gas within the tube. The ionized gas particles are attracted toward either the cylinder or the wire, depending on their way through the gas, these ionized gas particles collide with non-ionized gas molecules and ionize them. As a result, a large portion of the gas becomes ionized, thus producing a large current flow for only one initially created ion pair. Therefore the output from the tube is much greater per ionizing pulse than in the lower voltage ionization chamber. This current flow is quenched quickly, either by a small amount of organic vapor which is included in the gas mixture or by the use of external circuits which reduce the potential between the tube elements after conduction. As soon as tube conduction stops, the voltage across the tube is returned to the original pre-ignition value, and the tube

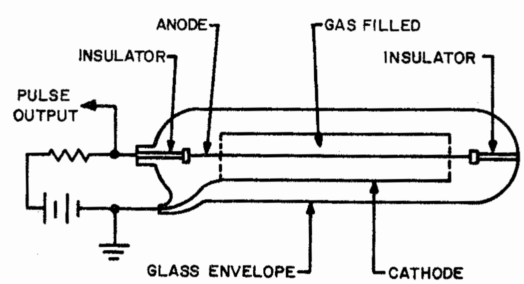


Figure 3-473. Typical Geiger-Mueller Tube

awaits the next ionizing event. The duration of tube conduction is short compared to the average time between ionizing events and, therefore, the tube output is in the form of a series of pulses. Because of the fluctuating intensity of the ionizing radiations, the random time interval between ionizing events, and the chance arrangement of the gas molecules in the G-M tube, the pulses produced by the tube vary in amplitude (one half volt to 50 volts) and duration (50 to 100 microseconds); and occur at random time intervals. These pulses are generally used to activate various indicating devices. The Navy's AN/PDR-27 and AN/PDR-43 series of radiation detectors employ the above methods of detection.

3-11.6.3 Scintillation Counters

Scintillation Counters (Figure 3-474) function in conjunction with a photomultiplier. This tube is activated by fluorescence produced when a charged particle strikes certain materials. The AN/PDR-56 Alpha radiation detector employs this method. The monitoring of the alpha radiation entails the detection of a highly ionizing particle having a short range. Alpha particles are easily absorbed, losing energy by excitation and ionization of the absorber atoms. This can be generally accomplished by a sheet of paper, by an aluminum foil 0.0015 inches thick, or by a few inches of air. If the particles emitted by an alpha source (in air) are counted by a scintillation method, it is found that their number decreases gradually up to a certain distance from the source, after which the reading falls off sharply. This distance is called the "range" and is related to the initial energy of the particle. The most energetic natural alpha particle, 10.5 MeV, has a range of only 4 inches in air. An alpha particle of greater than 7.5 MeV is necessary to penetrate the outer layer of dead skin. Since only a few short-lived isotopes emit alpha radiation of these energies, the biological hazard from alpha particles is strictly internal. For alpha radiation having energies between 4 and 7 MeV the particle range in air can be calculated from the formula:

$$R = 0.12E^{3/2}$$

Where R is the mean range of the particle in inches, and E is the energy of the particle in MeV. Typical ranges in air of the more important isotopes associated with nuclear energy are given in the following chart:

ISOTOPE ENERGY	MeV	RANGE INCHES
Productinium 231	5.00	1.35
Plutonium 239	5.15	1.4
Polonium 210	5.3	1.46
Thorium 230	4.68	1.21
Thorium 232	4.03	0.98
Uranium 234	4.76	1.25
Uranium 235	4.40	1.1
Uranium 238	4.2	1.03

The characteristics of alpha particles complicate their detection. Thin films of water or oil can partially or completely prevent alpha particles from being monitored. Porous surfaces (wood, concrete, soil, etc.)

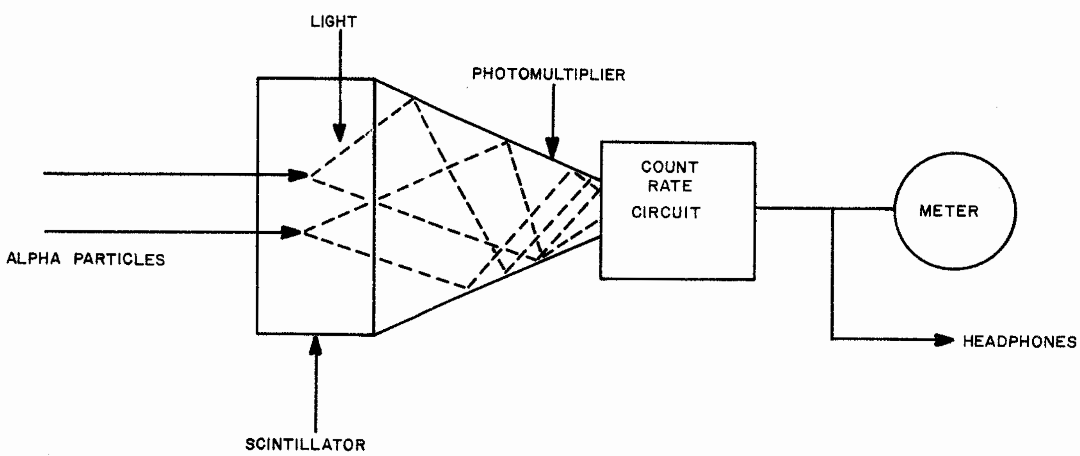


Figure 3-474. Typical Scintillation Counter

may harbor many times the activity measurable with monitoring instruments. Since the measuring instrument can only provide an indication of what it actually detects, appropriate consideration must be given in alpha surveys to the type and condition of surfaces being monitored in interpretation of meter reading. Most alpha survey instruments are calibrated for a 2 pi geometry under ideal conditions. A 2 pi geometry assumes that 50% of the radiations (in this case, alpha particles) are emitted in an upward direction toward the detecting probe, and that all the radiations so emitted under the sensitive area of the probe are detected and indicated on the meter. Since the efficiency of the instruments never attains 100%, the actual counts per minute as determined by the clicks in the headphone jack will always be less than the counts per minute indicated on the meter. Ideal conditions simply means that the source is thin, i.e., there is no significant self-absorption or absorption due to the other materials on or in the source.

3-11.6.4 Dosimeter

A typical radiacmeter (dosimeter) of the Radiacmeter-Dosimeter series IM-9C/PD and IM-C/PD and Radiacmeter-Dosimeter IM-9C/PD series is shown in Figure 3-475. Its function is to measure and indicate the accumulated dose of gamma radiation to which the wearer has been exposed. At one end of the radiacmeter is an optical eyepiece, and at the other end is the charging contract. The radiacmeter contains an ionization chamber into which is mounted a small electrometer. A scale, calibrated from zero to 200 milliroentgens, is mounted in such a manner that the amount of radiation to which the wearer was exposed since the charging of the electrometer can be read directly by holding the radiacmeter up to a source of light and looking into the eyepiece. A radiacdetector charger is required to charge the electrometer. The dosimeter is four inches long and of tubular construction. It is provided with a clip similar to those used on pencils and pens, and may be worn by personnel in a similar manner.

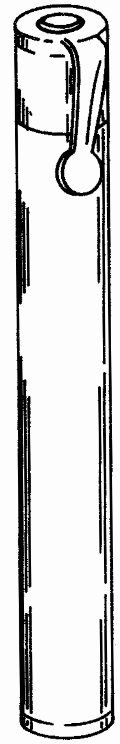


Figure 3-475. Radiacmeter-Dosimeter IM-9C/PD

3-12 SYNCHRO AND SERVO EQUIPMENT

Synchro equipment is used for remote indication or control by means of self-synchronizing motors. It consists of a series of synchro units which are used to electrically govern or follow the position of a mechanical indicator or device. The advantages of using an electrical synchro method rather than a mechanical arrangement include greater accuracy and simpler routing requirements for long-distance applications. The five general types of synchro units, classified according to function, are: transmitters, receivers, differential transmitters, differential receivers, and control transformers. However, if the power required to operate a device is large as compared with the power available from the controlling instrument (usually a synchro), power-amplifying means are provided. The term "servomechanism" refers to a large variety of power-amplifying devices. Servomechanisms are incorporated in synchro systems for such purposes as positioning guns, controlling radar antennas, and other automatic control applications where accuracy of remote reproduction is of primary importance.

3-12.1 GENERAL

General synchro types and servo equipment are discussed in the following subparagraphs.

3-12.1.1 Transmitter (Generator) Synchro

This unit, sometimes referred to as a synchro generator, consists of a rotor which carries a single winding, and a stator comprised of three windings displaced 120 degrees from one another.

voltages induced in the stator windings by the rotor windings represent the instantaneous angular position of the controlling shaft of the rotor. These voltages are used to control the position of the associated receiving synchro.

3-12.1.2 Receiver (Motor) Synchro

This unit, also known as a synchro motor, follower, or repeater, is similar electrically to the synchro transmitter, and is used in conjunction with a transmitter synchro. The receiver synchro takes the electrical signal voltage generated by the transmitter synchro. The receiver synchro rotor then follows in response to this voltage, so that its angular position corresponds to the postion of the transmitter synchro rotor. Mechanically, the receiver synchro differs from the transmitter synchro in that a damping device is incorporated in the former to prevent overshooting and hunting. For this reason, a transmitting synchro may not be used for receiving, although a receiver synchro may be used for transmitting.

3-12.1.3 Differential Synchros

Differential synchros are used in conjunction with transmitter and receiver synchros. A transmitting differential synchro inserts a correction voltage from the transmitter synchro to compensate for errors existing in various parts of the system. In effect, the angular position of the transmitter synchro rotor and the angular position of the differential synchro rotor are compared, and the sum or the difference of these two positions is transmitted to a receiver synchro. Whether the sum or the difference voltage of the differential synchro is employed depends upon the method used to connect the transmitter, differential, and receiver synchros. A receiving differential synchro indicates the angular sum or difference (depending upon the connections) between two transmitter positions.

3-12.1.4 Control Transformer (CT) Synchros

A control transformer synchro, known as a CT, is employed when it is desired to obtain a voltage proportional to the angle between the input position and the load position. As the angle between the input and the load becomes larger, the output voltage of the CT becomes larger. This output voltage is generally applied to a servo amplifier which provides the power to drive a servomotor. In addition, the CT output voltage is either in-phase, or out-of-phase, with the reference voltage, depending upon the direction (clockwise or counterclockwise) the input device is rotated. If the input device is rotated counterclockwise, the CT output voltage is generally in phase with the reference voltage. The phase relationship between the CT output voltage and the reference voltage determines the direction of rotation of the servomotor rotor to reposition the load. Several CTs may be connected to one synchro transmitter, and for this reason the stator windings of the CT must present a balanced, high impedance load to the transmitter. The rotor also serves as a high impedance circuit because the CT is not designed to deliver torque.

3-12.1.5 Synchro Capacitors

The differential synchro and the control transformer synchro both draw current from the synchro transmitter, even when the circuit reaches a position of electrical and mechanical balance. The differential synchro draws current because of the ratio of the step-up turns between the stator and the rotor. The control transformer synchro draws current from the transmitter because the control transformer synchro rotor is not energized and as a result, does not induce any voltage in the stator windings. To reduce current flow when either or both of these units are used, capacitors are connected into the circuit. Because the windings of the synchro are not a pure inductance, two currents exist in the windings: the resistive (loss) current, which represents the actual power loss in the circuit; and the inductive (out-of-phase) current, called the "magnetizing" (or exciting) current. When the loss current and the magnetizing current are added vectorally, the actual current lags by something less than 90 degrees. When the values of inductive reactance and resistance are known, a specific capacitance can be added to shunt the coils, since current leads in a capacitive circuit, so that the inductance-capacitance circuit will resonate at 60 Hzs, and the magnetizing current will be effectively cancelled by the capacitive current. In this state, the transmitter synchro will supply only the losses of the circuit.

3-12.1.6 Servo Circuits

A servo circuit is a more complex method of synchro control in which a controlled quantity is compared with an ordered quantity. The difference between the two quantities, known as the error, is used to govern the operation of a mechanical system. A great variety of servo methods are currently in use. These include electronic types, hydraulic types, amplidyne types, and many variations and combinations of these. All of these servo types are designed for a specific task; the applicable technical manual should therefore be referred to for specific testing and servicing instructions. To prevent oscillation, all servo methods have an anti-hunt feature embodied in their design. Electronic amplifiers used in servo circuits are classified according to the following four basic types:

Type 1: DC input, DC output
Type 2: DC input, AC output
Type 3: AC input, AC output
Type 4: AC input, DC output

For specialized uses, even these four basic types may be varied to suit a particular application.

3-12.2 SYNCHRO EQUIPMENT TESTING

Since synchros are employed to transfer angular shaft position to another synchro, usually some distance away, lengths of connecting bus and/or cable are used. Although the wiring may be clearly marked or color-coded, it is advisable to check these designations if a synchro device gives evidence of improper operation. This is particularly important if a new installation is being checked out, or if an installation has been repaired or overhauled.

3-12.2.1 Overload Indicators

An overload in a synchro circuit is usually caused by worn bearings or defective gears at the receiver synchro. This condition causes the receiver rotor to lag the transmitter rotor, allowing excess current flow in the stator leads. To detect this condition, it is necessary to measure the current in at least two of the stator leads. This is because synchro design makes it possible for one stator lead to indicate zero current while the other two leads are drawing excessive current. The usual procedures and precautions should be followed when making these measurements. Some synchro circuits may have an overload indicator included in the installation. This method uses two current transformers, with the primary of each being connected in series with a stator lead. The secondary windings of these two transformers are connected in series-aiding, and the two remaining secondary leads are connected to a neon bulb. The secondary windings of these two transformers are designed such that the neon bulb fires when a predetermined imbalance in current is present in the two stator leads. The neon bulbs are usually mounted on the control switchboard of the equipment.

3-12.2.2 Blown-Fuse Indicators

Some synchro circuits may incorporate a blown-fuse indicator, usually consisting of a transformer with two primary windings and one secondary winding. The primary power is connected to one primary winding, and the synchro excitation voltage is taken from the other primary winding. The leads of one primary are connected to the leads of the other primary by fused jumpers. The phasing is such that the voltages in the two windings oppose each other, thus normally no voltage is induced in the secondary. If one of the fuses blows, the primary winding connected to the primary power induces a voltage in the secondary winding. This secondary winding is connected to a neon bulb, which glows in response to the voltage, indicating a blown fuse.

3-12.2.3 Voltage and Resistance Measurements

The quickest method of locating opens and shorts in synchro units and their associated wiring is by performing resistance measurements. Since most synchros work in pairs, it can be assumed that, within close tolerances, the resistances of both the rotors and the stators will show the same reading. If the resistances should vary widely, the trouble may be easily located. Typical resistance values for synchros may run from a fraction of an ohm for the large synchros to a few hundred ohms for the smaller ones. Synchro resistance must not be measured without first shutting off all excitation voltage in the synchro rotors. An excellent method for detecting open or shorted stator windings utilizes a voltmeter connected across any two of the stator windings. As the angle of the transmitter is varied, a smooth variation between 0 and 90 volts should be indicated by the voltmeter (assuming a 110-volt synchro). Open or short-circuited stator or rotor leads may be detected by measuring with a voltmeter or by measuring the resistance of the suspected part.

3-12.2.4 Symptoms of Incorrect Wiring

In new installations and after repairs or overhauls in synchro circuits, the crossing of buses or wires is frequently the cause of improper synchro operation. Reference to Figure 3-476 should identify the wiring fault that is causing improper operation of the synchro circuit.

3-12.2.5 Symptoms of Open-Circuited and Short-Circuited Wiring

Typical troubles involving open-circuited and short-circuited wiring, with associated symptoms, are listed in Table 3-29.

3-12.3 SYNCHRO ZEROING METHODS

In any synchro circuit, it is important that all the synchros be electrically zeroed. Since different types of synchros must be zeroed by different methods, each type of synchro zeroing will be discussed separately.

FOR ALL WIRING TROUBLES SHOWN HERE, RECEIVER GIVES WRONG INDICATION OR TURNS IN A REVERSE DIRECTION. TORQUE NORMAL. OVERLOAD. NO OVERHEATING.

SPECIFIC SYMPTOMS		WIRING	TROUBLE	SPECIFIC SYMPTOMS		WIRING	TROUBLE
TRANSMITTER SET ON 0° AND TURNED CCW	RECEIVER READS AND TURNS AS INDICATED			TRANSMITTER SET ON 0° AND TURNED CCW	RECEIVER READS AND TURNS AS INDICATED		
0°	0°		NO TROUBLE SYSTEM OPERATES NORMALLY	0°	300°		R1 AND R2 REVERSED S2 AND S3 REVERSED
0°	240°		S1 AND S2 REVERSED	0°	180°		R1 AND R2 REVERSED S1 AND S3 REVERSED
0°	120°		S2 AND S3 REVERSED	0°	120°		S1 TO S2 S2 TO S3 S3 TO S1
0°	0°		S1 AND S3 REVERSED	0°	240°		S1 TO S3 S2 TO S1 S3 TO S2
0°	180°		R1 AND R2 REVERSED	0°	300°		R1 AND R2 REVERSED S1 TO S2 S2 TO S1 S3 TO S1
0°	60°		R1 AND R2 REVERSED S1 AND S2 REVERSED	0°	60°		R1 AND R2 REVERSED S1 TO S3 S2 TO S1 S3 TO S2

Figure 3-476. Incorrect Synchro Connections Causing Receiver to Operate in Wrong Direction or Give Improper Indication

Table 3-29. Trouble-Shooting Synchro Circuits

SYMPTOMS	POSSIBLE CAUSE OF TROUBLE	REMEDY
Receiver rotor is either in correspondence with transmitter or displaced 180 deg, but follows in proper direction. Stator voltages vary from 0 to 90 volts. Both rotor voltages are 115 volts. Same as above except that one rotor voltage is 15 volts and the other is 90 volts.	Rotor winding is open, connection to slip ring is open, or brush is not making contact. Supply line is open to rotor, and reading 90 volts; the 90 volts exists across the rotor by virtue of transformer action.	If sure that trouble is not in ring connection or brush, replace the unit. Locate open in supply line, and repair.
Voltages between pair of stator wires is zero for all transmitter positions. Other stator-lead voltages read from 0 to 90 volts. Both rotor voltages are 115 volts.	Pair of stator leads which read 0 volts is short-circuited. For motor behavior, see Figure 3-476. The three stator wires are short-circuited together.	Remove short circuit from wiring or inter-connecting switches. If the trouble is internal, the unit may require replacement. Locate deffective wiring or switches, and repair.
Both transmitter and receiver units hum and heat excessively. Receiver either does not follow, or may spin. Sudden change in transmitter rotor positon causes oscillation at receiver or a spinning effect.	Inertia damper is jammed tight on receiver rotor shaft. Absence of damper indicates that transmitter unit is being used as a receiver.	Free the damper if it is jammed. If transmitter has been used, replace with correct receiver unit.
Intermittent operation.	Corroded rings, defective brushes, loose connections. Stator wiring incorrect. See Figure 3-476 for specific symptoms.	Clean the rings, install new brushes, tighten loose terminals, etc. Correct stator wiring.
Torque normal. Receiver lags or leads transmitter, or may turn in either proper direction or reverse direction. Torque normal. Receiver follows transmitter, but is displaced 180 deg.	Rotor connections are reversed.	Correct wiring at proper unit.
Receiver shows large error, and lags transmitter. Connections normal, but excessive current flows, producing overload indication.	Bearings are frozen or partially frozen because of improper lubrication.	Replace unit, since bearing trouble usually damages other parts of unit.

CAUTION

In some of the methods of electrical zeroing described in the following paragraphs, it should be noted that 115 volts is applied directly to the stator coils. Since stator coils are normally designed for a maximum of 90 volts (across any two coils), the line voltage should be applied for only short periods of time. When it is necessary to apply voltage for a considerable length of time, 78 volts should be used.

3-12.3.1 Zeroing Receiver Synchros

Since a receiver synchro is usually free to turn, the jumper method of zeroing is usually employed. To zero a receiver synchro, the voltage between S1 and S3 must be made zero, and the phase of the voltage at S2 should be the same as the phase at R1. This is easily done by connecting S1 to S3, using a jumper wire, and connecting S2 to R1, using a jumper wire. This method is shown in Figure 3-477. When the power is applied, the rotor will line up in the zero position. If the indicator does not point to zero on the dial, the synchro should be loosened in its mounting and rotated until its dial reads zero. A second method, using a voltmeter, may also be employed for electrically zeroing the receiver synchro. Since this method is the preferred method for zeroing transmitter synchros, the procedure is described in the following paragraph and is illustrated in Figure 3-478.

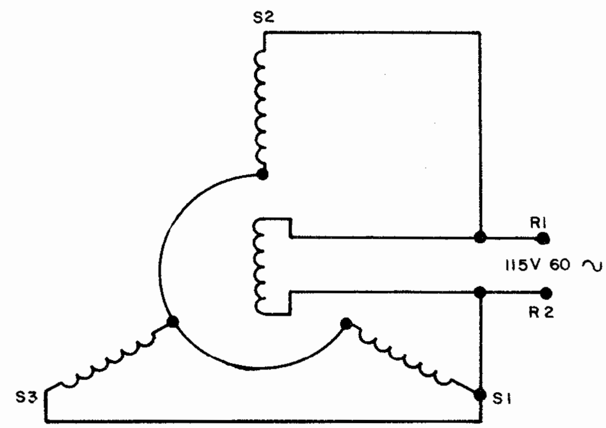


Figure 3-477. Electrically Zeroing a Receiver Synchro, Using the Jumper Method

3-12.3.2 Zeroing Transmitter Synchros

To zero a transmitter synchro, connect an AC voltmeter between S1 and S3, as shown in part (A) of Figure 3-478. Rotate the energized rotor until a zero reading is obtained on the voltmeter. Since the rotor at its zero-degree and 180-degree positions will produce this zero reading, it will be necessary to determine whether the phase of S2 is the same as that of R1. Make the connections shown in part (B) of Figure 3-478. If the proper polarity relationships exist, the voltmeter will indicate less than the line voltage being applied to the rotor. If the indication is greater than the line voltage, the rotor must be rotated 180 degrees

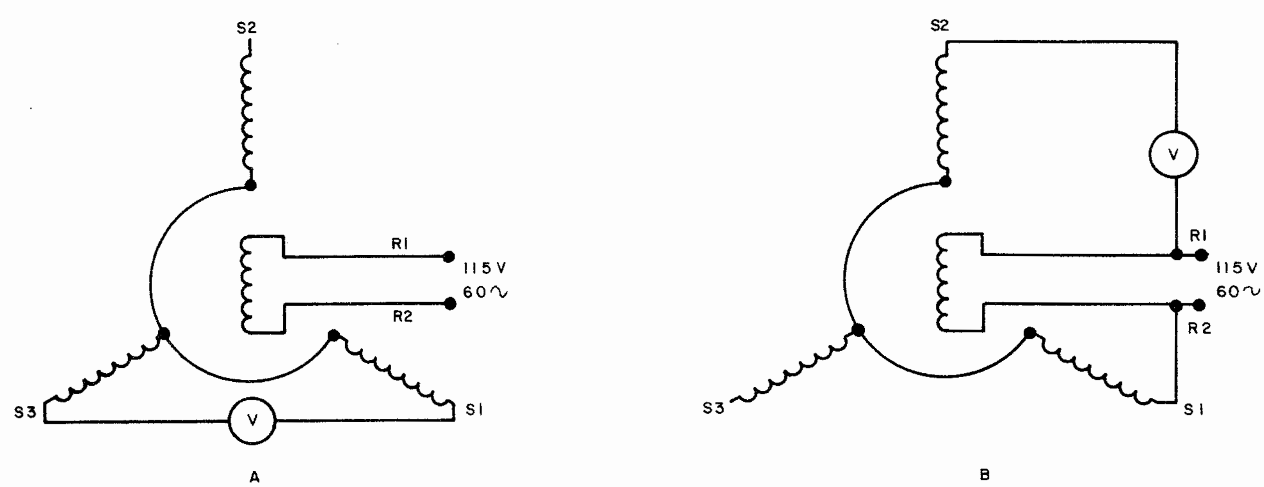


Figure 3-478. Electrically Zeroing Transmitter and Receiver Synchros, Using the Voltmeter Method

and the previous step. as shown in part (A) of Figure 3-478, performed again. The pointer connected to the rotor should be adjusted to indicate zero.

3-12.3.3 Zeroing Differential Transmitter Synchros

Because the differential transmitter synchro is usually employed to insert a correction voltage into a synchro circuit, it is normally driven either directly or through a gear train. Before zeroing the differential transmitter synchro, the unit whose position the differential synchro transmits should first be zeroed. After this has been done, connect the differential synchro as shown in part (A) of Figure 3-479. The synchro should then be turned in its mounting until the voltmeter shows a minimum indication. After completing this step, make the connections shown in part (B) of Figure 3-479. Again turn the synchro slightly in its mounting until minimum voltage is indicated by the voltmeter.

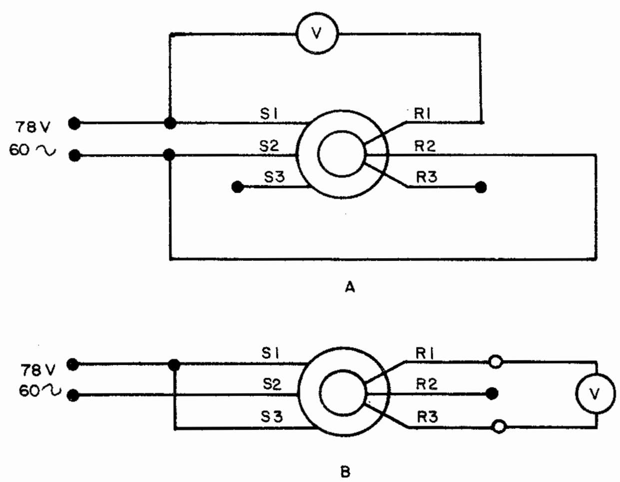


Figure 3-479. Electrically Zeroing a Differential Synchro Transmitter, Using the Voltmeter Method

3-12.3.4 Zeroing Differential Receiver Synchros

To zero a differential receiver synchro, make the connections shown in Figure 3-480. As soon as the power is applied to the synchro, the rotor will assume a position of electrical zero. The dial can then be set at zero and the unit reconnected to its circuit.

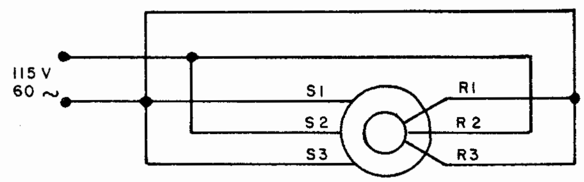


Figure 3-480. Electrically Zeroing a Differential Synchro Receiver

3-12.3.5 Zeroing Control Transformer Synchros

To zero a control transformer synchro, connect it as shown in part (A) of Figure 3-481. Apply power and turn the synchro in its mounting for minimum reading on the voltmeter. Then connect the control transformer synchro as shown in part (B) of Figure 3-481 and again turn the synchro slightly in its mounting, in either direction, for minimum indication on the voltmeter.

3-12.4 SYNCHRO TESTING

Two methods of synchro testing are described in the following subparagraphs in the order of preference.

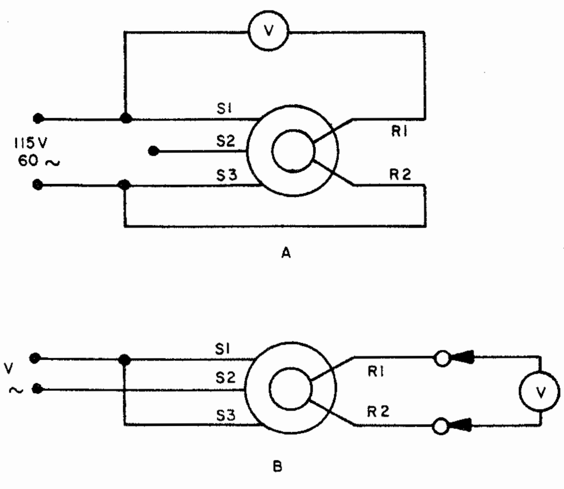


Figure 3-481. Electrically Zeroing A Control Transformer Synchro

3-12.4.1 Angle Position Indicator

Synchro testing and zeroing can be accomplished easily by using the angle position indicator (API). The API is a portable, solid state, electronic device that is connected to the synchro by a five-conductor cable (six conductors for resolvers) for terminals S1, S2, S3, R1 and R2. The API uses an accurate analog-to-digital converter to convert the analog angle of the synchro to an equivalent digital angle for display. The accuracy is typically $\pm$ 6 minutes of arc. When connected, the API will read zero for a properly zeroed synchro, and will increase from zero to 359° as the synchro shaft is rotated counterclockwise. This device greatly facilitates sychro troubleshooting.

3-12.4.2 Standard Test Synchros

A standard test synchro is used for performing various operational tests on synchro circuits. It may also be used for various kinds of checks and for troubleshooting. A standard test synchro is a small, precision synchro, mounted in an instrument case. It is equipped with a standard dial (numbers increasing in the clockwise direction), which moves past an engraved index. When the synchro is being used as a transmitter, a braking arrangement applies friction to the shaft. When the synchro is being used for receiving, the brake is released to allow the shaft to turn freely.

3-13 AUTOMATIC CONTROL SYSTEMS

There are a number of automatic control systems used for various purposes. Electronic systems are used, generally, to automatically control frequency, gain, load and other characteristics associated with receivers and transmitters. These types of automatic control devices are discussed in appropriate places in this manual with the equipment with which they are functionally used. This section is especially concerned with those automatic control systems primarily used for remote control of motors, generators and servos, usually located more remotely then the primary equipment it controls. Position and velocity are two quantities most often controlled by servo systems. Operation may be by electronic servomechanisms, hydraulic or other special type of control devices. Present discussion is limited to those types of equipment in most prevalent use, and does not include all types, but is intended for use as a general guide in the testing of this type of equipment.

3-13.1 ELECTRONIC SERVOMECHANISM CONTROLS

There are a great number of servo circuits (servomechanisms) in which power is supplied to a driving motor from the output of a servo amplifier. Electronic control methods are generally employed where large amounts of torque are not required. Both DC and AC servomotors may be controlled directly from the output of electronic servo amplifiers.

3-13.1.1 DC Servomotor Method

Permanent magnet DC servomotors are often used when the load to be positioned is light. The schematic in Figure 3-482 shows a DC motor whose direction is controlled by the application of a control voltage at either point 1 or point 2. If a positive input voltage is applied to point 1, transistors Q1 and Q3 become switched on, and current flows through the path a, b, c, d; causing the motor to turn clockwise. On the other hand, applying a positive input voltage to point 2, will cause transistors Q2 and Q4 to be switched on, and current will then flow through the motor in the opposite direction by the path a, c, b, d; causing counterclockwise rotation of the motor and associated load. In each case the load will be geared to an electromechanical device which will reduce the input (error) signal to zero volts as the load reaches the required position.

3-13.1.2 AC Servomotor Method

Servomotors employing AC are extensively used in positioning servo systems. The circuit shown in Figure 3-483 shows how one of these systems is connected and utilized. An explanation of circuit functions is given in the following subparagraphs.

3-13.1.2.1 Input Error Signal

The amplitude of the input error voltage from the CT determines how fast the AC servomotor drives toward null. The phase relationship between the input error voltage and the reference (in-phase or out-of-phase) voltage determines the direction of rotation of the servomotor toward the null (desired load position).

3-13.1.2.2 Diode Limiter Circuit

The error signal inputs to a diode limiter consisting of CR1 and CR2. Connected as shown, CR1 and CR2 scale the input error voltage to that voltage required to saturate output transistors Q5 and Q6, thereby driving the AC servomotor, B1, as rapidly as possible toward null.

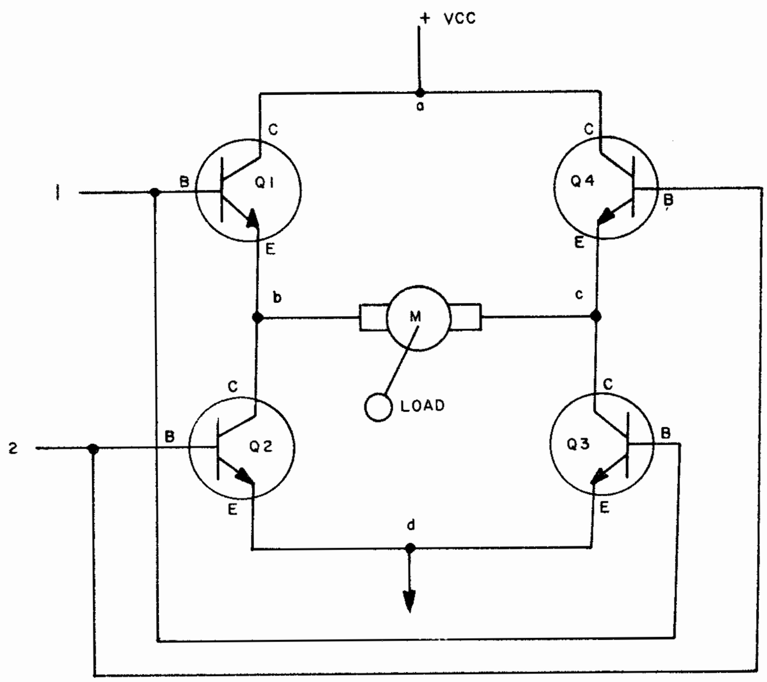


Figure 3-482. DC Servomotor Simplified Schematic

3-13.1.2.3 Differential Amplifier Circuit

As the error voltage on the base of Q1 swings positive with respect to signal ground, Q1 goes out of conduction and the current through R1 and R6 decreases. The decrease in Q1 conduction causes Q1 collector voltage (and hence the base of Q3) to go negative toward -Vcc and Q1 emitter voltage (and hence the emitter voltage of Q2) to go positive. Since Q2 is connected with its base of AC common, no phase reversal occurs between emitter and collector. The positive-going signal of the Q2 emitter is transferred to its collector as a positive-going signal, and hence to the base of the other push-pull transistor, Q4. Therefore, a positive-going pulse on the emitter of Q1 transfers to a negative-going pulse on the base of Q3 and a positive-going pulse to the base of Q4. These are the conditions necessary for proper action of the class B push-pull circuit consisting of Q3 and Q4.

3-13.1.2.4 Class B Push-Pull Circuit

As the pulse on the base of Q3 goes negative, Q3 conducts, thus drawing current from -Vcc2 through the upper half of T1. Transistor Q4 is not conducting at this time, due to the positive-going pulse on its base. However, when the input signal reverses direction and goes negative, Q3 is cut off and Q4 then conducts, drawing current -Vcc2 through the lower half of T1. This reciprocating action causes one half of transformer T1 to be drawing current at any time an error signal is present and represents the push-pull output. The transformer, T1 performs a 180° signal reversal from the push-pull circuit to the motor driver circuit consisting of Q5 and Q6.

3-13.1.2.5 Motor Drive Circuit

The driver circuit consisting of Q5 and Q6 is also a class B push-pull amplifier. However, these transistors have a high enough voltage and current rating to drive the AC servomotor control winding.

3-13.1.2.6 Gear Train

The load and input CT are connected to the motor (B1) by a gear train. This gear arrangement repositions the load to the desired angle and drives the CT rotor until the input signal goes to zero (null). When the input error signal goes to zero, no

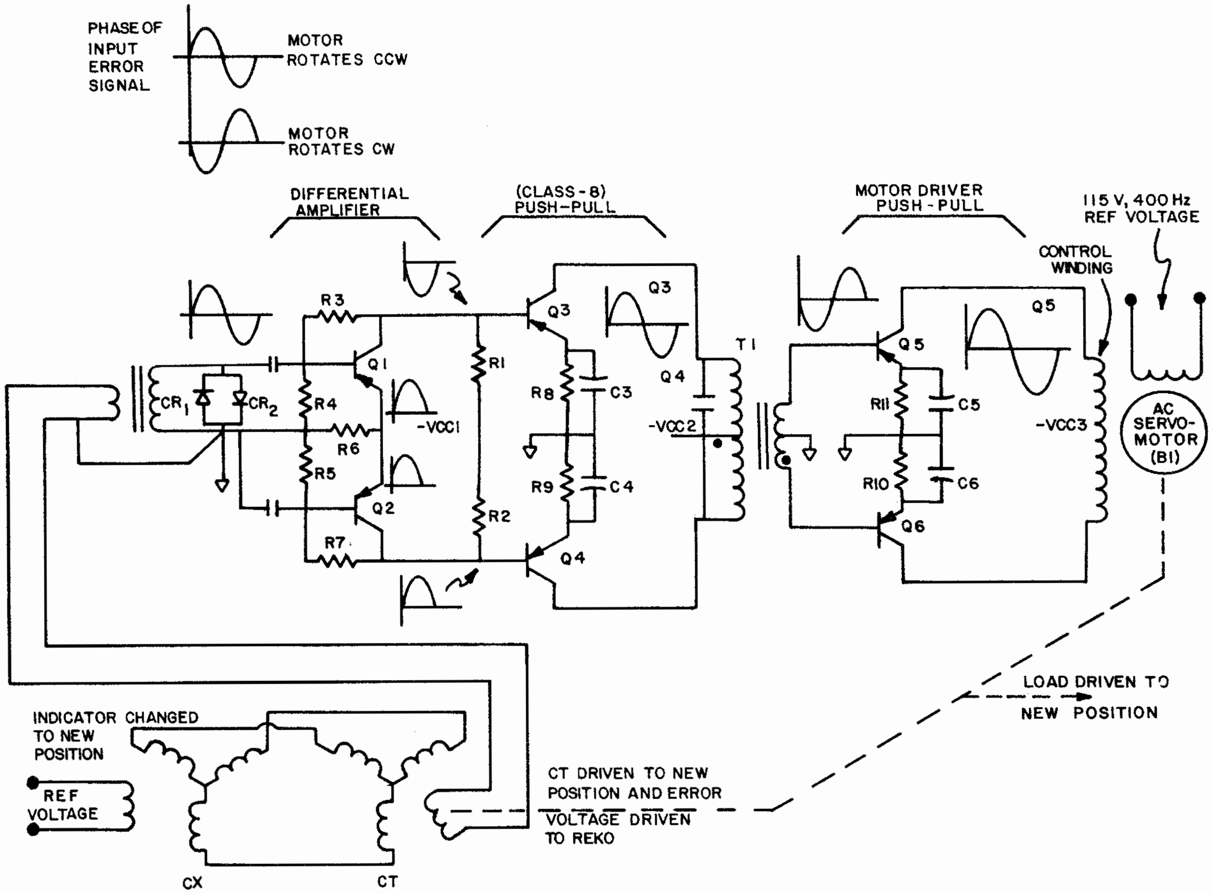


Figure 3-483. Typical Positioning Servo System

signal is applied to the push-pull circuits, therefore no drive signal is applied to the motor. The motor is usually internally damped to reduce hunting or oscillations when the motor is not being used.

3-13.2 AMPLIDYNE SERVOMECHANISM CONTROL METHOD

The amplidyne motor-generator consists of a constant-speed AC drive motor and a two-stage electromechanical power amplifier, contained in a single housing. The drive motor, which may be of the squirrel-cage induction type, has its rotor shaft coupled to the armature of the generator section. Since this motor drive mechanism is similar to that of other servo drive mechanisms, it can be considered as conventional and, therefore, self-explanatory. The amplidyne section, however, is radically different from the conventional generator in the unusual method it employes to obtain high power amplification. The step-by-step development of the amplidyne principle is described in the paragraphs that follow. A cross section of a conventional DC generator is diagrammed in Figure 3-484. In this representation, a load is shown as drawing a current of 60 amperes from the generator armature. To meet this demand, the armature must receive a voltage of sufficient magnitude to produce the necessary current. To provide the proper flux density to produce this current, a field excitation current of three amperes is necessary. The generator may now be considered as an amplifier having a current gain of 20. Since the direction of the excitation flux, ϕ^e, is from north to south (Figure 3-484), flux will pass through the amature core in a horizontal direction, as indicated by the arrows. When the external armature circuit is completed, causing a

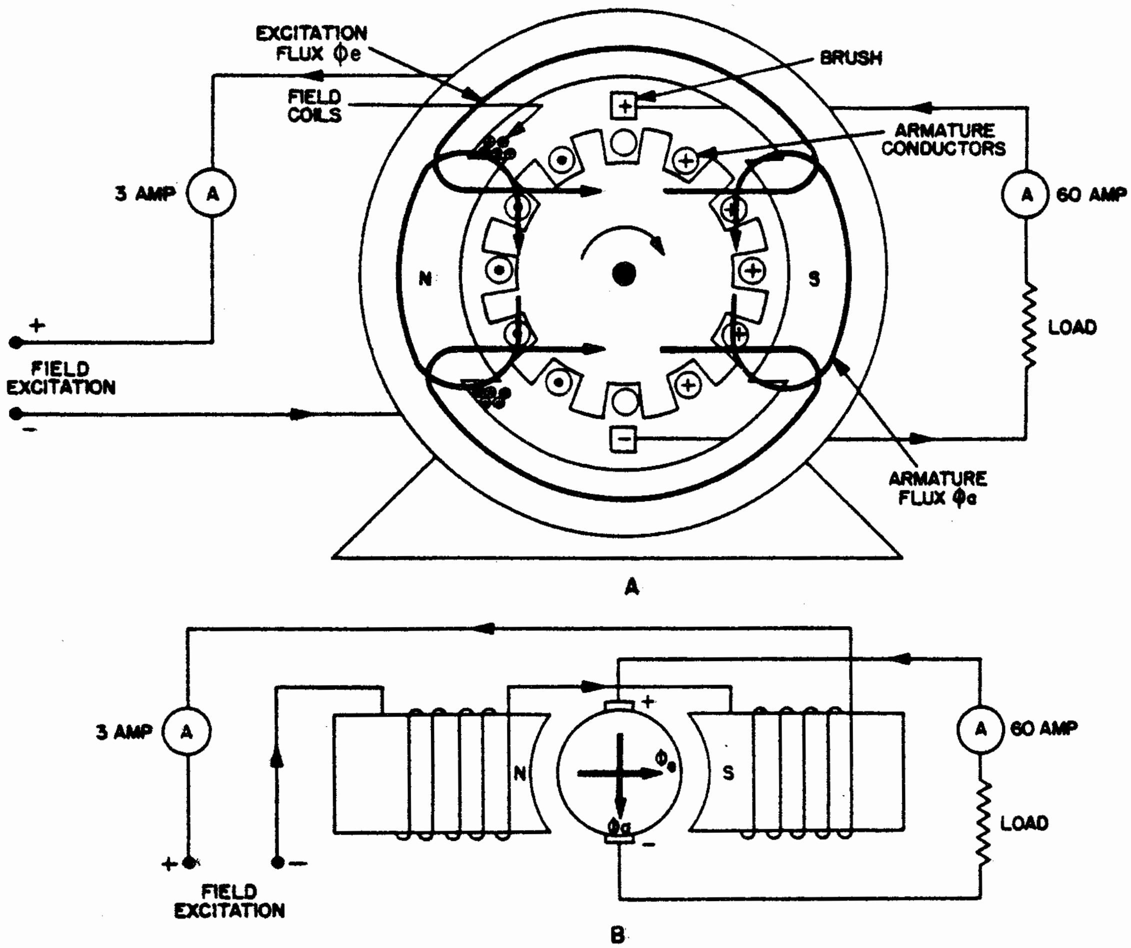


Figure 3-484. Magnetic Field and Current Relationship in Conventional DC Generator

current flow of 60 amperes, the armature (being wound on an iron core) acts as an electromagnet. This action gives rise to an armature reaction flux, designated as ϕ_a, in the diagram. If the left-hand rule is applied to the current flow in the armature conductors, the armature flux, ϕ_a, is shown to be at right angles to the excitation flux, ϕ_e. A simplified version of the circuit is shown in part (B) of Figure 3-484 with the flux directions and magnitudes indicated. Figure 3-485 is the same as Figure 3-484 except that, in this case, the load has been removed and the armature leads have been short-circuited. Since the resistance of the load is no longer a factor, the only opposition offered to current flow is the low resistance of the armature windings, plus the negligible resistance of the short-circuit wiring. The immediate result of such a short-circuiting procedure would be to increase the armature current to an abnormally high value. The end result would be a burned-out armature. However, one way of reducing the enormous armature current would be to reduce the excitation flux to a much lower level. It is apparent that this flux could be made weak enough so that the short-circuit current in the armature circuit could be reduced to 60 amperes (which it was in the previous example). Moreover, since the armature easily handled a 60-ampere drain in its loaded condition, a short-circuit current of the same value cannot damage it. A reduction in field excitation current will weaken the flux to the proper strength; in this case, the current has been dropped to

0.03 ampere. Consequently, it may be seen that a field current of 0.03 ampere controls a short-circuited armature current of 60 amperes, whereas 3 amperes were necessary to control the same amount of output current in the loaded state. It is clear, then, that the generator current gain has been increased to 2000. The problem now arises as to how the increased current gain can be put to use. Obviously, the load cannot be placed in series with the short circuit, since this would mean a return to the original status. The short circuit, therefore, must remain intact. From part (B) of Figure 3-485, it is evident that two fluxes exist- a weak field flux (ϕ_e) and a strong armature flux, (ϕ_a). The latter is created by a heavy current of 60 amperes. The cross sections in the figure show that the armature conductors are evenly spaced around the core; therefore, the conductors will cut the heavy armature (ϕ_a) at the same rate as they cut the excitation flux (ϕ_e). However, the maximum voltage induced in the conductors as they cut the armature flux appears across the armature at right angles to the voltage induced by the excitation flux. To take advantage of this new voltage, caused by the fact that the armature conductors cut their own reaction flux, a second pair of brushes (Figure 3-486) is placed on the commutator at right angles to the short-circuited brushes, and are connected to the load. Since a high voltage is realized as the armature conductors cut the strong reaction flux (ϕ_a), the voltage developed across the output brushes is sufficient to supply a large current (e.g., 60 amperes) to the load, despite the resistance of the load circuit. Another problem is encountered here, as shown by the diagram. Just as the armature current in the short-circuited section

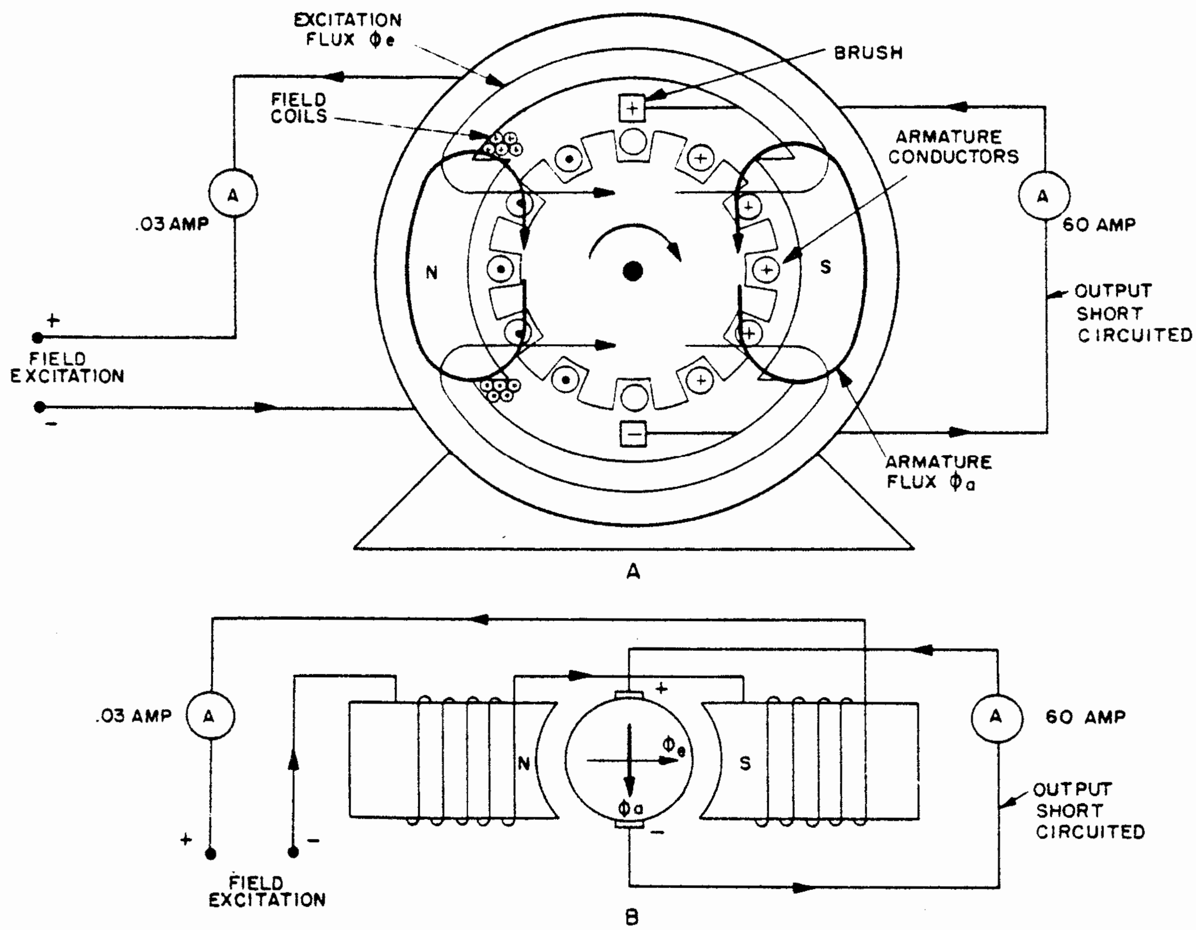


Figure 3-485. Magnetic Field and Current Relationship in Short-Circuited DC Generator

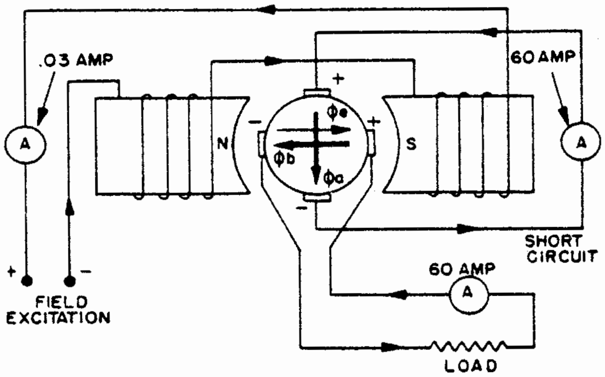


Figure 3-486. Short-Circuited DC Generator Supplied with Additional Brushes

creates a flux at 90 degrees to the armature flux. This new reaction flux (ϕ_b) is therefore removed 180 degrees from the original, much stronger excitation flux (ϕ_e). Since it is in opposition to ϕ_e, the excitation flux could no longer control the output. To overcome this obstacle, a compensating winding is placed on the field pole pieces, and is connected in series with the output to the load. In design considerations, the number of turns in this compensating winding is calculated and the direction of current flow determined, so that a compensating flux (ϕ_c) exactly equal and opposite to flux ϕ_b, is developed. The compensating winding and the four flux components are represented in Figure 3-487. Inasmuch as ϕ_b and ϕ_c cancel each other, the resulting fluxes are ϕe and ϕa, as indicated in the amplidyne equivalent circuit of Figure 3-488. The original build-up of a voltage in a conventional self-excited DC generator depends upon the presence of a certain amount of residual magnetism in the field pole-pieces. Some residual magnetism remains in the amplidyne field poles after the field excitation current is reduced to zero, which is the proper value when the input and output shafts of the servo mechanism are in correspondence. The presence of residual magnetism would cause a weak field flux, resulting in the induction of an appreciable voltage in the armature because of the amplidyne's high gain. This, in turn, would have the undesirable effect of causing rotation of the servomotor when no error exists. To eliminate the residual magnetism, a small AC generator is used. This generator may consist of a permanent magnet mounted on the amplidyne armature, and revolving in a small field-winding. The output of the AC generator is applied to two sets of opposed windings placed on the field pole-pieces. The effect of these windings is to neutralize the residual magnetism that exists when the field excitation current is zero. The amplidyne generator can be compared to a two-stage vacuum-tube amplifier. The development of the short-circuit current, with its accompanying armature reaction flux by means of a small excitation current, constitutes the first stage of amplification. The strong flux developed by the short-circuit current

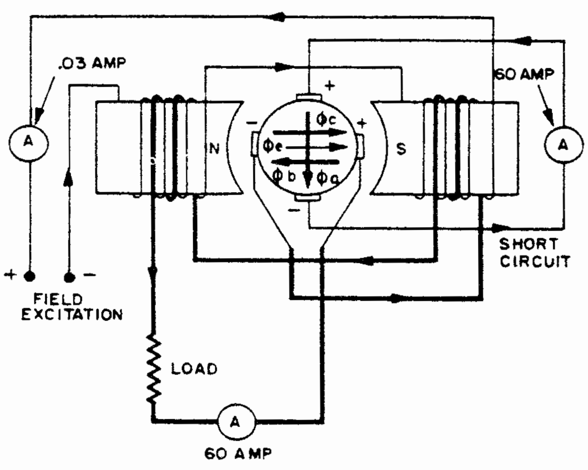


Figure 3-487. Short-Circuited DC Generator with Additional Brushes and Compensating Windings

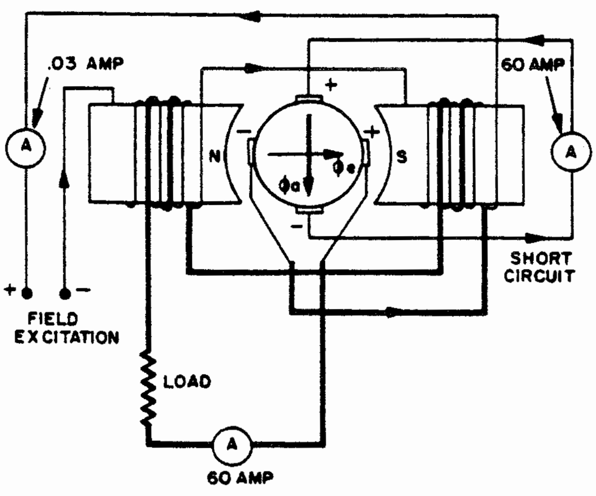


Figure 3-488. Amplidyne Generator Equivalent Circuit, Showing Effective Magnetic Field Amplification

produces a voltage high enough to cause an equally large current to flow through a load circuit. This represents the second stage of the system. The second reaction flux ϕ_b, is analogous to negative feedback, or degeneration, in a tube circuit. The compensating winding can be regarded as a regenerative circuit, designed to exactly balance the degeneration so that the net feedback in the amplidyne is zero. Because of its power gain of approximately 10,000, the amplidyne has extensive use in servomechanisms.

3-13.3 HYDRAULIC SERVOMECHANISM CONTROLS

Hydraulic servomechanisms are used in some radar equipment for antenna positioning of guns and other ordnance equipment. Smaller hydraulic servos find extensive use in specialized military control equipment. The hydraulic servomechanism is chosen for many applications where a rapid response, combined with smooth operation is required. For purposes of explanation, the hydraulic servo components mentioned in the following description are simplified. In actual practice such devices embody many complex refinements. Some of these include dither mechanisms, error correctors, coarse-fine data assemblies with accompanying contact-ring-relay transfer arrangements, limit switches, etc.

3-13.3.1 Variable-Flow Pump

Figure 3-489 shows a type of variable-flow pump similar to that illustrated in the block diagram of Figure 3-490. Figure 3-489 is a cutaway side view of pump mechanism A, a bottom view of cylinder-piston assembly B, and a bottom view of valve plate C. The solid-line portion of A shows the pump housing depressed 30 degrees below the drive shaft. In the dotted-line portion, the housing is raised 30 degrees above shaft axis. Only two pistons and cylinders are indicated in the side view of the sketch. However, there are actually six pistons (in this example), equally spaced in the cylinder block, as shown in B. In analyzing the operation of the pump, it is assumed that the AC motor is turning the circular drive plate and the rotating cylinder assembly in the clockwise direction, as indicated by the arrow. At the instant shown, piston 1 is at the top of its stroke and its cylinder is filled with oil while piston 4 is at the

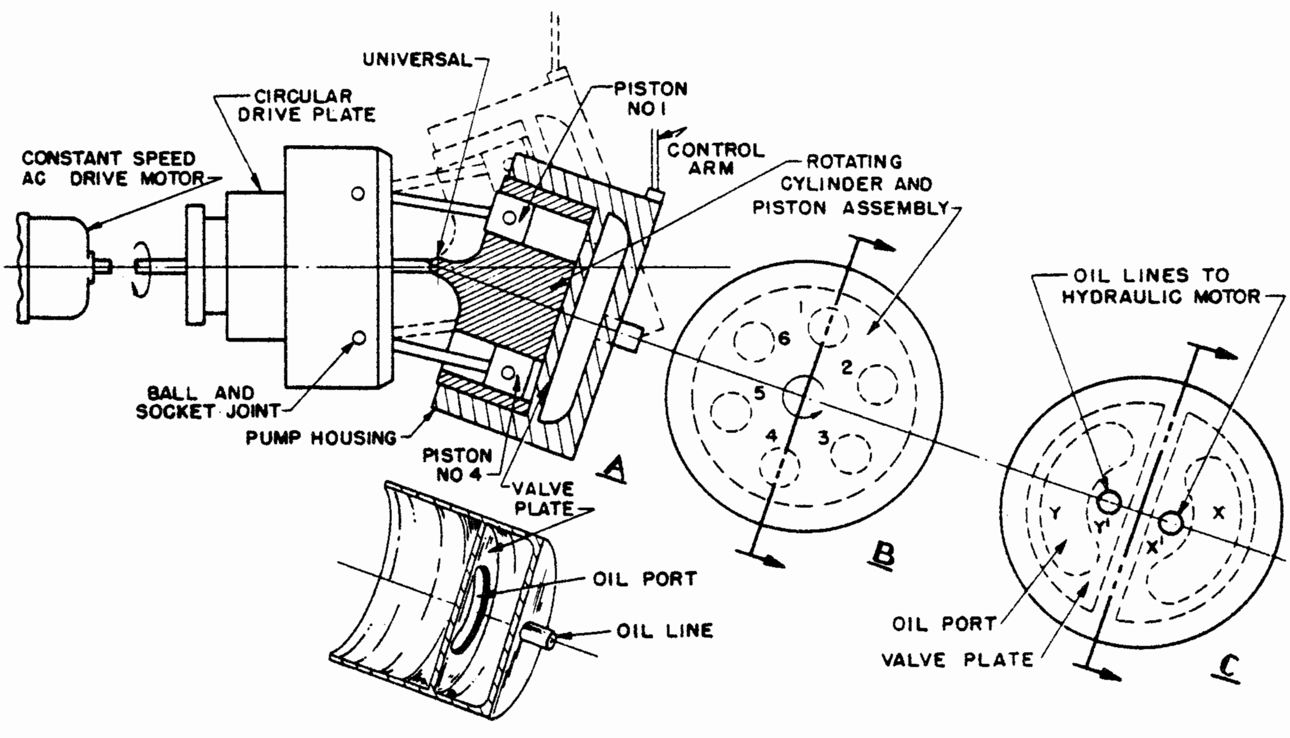


Figure 3-489. Hydraulic Variable Flow Pump

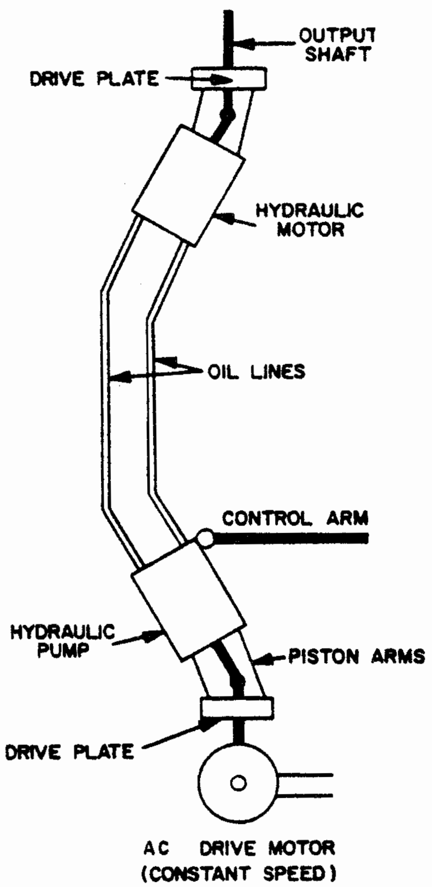


Figure 3-490. Basic Hydraulic Servomechanism

bottom of its stroke, having already pumped out its store of oil. By referring to drawing-plate projections B and C, it will be noted that neither piston is pumping at this instant. However, pistons 5 and 6 are now over port Y, and, since these pistons are moving downward and pushing into their cylinders, oil is pumped into port Y and out of oil line Y'. At the same time, pistons 2 and 3, which are aligned with oil port X, are moving upward and pulling out of their cylinders, thereby sucking oil from oil line X' into cylinders 2 and 3 through oil port X. In the solid-line pump position, then, Y' is the output line and X' is the input line. When the control arm pulls the mechanism through the zero-degree position into the 30-degree position (dotted lines), the pumping action takes place as the pistons move upward. Thus, as the cylinder assembly rotates, the oil is pumped into oil port X. In this condition, X' becomes the output oil line, oil is brought into the pump through input line Y', and the direction of oil flow is reversed. When the pump housing is set in the neutral position (where it does not form an angle with the drive shaft) it is apparent that the cylinder assembly rotates without piston action, since each of the pistons is now in the center of the cylinder. Without angular displacement, then, there is no in-and-out, or pumping, action of the pistons. A condition now exists where each piston cylinder is partially filled with oil, but there is no oil flow, either in or out, through X' or Y'. It must be stressed that the AC motor continues to turn the cylinder and piston assembly at all times, even when no pumping action takes place. The AC motor turns the variable-flow pump at a constant speed in one direction only. The amount of oil flow is controlled by the angular tilt of the pump housing, maximum flow being realized at 30 degrees. The direction of oil flow may be changed by reversing the pump-housing angle.

3-13.3.2 Hydraulic Motor

The hydraulic motor is a mechanism exactly like that of a variable-flow pump, except that the pump housing is held fixed at a 30-degree angle to the shaft axis. In the following analysis of the hydraulic motor action, the cylinder and piston assembly is assumed to be in the position indicated in Figure 3-491. Here again, B represents a bottom view of the rotating assembly, and C is a bottom view of the rotating assembly and the valve plate. If oil is pumped into the motor through oil line X' and oil port X, pressure is exerted on pistons 2 and 3, which, in turn, apply an upward force of the edge of the circular drive plate toward the reader. A clockwise rotation of the load results, as viewed from the end of the output shaft. Oil returns to the variable-flow pump through oil line Y'. When the pump angle is reversed, so that oil is pumped through Y into the motor, pistons 5 and 6, which are aligned with port Y, are under pressure. An upward force is again exerted on the drive plate, but, since the force is now applied to the edge away from the reader, the load now rotates in a counter-clockwise direction.

3-13.3.3 Oil Pressure

In the variable-speed transmission unit just described, the oil in the unit must be held under a certain pressure. This is accomplished by the combination of an oil-replenisher pump and valves which maintain an oil-supply pressure of approximately 75 psi to the transmission unit. The maximum hydraulic pressure in the unit, controlled by a high-pressure relief valve, is approximately 1400 psi.

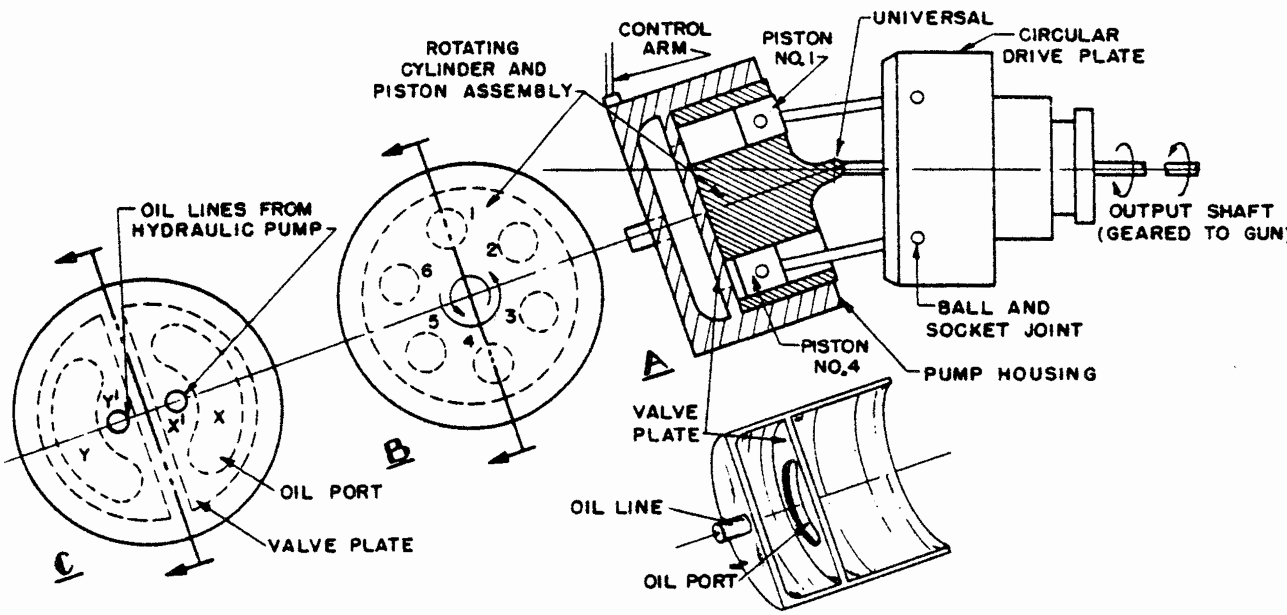


Figure 3-491. Hydraulic Motor

3-13.3.4 Error Measurements

It has been shown that the pump is in the neutral, or zero-degree, position when the gun is aimed according to the director order and a no-error condition exists. Furthermore, it is obvious that the angle assumed by the pump housing is a function of the control arm through an error-measuring device. As in the previous instance there are several error-measuring devices that may be used. In one method the position of the gun is ordered by a synchro transmitter, located at the gun director and wired to a synchro control transformer at the gun. Completing this type of loop, the shaft of the rotor winding feeds the error voltage to a servo amplifier. The amplifier then energizes a servo-motor, which actuates the pump-housing control arm by means of mechanical or hydraulic linkage. In the variation of this error circuit, a synchro differential is used as an error detector, supplying enough torque to mechanically position the input member of a hydraulic-booster arrangement. The output piston of the booster is then used to move the arm which controls the angular position of the pump housing. In this circuit, the differential receives its rotor and stator voltages from the synchro transmitters at the director and gun, respectively.

3-14 GYRO STABILIZATION SYSTEMS

Gyro stabilization systems provide a reference for pitch, roll, and bearing corrections for Navy automatic fire-control, radar, and sonar systems. Since the gyroscope is an important element in stabilization devices, a knowledge of gyro fundamentals is desirable.

3-14.1 GYRO FUNDAMENTALS

One type of gyroscope is a wheel supported such that it is free to spin about an axis called the "axis of spin" (see Figure 3-492) and is free to rotate about one or both of the other two axes which are perpendicular to each other and to the spin axis. These perpendicular axes are called the "input" and "output" axes. The wheel is often supported in a ring called a "gimbal," and this gimbal is usually supported by another gimbal, which is sometimes supported by a third gimbal. During operation, the wheel is driven continuously at a high rate of speed. The rotation of the wheel gives the gyroscope the characteristic of opposing any torque which tends to change the direction of the spin. A force applied along either the spin axis or the input axis will not

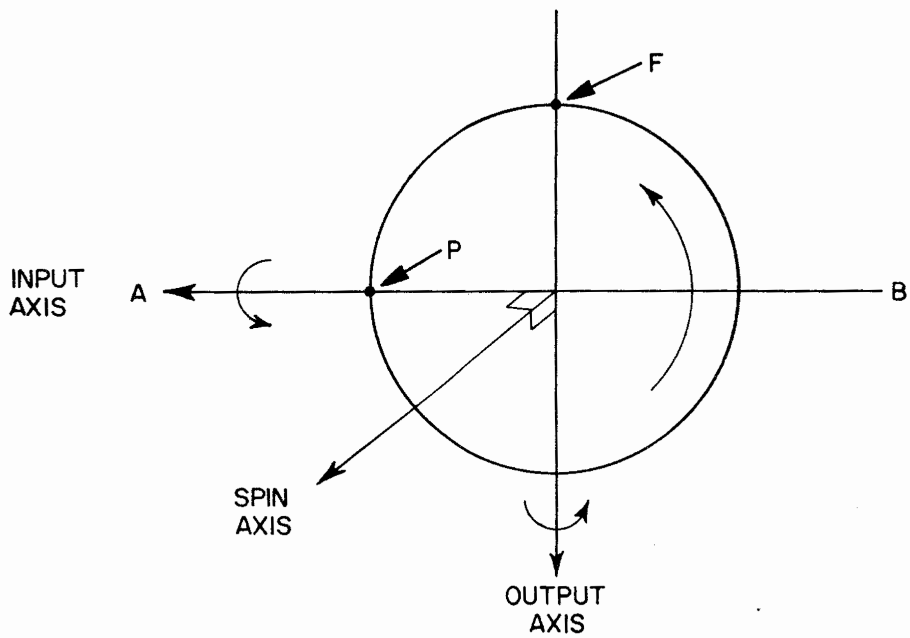


Figure 3-492. Gyroscope Precession

cause gyro action. The rotating wheel will move along the axes in response to the force just as any mass rotating or not. However, if a force, F, is applied as shown in Figure 3-492, a torque is applied clockwise about the input axis (as viewed from A to B), and the gyro precesses or rotates about the output axis as shown. It rotates about the output axis clockwise as if force P were applied to the wheel. In other words, force F is applied and the movement P occurs 90° ahead of F in the direction of spin. The gyro thus rotates about the output axis. The term "precession" comes from this action ("ahead of") or preceding the original applied force and in the direction of spin. Gyro stability derives from the resistance the gyro exhibits to any torque around its input axis. This stability and precession are two gyro characteristics utilized in every application of the gyroscope. For example, if a gyro were mounted with its input axis pointing for-and-aft on a ship, any roll of the ship would then produce a torque about the input axis of the gyro and the gyro would precess. This precession could be sensed by a synchro and an error signal developed. When applied to a servo motor, this error signal would restore the gyro to its original position by applying an equal but opposite torque about the input axis. The amplitude of the error signal is directly proportional to the amount of roll the ship experienced and could be distributed around the ship to isolate radar antennas, guns, etc. from the ship's rolling motion. Gyros could be added to sense other ship's motions, such as pitch and change in heading. These are the basic principles behind a stable platform. Stability in space is not, by itself, sufficient in a stabilization installation. Another requirement is imposed; that of determining the vertical (or horizontal) position with respect to the surface of the earth. In other words, it is necessary to measure to the true horizontal of the earth's surface. Part A of Figure 3-493 illustrates how a perfectly free gyro has stability (or rigidity) in space which is independent of the earth's horizontal. Should the gyro start at position with its spin axis normal (perpendicular) to the earth's horizontal, when it reached position D the relationship between the spin axis and the horizontal would be changed by 90 degrees with these changes in latitude. The spin axis has not changed. Rather, there is a 90-degree difference in horizontal to the earth at the two points. Also, due to the earth's "turnable" motion (rotation about the north-south axis each 24 hours), at any latitude between the poles on the equator with the spin axis initially set in any direction, the gyro will maintain this direction in space. However, the earth will move relative to it, giving an apparent tilt of the spin axis about the horizontal and a twist about the vertical

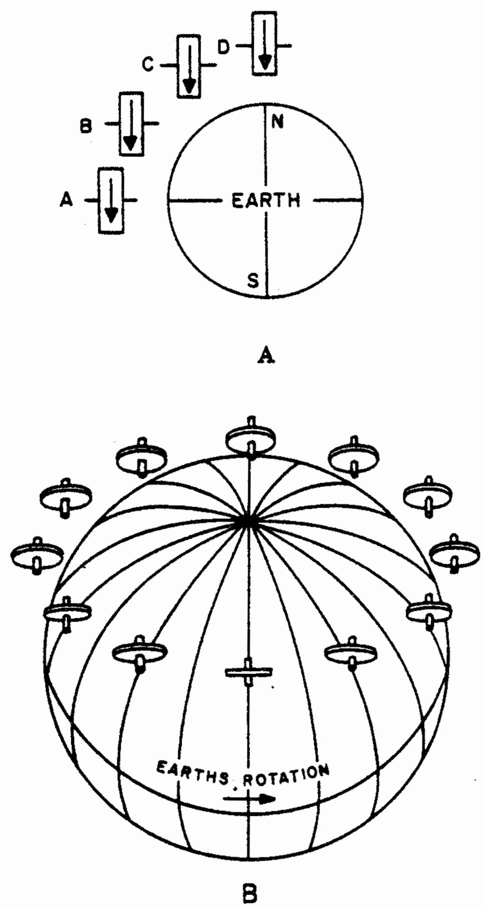


Figure 3-493. Gyro Stability (Rigidity in Space) as Related to Position Relative to the Earth

(B of Figure 3-493). Thus, a vertical or horizontal-seeking capability is often incorporated into gyro stabilization systems. Consider now the remaining elements of a stabilizing system. Whether used aboard ships, land vehicles, or other craft, only the details of the system vary. Basically, the problem is illustrated in Figure 3-494, and the solution is shown in the block diagram of Figure 3-495. In some cases, a platform is moved by the system to maintain its level while in other cases the platform is the craft itself. In this case, correction for tilt is applied directly to the part which is aimed. An antenna such as might be used for a radar height-finder or gun-laying set is shown in Figure 3-494. In lieu of the antenna, a gun, rocket launching pad, searchlight, telescope, or almost any device capable of being aimed could be substituted. Returning to Figure 3-494 angle "g" is measured to the true horizontal. If the craft were level (as could easily be the case for a fixed support), the angle measured would be true. In actual practice, however, the craft may be tilted because of roll, pitch, or a combination of the two, thus requiring the addition of angle "e". The antenna must then be pointed at the new angle if it is to continue proper aiming at the target. If no correction is applied to the indicator when the antenna is reaimed, the target echo appears at position B, rather than at the correct position which is A. Therefore, a gyro correction is needed to keep the echo at position A.

3-14.1.1 Electrically Suspended Gyro

The theoretically near-perfect gyro is a perfect sphere rotating very fast and surrounded by a vacuum. This concept is approached by the electrically-suspended gyro (ESG). The ESG consists of a solid beryllium ball approximately one centimeter in diameter. The ball (or rotor) is enclosed inside an evacuated spherical cavity having a high vacuum. To obtain gyro action, the ball is levitated (suspended) within the cavity by electrical signals and then spun up to approximately 2600 revolutions per second by an external induction torquing motor (see Figure 3-496). The levitation circuitry establishes essentially a fixed distance between the cavity and the rotor of approximately 300 millionths of an inch. Therefore, as the ship moves, the cavity moves with the ship but, due to rotational inertia, the ball tries to remain fixed in space. However, the levitation circuit senses this movement between the rotor and cavity through a change in capacitive reactance, and restores the rotor distance within the cavity. This restoring force appears as a modulation on the levitating signal and is interpreted as ship's motion. This motion could be roll, pitch or changes in heading. As with more conventional gyros, this modulation is used as an output to stabilize platforms, guns, antennas, etc.

3-14.1.2 System of Coordinates

In most stabilization systems, measurement of present target position is made either in stable coordinates where the true horizontal is used as the reference plane, or in unstable coordinates where the deck is used as the reference plane. To convert coordinates between these reference systems, the inclination of the deck plane with respect to the horizontal plane is measured, and from this data, correction for pitch and roll of the ship may be applied to the stabilization problem. In fire control, the position of a line in space, such as a gun bearing, a gun elevation, a sonar bearing or sonar depression, is measured about axes which are horizontal and vertical in

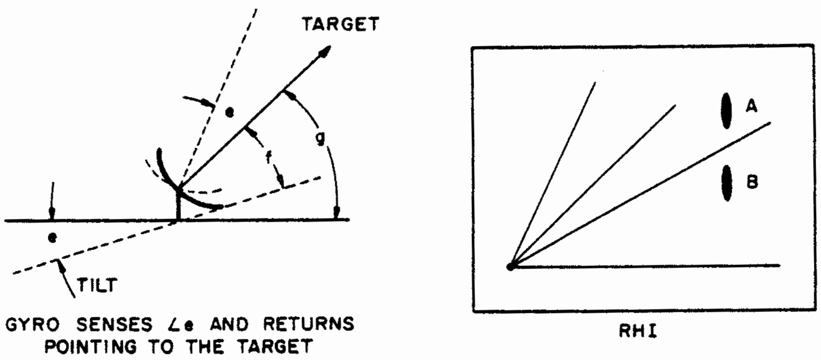


Figure 3-494. Gyro Corrections to Antenna and Indicator

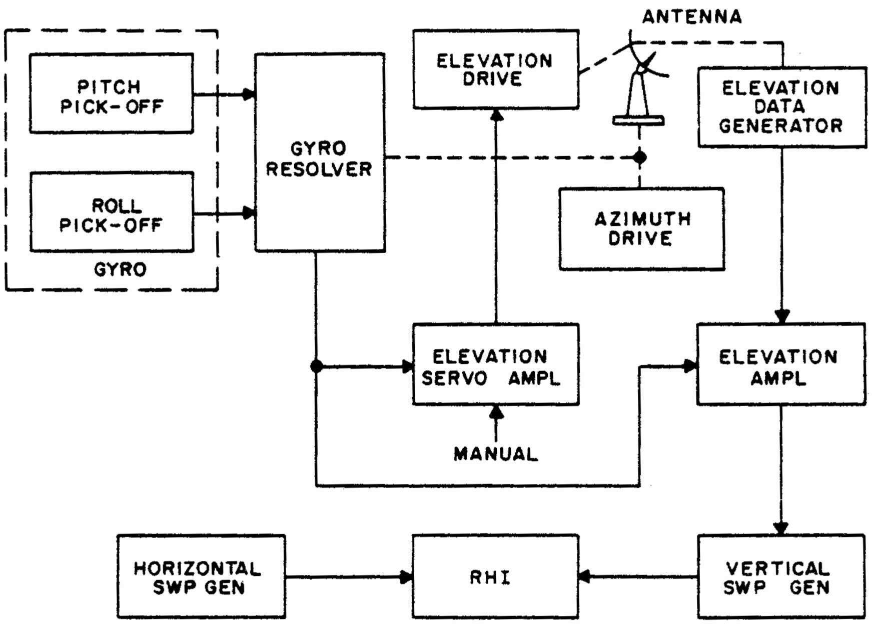


Figure 3-495. Pitch and Roll Stabilization of an Antenna

space. When these axes are relative to the plane of the deck of the ship, the bearing is said to be from the "Deck-Deck System of Coordinates." This is true for such angles as train and sonar transducer depression. The other broad classification of coordinates is the "Horizontal-True-Zenith" system. When a ship is at rest (i.e., the place of the deck and the horizontal plane coincide) then the true zenith and deck zenith axes are in alignment. Underway, the two sets of axes are being continuously displaced with respect to each other by the tilting motion of the deck (B of Figure 3-497). Electrical computations are used for target tracking or for similar purposes. The equipment itself is rarely stabilized physically; i.e., the equipment is not supported on a stabilized platform but is stabilized by electrical corrections that are entered into the computation of the tracking problem. For example, if a gun is tracking a target, the computations that take into account the speed and direction of target, course and speed of Own Ship, windage, ballistic, and other required data to provide solutions for lead angle and elevation of the gun, also incorporate pitch and roll data. This means that the gun train and elevation mechanisms are also used for correcting for pitch and roll. When the gun is not receiving a solution, the gun barrel will be maintained at a line in space. The two

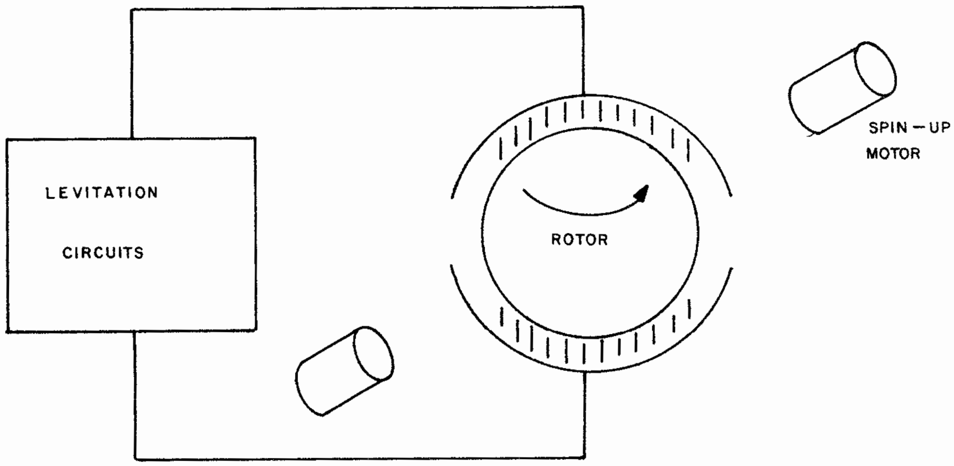


Figure 3-496. Simplified ESG

quantities that provide data for stabilization are "pitch" and "roll". These are defined as follows:

1. Roll. Measured about a horizontal axis in the deck plane, roll is the angle measured in the athwarthship plane perpendicular to the deck between its intersection with the horizontal plane and the deck plane (positive when the starboard side of the ship is tilted up). (See B of Figure 3-497).

2. Pitch. Measured about horizontal axis, pitch is the angle measured about the intersection of the horizontal plane with the athwarthship plane perpendicular to the deck, between the vertical plane and a plane perpendicular to the deck through this axis (positive when the bow of the ship is pointed down). (See C of Figure 3-497.)

Quantities that include pitch and roll considerations are "level angle" and " cross-level" angle. These are defined as follows:

1. Level Angle. Measured about an axis in the horizontal plane, level angle is the angle between the horizontal phase and the deck plane, measured in the vertical plane through the line of sight to the target (positive when the deck section toward the target is below the horizontal plane). (See C of Figure 3-497.)

2. Cross-Level Angle. Measured about an axis in the deck, cross-level is the angle which is measured about the intersection of the plane of the deck with the vertical plane through the line of sight, between the vertical plane and a plane perpendicular to the deck through this axis (positive if, when facing the target, the starboard deck of the ship is tilted up). (See C of Figure 3-497.)

3-14.2 STABLE ELEMENTS

Shipboard electronic systems such as gun-fire control, guided missile control, radar and sonar are usually gyro stabilized. Since more than one stabilized system is ordinarily installed aboard ship, and gyro units are expensive, a practice that is becoming more common is the provision of one master gyro in a ship or other craft to provide stabilization for more than one device. The units used for stabilization are termed "stable elements" and are mounted close to the center of gravity of the ship. The stable element is used to both measure movement of the deck in level and cross-level angles or in roll and pitch angles, depending on the connections of the stable element, and to transmit these angles as synchro signals. The principal part of the stable element is an electrically-driven gyroscope that establishes a horizontal reference plane from which the level and cross-level angles are measured. There are three follow-up systems in stable element equipments: the train; the cross-level; and the level follow-up systems. In some equipments the train is locked on zero, and the outputs are in terms of roll and pitch instead of level and cross-level. The follow-up system for train determines the error

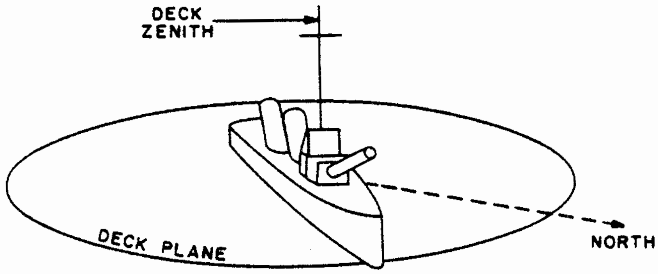


FIG. A

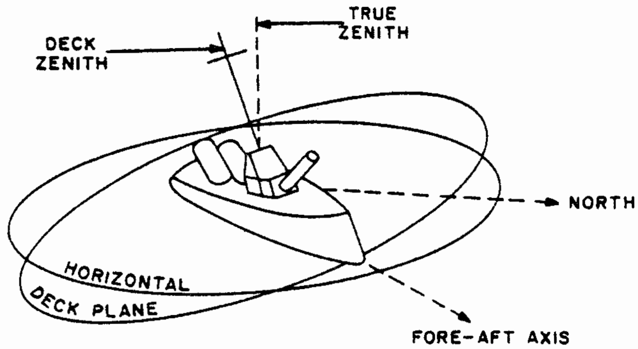


FIG. B

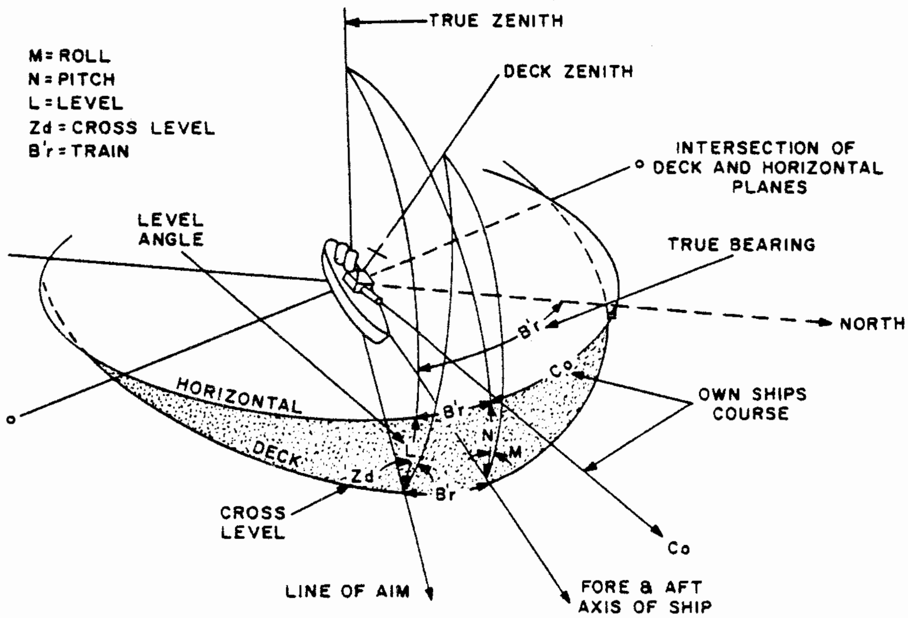


FIG. C

Figure 3-497. System of Coordinates as Related to the Gyro Stabilization Problem

between the bearing of the equipment being stabilized and the bearing of the training frame in the stable element. When the stable element supplies roll and pitch data, however, the train follow-up system is not used. The train input to the stable element is then locked in the zero position. In stable elements having outputs of level and cross-level, the train follow-up is used to rotate the stable element to the bearing of the equipment being stabilized. The follow-up systems for level and cross-level are identical and are actuated by electrical error signals originating in the gyro unit. These signals are amplified and then used for actuating the level and cross-level motors. These drive not only the synchros transmitting the level and cross-level angles to the equipment being stabilized, but also drive the level and cross-level follow-up circuits. If the train input is locked, the output is then roll and pitch data. The stable element is spatially referenced by a gyroscope rotating about a vertical axis, thus continuously maintaining an AC electromagnet in a fixed position in space. Above the electromagnet are two sets of follow-up coils whose fields are at right angles to each other and are supported on mutually perpendicular gimbals. When the motion of the ship displaces the coils with relation to the electromagnet, follow-up systems are actuated by the coils to move the gimbals in such a direction as to restore the original position of the coils with respect to the electromagnet. The angular movement of the follow-up controls causes the signals across synchro transmitters in the follow-up system to change in correspondence with the attitude of the gimbals. They thus measure the attitude of the deck with respect to the horizontal. Nontrainable types of stable elements provide roll and pitch signals; trainable types provide level and cross-level signals. When it is necessary to obtain roll and pitch signals from level and cross-level signals, they can be obtained automatically by computers according to the following mathematical relationships:

1. $\cos M \cos N = \cos L \cos Zd$
2. $\sin M \cos N = \cos L \sin Zd \cos B'r - \sin L \sin B'r$
3. $\sin N = \cos L \sin Zd \sin B'r + \sin L \cos B'r$
4. $\sin M = \dfrac{\cos L \sin Zd \cos B'r - \sin L \sin B'r}{\cos N}$

where:

M = Roll
N = Pitch
L = Level Angle
Zd = Cross-Level Angle
B'r = Train Angle

3-14.2.1 Pitch and Roll Stabilization

Figure 3-495 is a simplified block diagram of a typical system for pitch and roll stabilization of an antenna. The system applies also to other elements, such as a gun, transducer, etc. In this case, the pitch and roll signal is resolved into a single elevation correction which, with a bearing correction, provides stabilization. Other systems, like those which stabilize a platform, may have a separate roll and pitch servo system and thus provide direct corrections for roll and pitch. Regardless of the ultimate system, the gyroscope assembly provides the reference for pitch and roll. Its output might be in direct pitch and roll data, level and cross-level data, or elevation and train signals. Figure 3-498 is a diagram of a gyroscope assembly showing only the basic elements required to provide pitch and roll data. Actually, a stable element may have an AC electromagnet in a fixed position with follow-up coils supported on the gimbals above the electromagnet. When the motion of the ship displaces the coils with respect to the electromagnet, follow-up systems are actuated to correct the position of the gimbals. In this way the original position of the follow-up coils with respect to the electromagnets is restored. The follow-up systems control synchro transmitters which move in correspondence with the attitude of the gimbals and thus measure the attitude of the disk with respect to the horizontal. Figure 3-498 is not intended to show the actual mechanism, but serves to illustrate its basic operation. To correct for the effects of the roll and pitch of the ship upon the device being aimed (in this case a radar antenna), it is necessary to measure the degree and direction of the rotations from established references. The primary functions of the gyro assembly are to establish the required references and to make the necessary measurements of the ship motions. Gyro assemblies may be equipped with a single gyro vertical having separate pitch and roll gimbals, or of separate pitch and roll type verticals mounted in common gimbals. The assemblies are also equipped with an erection system and some type of earth rotation or latitude corrector. Bearing references are obtained from an azimuth directional gyro or from the ship's master gyro. The gyro erection mechanism may be either a mechanical or electrical servo loop which senses the direction of gravity and keeps the axis of spin of the gyros vertical. As the earth rotates, and as the ship travels, the gyro verticals would ordinarily appear to precess so that the axis of spin would not be vertical. Earth's rotational correction torques are applied to the gyros to cause them to precess toward the vertical at the same rate as the apparent precessions away from the vertical. Thus, the pitch and roll gimbals are stabilized in both pitch and roll. Roll and pitch data are obtained by means of separate pitch and roll synchro transmitters. Roll data is obtained from the displacement between the roll gimbal (outside gimbal) and the gyro assembly frame. Because the frame is secured to the ship, it rotates with the ship whenever the vessel rolls, pitches or changes heading. The roll gyro and its associated centering and earth rotation corrector circuits correct any roll-gimbal deviation from the horizontal about the roll servo axis. The angle through which the gyro assembly frame rotates about

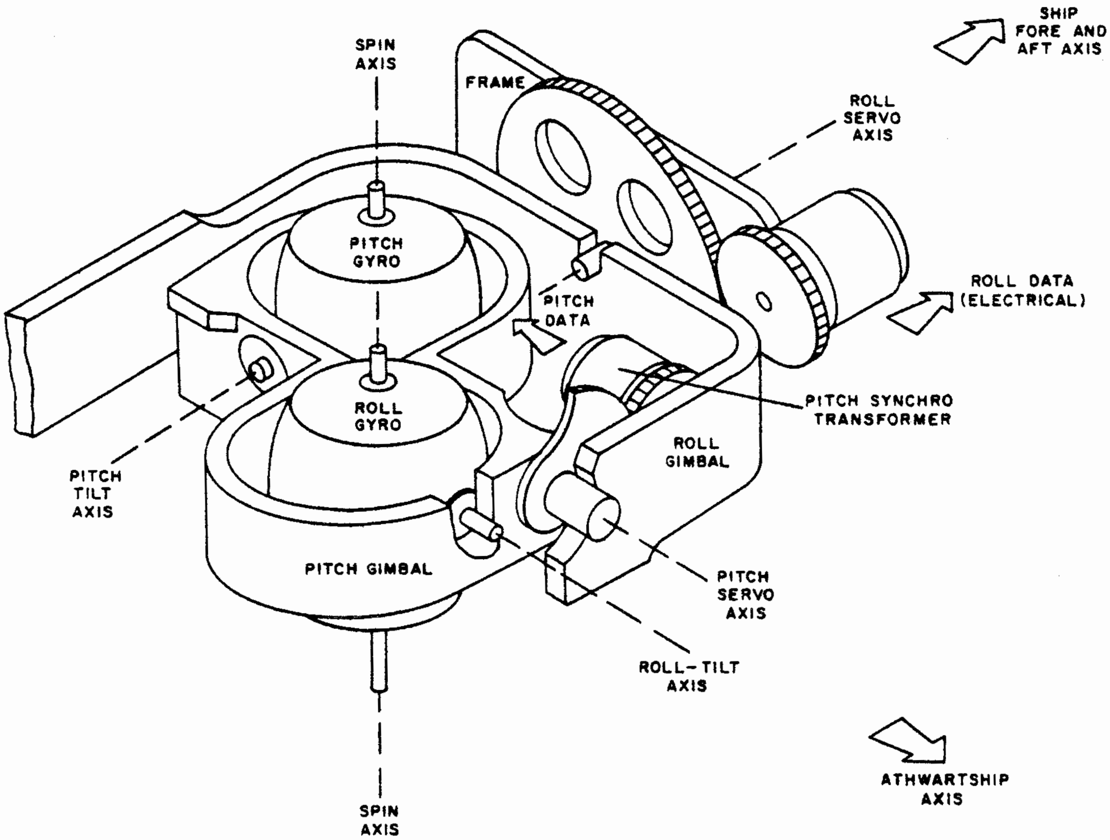


Figure 3-498. Gyro Reference Assembly Showing Roll and Pitch Data Measuring Transmitter Synchros

the roll servo axis, with respect to the horizontal of the gimbal, is the angle of roll of the ship. A large gear mounted on the roll gimbal about the roll servo axis engages the drive gear of a roll transmitter mounted on the gyro assembly frame with the shaft parallel to the roll servo axis. Since the rotor is geared to the roll gimbal and the roll gimbal is stabilized in roll, the synchro rotor is also stabilized in roll. As the ship rolls, the stator rotates about the rotor, thereby providing roll data which corresponds to the roll angle. Ordinarily, this data is fed to a differential transformer in the stabilizing drive system which produce error voltages, as required, for servo loops that stabilize the aimed element in roll. Pitch data is obtained from the relationship between the pitch gimbal and the roll gimbal. Although the roll gimbal is stabilized in roll, it is permitted to rotate about the pitch servo axis when the ship pitches. On the other hand, the pitch gimbal is stabilized both in roll and in pitch. This means that the roll and pitch gyros and their associated centering, erection, and earth rotation correction circuits, correct any pitch gimbal deviation from the horizontal about both the roll servo axis and the pitch servo axis. The angle through which the roll gimbal rotates about the pitch servo axis, with respect to the horizontal of the pitch and roll gimbal, is the angle of pitch of the ship. The pitch-measuring synchro transmitter is mounted on the roll gimbal with its rotor geared to the pitch gimbal. Any rotation of the gimbal about the pitch servo axis will displace the stator with respect to the pitch stabilized pitch gimbal. The pitch data is used in the stabilization system in a manner similar to that described for roll stabilization. Changes in ship heading center about the spin axis of the roll and pitch gyros and therefore have no effect on the pitch and roll gyros.

3-14.3 MAINTENANCE CONSIDERATIONS

Trouble in the stable element is usually indicated by failure of the device being stabilized to remain properly fixed on the aim point when the ship rolls, pitches, or changes heading. The first step in analyzing a stabilization trouble is to determine the operational status of the stable element unit so as to isolate the problem to either the unit being stablized or to the stable element itself. When the stable element is employed to stabilize more than one equipment, normal stabilization of the other devices will indicate normal operation of the gyroscope and indicate trouble in the pertinent servo loops of the equipment. Troubles in the stable element found to occur most frequently are as follows: defective erection mechanisms; defective gears; defective relays; faulty gyros; and defective synchros. Servicing and maintenance of the gyro units require the same degree of care as the servicing and maintenance of delicate instruments. For example, improper lubrication can actually cause malfunctioning of the component. Cleaning procedures set up in pertinent instruction books should be followed carefully. As little cleaning fluid as possible should be used, with all parts other than those being cleaned protected with a clean, non-linting splash cloth to prevent contamination from dirt or solvent. However, since many maintenance operations on the gyroscope require special test equipment and fixtures, repair should be rarely attempted in the field. Stabilization problems that do not originate in the gyro unit may be serviced by standard procedures. Special attention should be given to the servo systems comprising the stabilization circuits, and reference should be made to the particular service manual for proper alignment and test procedures. When locating trouble in a servo system, an understanding of the operating characteristics of the components and a familiarity with the system is an invaluable asset. When a servo is completely out of adjustment, the process of reasoning can be temporarily abandoned while a thorough check of the entire system is made. Such a check should start by zeroing the system, adjusting the gain setting of the servo amplifiers, and then testing the operation of each component. Assuming that such a procedure is unnecessary, the symptoms can usually be grouped under one of the following headings: no operation; sluggishness; instability; or inaccuracy. These symptoms may also occur in combination. A few causes of each of these types will be discussed, but because of the differences between systems, more complete information must be obtained from pertinent technical manuals. Failure of the servo system to operate may be caused by such obvious faults as no power, an open fuse, wiring open or short-circuited, loose or jammed gearing, or failure of such components as transistors, resistors, capacitors, transformers, or motors. Sluggish operation may be caused by incorrect gain setting, improper phasing, low amplifier gain, excessive mechanical friction, or an amplifier saturated with noise. Instability can be caused by improper gain settings, improper synchro zeroing, or improper operation of stabilizing components such as rate generators, networks or feedback circuits, or excessive mechanical backlash. Excessive, but constant error, is probably caused by improper synchro zeroing. When the error is not constant, the cause may involve sluggish operation of the servo itself. If the actuating signal is small when the error is large, the fault may be due to excessive error in the error detector, dial, gear or coupling, or from slippage on a shaft. The difficulty can often be localized by plotting the error as a function of the angle of rotation of the controlled shaft. By this means, the space-frequency of error can be determined and the error then localized to a specific component which rotates at a speed relative to the controlled shaft.

3-15 MAGNETIC AMPLIFIERS

Magnetic amplifiers are similar in operation to vacuum tube or transistor amplifiers in that all three amplifier types are devices whose output is controlled by an input current or voltage. Magnetic amplifier employs the principle that the impedance of a coil can be varied by changing the saturation of the core, thus varying the current through the coil. Some advantages of magnetic amplifiers are:

1. Ruggedness - like transformers
2. Durability - no moving parts
3. High Power Gain - gain of several million possible
4. Safety - inputs and outputs can be sealed and electrically isolated
5. No warm-up time required.

The major disadvantages of magnetic amplifiers are:

1. High initial cost
2. Frequency limited to about .5 MHz
3. Highly distorted output
4. Cannot amplify control signal having frequency higher than the supply frequency.

3-15.1 BASIC PRINCIPLES

The operation of magnetic amplifiers is based upon the nonlinear B/H characteristics of magnetic core material. When the characteristic is plotted, it is known as the B/H curve, magnetization curve, or hysteresis loop. The quantity (B) of Figure 3-499 represents flux density, which is a measure of the amount of magnetization; the H stands for the magnetizing force (or magnetic field intensity). These two quantities are related by an expression involving permeability (μ) which is expressed as: $B = \mu H$. If the magnetizing force, H, is increased from zero (point O) to a value K, as shown in figure A, the flux density, B, increases linearly. However, as H is further increased, the core material becomes saturated, and a point is reached where a further increase in H leads to no further increase in B. This condition is analogous to plate current saturation of an electron tube. If the magnetizing force is decreased to zero and then increased in the opposite direction, the flux density will follow the curve ABC; finally, if the magnetizing force is again decreased to zero and then increased in the original direction, the curve CDA results. The closed curve ABCDA is the B/H curve in this instance. It may have the form shown in part A or B of Figure 3-499, or it may assume other forms, depending on the material being magnetized. The core material used in the construction of magnetic amplifiers should possess a B/H characteristic curve that is as close to a rectangular curve (considered ideal) as possible; it should also have low eddy current losses and a high saturation flux density to permit maximum inductance. In addition, the material should be highly stable under conditions involving vibration, shock, and temperature extremes.

3-15.2 AMPLIFIER CIRCUITS WITHOUT FEEDBACK

Figure 3-500 shows a basic type of magnetic amplifier having current gain characteristics. This type of circuit is used in motor speed control applications and in lighting control for public places. The basic "reactor" (a name commonly used for the transformer of the magnetic amplifier) consists of three windings on an E-type core. The center winding is the control coil; this winding receives the control voltage, which must be DC, and is made variable by use of a rheostat. The two outside windings are the controlled coils. Connected in series-aiding, they are placed in series with the load as shown. The magnetization curve for the core material should have a shape similar to that shown in part (B) of Figure 3-499. With no current flowing in the DC control winding, the magnetizing current flowing in the AC windings through the load is small because of the large inductance exhibited by the reactor. This magnetizing current would provide the hysteresis loop shown in part (A) of Figure 3-501. Since the reactor is not saturated during either half of the AC cycle, the AC windings present maximum impedance in series with the load. When a direct current is fed to the control windings, a residual flux appears in the core, and the center of the core magnetizing force then shifts to the right as shown in part (B) of Figure 3-501. This magnetizing force is the resultant of both the AC and DC currents in the reactor. Since the hysteresis loop becomes smaller, the change in flux density decreases, resulting in a larger amount of AC magnetizing current and less impedance to the load. By varying the DC control current from zero to saturation, the impedance in the AC circuit may be made to vary from a high

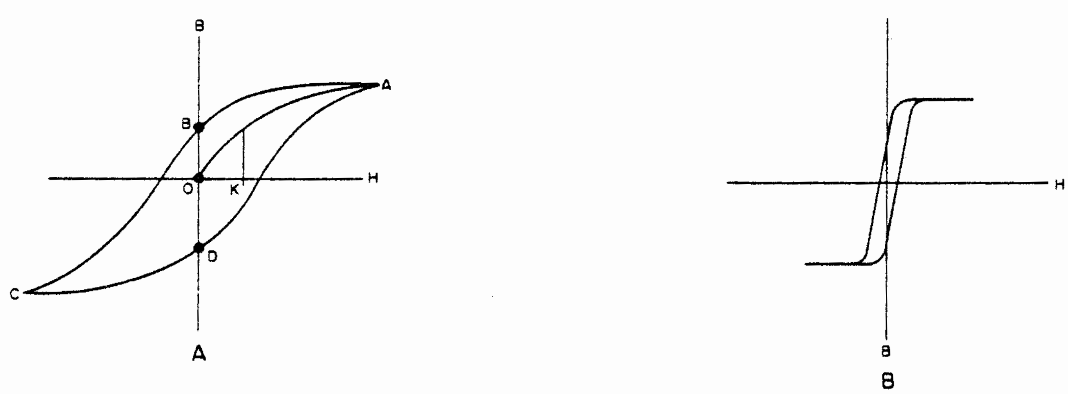


Figure 3-499. Typical B/H Characteristics Curves

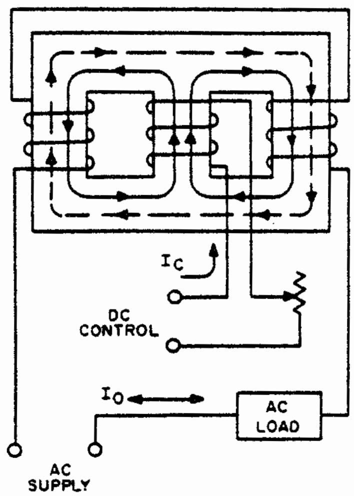


Figure 3-500. Basic Magnetic Amplifier, Using Series-Connected Windings

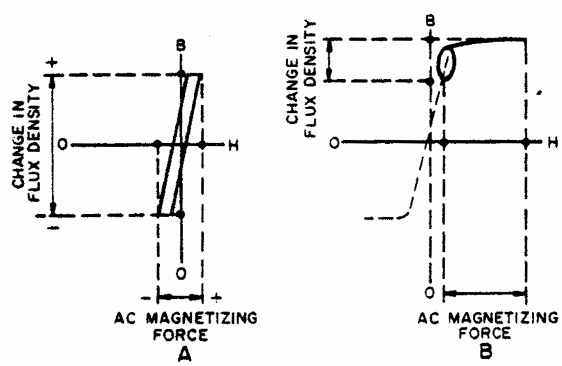


Figure 3-501. Operation of Magnetic Amplifier in the B/H Characteristic Curve

value to almost zero. Any DC current change in the control winding causes a corresponding AC current change of greater value. The circuit, therefore, exhibits current gain characteristics. By connecting the two AC windings in series-aiding, the core flux due to the alternating current will follow the path shown by the dotted line in Figure 3-500. Under these conditions, no AC voltage will be induced into the control winding. The flux due to the DC magnetizing current will follow the solid line path. During any half cycle, the flux due to the AC current in one leg of the core will be in phase with the flux due to the DC control current. It will thus provide a high degree of saturation in that leg, but the two magnetizing forces oppose each other in the opposite leg. This condition would tend to restrict the minimum impedance to a value too high for practical use. It can be overcome by using a high value of ampere-turns for the control winding. This prevents the peak AC magnetomotive force from driving the flux density too far down from the knee of the B/H curve, which is on the side where the fluxes are opposing. Part (A) of Figure 3-502 illustrates an arrangement with the AC windings connected in parallel. This circuit is usually employed to handle loads with medium to high current consumption, but has a somewhat slower response characteristic than the series circuit. Part (B) of Figure 3-502 shows a variation that permits the use of a DC load. In this amplifier, a bridge rectifier in the AC circuit provides DC to the load while keeping any DC component of current out of the AC windings, thus preventing self-saturation. Since a DC voltage is applied to the control winding, it is possible to connect two DC output

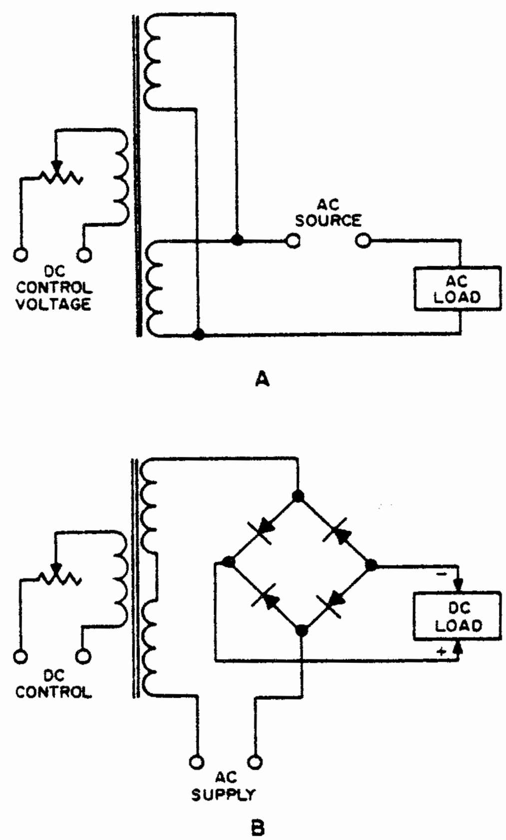


Figure 3-502. Magnetic Amplifiers Used for Control of AC and DC Loads

amplifiers in cascade to increase the gain as illustrated in Figure 3-503. This circuit has the disadvantage that an AC current always flows in the AC winding, and this rectified current is applied to the control coil of the succeeding stage. This factor limits the number of stages that can be connected in cascade, because the control windings of succeeding stages would saturate the core with the "no-signal" current, and the amplifier would thus effectively become blocked. This disadvantage can be overcome by using the AC winding of an amplifier in a bridge circuit to cancel out the residual flow of current in the load. Figure 3-504 illustrates this type of connection. In this circuit the windings (A and B) of the reactor form one leg of the bridge. Choke L1, which has the same inductance as the AC winding of the reactor without control current flowing, forms the second leg. The third and fourth legs (C and D) are formed by the secondary of a center-tapped isolation transformer which supplies AC voltage to the circuit. With no control current input, zero current

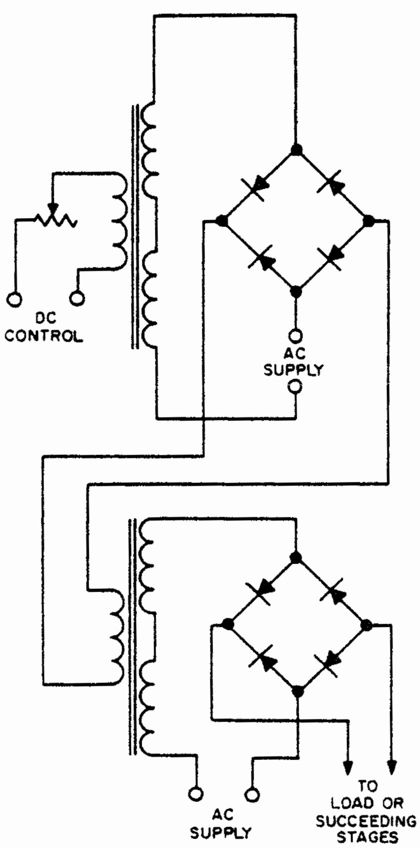


Figure 3-503. Two Stage Cascaded Magnetic Amplifier

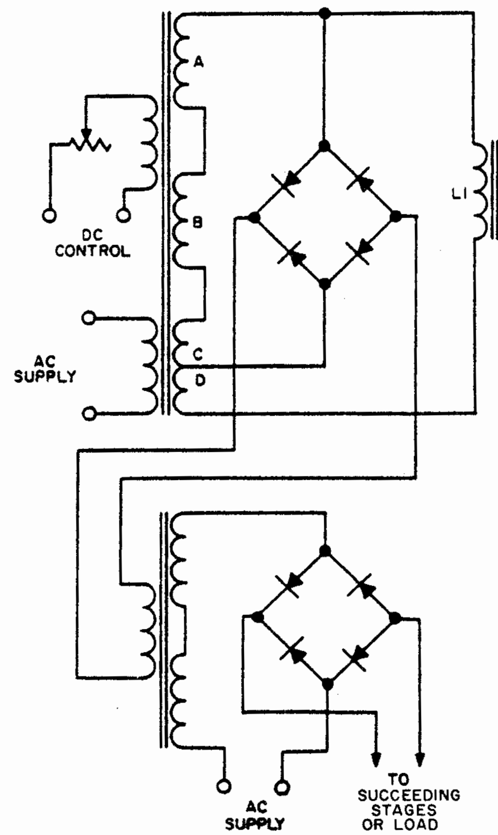


Figure 3-504. Magnetic Amplifier with Bridge Circuit Incorporated To Cancel Residual Current Flow in the Load

will flow in the control windings of the succeeding stage, making possible the use of many stages in cascade for increased current amplification. The control
current in this type of amplifier may be taken from a thermocouple or photocell to actuate motors, relays, etc. The operation and sensitivity of magnetic amplifiers can be improved by using push-pull operation and by biasing the magnetic cores to the knee of the B/H curve with a special bias winding, located on the same leg of the core as the control winding. In this type of amplifier, equal AC voltages are applied to the AC coils of two reactors. The outputs are rectified and applied to a load in the conventional manner. The circuit is polarized so that, when the inductances of both reactors are equal, the resultant output voltage is zero. The above is true when zero control current is flowing. A DC voltage applied to the bias windings causes current to flow in the same direction in both

of these windings. The bias current is adjustable so that the core may be brought to the high-sensitivity portion of the magnetization curve. The control coils are connected in such a manner that when current flows in the control circuit, the bias flux in one reactor will aid the control flux, thereby causing saturation, while the bias flux in the other reactor will cause an imbalance in the AC windings of the two reactors and result in a DC voltage output. If the polarity of the control voltage is reversed, the action just described will result in an output voltage of the opposite polarity. This type circuit, therefore, can be used as a DC amplifier, since the output polarity is a function of the control voltage polarity. Amplifiers of this type have been constructed which have gains in the vicinity of 23 dB with power inputs on the order of 400 microwatts.

3-15.3 AMPLIFIER CIRCUITS WITH FEEDBACK

By providing regenerative feedback in a single-stage magnetic amplifier, it is possible to obtain a considerable increase in power gain over that of a single-stage nonfeedback circuit. Basically, regeneration is accomplished by combining a portion of the output power of the single-stage with the input power, whereby, addition of the two results in a higher power gain. Positive feedback may be provided in two different ways. One method feeds a portion of the output power into a special feedback winding so that the feedback flux will aid the control flux. This type of feedback is usually referred to as external feedback. Another method makes use of a half-wave rectifier in series with the load so that a DC current component will flow in the AC winding circuit. This type of feedback is referred to as "self-saturation" or internal feedback. By the use of regenerative feedback, the power gain of a single-stage amplifier may reach 40 to 50 dB. Both types of feedback may also be used simultaneously, if desired. Part (A) of Figure 3-505 illustrates a basic type of circuit involving internal feedback. When the control current of this amplifier is increased, a larger value of pulsating load current will flow in the AC windings, bringing the core to saturation more quickly. This circuit is bidirectional as far as transfer characteristics are concerned. Replacing the load with a suitable meter would make the circuit adaptable as an extremely sensitive (although not very stable) zero-center DC voltmeter. Part (B) of Figure 3-505 shows an amplifier that incorporates an external feedback circuit. In this amplifier, a DC voltage from the bridge rectifier connected in the AC circuit is returned to a feedback winding, located on the same leg of the core as the control coil. As the control current is increased, the feedback voltage is also increased. Flux from the feedback winding reinforces the flux from the control winding, and saturation occurs rapidly. The control current must flow in the direction shown for the circuit to be regenerative; if the control current is reversed, degeneration will result. This circuit is representative of the polarized type of magnetic amplifier.

3-15.4 CONSTRUCTION

The construction of magnetic amplifiers (with respect to core assembly) differs considerably

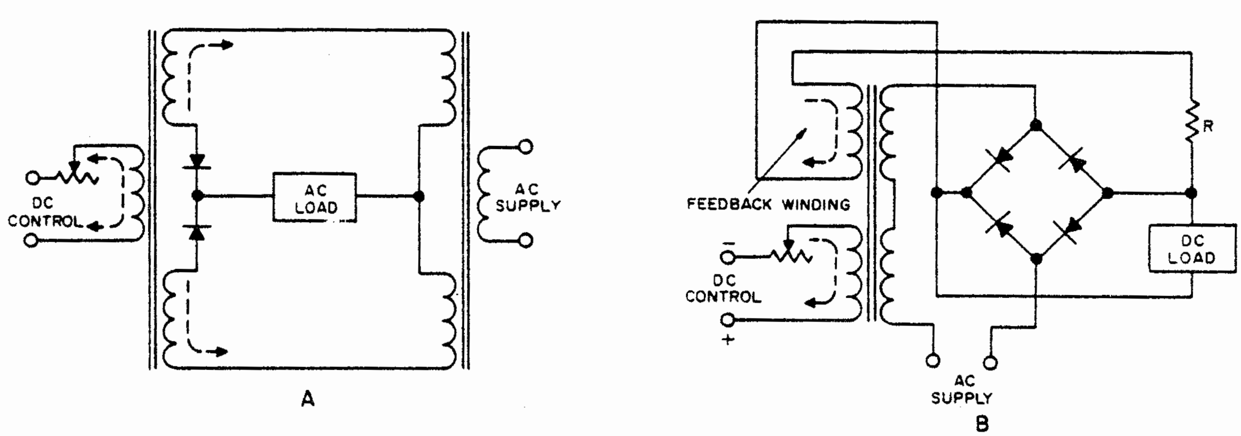


Figure 3-505. Magnetic Amplifiers Using Internal and External Feedback

from the practices utilized in ordinary transformer construction. Standard E-cores using staggered butt joints are unsatisfactory because much of the flux crosses the many small air gaps provided by the butt joints. Improvement can be gained by using the E-core arrangement shown in Figure 3-506. Here the laminations are overlapped for the greater portion of the magnetic path, resulting in a sharper knee in the magnetization curve. This construction lends itself to uniformity of core characteristics. Another type of construction, which is rapidly becoming popular, makes use of toroidal cores, spirally wound with silicone steel tape. Two spirally-wound cores are required in the average amplifier since there is no center leg. A circuit using these cores is illustrated in Figure 3-506. Two control windings are connected in series-aiding so that both cores can be saturated by a common current. The two AC windings are connected in series-opposing in order that a negligible alternating current will appear in the control circuit as a result of transformer action. With slight circuit variations, the twin-core method is adaptable to all of the previously described amplifier circuits.

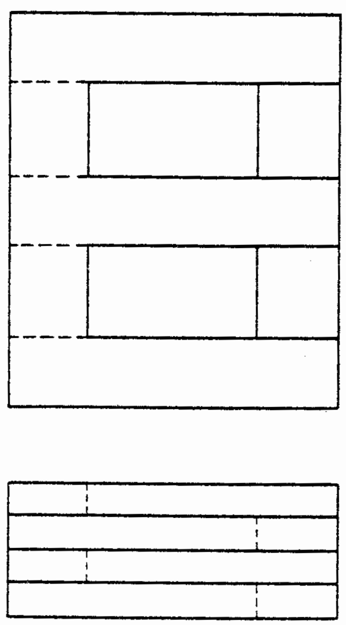


Figure 3-506. Construction of E-Core Assembly for Use with Magnetic Amplifiers

3-15.5 AMPLIFIERS USING TWIN TOROIDAL CORES

The full-wave amplifier shown in part (A) of Figure 3-507 may also be constructed with toroidal cores. The use of two cores requires that a high AC impedance be connected in series with the control winding. As in the case of the amplifier using the E-core, the output voltage across the load is AC, and one of the rectifiers is conducting on each half cycle.

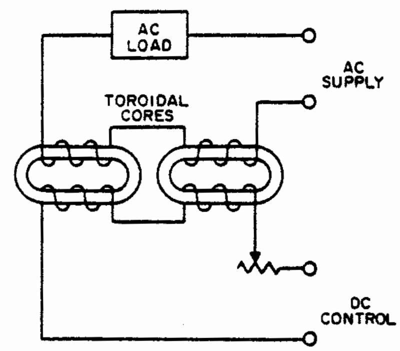


Figure 3-507. Spirally Wound Toroidal Cores Used in Magnetic Amplifier

3-16 POWER SUPPLIES

Power supplies employed to provide the voltage sources for operation of electronic equipments are of various types, such as Half-wave, Full-wave, and bridge circuit rectifiers. The type employed for a particular application depends on such factors as current and voltage load requirements, available space, weight, degree of voltage regulation acquired, etc. Electronic systems are generally complex, and contain one or more power supplies which provide several voltage outputs. The design performance of such equipments depends in large measure upon the rectifying characteristics of such substances as copper oxide, copper sulphide, selenium, germanium, silicon, etc. Semiconductor rectifiers have the advantages of light weight, high efficiency, ruggedness, and long life. The pulsating DC voltage is smoothed to a steady DC voltage by means of a suitably designed filter. Most power-supply circuits employ a power transformer, thereby allowing a step-up or step-down in the AC voltage input in order to produce the desired DC output voltage. In some power supplies, however, where weight, space, and cost are important considerations, the transformer is omitted and the rectifier is connected directly to the AC source. Rectifiers designed to furnish power greater than 1 kilowatt generally use three-phase AC power an a polyphase arrangement of vacuum tubes. Such multiphase rectifiers produce a smoother output waveform than single-circuits, because the ripple frequency is higher and, therefore, relatively easy to filter.

3-16.1 HALF-WAVE POWER SUPPLIES

Half-wave power supplies are the simplest of the various types of power supplies. The half-wave power supply uses a single rectifier (an electron tube, shown in part (A) of Figure 3-508, and a semiconductor rectifier, shown in part (B) of figure 3-508), which conducts essentially in only one direction. The rectifier conducts once during each cycle of the input voltage (during the interval when the plate is positive with respect to the cathode). A pulsating DC voltage is thus produced, and the pulsation frequency is the same as the frequency of the input voltage. Since the amplitude of the rectified voltage remains at zero for an appreciable interval between successive DC pulses, the output of a single-phase half-wave rectifier is difficult to filter, and requires a good filter network to produce a steady DC output voltage (with little ripple). This type of circuit has the additional disadvantage that the current flowing in one direction through the secondary coil magnetizes the core of the transformer. Because of these disadvantages, the half-wave power supply is rarely used. The circuit shown in part (A) of figure 3-508 employs a power transformer with two secondary windings; the low voltage winding, S2 supplies the correct filament voltage to the rectifier tube. Part (B) of figure 3-508 shows a resistor (R_s) in series with the recitifer. This resistor, called the surge resistor, serves to limit the peak current that flows through the rectifier, and is composed of the transformer resistance and any necessary external resistance. When the circuit is first energized, the input capacitor in the filter circuit is in a discharged condition. Thus, if power were applied with no resistor in the circuit, a very heavy current would flow to establish the charge on the capacitor. (This is not the case in electron-tube rectifiers because of the time required to heat the cathode and the relatively high internal resistance.) Since this current might damage the rectifier, R_s is included as protection. For 380-volt (peak-inverse) rectifiers, the value of R_3 is between 1 and 50 ohms, depending upon the peak current rating of the unit.

3-16.2 FULL-WAVE POWER SUPPLIES

The full-wave power supply, shown in part (C) of Figure 3-508 produces an output which is easier to filter and which has a higher average value than that of a half-wave power supply. However, this circuit requires two rectifiers (commonly operating in one envelope), and the transformer must have a center-tapped, high-voltage secondary with a step-up voltage approximately equal to twice the desired DC voltage output. The plate of one rectifier is connected to one side of the high-voltage secondary winding, and the plate of the other rectifier is connected to the other side. The voltage induced in each half of the secondary causes each tube to conduct on alternate half-cylces of the input voltage. Hence, two DC pulses (of the same polarity) occur during each cycle of the AC input voltage, thereby producing an output whose fundamental frequency is twice that of the input frequency. Since the DC pulses in the output of a full-wave rectifier are more closely spaced in time than those obtained from a half-wave circuit, the full-wave output is less difficult to filter. In other words, for the same type of filter, the full-wave circuit has a lower percentage of ripple than the half-wave circuit. In addition, DC saturation is not present in the core of the transformer, because the DC magnetization in the two halves of the transformer secondary are opposed to each other and therefore cancel the magnetizing effects.

3-16.3 BRIDGE-TYPE POWER SUPPLIES

A bridge-type power supply, shown in part (D) of Figure 3-508 produces an output similar to that of the conventional full-wave circuit (the ripple frequency is twice the input frequency). In the bridge circuit, four rectifier elements are needed; however, the power-transformer secondary need not be center-tapped. During each half-cycle that the AC voltage appears across the secondary, two tubes conduct in series, and produce one DC pulse in the output. During the interval when point "a" is positive with respect to to point "b", V_2 and V_3 conduct to produce an output pulse. Similarly, during the next half-cycle, when point "b" is positive with respect to point "a", V_4 and V_1 conduct, and produce another output pulse. Two DC output pulses are therefore produced for each input cycle (full-wave output). In a bridge-type rectifier, the peak inverse voltage appearing

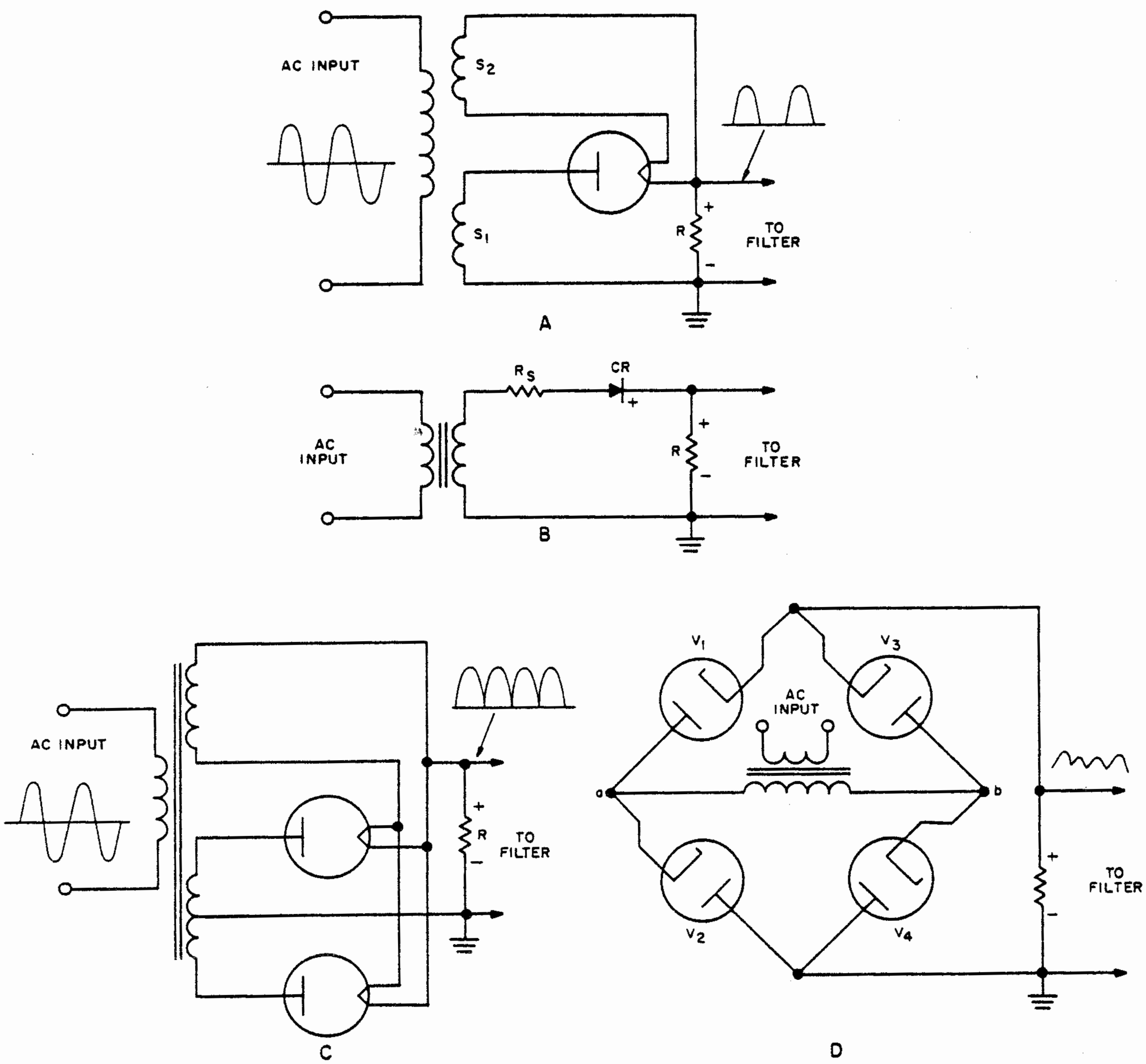


Figure 3-508. Basic Power Supply Circuits

across each tube does not exceed the peak transformer voltage. Hence, the bridge-type circuit can be used to obtain a higher output voltage than a center-tapped, full-wave rectifier using equivalent rectifiers. The bridge-type circuit is widely used with silicon and copper-oxide rectifiers, and less often with electron tube rectifiers. This is because the cathodes are at different potentials, and thus cannot be connected in parallel and supplied from a single filament transformer winding.

3-16.4 VOLTAGE MULTIPLIER CIRCUITS

Voltage multipliers are used to produce a higher DC output voltage than can be obtained from a conventional rectifier. Such multipliers are generally used in circuits requiring low current and also when use of a power transformer is not desired. Because of its ability to produce high voltage without the use of a power transformer, the voltage-multiplier circuit is frequently used in high-voltage applications in which the cost of construction, weight, and size of the power supply must be held to the minimum. If used with a transformer, the circuit can be designed to provide a much higher voltage than is attainable with a conventional power supply.

3-16.4.1 Full-Wave Voltage Doubler

This circuit, shown in part (A) of Figure 3-509, uses two rectifiers, each of which conducts on alternate half-cycles of the input voltage. On the positive half-cycles of the input voltage, V_1 conducts and charges capacitor C_1 to the peak value of the input voltage, with the polarity as shown. Likewise, on the negative half-cycles of the input voltage, V_2 conducts and charges capacitor C_2 to the peak value, with the polarity as indicated. The polarities of the charges on the two capacitors are such that the voltages add, and thus produce an output equal to their sum. Therefore, the voltage available at the output is approximately double that for a half-wave circuit.

3-16.4.2 Cascade Voltage Doubler

The operation of the cascade voltage doubler, shown in part (B) of Figure 3-509, is slightly different from that of the full-wave doubler, shown in part (A) of the same figure. During the negative half-cycle of the input voltage, V_2 conducts and charges C_1 to the peak value of the input voltage, with the polarity as indicated. The voltage at point X will then consist of the AC input in series with the charge on C_1, and will vary in amplitude from zero to twice the peak value of the AC input voltage. On the positive half-cycle of the input voltage, twice the peak value of the AC input voltage is effectively applied to V_1, causing it to conduct and charge C_2 to twice the peak value of the input voltage. The cascade voltage doubler has the advantages that no transformer is required and that there is a common connection between the input and output so that both may be grounded.

3-16.4.3 Voltage Tripler

The voltage tripler produces an output voltage approximately three times that produced by an equivalent half-wave power supply. This circuit, shown in part (C) of Figure 3-509 consists essentially of a cascade voltage doubler whose output is in series with the output of a half-wave rectifier. The portion of the circuit that includes CR1 and CR2 is a cascade voltage doubler, which charges C_2 to produce and output equal to twice the peak value of the input voltage. The portion of the circuit that includes CR3 is a conventional half-wave circuit, which charges C_3 to the peak input value. Since C_2 and C_3 are connected in series (aiding), the output voltage is equal to the sum of the two capacitor voltages, and is three times the peak input value. The voltage regulation of this circuit is poorer than that of a voltage doubler if the same filter capacitor values are used.

3-16.4.4 Voltage Quadrupler

The voltage quadrupler, shown in part (D) of Figure 3-509, produces an output which is four times the peak value of the input voltage. This circuit consists of two cascade voltage doublers whose outputs are in series. When V_4 conducts, capacitor C_1 charges to the peak input value. This charge on C_1, combined with the input voltage, causes twice the peak value of voltage to be applied to V_3, which conducts and charges C_4 to this value. The conduction of V_2 charges C_2 to the same value. The voltage on C_2 is then applied to V_1, which conducts and charges C_3 to twice the peak value of the input voltage. Since C_3 and C_4 are in series, the voltages across them add to produce the output voltage, which is equal to four times the peak value of the input.

3-16.5 MULTIPHASE POWER SUPPLIES

3-16.5.1 Three-Phase Half-Wave Power Supply

The three-phase half-wave rectifier, shown in part (A) of Figure 3-510, uses a power transformer with wye-connected secondaries. The primary connections can be either wye or delta. The voltages induced in the secondary windings differ in phase by 120 degrees. Hence, each diode conducts for a 120-degree interval of a complete input AC cycle, and must carry one-third of the average current supplied to the load. The frequency of the DC pulses in the output is three times the fundamental power frequency.

3-16.5.2 Six-Phase Half-Wave Power Supply

The double-three-phase rectifier, shown in part (B) of Figure 3-510, employs a transformer having delta-connected primaries and star-connected secondaries. The use of this transformer connection has the advantage over the delta-wye connection, shown in part (A) of the figure, of producing less saturation of the transformer cores. The voltages induced into the star-connected secondaries differ in phase by 60 degrees. Each diode therefore conducts for 60 electrical degrees, and passes a current that is one-sixth of the average load current. The output of this circuit contains a ripple frequency that is equal to six times the fundamental power frequency. Because of this high ripple frequency and low percentage ripple content, filtering is greatly simplified. In multiphase rectifiers that deliver very large values of current, mercury-pool rectifier tubes are generally used.

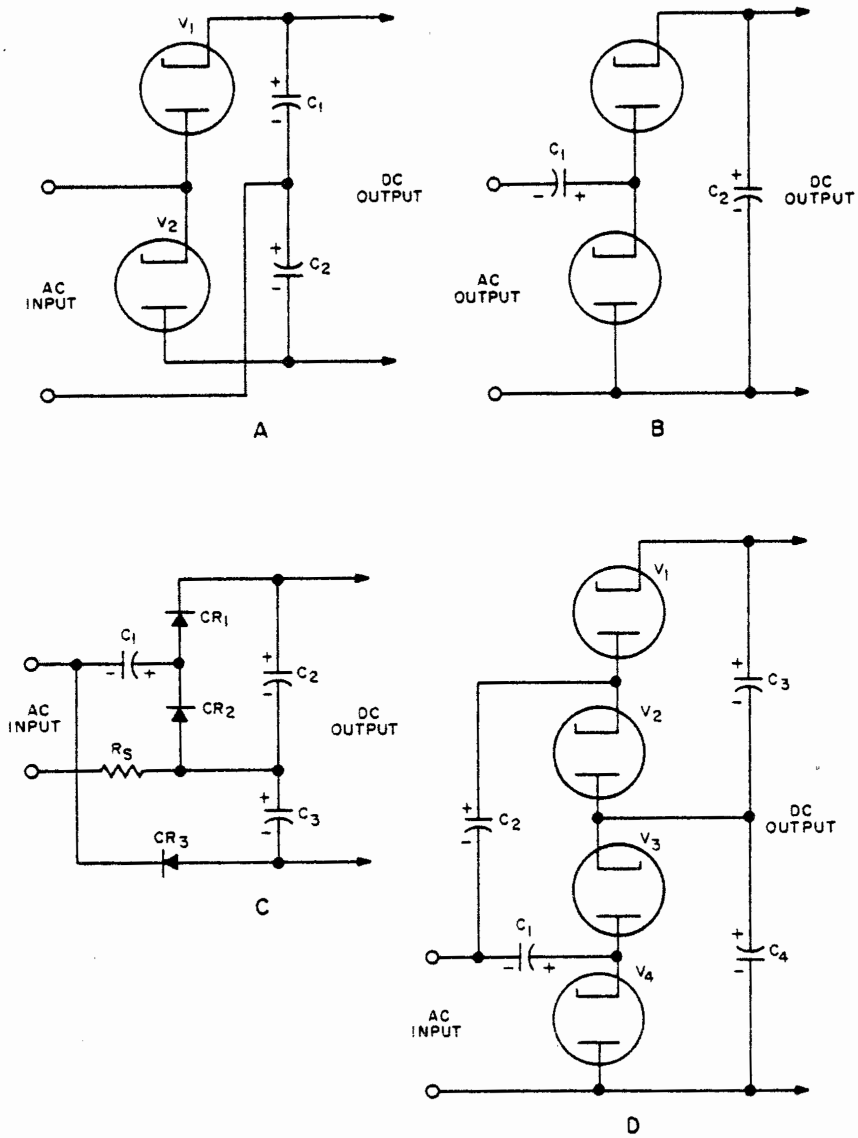
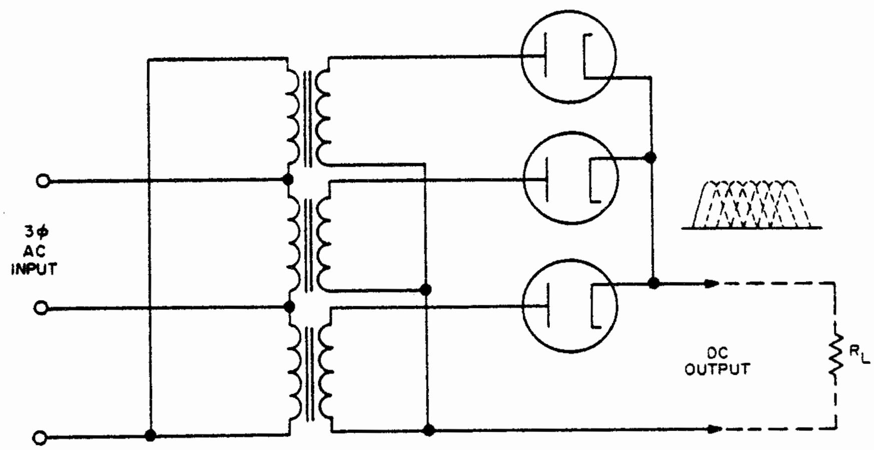


Figure 3-509. Voltage Multiplier Circuits

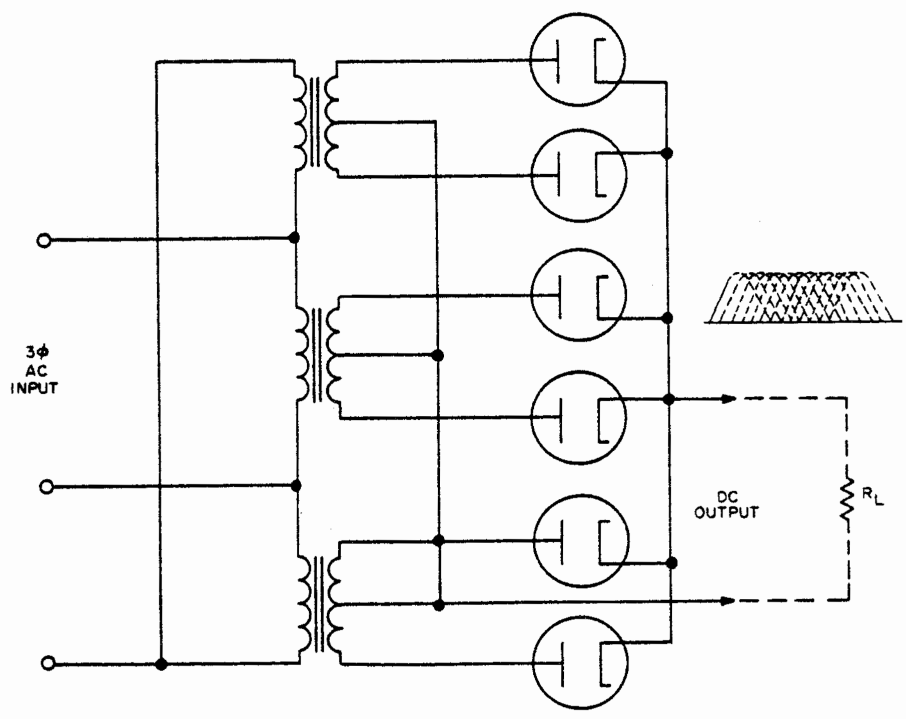
3-16.6 RF HIGH-VOLTAGE SUPPLY

The RF high-voltage power supply is a half-wave rectifier using a conventional LC power oscillator; it is used principally in cathode-ray-tube (CRT) applications where extremely high DC second-anode potentials at low current are required. The circuit shown in Figure 3-511 is typical. Tube V1 is connected in a modified Armstrong oscillator circuit, with tuning accomplished in the plate circuit. The oscillator output is transformer-coupled to the plate of V2 (connected as a half-wave rectifier). Since the step-up RF transformer (air core) has a tuned primary and usually a self-resonant secondary, high efficiency in terms of high voltage with low current is realized. Filament power for V2 is obtained from the oscillator tank circuit by the use of a small pickup loop. This method of obtaining filament power isolates the filament circuit from nearground potentials, as a precaution against voltage breakdown. Regenerative feedback for the oscillator circuit is provided by capacitive coupling between a metallic spring mounted around the glass envelope of the rectifier tube, V2,



A

THREE-PHASE HALF-WAVE



B

SIX-PHASE HALF-WAVE

Figure 3-510. Multiphase Power Circuits

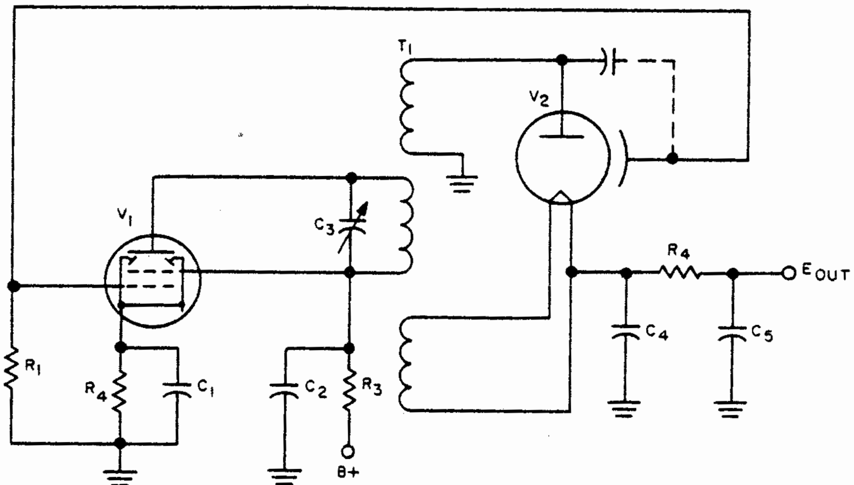


Figure 3-511. RF High Voltage Power Supply

and the plate of the tube. The output voltage level may be varied by C3; however, C3 is generally tuned to provide maximum stability rather than maximum voltage. The oscillator is normally tuned to operate at some frequency between 50 to 500 kilocycles. The output of the circuit is applied to an RC filter network. Since the ripple is in the radio-frequency range, small filter constants may be used, with a consequent saving in space, weight, and cost. Also, the shock hazard to persons coming in contact with this high potential is reduced because of the small amount of energy stored by these small filter components. The advantages mentioned above are the principal reasons for the use of this circuit in applications where its poor regulation and stability can be tolerated. Since the high DC voltage output is a result of the heavily stepped-up RF, failure of this circuit is often a result of failure of the oscillator. A fast check of this circuit can be made by measuring the grid voltage of V1, using an electronic voltmeter. Oscillator operation is indicated by a negative voltage reading.

3-16.7 VOLTAGE REGULATORS

Voltage regulators are used to maintain a constant voltage to the load, even though changes occur in input voltage or in load current. The regulators operate on the principle of the voltage divider: the voltage divider consisting essentially of a fixed resistance and a variable resistance.

3-16.7.1 Gas-Tube Regulator

The gas-tube regulator, shown in part (A) of Figure 3-512, is the simplest type of voltage regulator; it consists of a resistor in series with a gas tube, with the regulated output being taken from across the tube. This circuit operates satisfactorily to provide a definite output voltage (dependent upon the type of tube) if the variation in load current is within a certain range. Gas tubes are rated according to the voltage drop that appears across them, and according to the maximum current that can be allowed to flow through them. For example, a VR-105/30 gas tube maintains a 105-volt output, and has a maximum current of 30 ma. The internal resistance of the gas-tube depends upon the amount of current flowing through it. If a large current flows, the gas becomes highly ionized and hence has a low resistance. If a small current flows, the gas is only partly ionized and thus has a higher resistance. Therefore, the voltage drop across the tube is relatively constant over its entire operating range. In part (A) of figure 3-512, both the load and the gas-tube currents flow through the series resistor R. If the supply voltage decreases, the voltage across the gas-tube tends to decrease also. This effect causes the gas in the neon tube to deionize slightly, such that less current flows through the tube because of the increased resistance. The resulting decrease in current through resistance R lowers the voltage drop across R and thereby raises the output voltage to its normal value. If the output of a single voltage-regulator (V-R) tube is too low in value for certain applications, several gas-tubes may be connected in series and the output taken from the combination. The circuit shown in part (B) of Figure 3-512 is designed to provide three different regulated voltages (with low-current drain). If

VR-105/30 tubes are used, the circuit will apply 105 volts to load A, 210 volts to load B, and 315 volts to load C. For a gas-tube to conduct when a voltage is applied, the voltage must exceed the striking or firing level of the gas, which is the potential required to ionize the gas. For example, a 105-volt regulator must be provided with at least 133 volts to produce positive firing or starting. Such a tube will produce a 4-volt change over the current range of 5 to 40 ma, representing an AC impedance of about 115 ohms. If the gas-tube were replaced with a resistor designed for 105 volts at the mid-current rating of the tube (20 ma), the resistor would have an AC impedance of 5000 ohms. Thus, it can be seen that the V-R tube acts to prevent changes in voltage. Since it is possible to get negative-resistance effects in a V-R tube, it is recommended that the shunt capacitance for a tube of a given type be limited to the specific values indicated in a tube manual for that type of tube. The maximum value of capacitance that can be used effectively is usually less than 1 microfarad.

3-16.7.2 Electronic Regulator

The electronic regulator, shown in part (C) of Figure 3-512 is superior to a gas-tube regulator in that it maintains the output voltage more nearly constant and is also capable of operating over a much greater load-current range. Regulation is accomplished by using the plate-to-cathode resistance of an electron tube in series with the output of the power supply. The result is a variable resistance that provides the voltage drop necessary to compensate for any change in the output voltage. Since the plate-to-cathode resistance depends upon the plate current, the amount of resistance can be controlled automatically by connecting the circuit such that changes in the output voltage affect the grid bias. The value about which the resistance varies is determined by the initial bias level established for the tube employed. To best understand the operation of the electronic regulator circuit, assume that the output voltage supplied to the load is correct. (Refer to part (C) of Figure 3-512.) The cathode of the series tube, V1, is positive with respect to ground by the voltage across the load, while the grid is held somewhat less positive by the voltage drop across the gas-tube, V2. The difference between these two voltages is V1 bias voltage. This is then set at the proper value to cause the tube to have the required amount of plate-to-cathode resistance to produce the correct output voltage. If the output voltage increases by a small amount, the cathode potential of V1 increases too, but the grid potential changes very little because of the regulating action of the gas-tube, V2. Hence, the bias on V1 increases, and the tube resistance becomes greater. Therefore, a greater voltage drop occurs across V1 to compensate for the increase in the output voltages. A regulator with increased stability can be designed by using a pentode with a high amplification factor to control the resistance of the series regulator tube. The improved electronic regulator illustrated in part (D) of Figure 3-512 produces an output voltage that is relatively independent of input voltage and load changes over a wide range. The output voltage is developed across resistors R3, R4, and R5 in parallel with the load impedance. The load current flows through the regulator tube, V_1, which is in series with the output. The remaining elements in the circuit (R_2, V_2, and V_3) are used to control the resistance of V_1. The plate voltage of the control tube, V_2, is taken from the output of the regulator. The cathode potential of V_2 is taken from the output of gas-tube regulator V_3, and is therefore a constant voltage. The control-grid voltage is determined by the setting of potentiometer R_4, and is such as to provide the correct bias for V_2, which, in turn, determines the amount of plate current. The plate current of V_2 flows through plate load resistor R1, across which the produced voltage drop provides the bias for V_1. Therefore, the adjustment of R4 determines the resistance of V1. This adjustment is initially set for the desired regulated output. If the output voltage tends to increase (because of either a decrease in the load current or an increase in the rectifier voltage), the potential at the control grid of V2 increases. Since the cathode potential remains constant, a decrease in bias occurs, which allows a greater plate current in the tube. This increase plate current produces a greater drop across R1, which increases the bias on V1. This action, in turn, increases the voltage drop across V1. Therefore, a larger portion of the rectifier voltage appears across V1, and the regulator output is returned to the correct value. If the output voltage decreases, the action of the circuit is reversed. Since a pentode is used to control the resistance of V1, small variations in the output voltage are amplified sufficiently to operate the circuit. Because of its high sensitivity to voltage changes, an electronic regulator of this type will reduce any ripple that remains in the filtered output of the power supply. The anode of the gas-tube is returned to B+ through R2 to ionize the gas when the power supply is first turned on.

3-16.8 TIME DELAY CIRCUITS

In order to protect and lengthen the useful life of power supply rectifiers, a time-delay circuit to delay the application of plate voltage is incorporated in many equipments. The power transformers used with these circuits ordinarily employ separate filament and plate transformers. The most common time-delay circuits use a time-delay mechanism, usually a motor or escapement-driven

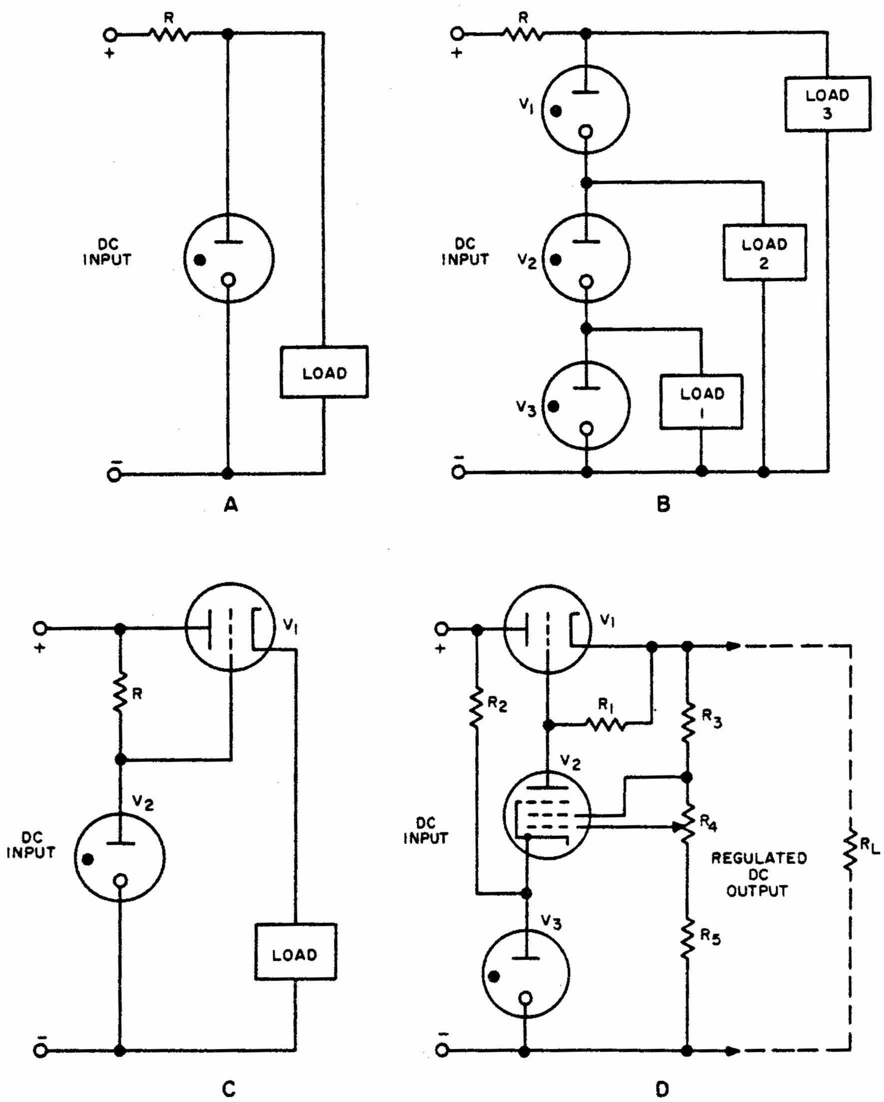


Figure 3-512. Voltage Regulators

relay. The filament transformer and the mechanism are energized simultaneously by the application of input power. The delay mechanism incorporates electrical contacts that close the input power to the plate transformer after a predetermined time (about 20 seconds). Another type of time-delay circuit employs a special power supply with a rectifier and plate-circuit relay for the purpose of controlling the input power to the equipment power supply. The filament windings of the rectifiers are connected in common with the special power supply. Input power is first applied to this power supply. As the rectifiers reach normal operating temperature, the plate current from the special rectifier actuates the time-delay relay in the plate circuit, thereby applying power to the equipment power supply. Variations of these delay circuits may employ components such as thermally-actuated relays, holding relays, etc; although the basic purpose is the same. The heater voltage and plate voltages are often controlled by separate manually-operated switches. In many situations, such as in locations of high humidity, heater voltage may be applied even though the equipment may not be in operation. When starting equipment thus equipped, it is good practice to apply filament power for 30 seconds or more before applying plate voltage, since longer heating times extend tube life. Hot-cathode gas-tubes, such as used in transmitters, public address amplifiers, etc, are used where high power is required. Tubes of this type must be brought to normal operating temperature prior to the application of plate voltage, or damage to the tube will result. High-vacuum rectifiers are much more rugged than gas-tube rectifiers because the cathodes are ordinarily made with a coating of thoriated tungsten. Under normal operating conditions, no damage to the tube occurs during the warm-up time. However, in certain applications, especially when high plate voltages are involved, preheating the tube will extend its life.

3-16.9 GENERATORS AND MOTORS

Rotating electrical machinery (motors, generators, dynamotors, amplidynes, etc.) are used for many purposes in almost every large electronic equipment. Proper service and repair of these components is essential for reliable operation of the equipment. All motors and generators consist of a rotatable member (rotor), usually an armature, enclosed in a magnetic field (stator). Various types of construction and electrical connections produce the different characteristics necessary for specific applications of the machine.

3-16.9.1 Generators

A generator is a machine that converts mechanical energy into electrical energy. A basic generator consists of a single-turn coil (armature) suspended in a magnetic field. When the coil is rotated at a constant speed, a sinusoidal voltage is induced in the coil. An alternating-current generator has a slip-ring connected to each end of the coil. The slip-rings are mounted on the shaft rotating the coil, and are insulated from each other. A stationary brush, made of either carbon or metal, rides on each slip-ring and transfers the resulting voltage to the generator load. One rotation of the single-turn coil produces one cycle of alternating current (360 electrical degrees). The basic direct-current generator is similar to the basic AC generator except in the method of taking the output from the rotating coil. In the DC generator, the rotating coil is connected to the external circuit by a device called a "commutator." This is effectively a switch that converts the alternating current from the rotating coil to current that flows in one direction through the external circuit. In general, a commutator consists of a metal ring, usually of copper, which is divided into a number of segments. These segments are insulated from each other and from the shaft upon which they and the rotating coil are mounted. For the basic DC generator described, the commutator consists of only two segments, with one segment connected to each end of the coil. Two brushes, mounted on opposite sides of the commutator, bear on its surface such that electrical contact is made between the coil and the external circuit. One rotation of the single-turn coil (360 electrical degrees) produces two pulses of direct current. The output waveform resembles the output of a full-wave rectifier.

3-16.9.2 Motors

A motor is a machine that converts electrical energy into mechanical energy. A magnetic field exists about any current-carrying conductor; the strength of this field is dependent upon the amount of current. When a current-carrying conductor is placed in a magnetic field, a force is exerted on the conductor, tending to move it out of the field. This is the fundamental principle of motor action. When current is passed through the coil of the rotor of the simple generator described above (and the rotor is in the proper position of the magnetic field), a turning of the rotor results. If the current of the coil is reversed at the proper time, the turning can be sustained to cause continuous rotation. Current reversal is accomplished by the nature of alternating current to operate AC motors; reversal is accomplished by means of commutator in the operation of DC motors. Motors can be considered the reverse of generators. Manufactured motors and generators are complex versions of the basic motor and generator described above. It is not intended in this text to enter the theoretical principles of operation of these machines. The subject is covered in physics text books and other obtainable publications.

3-16.9.3 Dynamotors

A dynamotor is a combined DC motor and generator operating in the same magnetic field. It may have two armatures on one shaft, or have two sets of windings and two commutators on one armature. Dynamotors are usually employed to generate high DC voltage when driven from a low-voltage supply. They are usually employed in mobile equipment when high-current and high-voltage stability is required. Dynamotors have generally been replaced by electronic power supplies or motor-generator sets, except in medium portable equipment employing vacuum tubes. As these latter systems become phased out, both motor-generator and dynamotors will be replaced by electronic power supplies.

3-16.9.4 Maintenance Considerations

Before placing into service a new generator, or one that has been shut down for a protracted period, a routine check should be made, according to the recommendations given below. The generator should be kept clean, inside and out. The interior can be cleaned with a vacuum cleaner, or by the use of clean, dry compressed air at low pressure (25 psi). The nozzle of the cleaning device should be nonmetallic. If a cleaning device with a nonmetallic nozzle is not available, a clean, dry brush should be used. Avoid personal injury while blowing out electrical equipment with air under pressure. Use goggles to avoid eye injury. Greasy dirt may be removed by an approved cleaning fluid. However, cleaning fluid should not be used on brushes or commutators. Any dents or cracks indicating rough handling should be investigated. If moisture appears to be present, it should be removed to avoid the possibility of insulation failure. Infrared lamps may be used for this purpose. In case of emergency, when no drying devices are available, the generator may be run with its armature short-circuited, and with just enough field current to warm the machine if possible (do not exceed 200 percent of the full load value); this should be done at a reduced speed. After drying, remove the armature short; then apply the load gradually, to prevent any sudden current surge that might result in a breakdown of the insulation. Disconnect the coupling between the generator and the power source, and rotate the generator shaft by hand. It should turn freely. Note the end play of the shaft, and the tightness of the mechanical and electrical connections. If the generator is rigidly connected to the power source, turn the crank or shaft, and check for binding. With the coupling connected, check again for freedom of rotation, as there may be misalignment between the two units. A belt-driven unit should have the pulleys in exact alignment, and the belts should depress 3/4 to 1 inch, depending upon the center distance and the pulley size. In no case should the belts be tight. A general visual inspection should be made for loose parts, frayed wire, cracked insulation, loose setscrews, frayed or cracked belts, etc. Electrical equipment operating in a salt atmosphere should be inspected daily for signs of corrosion or salt growths. Open or dustproof generators or motors should not be used outdoors, or where splashing or hosing down may occur. For operation under water, pressurized equipment should be used although totally enclosed motors may be used for short periods when flooded or submerged. Electrical equipment operating in dust-laden atmosphere should be inspected daily for signs of dust accumulation. Usually, desert operation does not shorten the life of electrical equipment if the equipment is protected from sand and dust storms. Severe damage can be caused by dust or sand which is carried in large amounts in the atmosphere, for the presence of these particles between moving surfaces is almost sure to cause failure. In some locations, atmospheric dust that is peculiar to that locality can cause corrosion or breakdown of insulation. Coral dust is a conductive material, and may cause arcing. Volcanic dust, in addition to its abrasive qualities, also causes corrosion when it is combined with moisture. Equipment for use in the tropics is usually tropicalized at the time of manufacture. Unfortunately, the tropicalization treatment given to electrical equipment is not effective over a long period of time, and it is necessary to renew the protective varnish or lacquer at frequent intervals. Since extremely humid atmosphere is the condition that causes fungus growth, it is necessary to bake the equipment in order to remove the moisture before revarnishing. This fungicidal varish does not take the place of the usual insulating varnish applied to coils, but is intended to supplement it. The presence of a heavy growth of fungus on or in equipment does not necessarily mean that the machinery is ruined. Usually, removal of the fungus, followed by baking and varnishing, will restore the power-generating equipment to service. Equipment must not be rebaked after the fungicidal coating has been applied, as heat destroys its effectiveness. Certain types of fungus do not require tropical conditions for growth; other types flourish even in arctic regions. If the electrical equipment is to be used where termite infestation is likely, chlorinate naphthalene should be added to the insulation varnish. Cellulose-base insulating materials are especiallly susceptible to attacks by termites.

3-16.9.5 Lubrication

The importance of proper lubrication cannot be overemphasized. Insufficient lubrication can cause serious wear in moving parts. Too much lubrication is also damaging, as the lubricant may get into the commutator and brush rigging where it will serve as a binder for copper and carbon dust collection, thus causing leakage paths and short circuits. Determining the correct amount of lubricant for ball bearings is one of the most important considerations in the maintenance of rotating machinery. Too much lubricant here can cause troubles just as readily as too little. Many machines have ball bearings designed to be greased with a pressure gun. Only a high grade of grease, having the following general characteristics, should be used for ball bearing lubrication:

1. Consistency a little stiffer than that of vaseline, maintained over operating temperature range.
2. Melting point preferably about 150 degrees.
3. Freedom from abrasive matter, acid, and alkali.

A sleeve-bearing generator must not be lubricated while it is running, as this will give a false indication of the oil level. Ball-bearing generators, however, should be greased while running, with the relief hole open. Approved cleaning fluid may be used to remove excess lubricant from the interior of the generator. Under ordinary conditions, greasing once a year is sufficient, while under severe service, every 3 months may not be too often. If the bearings have oil cups or reservoirs, any good oil with a viscosity approximating SAE 10 may be used. Turbine oil is excellent for this purpose, and enough should be added to bring the level to the proper point. (See Figure 3-513). On some generators, prepacked lifetime-type bearings are used, and no cups or fittings are provided; in such cases, the bearing caps are usually so marked.

3-16.9.6 Temperature Checks

Excessive temperature rise is usually the first indication of motor or generator trouble. A temperature rise of 55 degrees to 65 degrees centigrade is considered normal, with 40 degrees centigrade being considered the ambient temperature. Feeling the case or frame with the hand is not an accurate means of determining the temperature, but it will serve to indicate a hot spot on which to place a thermometer. A mercury or alcohol thermometer can be used to measure generator temperature if the bulb can be placed against the hot area. The thermometer can be insulated from the surrounding air by means of friction tape or other insulating material. It is advisable to check the temperature frequently over a 3- or 4-hour period, since a generator (if not overloaded) will not come up to, or exceed, its normal operating temperature for several hours. An excessively high temperature is always an indication of trouble. Overload should be suspected first. The load condition should be checked by means of an ammeter, and the reading compared with the current rating of the generator. Worn brushes and "high mica" contribute to a temperature rise. Clogged cooling vents, bent fan or impeller blades, and dirt are other causes. In two-pole generators, a frequently overlooked cause of temperature rise, brought about by sudden load changes, is eccentric oscillation of the armature about its axis. This causes excessive bearing load, and in a sleeve-type bearing, will cause bell-shaped wear of the bearings. Trouble of this type usually produces a rise in bearing temperature. Rubbing between the field pole and the armature may also be caused by sudden load changes, but such rubbing is usually indicated by a knocking noise. This condition may exist even when there is apparently sufficient clearance between the two moving parts. Knocking with sudden changes in load is due to a design fault, assuming that the shaft has not been sprung. The proper corrective measure is to reduce the load on the generator, and avoid imposing sudden, heavy loads on it. Commutators may overheat if the brushes are not on the proper load plane, if they are making poor contact, or if dirt is clogging the risers (in large commutators). Shorted turns in the coils or shorts to the metal parts will also cause excessive heating.

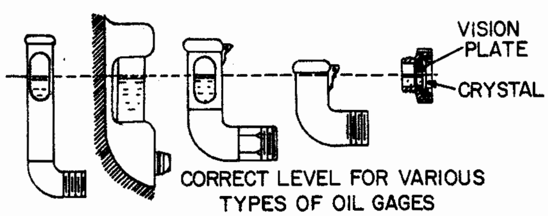


Figure 3-513. Proper Oil Level for Various Types of Oil Gages

3-16.9.7 Noise

Noise presents another indication of impending trouble. High mica, incorrect brush angle, an out-of-round commutator, excessive surface film, or too much clearance between the commutator and the brush holder with all cause abnormal noise. Noisy bearings may be caused by worn races balls or rollers or by lack of lubrication. Too much grease under pressure in a bearing, especially a semi-solid grease, will cause excessive noise. If sleeve-type bearings are worn out-of-round, they will make a knocking or pounding noise. Foreign objects or loose parts inside the generator may cause a clicking noise. Loose laminations or other parts will usually cause a humming or buzzing sound.

3-16.9.7.1 Electrical Noise

Electrical noise is usually caused by high mica, improper brush tension, shorted or open windings, defective shunt capacitors or excessively worn brushes. This type of noise can be isolated by monitoring the output voltage on an oscilloscope. Excessive electrical noise in a generator can cause RF interference or cause damage to electronic equipment, particularly where solid state devices are involved. The actual limits of "Spikes" developed in electrical noise must be derived from the technical manual for the motor/generator concerned.

3-16.9.8 Test Instruments

A milliammeter can be used in conjunction with a voltmeter to test an armature for shorted coils. The use of an ohmmeter is impracticable for this test, because the resistance of the coils is of such a low value. An ohmmeter is useful mainly for checking the continuity of shunt field coils. A voltmeter can be used to check voltages in any part of the circuit and can also be used for finding open coils. A megger is used to test insulation resistance to ground. The reading of insulation resistance must be considered in conjunction with other conditions if it is to be of any value. When certain machines come continually under the care of one person, is it advisable to keep records of insulation tests, since only in this way it is possible to have an accurate indication of insulation conditions. The original test may be plotted on graph paper at 15 second intervals for 60 seconds, or readings may be plotted once each minute for 10 minutes, using a motor-driven megger. Figure 3-514 shows typical readings; from these curves it can be seen that insulation resistance increases as the time of DC voltage application is increased. With a motor-drive megger, a steady reading should be obtained in 10 minutes. If a winding is wet or dirty, the steady reading will be reached sooner, as shown by curve 2 in the figure. After the windings have been cleaned and dried, the reading will probably approximate that indicated by curve 3 in the figure. Insulation resistance varies inversely with the temperature of 40 degrees (C). The resistance of insulation doubles for each 12 degrees (C) drop in temperature within the range normally encountered in the operation of these machines.

3-16.9.8.1 Motoring Test

For a motoring test, the generator must be disconnected from its prime mover. To test, connect the generator to a DC voltage source of some value near the rated output voltage (for example, a 150-volt generator to 100-volt source); include an ammeter of suitable range in the circuit. A generator that is in good condition will run as a motor current. A shunt generator with an open shunt field will not run, and may draw over three times as much current. A generator with a shorted or grounded armature will run unsteadily, and the current will fluctuate from 5 percent to 50 percent of its normal rated output current.

3-16.9.8.2 Testing for Insulation Breakdown

All electrical grounds in the generator must be disconnected. If the generator is of the open type, or if the brush rigging is accessible and the commutator can be reached, the generator can be tested for grounds without disassembly. A megger may be used, or a voltage source and a voltmeter or lamp may be used. Connect the ground lead of the megger to the shaft or the frame of the generator. Connect the test lead of the megger to the armature copper; then crank the megger and note the reading over a 60-second period. When testing with a

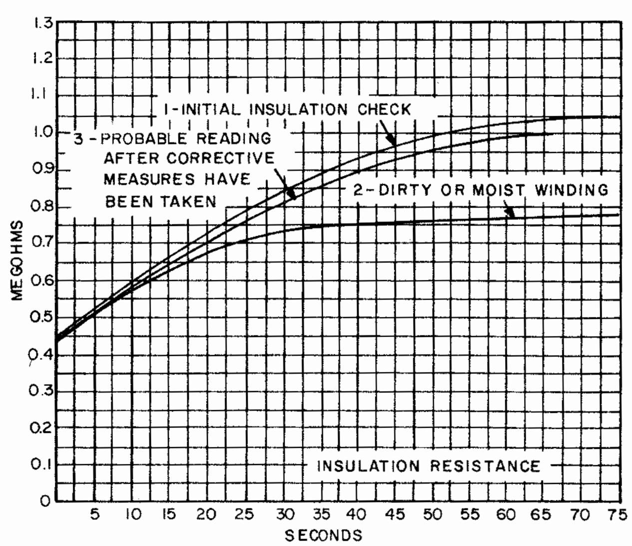


Figure 3-514. Insulation Resistance vs Time

megger, if the initial resistance is lower than approximately 1 megohm or is about 1 megohm or more and then suddenly drops to a low value, insulation break-down is imminent. A dead short (zero reading) indicates insulation breakdown. If a breakdown is indicated, isolate the armature from the fields by raising the brushes and placing a piece of insulation under them, to insulate them from the commutator. Connect one lead of the megger to the frame and one to the commutator copper. If an insulation breakdown is now indicated, the fault is in the armature. If not, the lead of the megger should be moved from the armature copper to a field-coil insulation breakdown. Generators or motors having defective armatures or field windings must be sent to an electrical repair shop for disassembly and repair.

3-16.9.8.3 Checking Commutator

Commutators should be checked for high mica, badly burned or pitted copper, etc. If these conditions exist, it will be necessary to remove the armature, "turn down" the commutator, and undercut the mica. If solder has been thrown out of the risers, this indicates excessive heat at the bars. Sometimes the mica between the segments of a commutator will burn out. There are two possible causes for this condition: 1) the mica may have become oil-soaked; 2) segments may have loosened, allowing foreign conducting material to become imbedded between them. Repairs can be made by scraping out the burned mica and then filling the holes with commutator cement. If high mica, flats, burned bars, or burned mica are in evidence, the generator or motor should be sent to an electrical repair shop for overhaul. The negative brushes of a generator continually deposit material, which is removed by the friction of the positive brushes. (In a motor, the action of the brushes is the opposite; the positive brushes deposit the material, and the negative brushes tend to remove it.) At some point a balance is reached, and the surface film is stabilized. A good commutator film is desirable and necessary. However, some grades of brushes may leave a deposit which is too thick and heavy for quiet operation. The degree of film thickness may be estimated by observing the color of the brush path on the commutator. The desired color when electrographitic or carbon brushes are used is glossy brown or light chocolate. A film that is too thick has a dull-black appearance. With this condition the brush friction increases, the temperature rises, and there is high-frequency brush chatter and sparking. If this condition is allowed to persist, rapid brush wear or brush destruction will follow. The film thickness may often be controlled by applying a more abrasive grade of brush. This remedy is not always practicable because brushes with added abrasive usually cause greater friction or have unsuitable commutating characteristics. However, when properly applied, they provide a means of film control. When brushes have too much abrasive action, brush chatter may result. In such cases, the brushes will not allow a film to form; it is rubbed off as fast as it is deposited. This is often true when graphite brushes are used. In this case the commutator will have a dull, brassy appearance. This should not be confused with the desirable lacquered-copper color formed by the proper grade of graphite brushes.

3-16.9.8.4 Checking Generator Under Load

If the cause of a fault is not obvious, and is not readily corrected, it will be necessary to shut a generator down at the first opportunity that permits further inspection. If these checks indicate that the generator is normal, start it up and slowly apply a load 25 percent in excess of the nameplate rating. Continue operation at this load for 1 1/2 hours, observing the machine carefully for signs of trouble. Throughout this test check the temperature carefully at any questionable spot. After 1 1/2 hours, reduce the load to normal or less. If trouble appears, shut the generator down immediately and remove it for disassembly and repair.

3-16.9.8.5 Testing and Servicing Brushes

In a direct-current generator, brushes usually give the first visual indication of trouble, both in the machine itself and in the equipment to which it is connected. A daily visual inspection of the brushes, while the generator is under load, will often highlight trouble long before it becomes too serious. The proper location for the brush rigging is usually marked with punch marks and paint by the manufacturer. The brushes should not be moved from this position until all other methods of correcting excessive sparking have failed. Brushes may have been displaced from their proper locations with respect to each other because the rigging or holder has been bent out of position. When trouble is indicated by excessive sparking at the brushes, these two conditions should be checked before proceeding further. The brush holder should not be more than 1/16-inch away from the commutator copper. Brush holders are usually staggered, and a slight longitudinal oscillation of the armature is desirable in order that the brushes will not always ride in the same path. This oscillation prevents ridges from forming on the commutator, and also exerts a scrubbing effect that helps to dislodge foreign particles. If the brush holder locator mark on the rigging does not match the mark on the frame, adjust the position of the holder assembly. Check the locations of the brushes with respect to each other by placing a piece of paper tightly about the commutator and then marking the positions of the brushes on the paper. When the paper is removed and laid out flat, the distance between the marks should be equal. Brushes that are binding in their holders cannot exert the proper pressure on the commutator, and the carbon in the brushes can be heated red-hot by the lack of proper pressure; this often causes an incandescent arc. The brush holders and the brushes should be clean, and the brushes should fit freely, but not loosely. If the brush holders are clean, sanding of the brushes to make them fit is permissible. Under no conditions should any lubrication be used on the brushes or commutator. Some brushes have a waxy composition, which softens under heat and helps lower friction (but at the same time introduces contact resistance). It is permissible to lubricate slip-rings and their brushes sparingly in AC machines.

3-16.9.8.6 Fire and Flashing

Oil will cause the mica between the segments to break down from bar to bar, or a shorted coil will cause a bar-to-bar flash. Bright flashes are caused by copper particles imbedded in the brushes; the lower resistance of this copper contact causes a heavy current to flow at this point, producing a severe, bright spark when the segment leaves the brush. Copper imbedded in the brushes by "copper picking" will also score the commutator. In such cases the brush should be replaced, or the copper picked out of the brush surface and the brush refitted to the commutator. (Copper picking is defined as the action in which the negative brush in a generator, or the positive brush in a motor, takes on a deposit of copper from the commutator, because of electrolysis. If the commutator risers designed for cooling are logged, an excessive temperature will build up at the brushes. High commutator temperature brings about the conditions most conductive to brush disintegration, with excessive sparking and possible fire. Fire and flashing are not to be confused with normal brush sparking. Normal brush sparking can occur at either the leading or the trailing edges of the brushes, depending upon the load plane. Occasionally, when a generator is operating under a variable load, heavy sparking of short duration may be observed. Such effects are caused by sudden changes in the magnitude of the load; the removal of a load can cause sparking just as readily as the sudden application of a load. The removal of a load causes a momentary surge of current during the time the field is relaxing. Before removing any brush mounting, always be sure that it has been properly marked, and that it has a corresponding mark on the frame, so that it may be replaced in its original position. When a generator is being operated under overload conditions, the commutator and brushes should be checked every 5 minutes for signs of overheating; if severe sparking occurs, the load must be reduced, or commutator failure will result. Only under emergency conditions should a generator be operated continuously under overload. Brushes should be replaced when worn down to the mark molded in the surface of the brush. This may be an arrow, a line, or both. If the brush does not have a maximum-weak mark, and if there is any doubt as to its condition, it should be replaced.

3-16.9.8.7 Replacing Brushes

When replacing brushes, the following points should be observed. If possible, always use complete sets, as one or several abrasive brushes are usually included in a set to aid in keeping the commutator film at the proper density. The abrasive brush is necessary for proper commutator action, but more than the required number of abrasive brushes would cause excessive commutator wear and would thus remove too much of the film. Make certain that the pigtail connection to each brush is secure. Remove any roughness inside the brush holders with sandpaper; also, make sure that the brushes and the holders are clean and of the right kind. Insert the brushes in the holders, making sure that they move freely, but not loosely. The brushes should be worked down to the proper fit by the use of sandpaper. In large machines, fit the brushes, one at a time, with sandpaper wrapped half-way around the commutator, or with a seating stone. A seating stone is made of soft, abrasive material. When the stone is held against the commutator and in front of the brush, abrasive material is carried under the brush. In small machines, the sandpaper should be wrapped around the commutator, secured with a rubber band, and the brushes worked down by rocking the armature back and forth. Use only sandpaper or a seating stone, since other abrasive papers may cause short-circuiting of the commutator. Make certain that the pigtails are securely connected to their terminals. Measure and adjust brush spring pressure. The proper method of making the measurements is shown in Figure 3-515. The following brush pressures are recommended for average use when the manufacturer's specifications are not available:

Electrographitic grade - 1-3/4 to 2-1/2 psi
Carbon and carbon-graphite grade - 1-1/4 to 1-1/2 psi
Graphite grade - 3 to 4 psi
Metal-graphite grade - 2-1/2 to 4 psi

(The tension on the lower brushes of large machines should be made slightly greater than the tension on the upper brushes, to compensate for the effects of gravity).

3-17 SWEPT FREQUENCY TESTING EQUIPMENT

Swept frequency is used to determine the bandwidth, alignment, frequency response, impedance matching and attenuation of various circuits, systems,

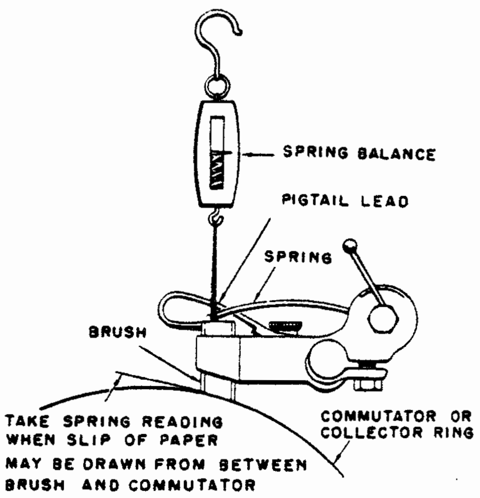


Figure 3-515. Method of Measuring Brush Tension

or components. It is usually exposed to quickly determine the broadband response of a device which otherwise would require a number of separate measurements and manual plotting of the response curve. Swept frequency techniques are applicable over the entire electronic spectrum from VLF to EHF, and are generally limited only by the ingenuity of the user and the limitations of the equipment employed. The basic sweep frequency arrangements are shown in Figure 3-315. The technique can effectively determine the frequency response of an amplifier or filter, and is useful in the alignment or bandwidth determination of an EF or RF stage. The test equipment employed permits direct visual readout of the effects produced by a change in adjustment or circuit component substitution. Figure 3-516 shows the test arrangement employed to check the frequency response of a termination device or antenna. The tracking generator used must be capable of sweeping the desired frequency range of the device under test (DUT).

3-17.1 TRACKING GENERATOR

The tracking generator shown in Figure 3-517 is basically a sweep generator whose sweep rate is matched to that of the spectrum analyzer, as illustrated in Figure 3-516. The output circuitry of the tracking generator contains a network that insures a constant output over the range swept. When the FM signal produced by the tracking generator is applied to a device or circuit under test, the instantaneous output amplitude is always proportional to the response of the circuit to the frequency at that instant. Thus, the original frequency-modulated input signal is changed in passing through the circuit under test. The output signal, therefore, now consists of an FM signal which is also amplitude modulated. For equal deviations, the positive and negative portions of this envelope are symmetrical, consequently it is only necessary to observe one side of the envelope. After the detection stage in the spectrum analyzer, only the modulation remains to appear on the face of the CRT. This presentation will appear as a continuous curve, due to the persistence of vision and the phosphor characteristic of the CRT. The detector's polarity determines whether a positive or a negative output is displayed. The frequency at any point on the CRT display can be analyzed by arresting the spectrum analyzer's scan either electronically or manually at the point of interest. For greater accuracy in frequency determination, a frequency counter may be attached to the output of the tracking generator at the point of the arrested scan.

3-17.2 IMPEDANCE MATCHING

Conventional tuners cannot be used successfully to cancel source or load reflections in swept frequency measurements. This is because the tuning is effective only at single frequencies, therefore pads or isolators are required. However, by the use of automatic level control, the power output of the sweeping generator can be maintained relatively constant at the point of measurement. The source impedance may thus be maintained very close to the nominal value. With this arrangement, any impedance variation in the connecting cables, connectors, and adapters are effectively cancelled since they are within the leveling loop. Thus, the attenuation of a DUT will be displayed in the associated CRT as a continuous response curve as it becomes scanned. This will result in an attenuation versus frequency plot of the DUT only.

3-17.3 OTHER SWEEP FREQUENCY TECHNIQUES

3-17.3.1 Impedance

Circuit impedance is measured conveniently by using the reflectometer principle. The individual values of the incident and reflected signals (SWR) in a transmission line feeding an unknown impedance are measured. The ratio between these signals indicates how closely the load impedance matches that of the transmission line. Another method used in waveguide holds the forward power constant by means of automechanic load control, and the

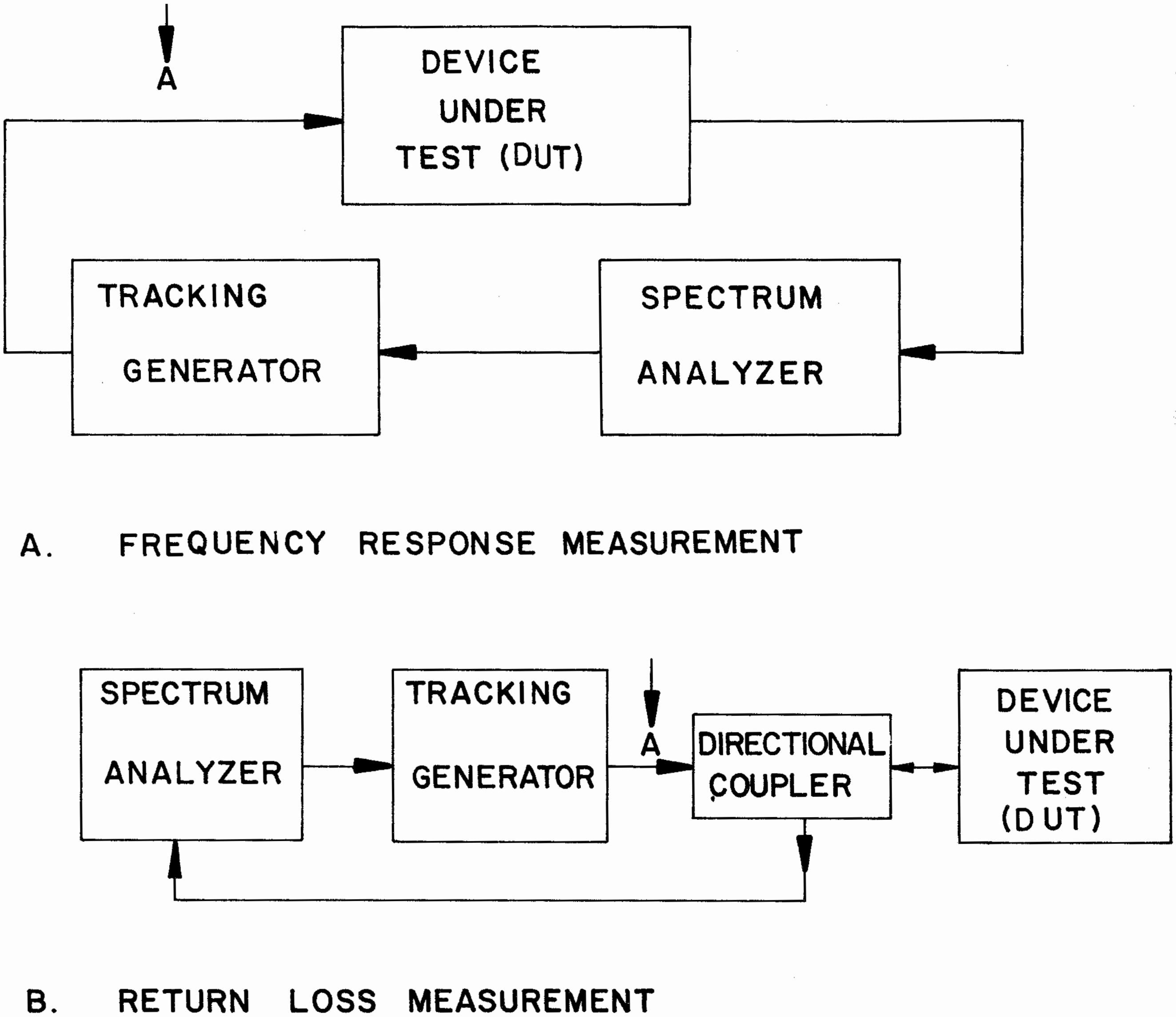


Figure 3-516. Swept Measurement Technique

return loss of a specific load is measured, rather than the ratio of the incident and reflected power. A short is then placed in the circuit and 100 percent reflected power is measured. The loss thus ascertained is then converted into SWR figures by calculation.

3-17.3.2 Noise Figure

By using a frequency-sweeping receiver and an automatic noise figure meter it is possible to make noise figure measurements on broadband microwave devices, such as a traveling wave tube amplifier. To conduct such a test properly, it is necessary first to check the receiver noise figure.

3-17.3.3 Additional Tests

Reflection losses and discontinuities in long waveguide and transmission line runs can be determined to an accuracy of about 1 percent by sweeping the line output through a mixer. The reflected signal returns and creates a heterodyne, since its frequency is now different than that of the incident signal. This is because of the time difference in the swept frequency due to the time required to travel down and then back along the line. This type of testing is discussed more fully in the time domain reflectometry testing procedures in section 5 of this Handbook.

3-18 COMPUTER EQUIPMENT TESTING

3-18.1 COMPUTER TYPES

Computers are used extensively by the Navy for such applications as air defense data processing,

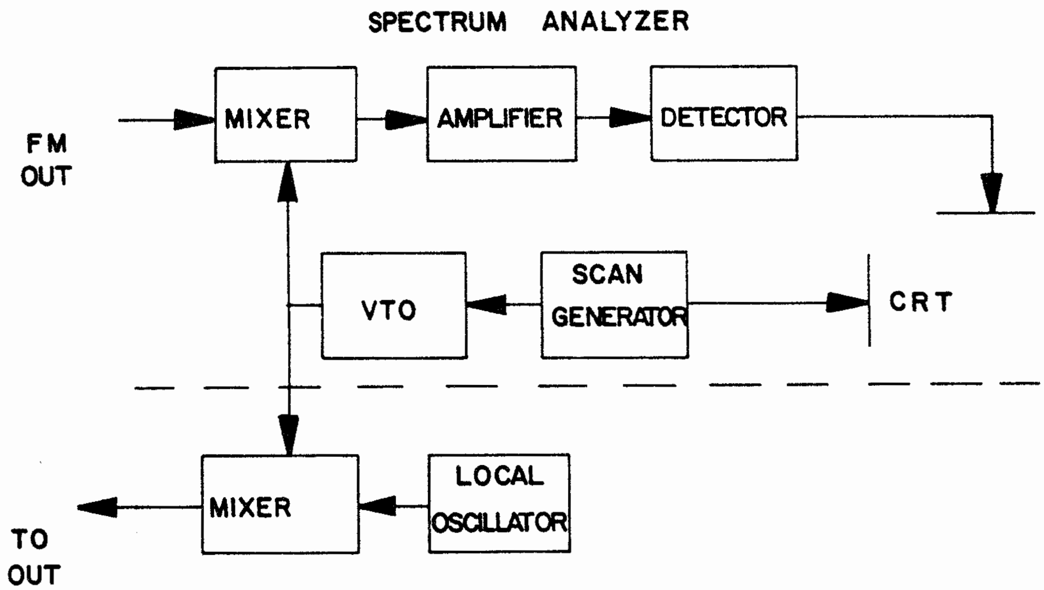


Figure 3-517. Tracking Generator

supply data processing, weapons control, ground control of aircraft, and telemetering. A computer may be a small unit consisting of a few circuits, or it may be a large complex equipment involving thousands of circuits. Two general types of computers are used: digital, and analog computers. Information applied to a digital computer is separated into discrete units for processing, whereas the information input to an analog computer is processed as continuously variable quantities. The digital computer performs mathematical operations by repeated addition, while several methods are used in analog computers. Large computers which require a high degree of accuracy are normally of the digital type. Although analog computers are not accurate enough for many applications, voltage or current representing specific conditions may be developed by analog circuits or devices and then converted to a digital form. After mathematical operations are completed, the resulting data may be converted to an analog voltage or current to operate certain controls or indicators.

3-18.1.1 Analog Computers

Analog computers are used for such applications as determining the altitude of targets detected by radar equipment, and solving specific types of mathematical equations for design calculations. Such computers are usually composed of summing, multiplication, differentiating, and integrating circuits. Figure 3-518 is a block diagram of the type of computer that can be used to calculate the altitude of a radar target. In this example an electromechanical computer receives values of slant range from the radar receiver and values of elevation angle from a synchro connected to the antenna, and converts this data into altitude information.

3-18.1.2 Digital Computers

The basic sections of a digital computer are shown in Figure 3-519. These perform the input, output, memory, arithmetic, and control functions. All data enters the computer through the input section, where it may be converted to a digital form and placed in a buffer storage device. The memory section accepts

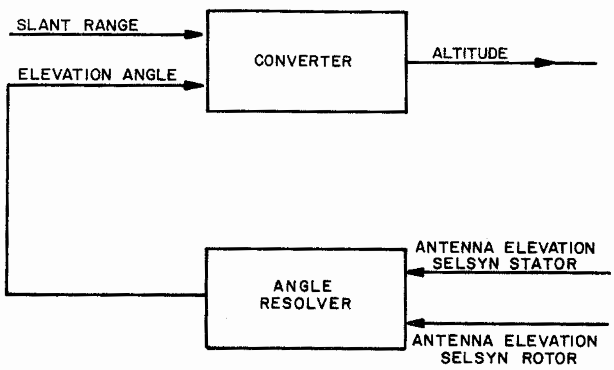


Figure 3-518. Analog Computer for Calculating Altitude of Radar Targets, Block Diagram

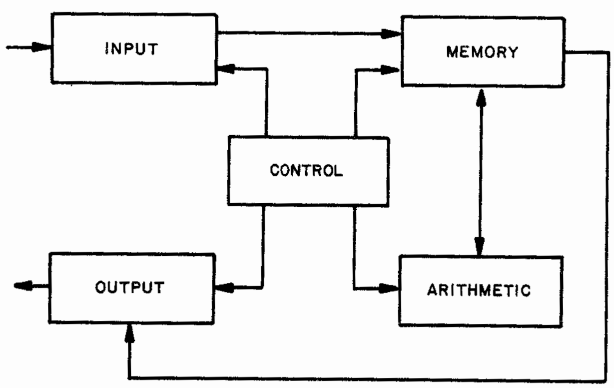


Figure 3-519. Basic Digital Computer Sections, Block Diagram

this data at specific intervals and places it in the proper storage location. Two paths lead out of the memory element: one to the control section and one to the arithmetic section. Instructions are transferred from the memory to the control section, where they are decoded and certain commands are set up for operation of the computer.

3-18.1.2.1 Input

The input section accepts information in various forms and converts it to a form which can be used by other sections of the computer. Input information is applied to computers by such methods as punched card readers, magnetic tapes, paper tapes, typewriters, and telephone lines that transmit data from distant sources.

3-18.1.2.2 Output

The results of the computer are fed to the output for either local or distant applications. Some examples of output units are line printers, card punchers, magnetic tapes, counters, cathode-ray tubes, and telephone lines that are connected to distant equipment. The output section varies widely from one computer to another.

3-18.1.2.3 Memory

The memory section stores information until it is needed by one or more of the other sections of the computer. Memory devices are assigned adresses, and the appropriate address is specified in the computer signal when information is needed or is sent to be stored. Several types of memory devices are used in computers; the most common are magnetic cores, magnetic tape, multivibrator circuits, and magnetic drums. At present magnetic cores are the most popular device, primarily because of their high speed, stability, and ability to retain information if there is a power failure.

3-18.1.2.4 Arithmetic

Addition is the basic operation of the arithmetic section of a digital computer; other arithmetic functions are simply variations of the addition function. For example, multiplication in the arithmetic section is simply a repetitive addition. The basic circuit in the arithmetic section is usually the multivibrator, which is used to store or transfer the results of computation.

3-18.1.2.5 Control

The control section generates signals at the proper time to cause a desired action to take place within the computer. This is normally accomplished by decoder circuits, along with multivibrators and other switching devices. The control section keeps track of the instructions that are to be decoded and may even perform part of the decoding. In addition, this section provides timing pulses that synchronize all sections of the computer.

3-18.2 MAINTENANCE TECHNIQUES

To meet reliability requirements and provide efficient troubleshooting procedures, special maintenance techniques have been developed for large computers. Maintenance program tests provide loop circulation of a simulated data pattern. The type of testing will determine whether the test signal becomes distorted in the circulation process and whether the control circuits in the loop allow passage of the entire succession of signals. Marginal checking circuits are included in some computers to detect aging of the circuit parts before a failure occurs. In addition to these special techniques, normal signal-tracing methods are used to locate the specific circuits in which malfunctions exist. Basic measurements are used to detect faulty parts.

3-18.2.1 Maintenance Programs

Computers can be given an overall check by means of maintenance programs. A maintenance program provides a thorough and rapid method of detecting failure in a special portion of a computer. This type of overall maintenance check is very flexible and efficient. These programs use the same type of tape, memory, computing, and drum circuits as the operational programs. A program can be changed when the computer or auxiliary components are changed, and the program can be constantly improved. No extra test equipment is required since the computer circuits are themselves used to perform the test. Testing by means of maintenance programs also results in the computer circuits being used in a more normal manner than during signal-tracing procedures. When a program has been checked and accepted as a good maintenance tool, it is not subject to deterioration. In contrast, test equipment may be checked and accepted only to become unreliable shortly after being placed in actual use. Maintenance programs are divided into three main classes: reliability, diagnostic, and utility programs. Maintenance programs employed to detect the existence of errors are called "reliability" programs. Reliability programs should be designed to check as many computer circuits as possible. Maintenance programs employed to locate the circuits in which computer malfunctions originate are called "diagnostic" programs. An effective diagnostic program should locate the source of trouble as precisely as possible. Actually, in many cases reliability programs have diagnostic features, and diagnostic programs have reliablity features. For convenience, a program is called either a reliability or diagnostic program depending on its intended emphasis. In general, programs that check rather than diagnose are shorter and simpler.

3-18.2.2 Basic Programs

A "program" is a series of instructions which control the operations of a computer. Each instruction is used to cause some action which is a part of the overall task the computer must perform. Therefore, an instruction may be considered to be the basic building block of a computer program. An efficient program makes full use of the instructions which are available to accomplish the task in the shortest possible time, and uses the least number of instructions. In most cases, one criterion, either time or the number of instructions, must be chosen over the other, and the program is developed along this line. If time is important, a maintenance program should be written which uses instructions of short duration, but may use quite a few memory locations for storage. On the other hand, if time is relatively unimportant, but only a few locations are available, instructions must be chosen which perform a number of operations or cause the computer program to run through the same program more than once. To write a satisfactory maintenance program it is necessary to have a thorough knowledge of the instructions that can be used. This includes execution time, the overall purpose of the instruction, when the instruction may be used, and the state of the computer after the instruction has been carried out. In addition, the programmer must know whether the instruction can be indexed and what internal conditions must be satisfied before it can be executed.

3-18.2.3 Reliability Programs

Reliability programs are used in both preventive and corrective maintenance tests to detect circuit failures rapidly and to discover failures that may occur only under particular operating conditions. Examples of troubles that are not evident at all times are failures that appear at specific repetition rates or for certain combinations of bits. To detect such failures, it is necessary to use reliability programs which check logical operation, paths of information flow, timing, ability of the computer to perform all functions, execution of instructions, etc.

3-18.2.3.1 Types

Reliability programs check either the logical functioning of an entire computer section or the logical functioning of individual circuit groups within a section. Whichever method is used, it is assumed that associated circuits which are not directly checked by the program are in satisfactory operating condition. Thus these programs can be considered to fall into two categories: first-order, and second-order. First-order reliability programs check the operation of an entire computer section, whereas second-order programs check the operation of assemblies or circuit groups, such as registers, counters, etc. In most cases, first-order programs are merely a combination of several second-order programs.

3-18.2.3.2 Interpretation

A reliability program provides a good-or-bad indication regarding the ability of the tested computer section or circuit to perform its design operating functions. For example, consider a reliability program that checks the switching time of relays with a specific section of a computer. As long as the switching time of the relays is within normal limits, the reliability program will indicate satisfactory operation. If switching time is excessive, however, there is an indication that maintenance is required. If the program runs successfully, there are no failures within the checked area. In event of a failure indication, the failure may be in the area being checked or in another area that has been assumed free of trouble. Diagnostic maintenance programs should then be used to pinpoint the source of trouble.

3-18.2.4 Diagnostic Programs

To be efficient, maintenance programs for diagnostic applications must narrow the area of a failure down to the smallest possible number of circuits. This can be accomplished by employing increasing-area, decreasing-area, overlapping-area, and large-area checks. The most effective method will depend on the particular type of computer being tested.

3-18.2.4.1 Increasing-Area Check

A maintenance program using the increasing-area check initially tests a small number of circuits. If a check indicates that all tested circuits are operating properly, successive checks are run in which progressively greater numbers of circuits are added. By this method, circuits which are found to be operating correctly are used to check other circuits. This process is continued until all circuits that can be checked by a maintenance program have been tested.

3-18.2.4.2 Decreasing-Area Check

When this method is used to find the source of a trouble, a large number of circuits are initially checked by the maintenance program. If trouble is detected in a large area, additional checks are made of successively smaller portions of the equipment until the stages affected by the failure are not included in the test area. It should then be possible to determine which stages are defective. If the check of a single large area reveals no error, the remaining large areas of the equipment are checked until the trouble is detected. In many cases, trouble can be located more rapidly by this procedure than by the increasing-area method.

3-18.2.4.3 Overlapping-Area Check

Another efficient method of narrowing trouble detection to within a small section of the equipment is the overlapping-area check. The routines of this type of maintenance program overlap each other. Thus, a failure will be located at the overlapping portions of those routines indicating the presence of trouble.

3-18.2.4.4 Large-Area Check

It may not prove feasible to program an effective maintenance test for certain small sections of a computer. A maintenance program is useful only to detect the general area in which the malfunction occurs. When the general area is located, conventional troubleshooting will be necessary to find the circuit in which a failure has occured.

3-18.2.5 Utility Programs

Utility programs are used as aids for both operation and maintenance programming procedures. This type of program is used to print out information from magnetic cores, magnetic drums, or other storage devices within the computer memory section. It is also used to transfer maintenance programs from punched cards or magnetic tape into the computer memory section. Utility tracing programs provide a printed record of the contents of various computer registers to enable follow-up on maintenance program operations.

3-18.2.6 Marginal Checking

Marginal checking is a preventive-maintenance technique used for some Navy and commercial computer equipment to detect the decrease in reliability of circuit parts due to aging. Aging circuit parts almost invariably change in value, current-handling capabilities, or in voltage limitations. Generally, the changes brought about by aging are gradual and without alarming variation in the normal operation of the equipment. For maximum equipment reliability, parts that are beginning to deteriorate must be detected and replaced before a failure occurs. Marginal checking is usually controlled by a maintenance program. The program directs the computer to perform the normal computer operations of addition, subtraction, etc., while the program varies certain circuit parameters about their normal values. In this way, the computer is made to perform normal functions under adverse operating conditions. To accomplish marginal checking, certain operating conditions are changed from their normal values. Since circuit-part values normally change with age, the variations that can be introduced before a failure occurs become less as the actural parts age. The amount of variation from the normal value that can be introduced before equipment failure occurs, is called the "margin of reliability" of the circuit or group of circuits being tested.

3-18.2.6.1 DC Supply Voltage Variation

The most versatile method of marginal checking is by variation (excursion) of the DC supply voltage for one or more circuits. Causing an excursion of a circuit's DC supply voltage will simulate the changes that normally result from the aging of circuit parts. Gradually increasing the excursion of the supply voltage to a circuit will eventually result in a circuit failure regardless of the circuit's age. Figure 3-520 shows the relationship between circuit reliability and the excursion voltage required to cause a circuit failure. The magnitude of the voltage excursion necessary to cause a failure is called the "margin of the voltage" on the circuit. This margin becomes smaller as the circuit ages. When the circuit fails at the normal operating voltage, the margin is zero. As long as the possibility of circuit failure is low the circuit is considered satisfactory. For the example shown in Figure 3-520 the circuit reliability of 80 percent is acceptable. When the voltage excursion necessary to cause a circuit failure decreases such that the circuit reliability is below the 90-percent value, maintenance must be performed to replace parts or an entire plug-in assembly. The level at which the circuit reliability is acceptable must be determined for each circuit or circuit group that is tested by marginal-checking methods.

3-18.2.6.2 Circuit-Part Value Variation

Failure of circuit parts can be anticipated by periodically simulating the aging of the parts. Figure 3-521 represents a circuit that can be selected by a maintenance program for marginal checking. The operation of this circuit is such that successive pulses place a "1" in FF1, transfer it to FF2, and then clear both flip-flop stages. During the marginal checking operation, an excursion is applied to the supply voltage line of FF1. Assume that, at some voltage value, the computer senses that a "1" was not transferred to FF2. The excursion is stopped and the margin noted. To determine which stage has failed, voltage and resistance checks can be made of the circuit parts of FF1. If this check indicates that FF1 is functioning correctly, it is possible that the output pulse from gate AG1 is so small that a slight change in the supply voltage applied to FF1 causes the circuit to fail. It is also possible that gate AG2 has aged to the point where any decrease in the signal output level of FF1 will result in failure of the pulse to be transferred to FF2.

3-18.2.7 BITE Testing

Most of the modern Navy computers use a maintenance system termed "Built-In Test Equipment" (BITE). This program periodically checks the condition of the computer. Some of the parameters that can be thus tested are:

a. circuit response time
b. component use time
c. component full time
d. pulse duration
e. check-sum
f. pulse amplitude

Part A of Figure 3-522 shows a typical pulse that can be measured by a BITE program. Detail A_1 in the figure shows the circuit's response time; i.e. the speed with which the circuit in question turns on and off. The computer program will open up a "window" during which the circuit must respond before the "window" closes. As a circuit ages, its response time will normally decrease to a point where the circuit will fail the test. Component use and full time is the interval the circuit requires to change its logic state. Pulse duration is a measure of pulse width at the

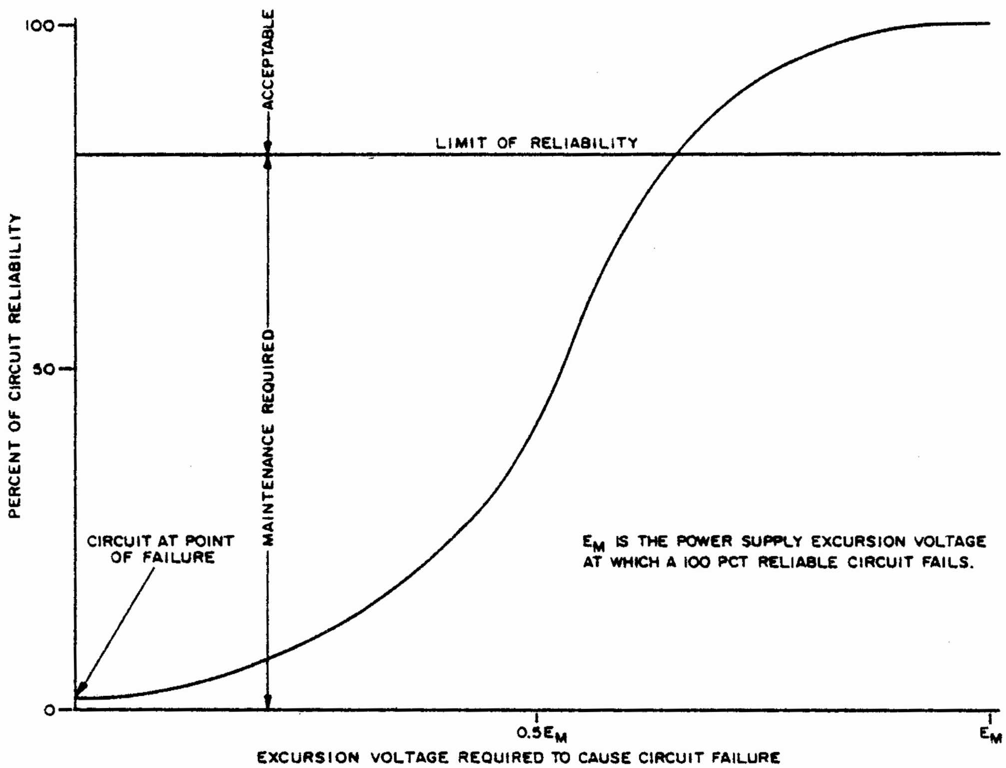


Figure 3-520. Circuit Reliability Versus Excursion Voltage Required to Cause Circuit Failure

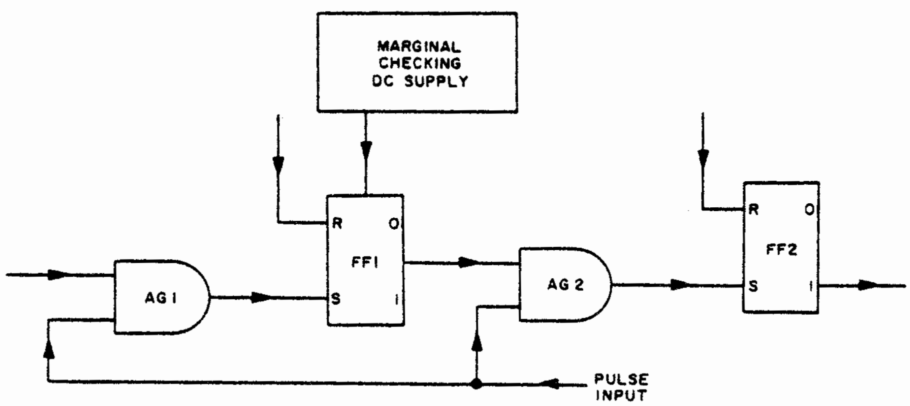
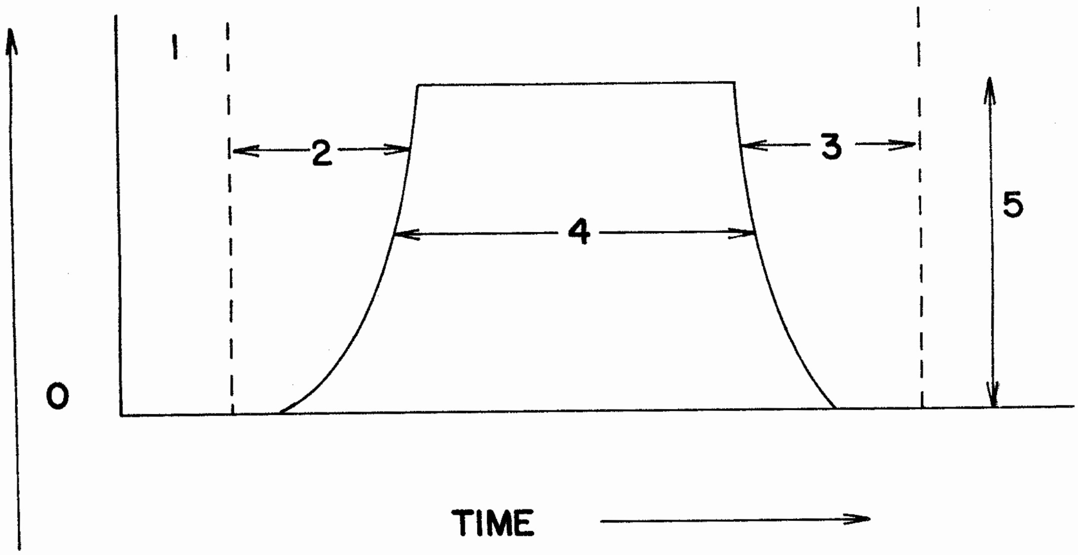
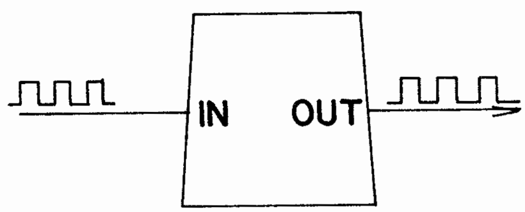


Figure 3-521. Typical Circuit Selected for Marginal Checking, Logic Diagram



A
PULSE PARAMETERS



B
CHECK SUM

Figure 3-522. BITE Parameters

half-power points. Pulse amplitude serves as a check of the pulse voltage level. If the circuit has deteriorated, the computer may not be able to recognize the pulse. If the pulse is supposed to be a logic "1", the computer may recognize the pulse as a logic "0" due to insufficient voltage level. Part "B" of Figure 3-522 illustrates the check-sum process. If the BITE program puts 3 pulses into a circuit, it should get 3 pulses out. This test is repeated frequently in the course of testing the computer circuits.

3-18.3 COMPUTER DIAGRAMS

Logic diagrams are usually included with computer maintenance instructions to show the paths of data flow through the computer circuits. The use of logic diagrams will often simplify overall circuits. The use of logic diagrams will often simplify overall circuit tracing and thus help the technician to locate trouble in equipment that has many circuits. When it has been determined that a trouble is in a specific circuit or group of circuits, conventional schematic diagrams should be used as an aid to locate faulty parts.

3-18.3.1 Logic Symbols

Symbols used in logic diagrams are geometric figures that represent the various types of computer circuits. Common electrical symbols are used to represent circuits and devices such as amplifiers, rotating machinery, and test points. Symbols peculiar to computer applications are used for circuits and devices such as magnetic drums, flip-flop stages, and gating circuits. Each symbol represents a function or, in special cases, a combination of functions. The symbols are interconnected by lines that indicate information or control signal paths, but not actual wire connections. As a rule, there is no correlation between the logical functions and the individual modules of a computer. One module may be used to perform several logical functions, or a logical function may require several modules. Summarization logic diagrams are sometimes included in a technical manual to further simplify explanations of equipment operation. Such a diagram may show groups of circuits that perform a related function, such as an accumulator, storage register, or full adder, as a single block. All signal inputs and outputs of the block are identified, and the function may also be marked on the block.

3-18.4 LOGIC GATES

In the field of digital computation, repeated use of a few basic circuits, sometimes referred to as building blocks or logic blocks, go to make up about 80% of a digital computer. The principles upon which logic gate operation is based (Boolean Algebra and computer arithmetic) will be employed here to help describe the circuit operation of the various logic gates. The technician should understand the binary system of numbers, where the only concern is with two values: "1" and "0" (zero). Any base-ten number can be represented in a binary form by a series of ones and zeros. In the digital computer, the ones are represented by a certain voltage level and the zeros are represented by a different voltage level. For example: Let +10 volts represent the binary one and +5 volts represent the binary zero. Another example can be seen in Figure 3-523 (Schmitt trigger output). The -5 volt level could represent the binary numeral "1", with the -12 volt level representing the binary zero. In the case of switching transistors, it is convenient to allow the "ON" condition of the switch to represent one of the binary levels and to allow the "OFF" condition of the switch to represent the other binary level. Pulse logic may be either positive or negative, and is defined as the

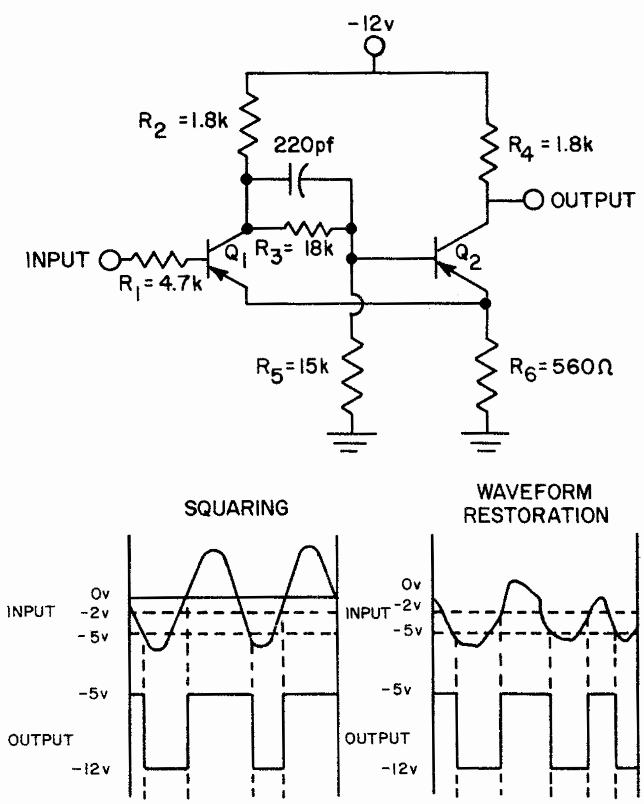


Figure 3-523. Schmitt Trigger

level of voltage difference between the one and the zero, using the zero as an actual zero. The choice of pulse logic is a manufacturer's design consideration. It is also true that once a type of pulse logic is selected, the same pulse logic is used throughout the computer. Figure 3-524 shows examples of both types of pulse logic, with binary ones ("1") and binary zeros ("0") placed above or below the voltage level used to represent them in the computer. The principles of Boolean Algebra must also be taken into consideration when determining the type of logic gate operation. The input terminals to the various gates are marked with letter designations such as A; B; X; etc. These designations represent the signal source, which may be a flip-flop circuit or an emitter follower. In any case, it must be understood that the signal source, such as "A" may be either a binary one or a binary zero, depending on the condition of that particular flip-flop or signal source. Figure 3-525 shows the four binary conditions that could exist in a two-input "AND" gate. This is the familiar "TRUTH TABLE" arrived at through the application of the laws of Boolean Algebra.

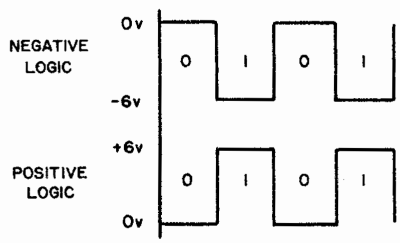


Figure 3-524. Examples of Pulse Logic

INPUT		OUTPUT
A	B	AB
0	0	0
1	0	0
0	1	0
1	1	1

Figure 3-525. Boolean Inputs and Outputs

3-18.4.1 The Inverter Circuit

Many computer circuits, such as AND gates and OR gates are basically common emitter amplifiers. They therefore cause the output signal to be 180 degrees out of phase with the input signal. A basic single stage inverter is shown in Figure 3-526. When the input to the base of the transistor is a logic "0" level, the positive voltage (called "pre-emphasis voltage") holds the PNP transistor cut off. This allows the output voltage level to hold at a logic "1" value (negative voltage). Likewise, when a logic "1" appears on the base (negative voltage input), the transistor saturates and E_c goes to effective zero volts, which represents a binary "0" output. The symbol used to represent the inverter in logic diagrams is shown in Figure 3-527. The diagram shows the Boolean value "A" as the input, since the value "A" becomes inverted through the circuit, the output is represented as $\bar{A}$, the bar or vinculum over the quantity denoting inversion. If two inverters are connected in cascade, the output of the second inverter will equal the input to the first inverter, as shown in Figure 3-528.

3-18.4.2 The AND Gate

In digital computers the primary function of the AND gate is that of coincidence. The only time that a binary "1" will appear on the output is when all of its inputs are binary "1" at the same time. Figure 3-529 shows the military standard symbol for a two-input AND gate. In this example, Figure 3-529, the "B" input has all "1's" applied, while the "A" input has "1101": reading LSD bit to the right. In this condition, the gate is said to be "qualified", because the output will follow the "A" input regardless of its binary value content. In Figure 3-530 the "B" input has all binary zeros applied. In this case the gate is said to be "inhibited" or "disqualified", because the output will always be binary zeros regardless of the binary values applied to input "A". There are several methods of connecting semi-conductor devices to produce the AND function. Figure 3-531 shows a two-input diode AND gate using positive logic. When a binary "1" and a binary "0", applied at the same time, cause the output level to be at a binary "0" value, then the gate is performing the AND function. (Refer to the AND truth table, Figure 3-525). Figure 3-532 shows a two-input diode AND gate that employs negative logic. Notice again that a binary "0" voltage level applied to one input at the same time that a binary "1" voltage level is applied to the other input will produce a binary "0" voltage level on the output. This is the action necessary to produce the AND function. Figure 3-533 shows transistors connected in a circuit so as to produce the AND function, using

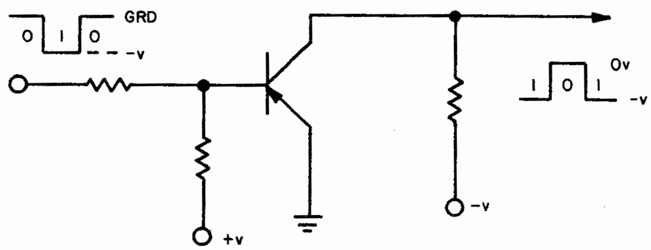


Figure 3-526. Basic Single Stage Inverter

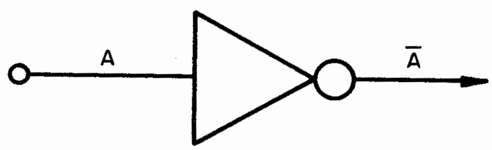


Figure 3-327. Inverter Symbol

negative logic. When both Q1 and Q2 are forward-biased by negative input pulses (representing binary ones applied), current will flow from the negative supply, through resistor R_c and Q1 and Q2 to ground. The transistors are thus acting as a switch, having maximum current and minimum resistance. All of the supply voltage is dropped across resistor R_c, which leaves the input to Q3 at ground (binary zero) level. Q3 in this case is simply an inverter. With binary zero input, the stage is cut off, allowing the output to ride at a negative (binary one) level. If either Q1 or Q2 were not forward-biased (binary zero applied), current would not flow through R_c. As a result, Q3 would be forward-biased, producing an output level of ground potential (Binary zero). Throughout the foregoing presentation only two inputs to the AND gate have been considered. In actual practice, the number of inputs to the AND circuit is variable between 2 and 5. The limitation to 5 inputs is imposed by loading considerations. Figure 3-534 shows how a "single" AND gate may appear if multiple inputs are necessary.

3-18.4.3 The OR Gate

The function of an OR gate in a digital computer is twofold. First, its purpose is to accept inputs from several sources and to join them such that the output appearing on one wire represents data from several sources. The output thus represents a union (collection) of information from these several sources. Second, the OR gate will act as a buffer or isolation stage, preventing interaction between signals applied to the various input legs. The military standard symbol for a two-input OR gate along with its logical operation in truth table form is shown in Figure 3-535. A "qualified" OR gate is shown in part A of Figure 3-536. The gate is said to be qualified if one of its input legs has a binary "0" level applied. This will allow the output to follow or reproduce the data applied to the other input. A binary "1" level applied to any leg of an OR gate will inhibit or "disqualify" the gate. Detail B of the figure shows a constant "1" level applied to the "B" input, thus the output will be a constant "1" level, preventing the "A" input data

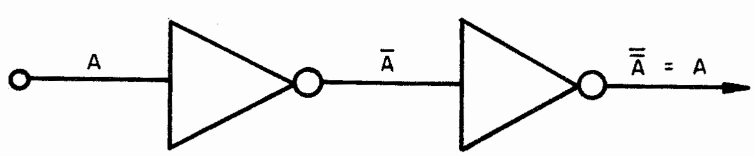


Figure 3-528. Inverter Cascading

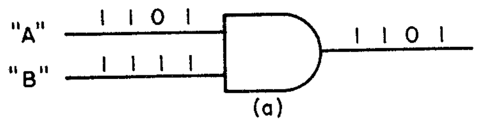


Figure 3-529. Two-Input AND Gates

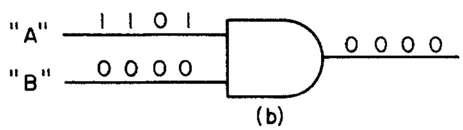


Figure 3-530. Two-Input "Inhibited" AND Gate

from appearing on the output. A simple diode OR gate utilizing negative pulse logic is shown in part A of Figure 3-537. Positive logic application is shown in part B of the figure. Two variations of transistorized OR gates are illustrated in Figure 3-538. An OR gate can theoretically have any number of inputs. In practice, however, the number of inputs is normally limited to 5, due to excessive loading encountered. To provide drafting room for multiple inputs, a single "OR" gate may be drawn as shown in Figure 3-539.

3-18.4.4 The NOR Gate

Another type of gating circuit employed in digital computers is the NOR gate. The NOR gate operation is similar to that of an AND gate in that it,

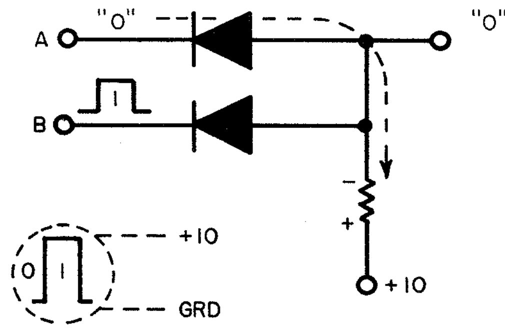


Figure 3-531. Positive Logic Diode AND Gate

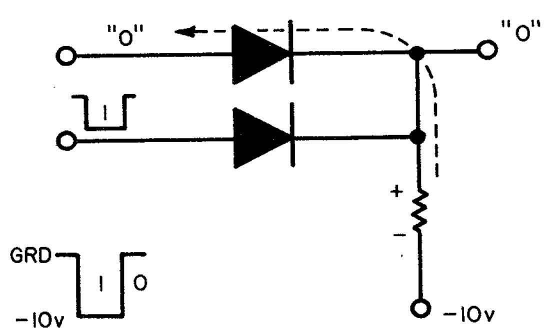


Figure 3-532. Negative Logic Diode AND Gate

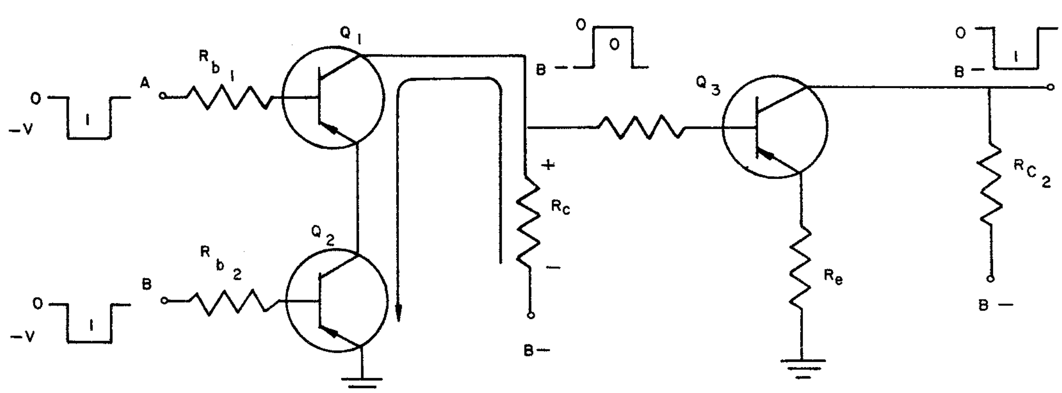


Figure 3-533. Transistor Connection to Produce AND Function, Using Negative Logic

also, will allow data to pass through the gate under certain input conditions, but not under all input conditions. The NOR gate is sometimes referred to as the "NOT" OR function. Symbolically it is represented by an OR symbol, with a small circle on the output side to denote signal inversion. The output is thus said to be the "NOTTED" OR function. The logic symbol for the NOR gate, along with its logical operation in truth table form, is shown in Figure 3-540.

The NOR truth table (part B in Figure 3-540) shows

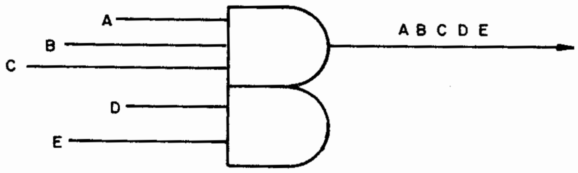
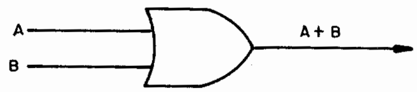


Figure 3-534. Single AND Gate with Multiple Inputs



OR GATE LOGIC SYMBOL

INPUTS		OUTPUT
A	B	A+B
0	0	0
1	0	1
0	1	1
1	1	1

OR GATE TRUTH TABLE

Figure 3-535. OR Gate Symbol and Truth Table

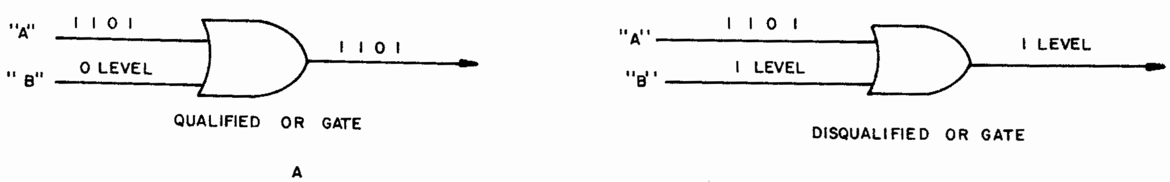


Figure 3-356. OR Gate Levels

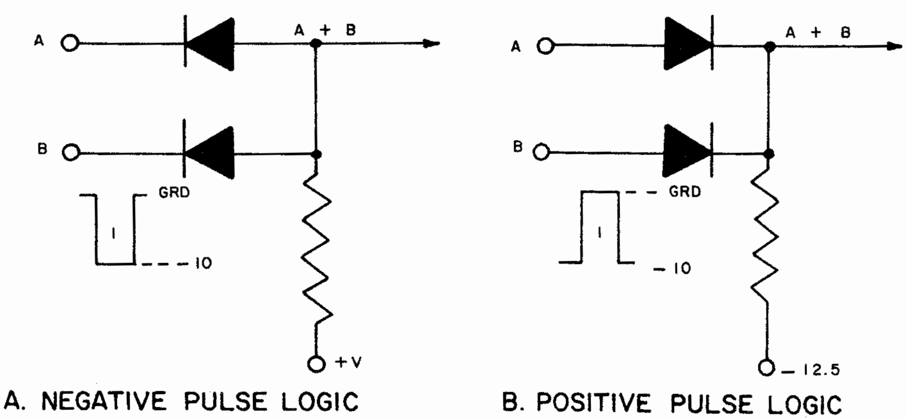


Figure 3-357. Diode OR Gate Pulse Logic

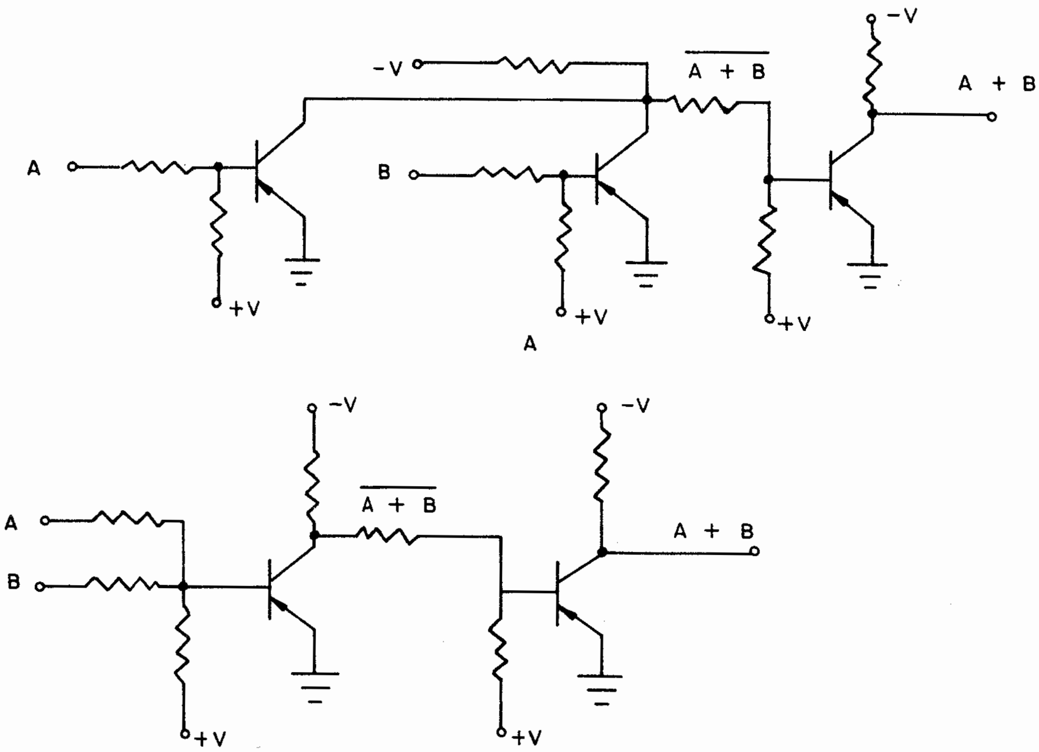


Figure 3-538. Two Versions of Transistorized OR Gates

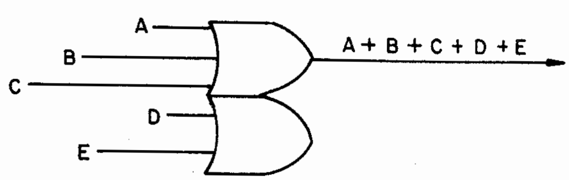
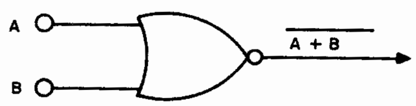


Figure 3-539. Five-Input OR Gate in Simplified Form

that the only time the circuit will produce a logic "1" on the output is when all inputs are simultaneously logic "0". In the case of the two-input NOR gate shown in Figure 3-540 the gate will produce a logic "1" output when neither of the A or B inputs is a logic "1". One of the simplest NOR gate circuits consists of a single inverter stage with multiple resistive inputs to its base. Operation with various input combinations to this type NOR circuit is shown in part A of Figure 3-540. Figure 3-541 shows a positive logic NOR gate, using diodes on the input. When +10 volts (binary "1" level) is applied to both input "A" and input "B", each diode will conduct, placing forward bias on the transistor. The potential on the output will be, effectively, zero volts (binary "0" level). When both diodes have zero volts applied at the input (binary "O" level), they will be forward-biased and will thus provide a low resistance voltage divider in the base circuit. The transistor will then be reverse-biased by the negative potential applied across the voltage divider. With the transistor cut off, the output voltage level will effectively be at source potential (binary "1" level).

3-18.4.5 The NAND Gate

The NAND function is quite similar to that of the OR function in that the gate will produce a logic "1" on the output if either input "A" or input "B": is a logic "0". The NAND gate is often called the "NOT" AND circuit because its output is the inversion or "NOTTED" form for similar inputs to an AND gate. The military standard symbol for the



A. NOR Gate Logic Symbol

INPUT		OUTPUT
A	B	$\overline{A + B}$
0	0	1
1	0	0
0	1	0
1	1	0

B. NOR Truth Table

Figure 3-540. NOR Gate and Truth Table

NAND gate, shown in Figure 3-542, consists of the standard AND gate will a small circle attached to the output side. The small circle denotes that all inputs or combinations of inputs will be inverted on the output. The figure also shows the logical operation of the NAND gate in truth table form. A simple NAND gate used with negative pulse logic (Figure 3-543) operates in the following manner. Transistors Q_1 and Q_2 are in the cutoff condition until a negative pulse is applied to both inputs to the circuit. The output voltage is therefore at a negative potential equal to source value until a negative pulse is applied to both inputs. When this occurs, output voltage essentially drops to zero. A positive logic NAND gate (shown in Figure 3-544) is constructed by adding an additional transistor (inverter stage Q4) to the AND circuit formed by the combination of Q1, Q2 and Q3.

3-18.4.6 The Emitter Follower

The emitter follower circuit (Figure 3-545) is analogous to the vacuum tube cathode follower and serves essentially the same function as does the cathode follower: namely, isolation and circuit matching. As with its vacuum tube counterpart, there is no polarity reversal at the output of this circuit with reference to its input, and the voltage gain is less than unity. There is, however, a power gain (Figure 3-545). The symbol representing the EMITTER FOLLOWER in this handbook is a plain block, with arrows to indicate the input and the output, as shown in Figure 3-546.

3-18.5 DIGITAL BINARY REGISTERS

Any bi-stable device is a "one bit" storage cell. Since it has two stable conditions, it can be used to represent either a "one" or a "zero". When using a group of two or more cells together, a binary word of two or more bits can be represented. This combination of cells describes three (3) important digital building blocks: the counter, shift register, and parallel. The method of changing the information they contain serves to differentiate between the three.

a. Counter: When the binary number in a register has its value changed by a discrete amount each time a pulse is applied at the LSD (least significant digit), it is functioning as a counter.

b. Shift Register: When the binary word in a register moves from right to left or from left to right one bit at a time, it is functioning as a shift register. (The input is generally at the MSD.)

c. Parallel Register: When a register has its configuration changed simultaneously in all cells by application of inputs to each cell, permitting a binary word to be moved into the register all bits at one time, with each cell in the register having an output permitting the binary word to be moved out all bits at once, it is functioning as a parallel register.

d. A single register may be designed to operate in either the serial mode described in (b) above, or in the parallel mode, as described in (c). It may be used as a buffer to convert from one mode of data flow to the other, with the bits moving into the register in parallel and then out in serial, or into the register serially and then out in parallel.

Figure 3-547 is a truth table that includes all possible configurations a three stage register can contain. (Note that eight is the third power of two (2^3).) Thus, when 16 configurations are desired four cells are

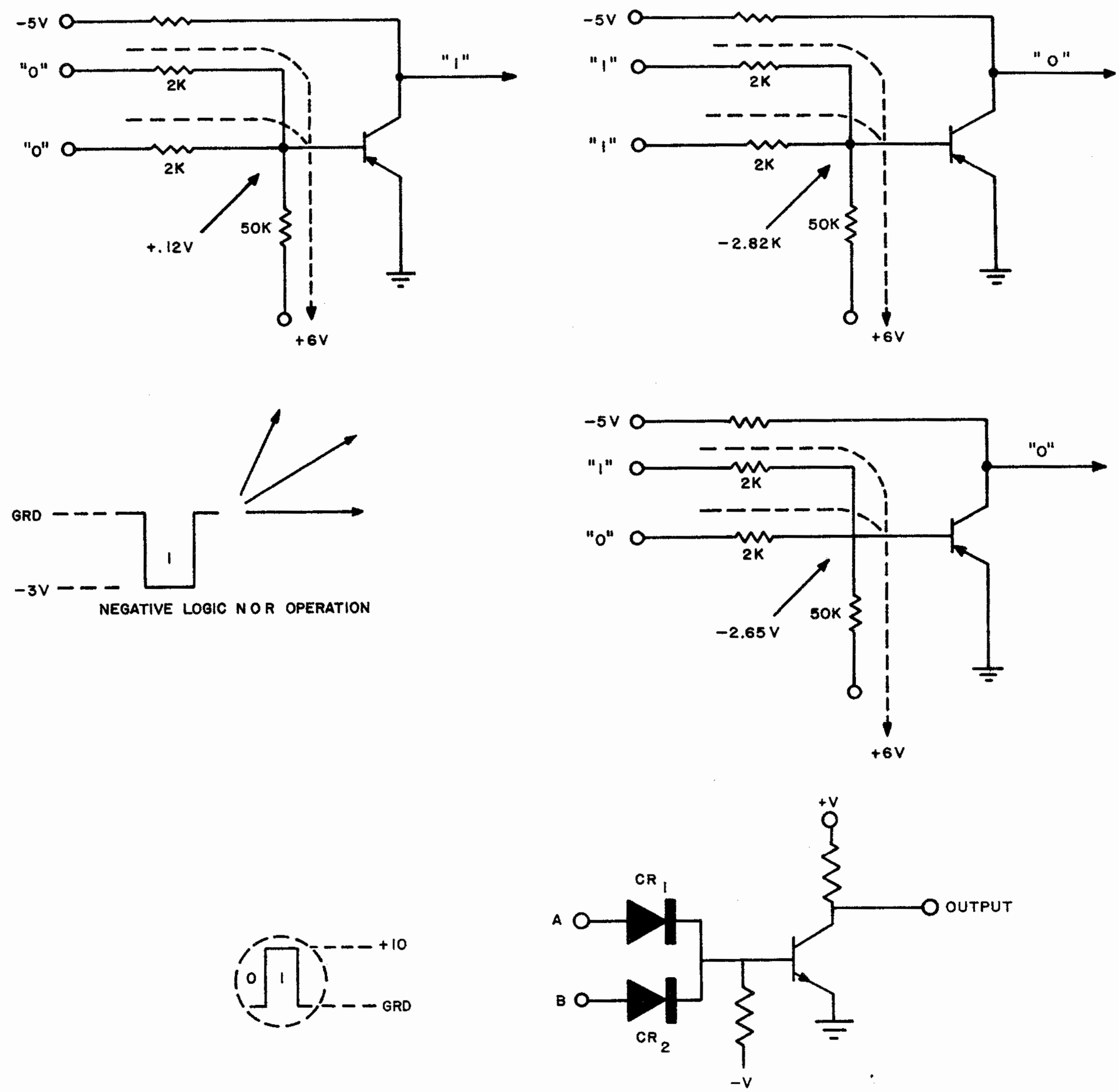
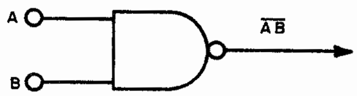


Figure 3-541. Types of NOR Gates

required: i.e., (2^4), etc. The power, then, determines the number of cells required. The basic approach to designing computer logic circuitry is to construct a truth table or chart of all possible variations of the units within the system. Such a truth table for a counter is illustrated in Figure 3-547. If read from top to bottom, it represents a "down" counter. When the count has progressed through the whole table, the next input pulse returns to the first condition. This table relates to a three-stage (bit or cell) counter which goes from 0 to 7 and then returns to 0, or goes from 7 to 0 and then returns to 7. This table thus shows the conditions for changing each stage, and will be used in the following example of how an "up" and a "down" counter are designed.



NAND GATE LOGIC SYMBOL

INPUT		OUTPUT
A	B	$\overline{AB}$
0	0	1
1	0	1
0	1	1
1	1	0

NAND TRUTH TABLE

Figure 3-542. NAND Gate and Truth Table

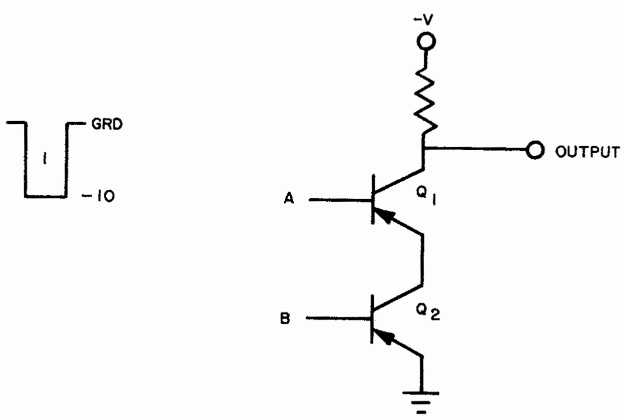


Figure 3-543. Simplified NAND Gate

3-18.5.1 UP Counter Design

The progression of the truth table (Figure 3-547) from top to bottom reveals the conditions for which the A stage is changed from 0 to 1 (set) to be a minus as follows. When A is O, B is O, C is O and a clock pulse occurs, or is expressed in Boolean $\bar{A}\,\bar{B}\,\bar{C}\,X$; also $\bar{A}\,B\,\bar{C}\,X$, $\bar{A}\,\bar{B}\,C\,X$, or $\bar{A}\,B\,C\,X$. This forms the minus expression - $\bar{A}\,\bar{B}\,\bar{C}\,X + \bar{A}\,B\,\bar{C}\,X + \bar{A}\,\bar{B}\,C\,X + \bar{A}\,B\,C\,X$ which, when reduced by standard Boolean manipulation, becomes $\bar{A}\,X$. The expression for changing A from 1 to 0 (resetting), developed in the same manner, is $A\,\bar{B}\,\bar{C}\,X + A\,B\,\bar{C}\,X + A\,\bar{B}\,C\,X + A\,B\,C\,X = A\,X$. The logic circuitry has thus been developed to control the A stage, illustrated symbolically in Figure 3-548A. The configuration illustrated in Figure 3-548A is used so generally in computer design that a special symbol has been

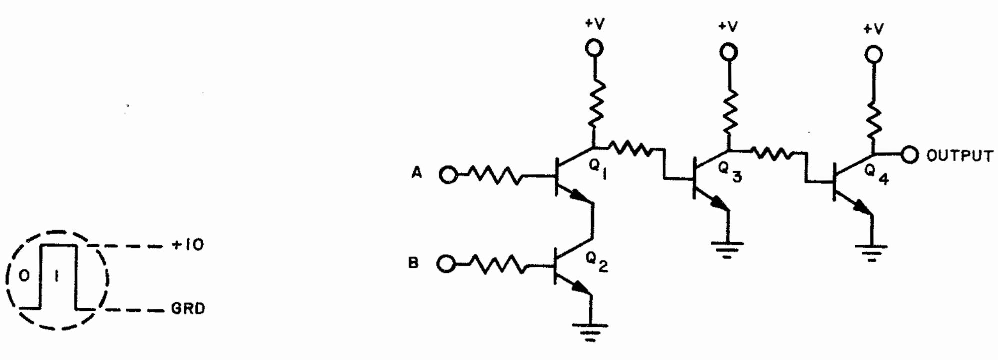


Figure 3-544. Positive Logic NAND Gate

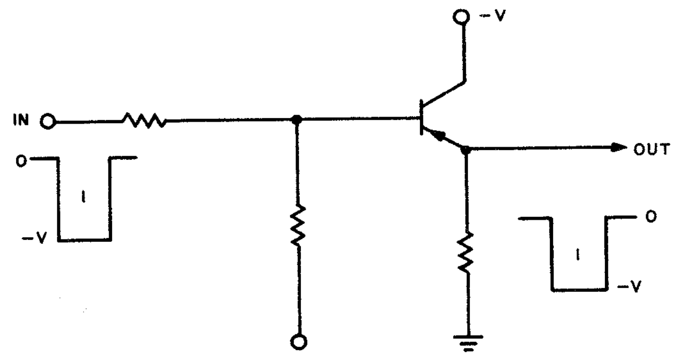


Figure 3-545. Emitter Follower Circuit, Showing Power Gain

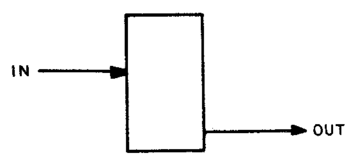


Figure 3-546. Emitter Follower Chart Symbol

developed to simplify drafting. Since the X input always changes the state from its current condition to the opposite, it is called a "toggle", being similar in action to a toggle switch. The "AND" functions are included with the F/F block, and the pulse is applied to an input, designated "T" as in Figure 3-548B. Hereafter, the symbol shown in Figure 3-548B will be used in this discussion. The next step in computer design is to develop the logic circuitry to control the "B" stage. Progressing from the top to the bottom of the truth table, you find the conditions that change B from 0 to 1 (set) condition when clock

C	B	A	OCTAL EQUIVALENT	BOOLEAN
0	0	0	0	$\bar{A}\ \bar{B}\ \bar{C}$
0	0	1	1	$A\ \bar{B}\ \bar{C}$
0	1	0	2	$\bar{A}\ B\ \bar{C}$
0	1	1	3	$A\ B\ \bar{C}$
1	0	0	4	$\bar{A}\ \bar{B}\ C$
1	0	1	5	$A\ \bar{B}\ C$
1	1	0	6	$\bar{A}\ B\ C$
1	1	1	7	$A\ B\ C$

Figure 3-547. Three-Stage Register Truth Table

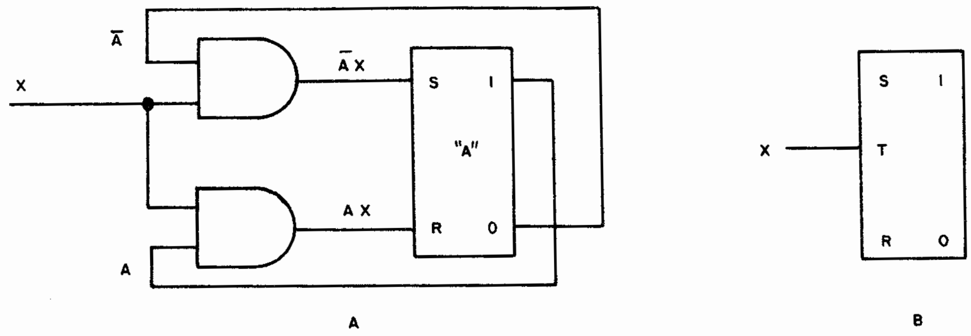


Figure 3-548. Control Logic Circuitry for an UP Counter

(X) occurs are shown in Figure 3-547, reading from top to bottom of the table. These conditions are $A\,\bar{B}\,\bar{C}\,X$ + $A\,\bar{B}\,C\,X$. When this expression is reduced by Boolean algebra, the expression becomes $A\,\bar{B}\,X$. The expression for resetting B is derived by the same method. The logic circuitry for setting and resetting the C stage is also found by use of the truth table (Figure 3-547). As shown on the table, C only goes from 0 to 1 (set) once, therefore the expression $A\,B\,\bar{C}\,X$ will set C, and the expression that resets C is A B C X. The following summarizes the Boolean expression just developed.

Set A $= \bar{A}\,X$ Set B $= A\,\bar{B}\,X$ Set C $= A\,B\,\bar{C}\,X$
Reset A = A X Reset B = A B X Reset C = A B C X

According to the above, "X" is common to the set and reset expressions for the A stage. Expression A X is common to the set and reset expressions for the B stage, and A B X is common to the set and reset expressions for the C stage. Using this information and the circuit of Figure 3-548, the UP-counter in Figure 3-549 is drafted. Observing the circuitry in the UP counter, it can be seen that the A stage will change state each time the clock pulse occurs. The B stage will change state only when the A stage is set and a clock pulse occurs. The B stage therefore changes state with every second clock pulse. The C stage will change state when A and B are set and a clock pulse occurs. The C stage then changes state with every fourth clock pulse. Comparison of this information with the truth table in Figure 3-547 shows that the conditions for an UP counter have thus been met.

3-18.5.2 DOWN Counter

A three-stage counter that starts from 7, progresses to 0, then goes back to 7 is called a DOWN counter. Inspection of the truth table in Figure 3-547, progressing from bottom to top, reveals the conditions

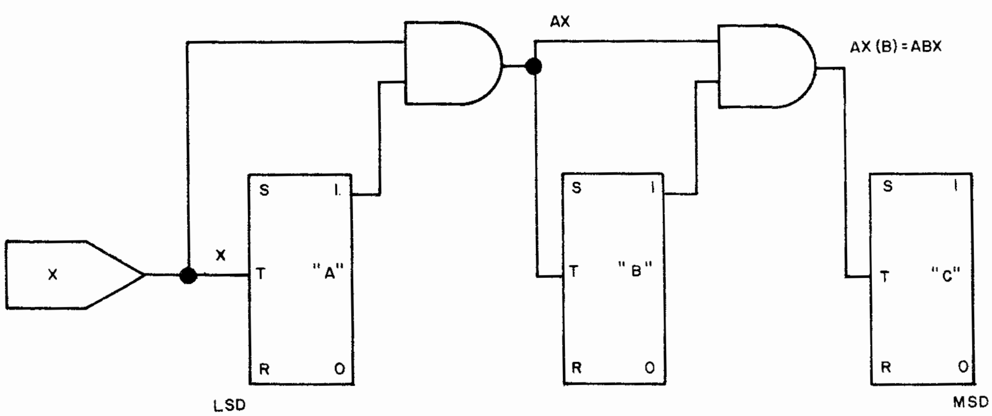


Figure 3-549. UP Counter

for which the A stage is changed from 1 to 0 (reset) when the clock (X) occurs. Taking these conditions from the truth table, the Boolean expression is $A\,B\,C\,X + A\,\overline{B}\,C\,X + A\,B\,\overline{C}\,X + A\,\overline{B}\,\overline{C}\,X$. Reducing this expression by Boolean algebra, it becomes AX. The expression for setting A is found in the same way. From the truth table, the expressions that reset the B stage when clock occurs is $\overline{A}\,B\,C\,X + \overline{A}\,B\,\overline{C}\,X$. This reduces by Boolean to $\overline{A}\,B\,X$. By the same reasoning, $\overline{A}\,\overline{B}\,X$ will set the B stage. The expression $\overline{A}\,\overline{B}\,C\,X$ is the reset expression for the C stage, whereas $\overline{A}\,\overline{B}\,\overline{C}\,X$ is the set expression. The following summarizes the Boolean expressions just developed.

Set A $= \overline{A}\,X$ Set B $= \overline{A}\,\overline{B}\,X$ Set C $= \overline{A}\,\overline{B}\,\overline{C}\,X$
Reset A $= A\,X$ Reset B $= \overline{A}\,B\,X$ Reset C $= \overline{A}\,\overline{B}\,C\,X$

Collecting the common expressions for setting and resetting the stages, we find X applies for the A stage, $\overline{A}\,X$ for the B stage, and $\overline{A}\,\overline{B}\,X$ for the C stage. Using this information and the circuitry of Figure 3-548, the DOWN counter in Figure 3-550 is drafted. Observing the circuitry in the DOWN counter, it can be seen that the A stage will change state each time a clock pulse occurs. The B stage will change state only when the A stage is reset and a clock pulse occurs. The B stage therefore changes state with every second clock pulse. The C stage will change state when A and B are reset and clock occurs. The C stage then changes state every fourth clock pulse. Comparison of this information with the truth table in Figure 3-547 shows that the conditions for a DOWN counter have been met.

3-18.5.3 Double Rank Counter

A Double Rank Counter (DRC) is a form of counter using two flip flops for each bit to increase dependability. The F/Fs are arranged in ranks, the first rank advancing the count and the second rank storing the state until the pulse has passed. The first rank has the same inputs as a conventional counter but they are taken from the second rank. The second rank does not change state (count) until application of a "Transfer" pulse which occurs between the "Advance" pulses. The levels on the gates are thereby prevented from changing while the Advance pulse is present, no matter how fast the F/F is changing states. The count in Rank 1 is parallel dropped into Rank 2 by the Transfer pulse. Advance and Transfer pulses occur alternately. As shown in Figure 3-551, the output levels of flip flops 1A, 1B and 1C are applied to the inputs of 2A, 2B and 2C respectively, ANDed with Transfer. This causes a parallel drop of the count in Rank 1 into Rank 2 at Transfer pulse time. Rank 2 duplicates 1 at the time of Advance. The logic inputs to Rank 1 are:

Set 1 A - $\overline{(2A)}$ (Advance)
Reset 1A - (2A) (Advance)

Set 1B - (2A) $\overline{(2B)}$ (Advance)
Reset 1B - (2A) (2B) (Advance)

Set 1C - (2A) (2B) $\overline{(2C)}$ (Advance)
Reset 1C - (2A) (2B) (2C) (Advance)

The above is the normal logic for a binary counter, considering that Rank 2 is exactly like Rank 1 at the time Advance occurs. However, the second Rank does not change state until Transfer time, and cannot change state while Advance is present, however fast the flip flops may be. The Advance and Transfer pulses are generated by synchronized clocks occuring alternately. By the appropriate-logic inputs, the DRC may

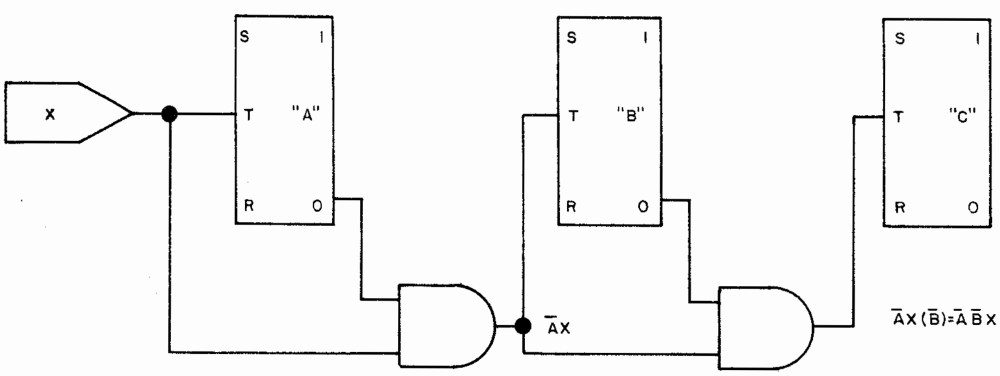


Figure 3-550. DOWN Counter

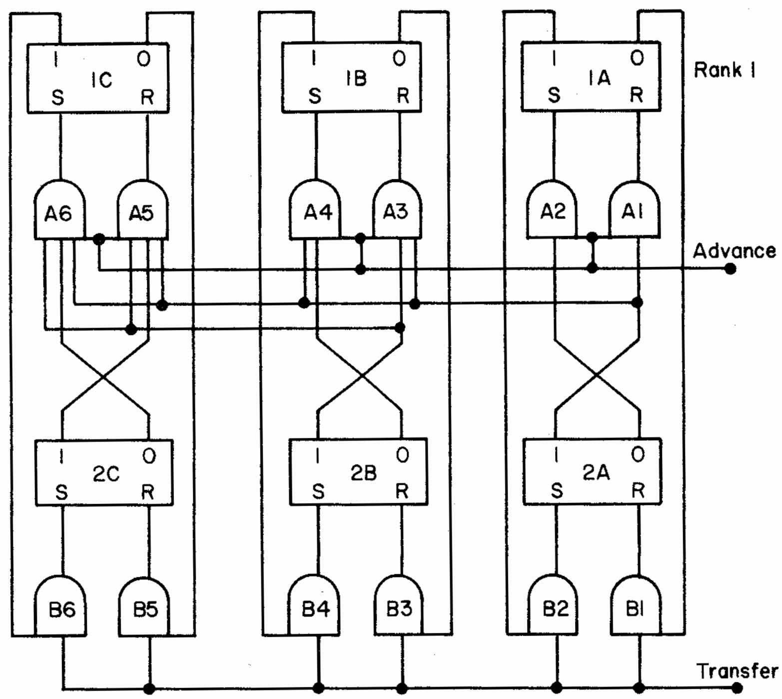


Figure 3-551. Double Rank Counter

be made to advance in accordance with the progression of Grey Code, BDC, or any other system of counting. Figure 3-552 is a timing chart showing the sequence of levels on the function output side of the three-stage DRC illustrated in Figure 3-552, using a 2.5 mc clock pulse. In summation, then, the DRC is a development of the binary counter designed to be highly reliable at high counting frequencies.

3-18.6 TIMING

In the process of comparing words in a computer, such as in addition, the words must have each unit compared in the proper sequence. The user must be able to determine time at any given instant with respect to the overall computer cycle or frame. Timing is based on the operating frequency, or "clock". In a serialized computer, each bit of every binary number occurs in coincidence with clock and can be gated through the circuits by clock. Each clock pulse is identical, repeating indefinitely, therefore we must have some method of identifying a particular clock pulse. An identified clock pulse is called a timing pulse (TP).

3-18.6.1 Timing Pulse Generation

The counters discussed in the previous section progressed through eight different configurations for each group of eight clock pulses applied. By designating each of these configurations as a timing pulse, eight TPs will be generated each time the counter progresses through its count sequence. The table in Figure 3-553 shows the relationship between TPs, binary configuration, and actual count. It should be noted that the TP count is one ahead of the actual count. The TPs generated by ANDing the individual bits in a configuration are in NRZ (non return to zero) form. A TP generator of this type is shown in Figure 3-554A. The relationship between clock and the TPs generated are shown in Figure 3-554B. The TPs generated have a pulse width equal to the time between clock pulses. Under certain circumstances, this could be a disadvantage in that the same clock pulse that ends one TP initiates the next. It may therefore be desirable to generate TPs in RZ (return to zero) form. Generation of TPs in RZ form is accomplished by applying clock to each TP AND gate, as shown in Figure 3-555A. The delay shown between clock and the counter is required to prevent the generation of two TPs with the same clock pulse; i.e., without the delay, the counter would be changing its configuration in coincidence with the application of clock to the TP gates. It may therefore be possible to have the counter change state before clock has gone, generating an output from two TP gates. The relationship between clock, delayed clock and TPs are shown in Figure 3-555B.

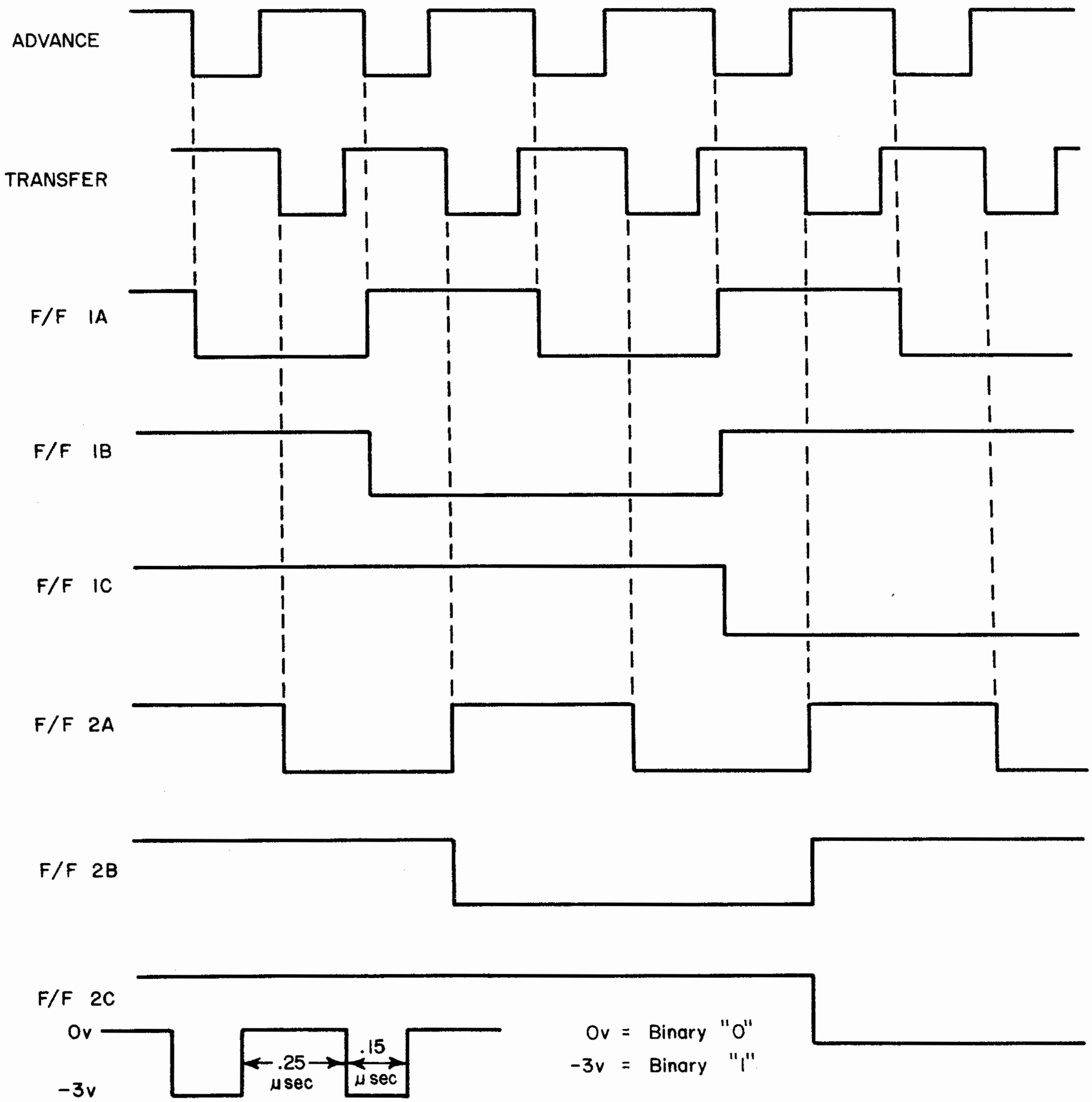


Figure 3-552. Timing Chart

3-18.6.2 Word Time Generation

Digital computers may be classified in general as to the way they are timed, either synchronous or asynchronous, and the way that data (binary configurations) is moved, either serial or parallel. Since asynchronous operation is more adaptable to parallel data flow, it will be discussed later. In the synchronous computer the movement of each bit is synchronized to a specific clock or timing pulse. This mode of operation will be used to develop word time generation. In order that the following discussion be best understood, the following terms are defined:

Serial operation - all bits in a word are moved sequentially (bit by bit) through the gates.

Word - a group of bits treated as a unit.

TP's	Binary Configuration			Actual Count
	A	B	C	
TP1	0	0	0	0
TP2	1	0	0	1
TP3	0	1	0	2
TP4	1	1	0	3
TP5	0	0	1	4
TP6	1	0	1	5
TP7	0	1	1	6
TP8	1	1	1	7

Figure 3-553. Clock Relationships

Word Time - the time required to shift word through a gate.

The length of a word (number of bits making up the unit) is determined by the size of the registers used in the computer design. In a synchronous computer, all words will be the same length, and each bit in a word will move in coincidence with a TP or clock pulse; e.g., the LSD with TP-1, the next pulse with TP-2, etc. With the word time fixed by computer design, the counter or timer must be designed so that the time required to progress through its count sequence corresponds in time to the word time. Also a method must be obtained to make the end of the count sequence in coincidence with the end of the word time. For example, assume that the word is eight output pulses will be required. A three-stage counter will satisfy this requirement. By the addition of a fourth stage, the end of the word time can be marked by changing the state of the fourth stage each time the counter goes through a count of eight. Two-word times are identified by this method when the fourth stage is reset and the counter progresses through its count, and a second when the fourth stage is set and the counter progresses through its count. The example and associated waveforms are shown in Figure 3-556. In the previous example, two word time levels were generated by the addition of a fourth stage to the counter. The word time levels may be used to gate words through a computer, e.g., transfer from one register to another during word time one (WT1). Add during word time two (WT2), etc. When more than two word time levels are required, more stages are added to the counter. A fifth stage will generate four word times, a sixth stage, eight word times, etc. In the previous example, the number of TPs generated by the counter was the same as the number of bits in the word. When the count sequence is different from the number of bits in a word, some method must be developed to alter the count. This can be accomplished in any of several ways, e.g., suppose the word contains 12 bits, then the counter should have a count of 12. A three-stage counter, however, has a count of eight, and one of four stages has a count of 16. A solution then, is to use four stages in the counter and modify the count as shown in Figure 3-557. In this illustration, each time the counter progresses through its count to the configuration $AB\bar{C}D$, delayed clock ANDed with this configuration in AND gate #1 will produce an output that will toggle the word time stage (E) to a set condition, initiating WT2, and through a delay, will reset the counter. After 12 more clock pulses, the counter again has the configuration ABCD. When this ANDed with delayed clock, the (E) stage resets, the counter resets after the delay and WT1 is started again. Another method of word time generation for a 12 bit word is shown in Figure 3-558. In this illustration, the counter is preset to $\bar{A}\ \bar{B}\ C\ \bar{D}$ and the word time stage (E) reset. As clock pulses are applied, the counter progresses through its count sequence until the configuration in the counter is A B C D. The next delayed clock pulses sets the (E) stage, starting WT2. The same delayed clock pulse sets the (E) stage, starting WT2. The same delayed clock pulse is applied to AND gate #1. The output from this gate presets the counter to $\bar{A}\ \bar{B}\ C\ \bar{D}$ and the count is repeated, starting from the preset condition and progressing to the configuration A B C D with the (E) stage alternately reset (WT1) and set (WT2). UP and DOWN counters may be designed to count any number

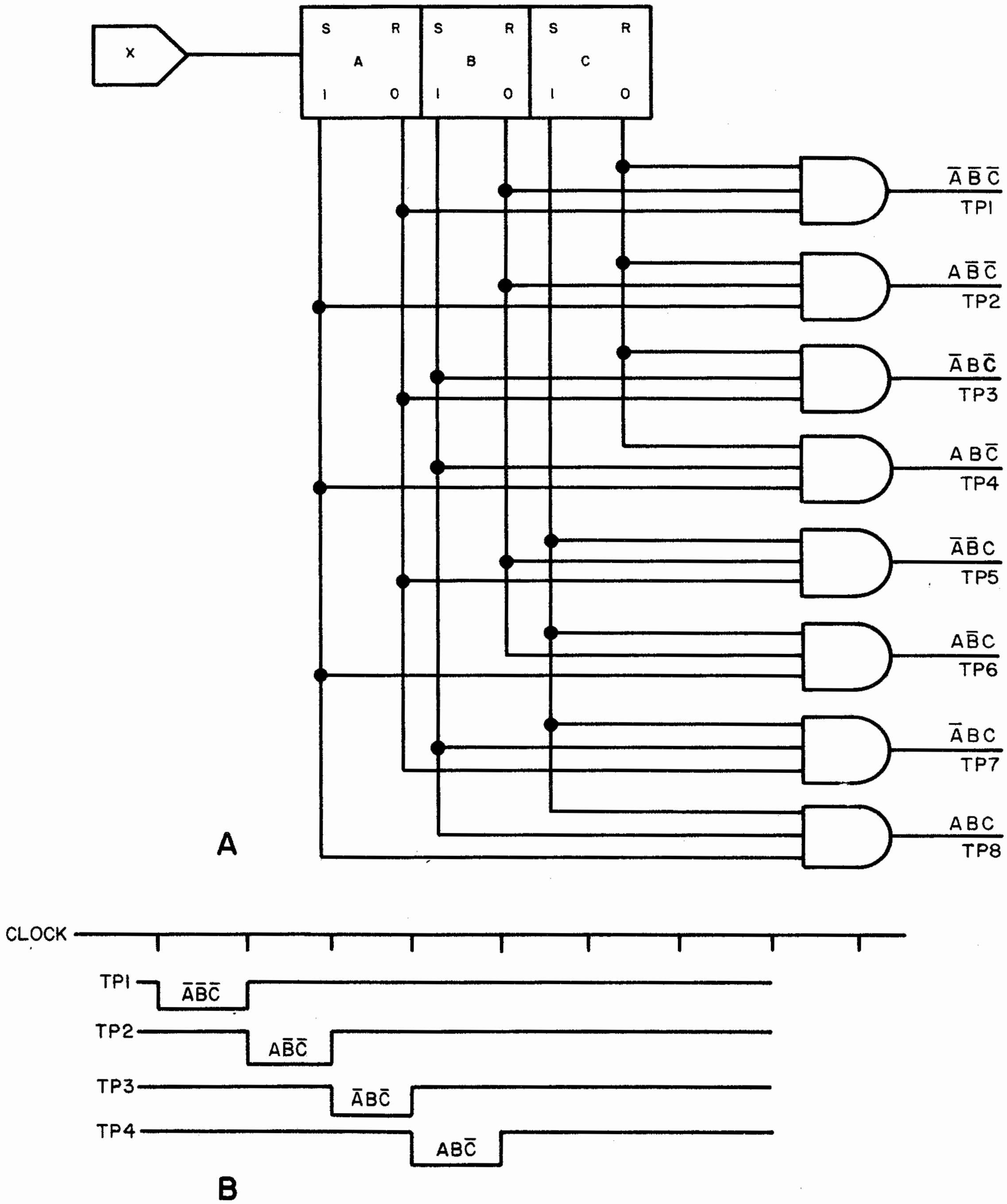


Figure 3-554. Word Time Values and Waveforms

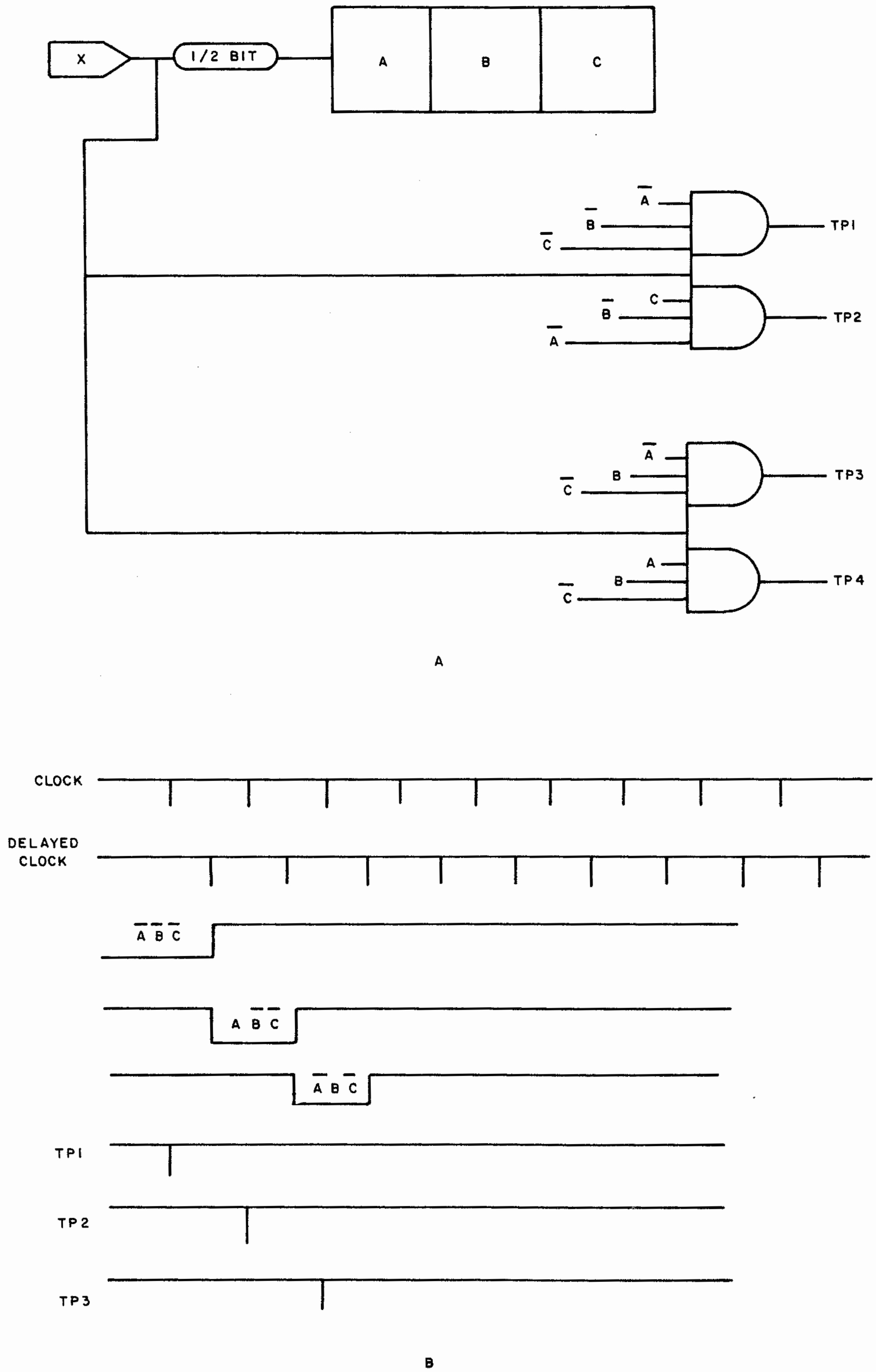


Figure 3-555. Three-Stage Counter and Timing Diagram

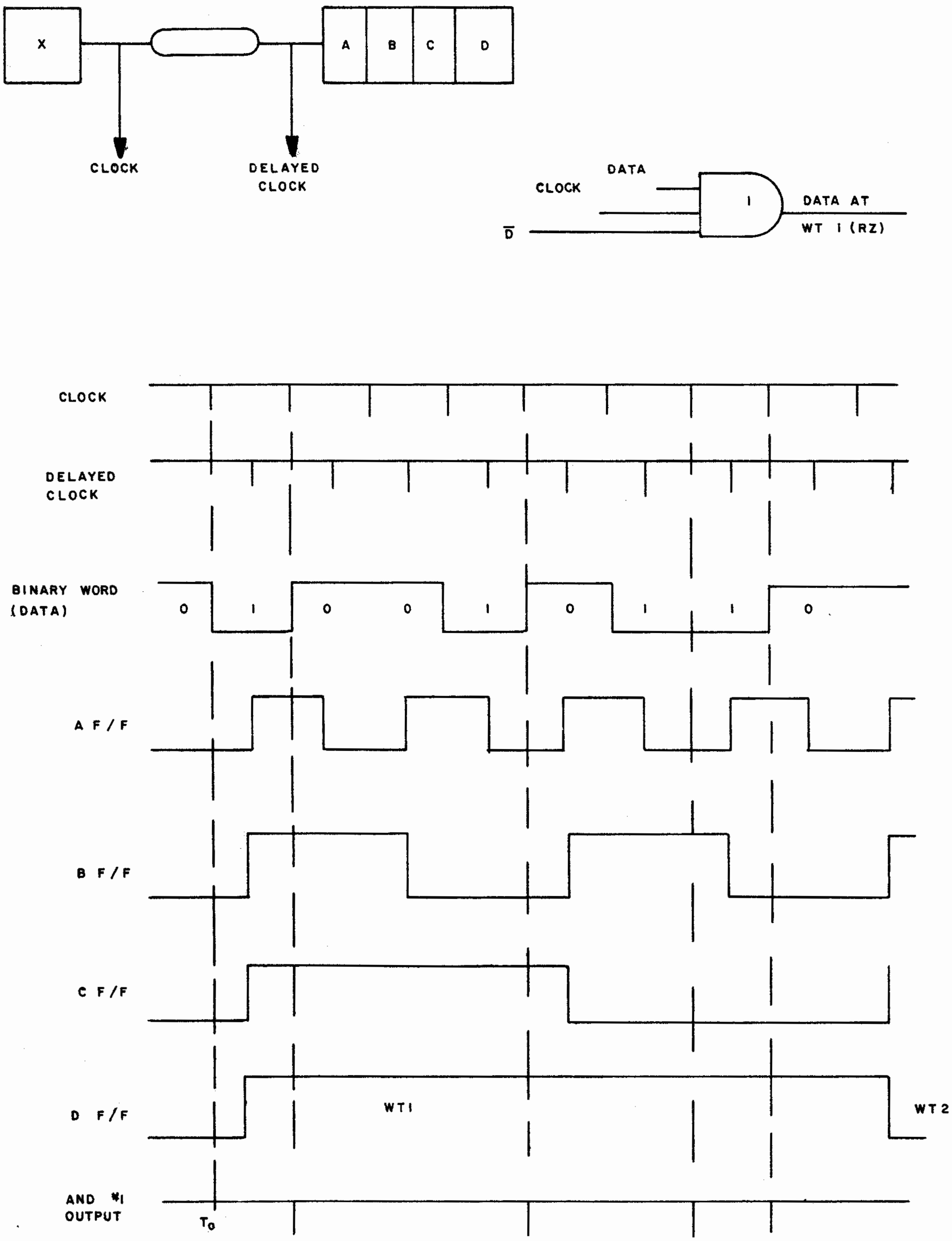
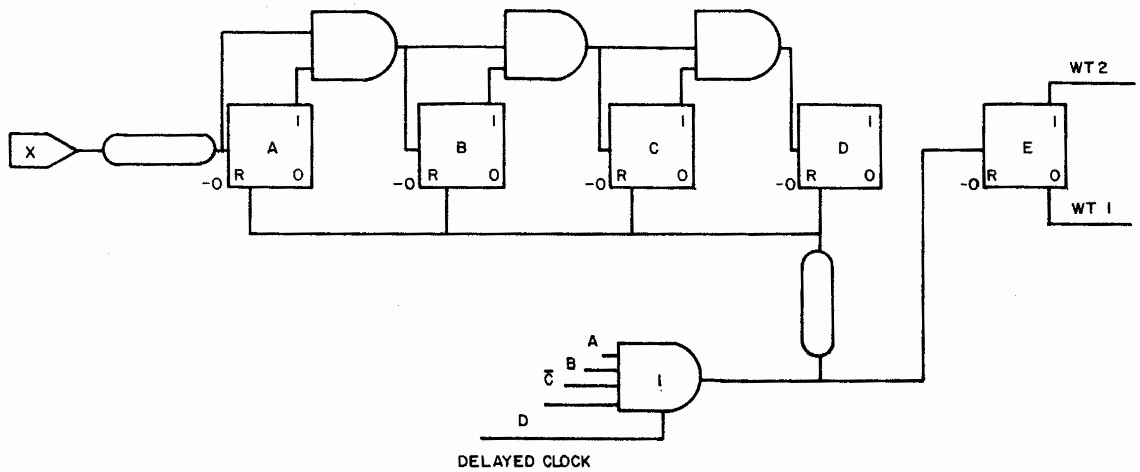


Figure 3-556. Four-Stage Counter and Timing Diagram



A	B	C	D	E
0	0	0	0	0
1	0	0	0	0
0	1	0	0	0
1	1	0	0	0
0	0	1	0	0
1	0	1	0	0
0	1	1	0	0
1	1	1	0	0
0	0	0	1	0
1	0	0	1	0
0	1	0	1	0
1	1	0	1	0

WT 1

A	B	C	D	E
0	0	0	0	1
1	0	0	0	1
0	1	0	0	1
1	1	0	0	1
0	0	1	0	1
1	0	1	0	1
0	1	1	0	1
1	1	1	0	1
0	0	0	1	1
1	0	0	1	1
0	1	0	1	1
1	1	0	1	1

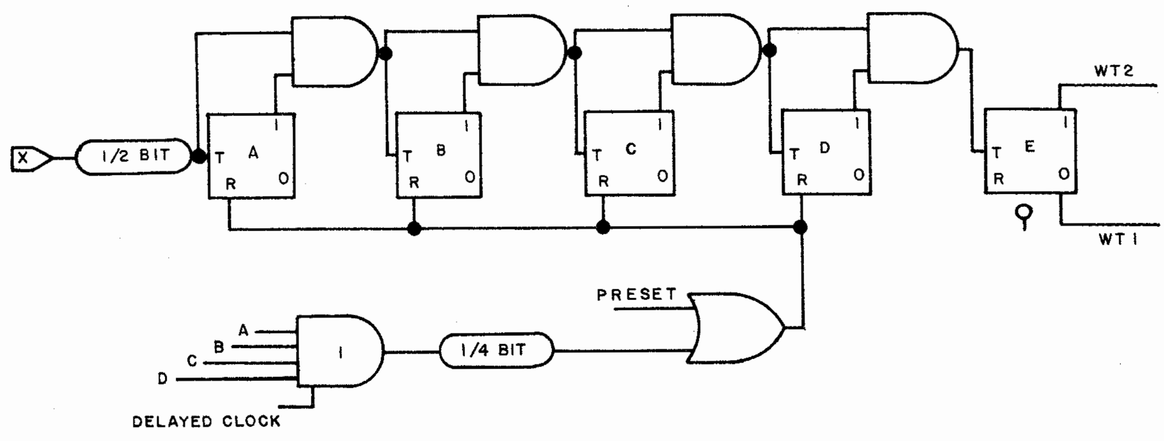
WT 2

Figure 3-557. Four-Stage Counter with Overflow

of input pulses during a count sequence. The maximum number of configurations in a counter's sequence is two raised to the power corresponding to the number of stages used. The count can be identified at any time during the sequence by sampling the configuration in the counter. The length of the count can be altered by presetting the counter to a starting configuration, or by resetting the counter with a configuration in the sequence.

3-18.7 SHIFT REGISTERS

For many of the operations performed by a digital computer, it is necessary to temporarily store or hold information within the working section of the computer. The circuits in the computer used for this purpose are called registers. In addition to their use as temporary storage devices, registers can be used to change the value of a number held in the register by a power of two. Registers may be used as a buffer between memory and the rest of the computer. They may be used to change the rate of data flow in a computer by changing from serial data flow to parallel and vice-versa. Since all information in a binary computer is represented by binary numbers, bi-stable devices can be used to represent individual digits. Each bi-stable device is capable of representing only one binary digit, therefore a register must have one bi-stable device for each digit in the word; e.g., a register used to temporarily store a 16 bit word would require 16 bi-stable cells, etc. Registers used for temporary storage are broken into two categories, parallel registers and (serial) shift registers. Shift registers may be classified as "shift to zero" and "forced transfer." Shift registers are further identified by the direction of shift, shift right, shift left or both.



A	B	C	D	E	
0	0	1	0	0	WT 1
1	0	1	0	0	
0	1	1	0	0	
1	1	1	0	0	
0	0	0	1	0	
1	0	0	1	0	
0	1	0	1	0	
1	1	0	1	0	
0	0	1	1	0	
1	0	1	1	0	
0	1	1	1	0	
1	1	1	1	0	

A	B	C	D	E	
0	0	1	0	1	WT 2
1	0	1	0	1	
0	1	1	0	1	
1	1	1	0	1	
0	0	0	1	1	
1	0	0	1	1	
0	1	0	1	1	
1	1	0	1	1	
0	0	1	1	1	
1	0	1	1	1	
0	1	1	1	1	
1	1	1	1	1	

A	B	C	D	E	
0	0	1	0	0	WT 1
etc.					

Figure 3-558. Word-Time Generator for a 12 Bit Word

A forced transfer shift right shift register will have information applied serially to the MSD and will move from left to right one bit at a time.

3-18.7.1 "Shift to Zero" Shift Registers

Figure 3-559 illustrates a shift right "shift to zeros" shift register. To satisfy the conditions for a serial shift register, the binary word must move from left to right (towards the LSD) one bit at a time. Referring to the illustration, it is desired to shift the data word 11011(2) into the register. Assume the register intially cleared (all zeros). Every stage is reset with the application of the first clock pulse. Since they were originally set, the inter-stage delays hold zeros, inhibiting the AND gates until after delayed clock has gone. During this time the first (LSD) data bit from the input delay sets the "D" stage. The second clock pulse resets all stages to zero. Now the delay on the function side of the "D" stage holds a "one" as a result of the "D" stage having been set. When delayed clock is applied, AND "1" will have an output setting the "C" stage. The "D" stage is also set at this time by the application of the next data bit. The third clock pulse is applied to the shift line and resets all stages. The delays on the function side of the "C" and "D" stages hold "ones", and when delayed clock is applied the "B" and "C" stages will be set. The "D" stage is not set at this time since the data input bit is a zero. The fourth clock pulse resets all stages to zero. The delays at the output (function side) of "B" and "C" stages hold "ones", setting "A" and "B" stages when delayed clock arrives. The "D" stage is set by the input data bit. The register now holds 1101(2). Note that each shift pulse moves the data toward the LSD one stage (bit), therefore the four stage register required four shift pulses to

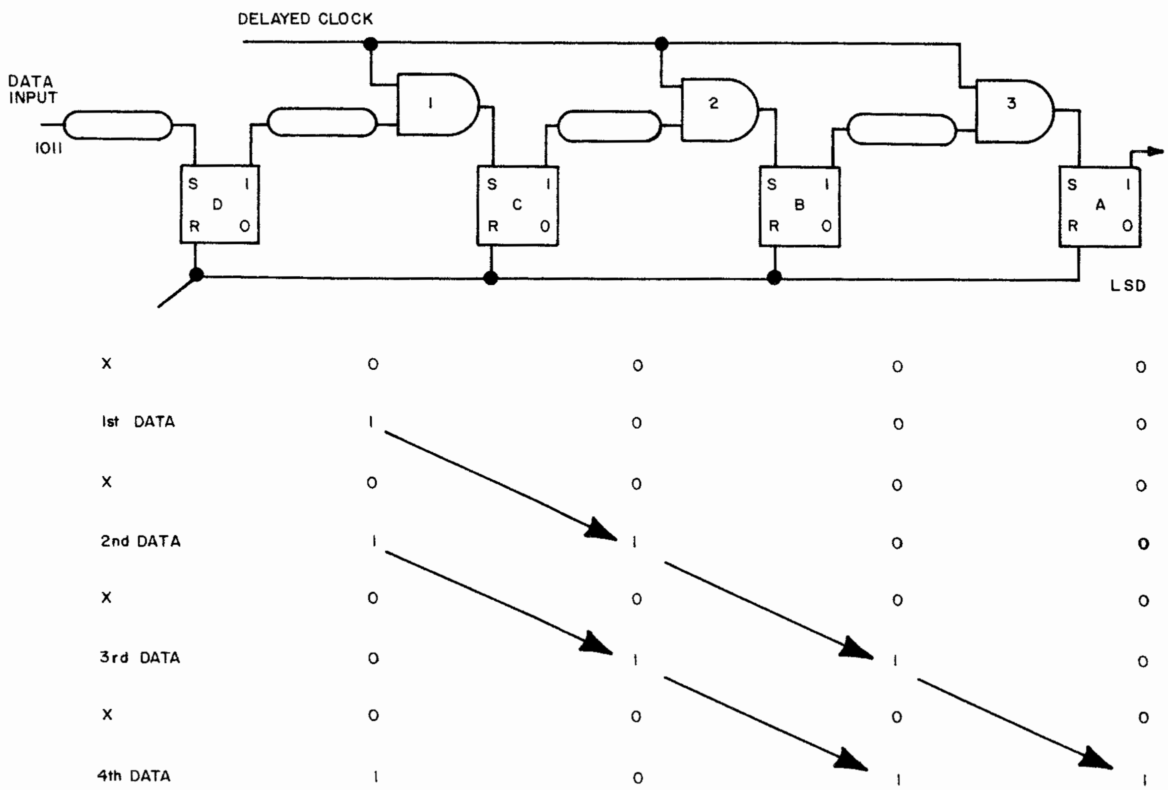


Figure 3-559. Shift-Right Shift Register

shift a word in. Sixteen stages would require 16 pulses, etc. The inter-stage delays must have a delay slightly longer in duration than delayed clock. This is required to ensure that the AND gate is enabled when delayed clock arrives.

3-18.7.2 The Force Transfer Shift Register

The "force transfer" register is a high speed transfer register. Unlike the "shift to zeros" register, all stages of the "force transfer" register are not reset by the shift pulses. However, more components are used in building this type of register. This shift register operates because no flip-flop can change its state in zero time. A finite time is required, therefore shift pulses will be allowed to pass through enabled AND gates before the flip-flop furnishing the enabling level can change its state. The inputs to the "force transfer" shift register (shown in Figure 3-560) pass through OR gates 1 and 2. This may be a new word, when AND gates 11 and 12 are enabled, or a recirculated word, when AND gates 9 and 10 are enabled. With the initial conditions as shown in Figure 3-560, WT 4 will recirculate the data from the LSD to the MSD. At the same time, data are shifted from stage to stage in the register. When the clock at TP 1 time of WT 4 probes all the AND gates, it finds gates 1, 3, 5, 8 and 9 enabled. Since it is already set, stage E will not be affected by the output pulse from AND gate 9 which passes through OR gate 1. However, stage D will be set by the output of AND 1, while stage C will be reset by the output of AND gate 4.

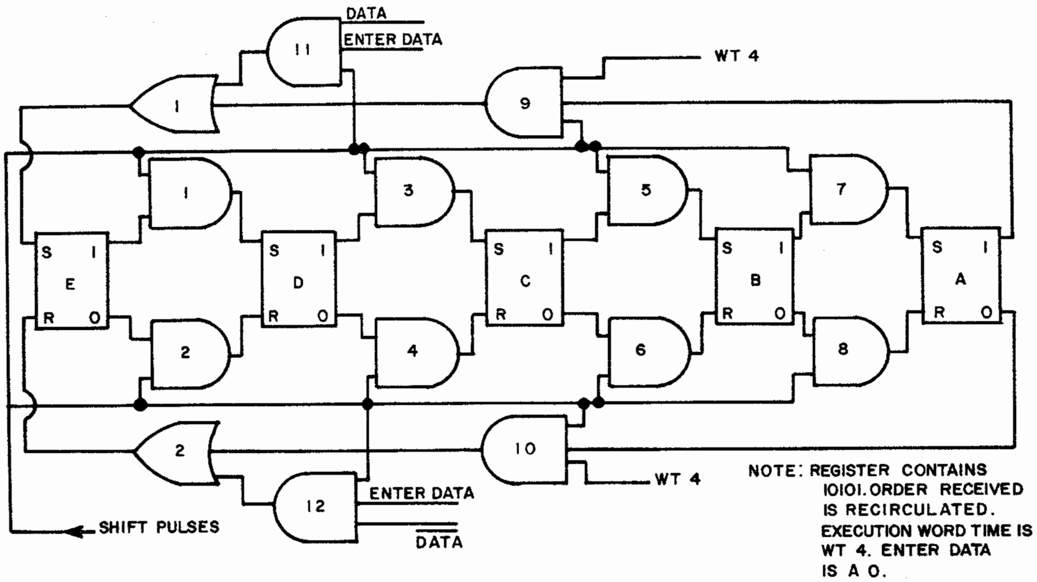


Figure 3-560. Force Transfer Shift Register

Stage B will be set by AND gate 5's output and stage A will be reset by the output from AND gate 8. The register now contains 11010 and has been circular shifted one place. The clock pulse at TP 2 is applied to the shift pulse input and finds AND gates 1, 3, 6, 7, and 10 enabled. This will cause stages D, C, and A to be set while stages E and B will be reset. The clock pulse at TP 3 will enable AND gates 2, 3, 5, 8 and 9. This will cause stages E, C, and B to be set while stages D and A are reset. The register now contains 10110 and has been circular shifted three places. At TP 4 time the clock pulse will qualify AND gates 1, 4, 5, 7 and 10. The outputs from these gates will set stages D, B, and A, and reset stages E and C. The register now contains 01011 and has been circular shifted four places. The clock pulse applied at TP 5 time will qualify AND gates 2, 3, 6, 7, and 9. The outputs from these gates will set stages E, C, and A. Stages D and B will be reset. The number in the register is now 10101, our original number. Although not shown on this diagram the number could be shifted to another circuit during this operation, as was done in Figure 3-558. The operation of this circuit would be the same for shifting in new numbers, with one exception: to shift a new number into the register AND gates 9 and 10 would be disqualified while AND gates 11 and 12 would be enabled.

3-18.7.3 Magnetic Core Shift Registers

A shift register may use magnetic cores instead of flip-flops to store individual bits. The cores used have a rectangular hysteresis loop. Once the core is magnetized, it will remain in the same condition until an mmf large enough to overcome this condition is applied. Each core in the register has three windings, as shown in Figure 3-561. When a pulse is applied to the WRITE winding, current will flow in the direction shown, and the core will be magnetized in the counter-clockwise direction. This condition represents a binary "1" (for this particular circuit). When a pulse is

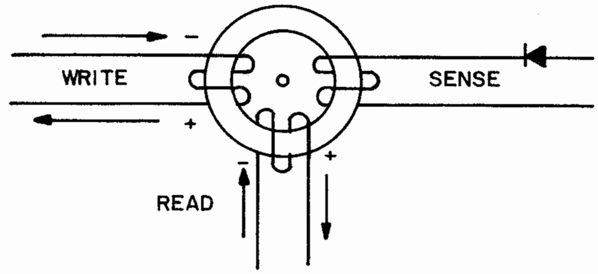


Figure 3-561. Magnetic Core Method

applied to the READ winding, current will flow in the direction of the arrows, and the core will be magnetized in the clockwise direction. This condition represents a binary "0" (for this particular circuit). The WRITE winding always switches the core to the "1" stage when energized. The READ winding always switches it to the "0" state when energized. The SENSE winding is used to sense a change of flux in the core. Any change in flux would induce a voltage into the winding, but the diode is placed so that current will flow only when the core is switched from the "1" state to the "0" state by the READ winding. The shift register shown in Figure 3-562 is a four stage register. The operation of the register is somewhat similar to that of the "shift to zeros" shift register, discussed earlier, in that the shift pulses will switch each stage to the "0" state. With the WT levels shown in Figure 3-562, a new word will be shifted into the register during WT 1 from AND gate 1. The word will be shifted out AND gate 4 and recirculated during WT2. As an example of the above, assume the conditions stated in Figure 3-562. At TP 1 the clock pulse will pass through AND gate 2 to the READ windings of all cores. The cores are in the "0" state so no switching will occur and no voltage will be induced into the sense windings. The data input at TP 1 is a "1", so the clock pulse will pass through AND gate 1 through OR gate 1 to trigger delay #1. Two microseconds later the output of delay #1 applied to the WRITE winding of stage D will switch that core to the "1" state. The register now contains 1000. At TP 2 time, the clock pulse applied to AND gates 1 and 2 will pass through AND gate 2 only, because the data input is a "0" state. The changing flux in D will induce voltage in the sense winding which will trigger delay #2. After 2 microseconds, the output from delay #2 will be applied to the write winding of core C switching it to the "1" state. The number in the register is now 0100. At TP 3 time, the clock applied to AND gates 1 and 2 will qualify these gates. The data input at this time is a "1." The pulse from AND gate 1 will pass through OR gate 1 and trigger delay #1. The pulse from AND gate 2 will switch core C to the "0" stage. The resultant change in flux in core C will induce a voltage in its sense winding. This voltage will trigger delay #3. After 2 microseconds, the outputs of delays #1 and #3 will be applied to the WRITE windings of cores D and B switching them to the "1" state. The register now contains 1010. The data input to AND gate 1 is a "0" at TP 4 time so the clock pulse to AND gates 1 and 2 at this time will pass through AND gate 2 only. This pulse will switch cores D and B to the "0" state, and the switching action will induce a voltage in the sense windings of those two cores. Delays #2 and #4 will be triggered, and after the two microsecond delay, the outputs from these delays will be applied to the WRITE windings of cores C and A switching them to the "1" state. At the end of the WT, the number in the register is 0101. During WT 2, AND gates 3 and 4 will be enabled by the "1" level at their WT 2 inputs. The shifting operation will be the same as before, except when a "1" is shifted from core A its sense winding's output will qualify AND gate 4 as the output, and the recirculate gate, AND gate 3, to trigger the input delay #1. The magnetic core memory has several advantages over the flip-flop

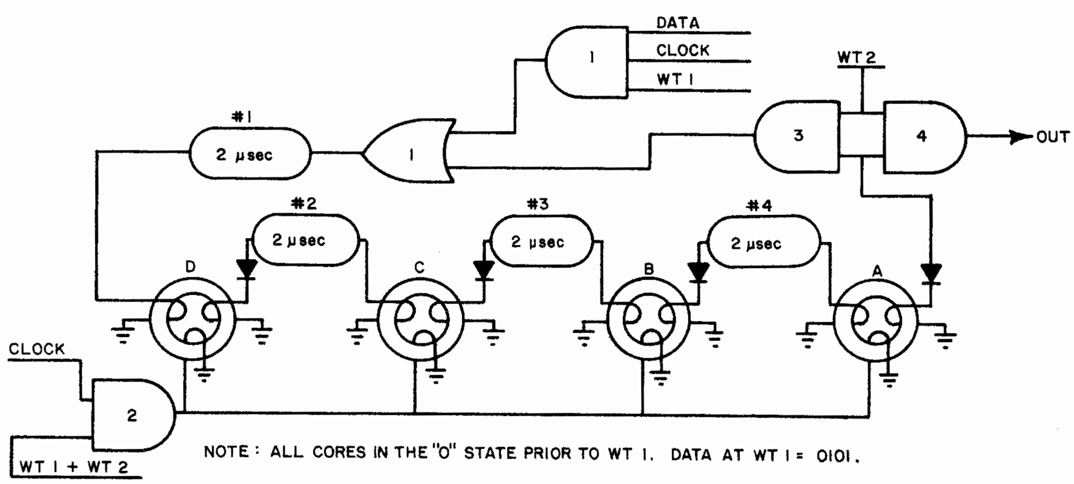


Figure 3-562. Four-Stage Register

register. The cores do not require power when in a static condition, and even if power is removed from the circuit entirely, the information will be retained by the cores. The magnetic core register is small in size and uses relatively simple circuitry. The major disadvantage of the "magnetic core" shift register is that the data contained by it cannot be sampled unless it is shifted out. Again, it must be realized that the WT levels shown here are used only as examples, and the actual levels used will be selected to suit the individual computer.

3-18.7.4 The Parallel Binary Register

The registers discussed to this point have all been registers which are capable of receiving data only if it is sent to the register serially. The register in Figure 3-563 is capable of receiving parallel data only and the outputs from this register will be parallel. TP 1 is applied to the reset input of all stages to clear the register. The number in the register after TP 1 is 00000. TP 2 is applied to AND gates 01, 02, 03, 04, and 05. The data inputs to AND gates 01, 03, and 04 are "1s" so TP 2 will qualify these gates to set stages A, C, and D. The data inputs to AND gate 02 and 05 are "0s", therefore, there will be no outputs from these gates and stages B and E will remain reset. After TP 2 the number in the register is 01101. The entire word was brought into the register in one TP time. This illustrates the advantage of parallel operation speed. The word is held by the register until TP 4. TP 4 is applied to AND gates 11, 12, 13, 14, and 15. Stages A, C, and D are set so AND gate 11, 13, and 14 will be qualified, and a "1" will be sent out the LSD, 3rd bit, and 4th bit lines. Stages B and E are reset and the "0" level applied to AND gates 12 and 14 will prevent these gates from having an output at this time. The 2nd and 5th bit outputs will be "0"s at TP4. In this example, no stages were reset or set during TP 4, and the number 01101 is still held by the register; therefore, no recirculation is necessary. Since no shifting occurs, this register is not called a "shift" register.

3-18.7.5 Military Standard Symbols

The military standard symbol for a serial shift register with a single data input, such as the "shift to zeros" register shown in Figure 3-564, would be as shown below. The number in parentheses (5) is the number of stages in the register. The symbol for a register with two data inputs, such as the "forced transfer" register shown in Figure 3-560, is shown in Figure 3-565. Note that in each of the two figures shown, the shift pulse line is at the lower left hand corner and is labeled "shift right input". Shift inputs may be shown in the upper corner. Each of these registers shifted the numbers to the right (toward the LSD). Many shift registers are built to shift numbers in either direction; either to the right or to the left (toward the MSD). The symbol for this type of register is shown in Figure 3-566. The symbol for the parallel binary register shown in Figure 3-561 is illustrated in Figure 3-567 below. Many registers must be capable of serial and parallel operation, and of shifting either

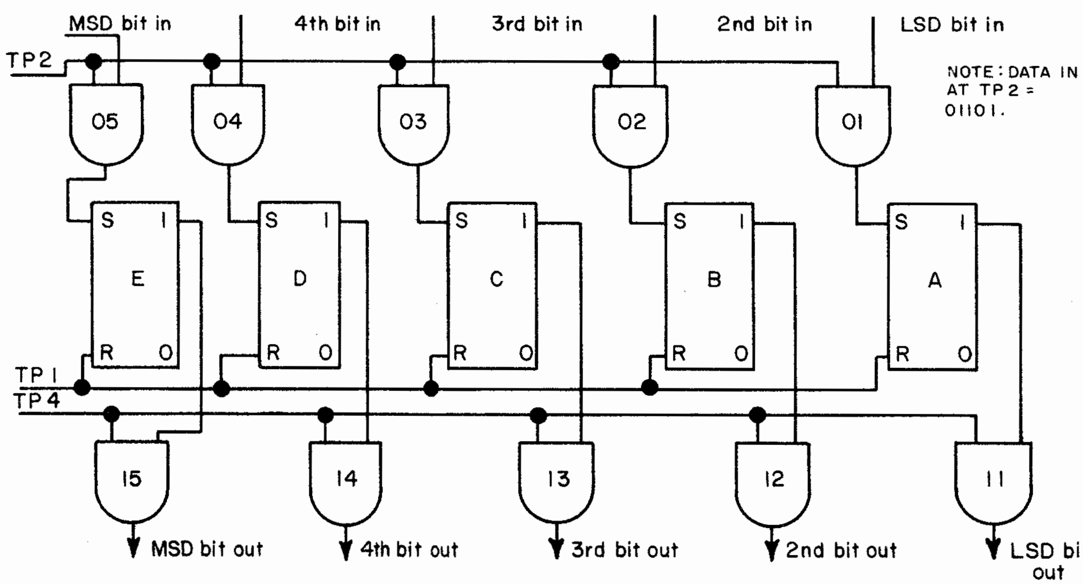


Figure 3-563. Parallel Register

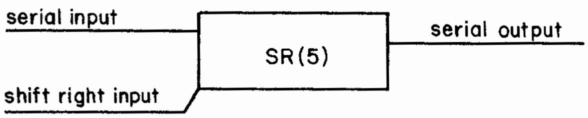


Figure 3-564. Shift to Zeros Register

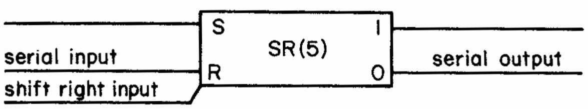


Figure 3-565. Forced Transfer Register

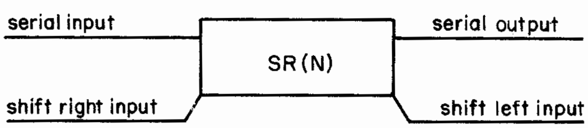


Figure 3-566. Shift Register

left or right. The symbol for such a register is shown in Figure 3-568. There are many different types of registers, using various combinations of inputs and outputs, each with a slightly different symbol. The symbols shown here are basic, and other types will have only slight variations.

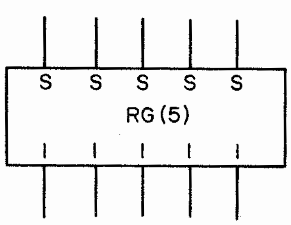
3-18.7.6 Multiplication and Division, Using Shift Registers

By applying a controlled number of shift pulses to a shift register, the numbers stored in them can be manipulated to perform mathematical operations. The number in the register in Figure 3-569 is equal to $24_{(10)}$. If one pulse is applied to the shift-right input, the entire number is moved one place toward the LSD and will therefore appear as in Figure 3-570. This number is equal to $12_{(10)}$ or one-half of the original number. A second pulse would cause the number to be as shown in Figure 3-571 equal to $6_{(10)}$. Each time any number is shifted one place to the right, that number is halved. A computer can therefore divide any power of 2 (2, 4, 8, 16, 32) by shifting right the correct number of places. A number in a register may be multiplied by shifting left. The register in Figure 3-572 holds the binary equivalent of $13_{(10)}$. One pulse applied to the shift-left input will cause the entire number to move one place to the left or toward the MSD. The binary number in the register in Figure 3-573 is now equal to $26_{(10)}$ (it has been doubled). Further shift-left pulses will double the number each time it is shifted. A computer can thus multiply any number by any power of two by shifting left the correct number of places. In shift registers which are designed to shift right only, a shift-left operation can be performed by using two registers as shown in Figure 3-574, (A) and (B). For example, if it is desired to shift the number in register P (A) to the left two places, the entire number is first shifted out, to the right, and then into register Q (B). During the next word time, six pulses are applied to the shift right inputs of the P and Q registers shifting the six LSD bits from Q into the six MSD bits of P as shown in Figure 3-575. The effect of the entire operation has been to displace the number in the P register to the left two places. There are many other types of shift registers not described in this section, but all function in much the same manner and are used for the same operations, as the registers described here.

3-18.8 ADDERS

3-18.8.1 Fundamentals of Addition.

Additions in any numbering system require that very definite rules be followed to arrive at the proper sums. The set of rules for digit by digit



NOTE: The letters RG are used in place of SR to indicate that this is a binary register and not a binary shift register.

Figure 3-567. Parallel Register

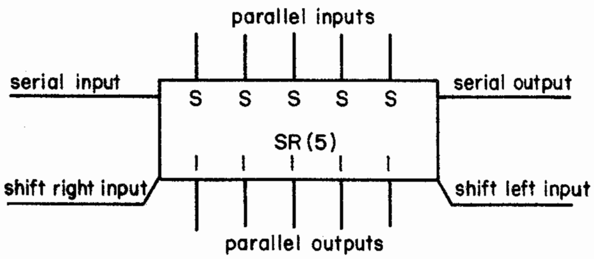


Figure 3-568. Series Parallel Register

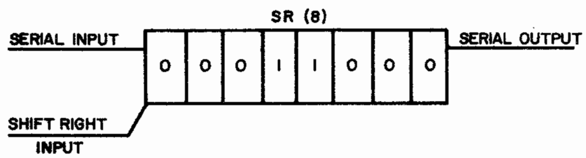


Figure 3-569. Shift Register

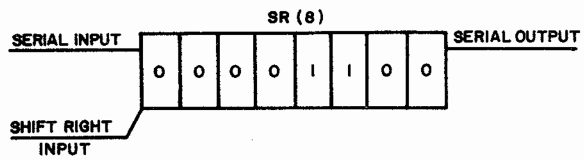


Figure 3-570. Shift Register

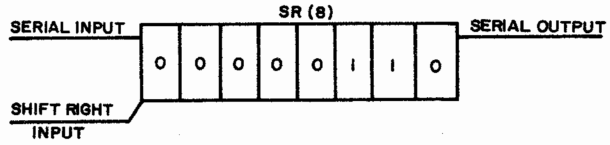


Figure 3-571. Shift Register

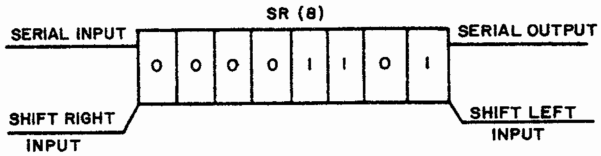


Figure 3-572. Shift Register

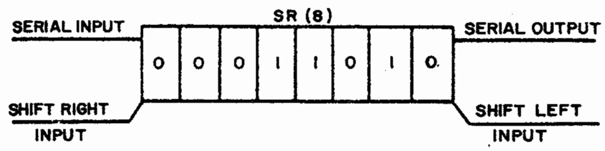


Figure 3-573. Shift Register

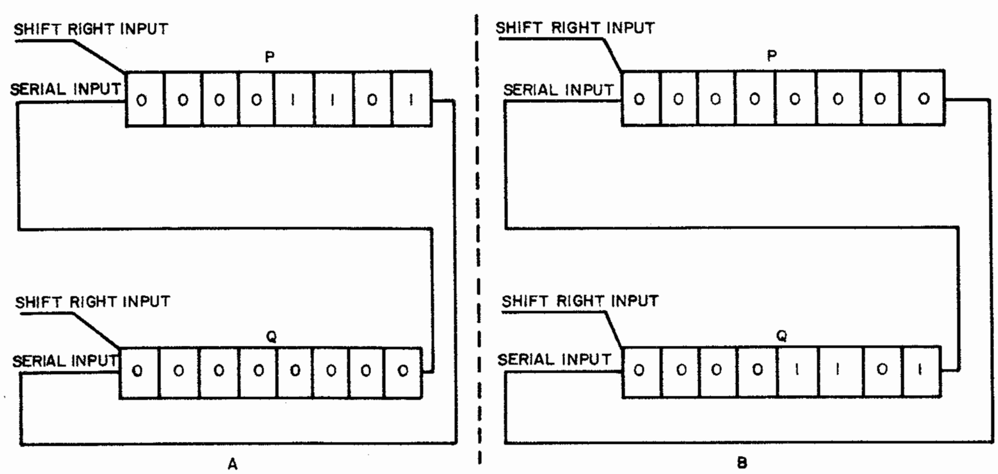


Figure 3-574. Right-Shift Register

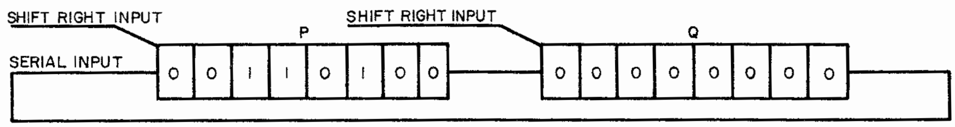


Figure 3-575. Shift Register

additions may be stated in the following form:

a. Generate an intermediate sum as necessary
b. Generate an intermediate carry as necessary
c. Consider previous carry
d. Generate a final sum and carry as necessary.

When the above rules are properly applied, digit by digit additions are accomplished with ease in any numbering system. Digit by digit additions result in the sum of any numerical values by beginning with the LSDs (least significant digits) and progressing to the MSDs (most significant digits). Since the decimal or Arabic numbering system is the one most familiar to the average person, the rules will be applied to an example in this system (Refer to Figure 3-575). Adding 1854 and 748 in the decimal numbering system illustrates the above four rules. One of the values is designated in the X number, while the other is designated as the Y number. In this example, the letter "K" is used to show previously generated carries when they are considered in the addition. No previous carry is present when the LSDs are added in the exampled problem. Rule (b) is used to generate carries when the first and third digits are added, while rule (d) is used to generate a carry in the second digit addition. The method of addition illustrated is the method commonly employed by serial digital computers. The foregoing example has been presented to emphasize the fundamental relationships in addtions. Carries are generated in one digit addition and are considered in the next more significant digit addition.

3-18.8.2 Additions Using Bi-Stable Devices

In the following presentation, the rules for addition will be used to develop adders for binary digital computers. Serial binary adders will be developed first. These adders begin additions with the LSDs and progress to the MSDs in approximately the same manner as human beings add two numbers. Shift

	X	=	1	8	5	4
	Y	=		7	4	8
INTERMEDIATE SUM		=	1	5	9	2
	K	=	1	1	1	0
FINAL SUM		=	2	6	0	2

Figure 3-576. Addition

registers were developed and explained in an earlier chapter. Their lowest order stages may be used as the X and Y inputs to serial adders. When $\overline{X}$, X, Y or $\overline{Y}$ are shown in logic diagrams and truth tables, these functions are originating at the LSDs of two shift registers. As the additions progress, the contents of the two registers are shifted toward the LSDs. Two binary numbers $11100_{(2)}$ and $111010_{(2)}$ are added in the following example (refer to Figure 3-577).

3-18.8.2.1 Serial Quarter-Adders

Rule (a) may be used to develop a truth table to perform approximately one-fourth of the full job of addition of binary numbers. A quarter-adder truth table is shown in Figure 3-578. Assume that it is desired to design a circuit to perform the functions indicated by this truth table. This is done by inspecting the "S" (Sum) column and designing the circuit to have a numeral one on its output when a one appears

	X	=	0	0	1	1	1	0	0
	Y	=	0	1	1	1	0	1	0
INTERMEDIATE SUM		=	0	1	0	0	1	1	0
	K	=	1	1	1	0	0	0	0
FINAL SUM		=	1	0	1	0	1	1	0

Figure 3-577. Binary Addition

X	Y	S
0	0	0
1	0	1
0	1	1
1	1	0

Figure 3-578. Quarter-Adder Truth Table

in this column. The Boolean expression for generating these ones is $X\overline{Y} + \overline{X}Y$. When two bits are different in binary addition, the intermediate sum is a one. "Quarter-adder" is the name given to the circuit shown in Figure 3-579. Because it performs approximately one-fourth of an addition. Since X does not equal Y when this circuit produces an output, the quarter-adder is sometimes used to compare two binary numbers for equality. If used as a comparator, there is no output from this circuit when X equals Y.

3-18.8.2.2 Serial Half-Adders

Expansion of the truth table to include a column for generation of carries is illustrated in Figure 3-580. The Boolean expression for the sum is the same as before; but this truth table also shows the function necessary to satisfy the (b) rule of addition (see paragraph 3-18.8.1). When the circuit to perform these different functions is designed, a separate AND gate is used for generating the carries. Inputs to this AND gate are taken directly from the truth table. These inputs are X and Y, as shown in the figure. Since this circuit performs approximately one-half of a complete binary addition, it is called a "half-adder."

3-18.8.2.3 Serial Adders

The circuits already described have no provision for remembering carries when generated or when considering them later. In serial binary additions, carries must be stored when generated and then considered when the next bits are added. A flip-flop lends itself to this function and performs the job well. The truth table for a full binary addition is shown in Figure 3-581, with the full-adder circuit. This circuit was designed by extracting the necessary Boolean expressions from the truth table and then designing the circuits to satisfy them. Any time both X and Y are ones, the "K" flip-flop should be set and the $\overline{K}$ may be dropped from the set "K" expression. Finally, the set "K" control circuit is one AND gate with X and Y as its inputs. Clock will be explained in detail later.

X	Y	S
0	0	0
1	0	1
0	1	1
1	1	0

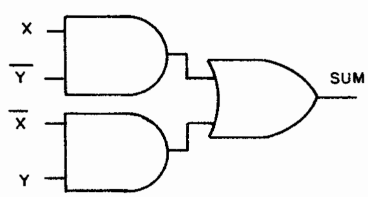


Figure 3-579. Quarter Adder and Truth Table

X	Y	S	C
0	0	0	0
1	0	1	0
0	1	1	0
1	1	0	1

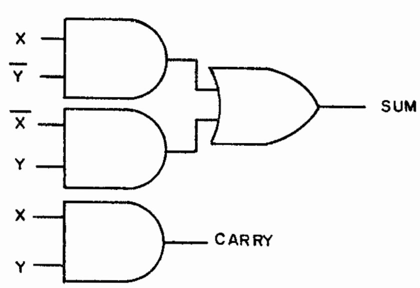


Figure 3-580. Half Adder

X	Y	K	S	C
0	0	0	0	0
1	0	0	1	0
0	1	0	1	0
1	1	0	0	1
0	0	1	1	0
1	0	1	0	1
0	1	1	0	1
1	1	1	1	1

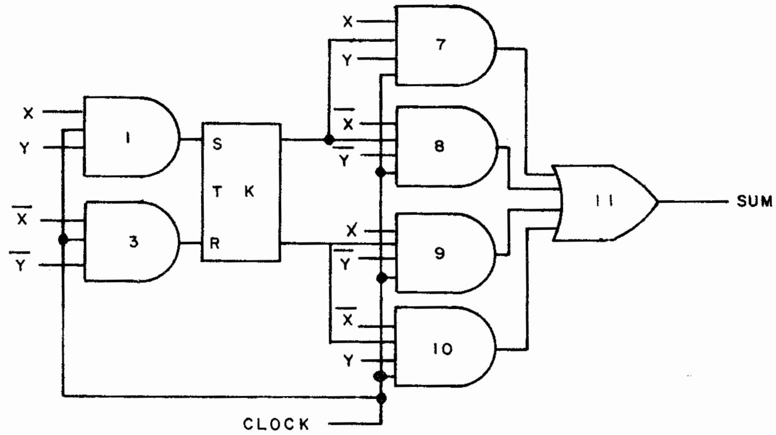


Figure 3-581. Full Adder

Provision to reset the "K" flip-flop must be made for conditions when no carry is desired. From the "C" column, the four zeros indicate the configurations for which no carry is desired. Extracting the Boolean expression for resetting the "K" flip-flop results in $\overline{XYK}$ + $X\overline{YK}$ + $\overline{X}Y\overline{K}$ + $\overline{XY}K$. Three of the terms in this expression require that "K" be reset to satisfy them and these terms may be eliminated. Since it is desired to reset the "K" flip-flop every time $\overline{X}$ and $\overline{Y}$ are present, K may be eliminated from the remaining term. The circuit to reset the "K" flip-flop becomes one AND gate, with $\overline{X}$ and $\overline{Y}$ as inputs. Clock pulses are used in the gates to control timing relationships between the sum output and the generation of carries. Due to inherent delay in the "K" flip-flop, sums may be generated at the same time an attempt is made to change the state of this flip-flop. The clock pulse will have disappeared before the "K" flip-flop actually changes state. This condition may not hold true in some systems, but other timing arrangements can be made to satisfy the requirements. All of the rules for addition are satisfied with each clock pulse applied to the full-adder. The adder just developed, and other circuits developed to be later, use the "two's complement" method when dealing with negative numbers. Negative numbers are in two's complement form when added. Methods of binary number complementing were discussed earlier. Ones appear in the sum column in four places, indicating that a sum must be generated for these four combinatons of X, Y, and K. Therefore, the Boolean expressions must be $X\overline{YK}$, $\overline{X}Y\overline{K}$, $\overline{XY}K$, and XYK. Since a sum must be generated for any of these conditions, they must be ORed together to become a true miniterm expression. The sum expression then becomes $\overline{XY}K + \overline{X}Y\overline{K} + X\overline{YK}$ + XYK. In the carry generation column headed "C", four ones are indicated, and carries must be remembered when these four conditions exist. The Boolean expression for generation of these ones becomes $XY\overline{K}$ + $X\overline{Y}K$ + $\overline{X}YK$ + XYK. Three of the terms in this expression require that "K" be set to satisfy them. If "K" is already set, there is no point in trying to set it again. This logical reasoning elminates the necessity for building gates to satisfy the terms including "K" and the set "K" input for carry storage becomes $XY\overline{K}$.

3-18.9 SUBTRACTORS

3-18.9.1 Fundamentals of Subtraction

Subtractions may be performed in the binary system with the same conditions as other numbering systems. Subtrahends and minuends must be designated for a correct difference to be arrived at. For the purpose of designing circuits to perform subtractions, the X input is the minuend and the Y input is the subtrahend. Having specified the minuend and subtrahend, the configurations to generate differences and borrows are now fixed. Substitution of the word "difference" in every place the word "sum" appears in the rules for addition, and substituting the word "borrow" for the word "carry" results in the following rules for subtraction:

a. Generate an intermediate difference as necessary

b. Generate an intermediate borrow as necessary

c. Consider previous borrow

d. Generate a final difference and borrow as necessary.

3-18.9.2 Subtractions Using Bi-Stable Devices

Serial subtractors will accomplish subtractions by beginning at the LSDs of two numbers and progressing to the MSDs in the same manner that serial adders add two numbers. Negative numbers will be in two's complements form for the subtractors developed here.

3-18.9.2.1 Quarter-Subtractors

Quarter-Subtractors are shown in Figure 3-582 and in the truth table. They perform the function required by rule (a). This quarter-subtractor is precisely the same circuit as the quarter-adder previously discussed. Ones appear in the difference column for the same configurations where ones appeared in the sum column for the quarter-adder.

3-18.9.2.2 Half-Subtractors

A "B" (borrow) column is shown in Figure 3-583 so that the circuitry necessary to perform the functions required by rules (a) and (b) may be designed. This column has a one in it when $\overline{X}$ and Y are present. Since this same configuration requires that a difference be generated, an AND gate supplies the borrow function while supplying one of the difference functions. The Boolean expression for the difference is $X\overline{Y} + \overline{X}Y$ and the expression for the borrow is $\overline{X}Y$. A half-subtractor circuit is designed as shown in Figure 3-538. Full binary subtractions require that borrows be generated and retained until satisfied. Another requirement is that borrows must be considered, whether present or not, in each bit subtraction. The following example (Figure 3-584) shows the full subtraction of one binary number from another. Borrows are shown as "B" when generated and are shown as "K" in succeeding digit subtractions. Boolean expressions for generating the final difference and borrows may be extracted from the truth table in Figure 3-584. Note that the expression for the difference is identical to the expression for the sum in the full-adder. This expression is $X\overline{Y}\overline{K} + \overline{X}Y\overline{K} + \overline{X}\overline{Y}K + XYK$. Generation of borrows must be accomplished when ones appear in the "B" column of the truth table. However, when the Boolean expression is written $\overline{X}Y\overline{K} + \overline{X}\overline{Y}K + \overline{X}YK + XYK$, the "K" flip-flop must be set to satisfy three of the terms. These three expressions may be eliminated for the purpose of designing circuits to set

X	Y	D
0	0	0
1	0	1
0	1	1
1	1	0

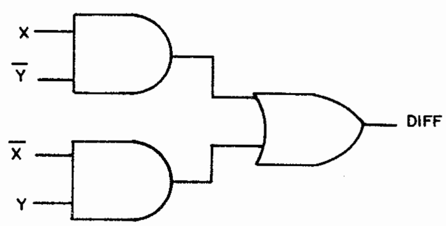


Figure 3-582. Quarter Subtractor

X	Y	D	B
0	0	0	0
1	0	1	0
0	1	1	1
1	1	0	0

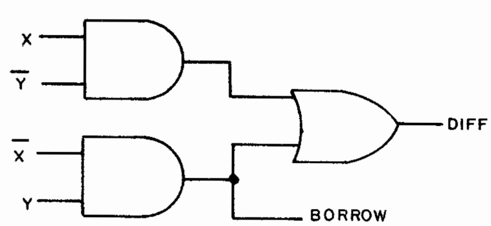


Figure 3-583. Half Subtraction

	X	=	0 1 0 1 0 1 1 0
Intermediate	Y	=	0 0 1 1 1 0 1 0
Difference			0 1 1 0 1 1 0 0
Final	K	=	0 1 1 1 0 0 0 0
Difference			0 0 0 1 1 1 0 0

x	y	k	D	B
0	0	0	0	0
1	0	0	1	0
0	1	0	1	1
1	1	0	0	0
0	0	1	1	1
1	0	1	0	0
0	1	1	0	1
1	1	1	1	1

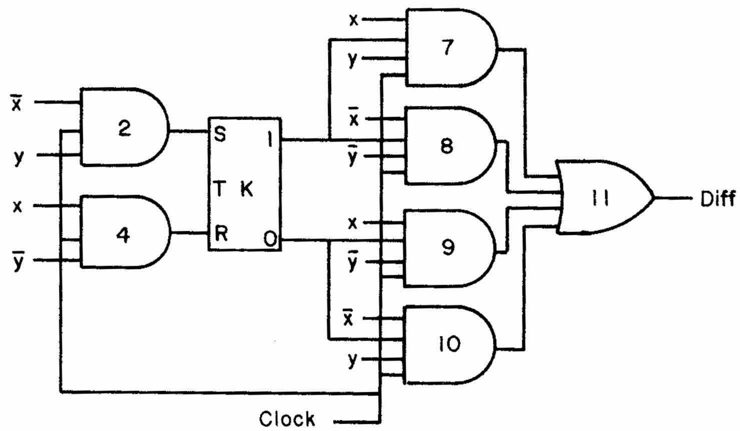


Figure 3-584. Full Subtraction

the "K" flip-flop because it is already set if they are satisfied. Zeros in the "B" column indicate the configuration for which no borrow from the next higher order digit is desired. Extracting the Boolean expression for these conditions results in $\bar{X}\bar{Y}\bar{K}$ + $X\bar{Y}\bar{K}$ + $XY\bar{K}$ + $X\bar{Y}K$. Three of the terms require the "K" flip-flop to be already reset for the terms to be satisfied. These terms may be eliminated from the reset "K" expression. Setting the "K" flip-flop for subtraction may be done by $\bar{X}Y\bar{K}$, and resetting it may be done by $X\bar{Y}K$. The K and $\bar{K}$ factors may thus be eliminated from these expressions. It is always desired to set "K" if $\bar{X}$ and Y are present for subtraction and the $\bar{K}$ input to the set AND circuit is unnecessary. When the configuration X and $\bar{Y}$ is present, it is always desired to reset the "K" flip-flop.

3-18.9.3 Symbolic Quarter-Adder

Exclusive OR circuits are represented symbolically as shown in Figure 3-585 with the quarter-adder. This circuit's function is exactly the same as the function of the quarter-adder in Figure 3-579. Note that the inputs are X and Y. When the inputs to an exclusive OR are different, it will produce a function output so that its output is $X\bar{Y} + \bar{X}Y$.

3-18.9.4 Full Serial Add-Subtract Units

Having developed the full-adder and the full-subtractor, these two units may be combined into one unit to make a serial full add-subtract unit. Control of the gates used for setting and resetting the "K" flip-flop is essential if both additions and subtractions are accomplished with one unit. ADD or SUBTRACT orders may be given to the unit from various sources. A flip-flop is used in Figure 3-586 for simplicity. If the A/S (ADD - SUBTRACT) flip-flop is set, the order ADD will originate from its function output. The order SUBTRACT originates from the A/S flip-flop's not-function output when the flip-flop is reset. Setting and resetting the A/S flip-flop may be done in many ways but a manually controlled pulse generator is used here. Truth tables were developed for full adders and full subtractors. These may be combined into a single table, as shown in Figure 3-586. The sum column in the adder table is identical to the difference column in the subtractor table. A single column may be headed as the S/D (Sum or Difference) column, as shown , to perform the functions. Boolean expressions for additions and subtractions may be extracted from the truth table. There will be one term to set the "K" flip-flop

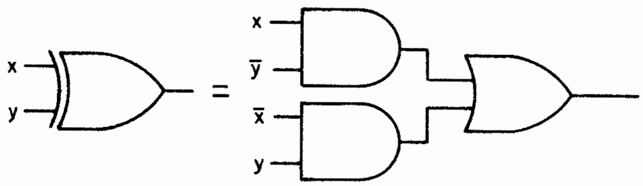


Figure 3-585. Quarter-Adder Using an Exclusive OR Circuit

X	Y	K	S/D	C	B
0	0	0	0	0	0
1	0	0	1	0	0
0	1	0	1	0	1
1	1	0	0	1	0
0	0	1	1	0	1
1	0	1	0	1	0
0	1	1	0	1	1
1	1	1	1	1	1

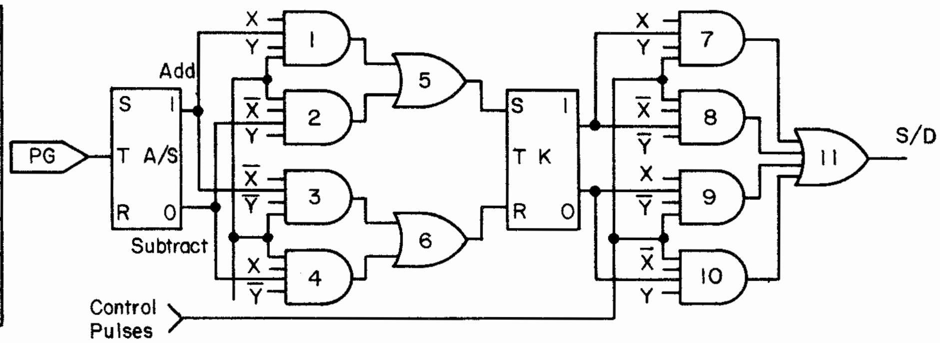


Figure 3-586. Full Serial Add-Subtract Unit

for either add or subtract, and one term to reset "K" for either condition. If the symbol A is used for the order ADD and the symbol S is used for the order SUBTRACT, the Boolean expressions for setting and resetting "K" are as follows:

Set "K" = $(AXY + S\overline{X}Y)$ clock
Reset "K" = $(A\overline{X}\overline{Y} + SX\overline{Y})$ clock

Sums and differences are generated by the expression $X\overline{Y}\overline{K} + \overline{X}Y\overline{K} + \overline{X}\overline{Y}K + XYK$ as in the full adder or full subtractor.

3-18.9.5 Parallel Additions and Subtractions in Binary Digital Computers

Comparison of Serial and Parallel Operation. Additions and subtractions may be accomplished in various ways for binary numbers. The serial adders and subtractors require that two binary configurations be shifted to their inputs and added bit by bit. Comparatively long periods of time are required to perform arithmetic operations in this manner. Each individual computer system will have its own unique requirements for the arithmetic operations it is to accomplish. If speed of operation is of prime importance, a system could use parallel addition and subtraction circuits. A system could use serial arithmetic operation circuits if power consumption or production cost of the system took precedence over speed of operation.

3-18.9.5.1 Parallel Adders-Subtractors

Half-adders have a special AND circuit for the generation of carriers. An ADD order could be added to the input of this gate so that a carry could only be generated when an ADD order was present. The borrow output could be ANDED with the SUBTRACT order for control of borrows. These changes make the half-adder into a half-add-subtract unit. Many of this type of unit are used together, as in Figure 3-587 to accomplish parallel additions or subtractions. Outputs at the C/B function are controlled by the proper order. Inputs to the units are NRZ (Non Return to Zero) and their outputs are the same. A clock pulse is controlled by ADD or SUBTRACT orders and applied to the pulse input to the S/D (sum or difference) output gates. With the ADD order

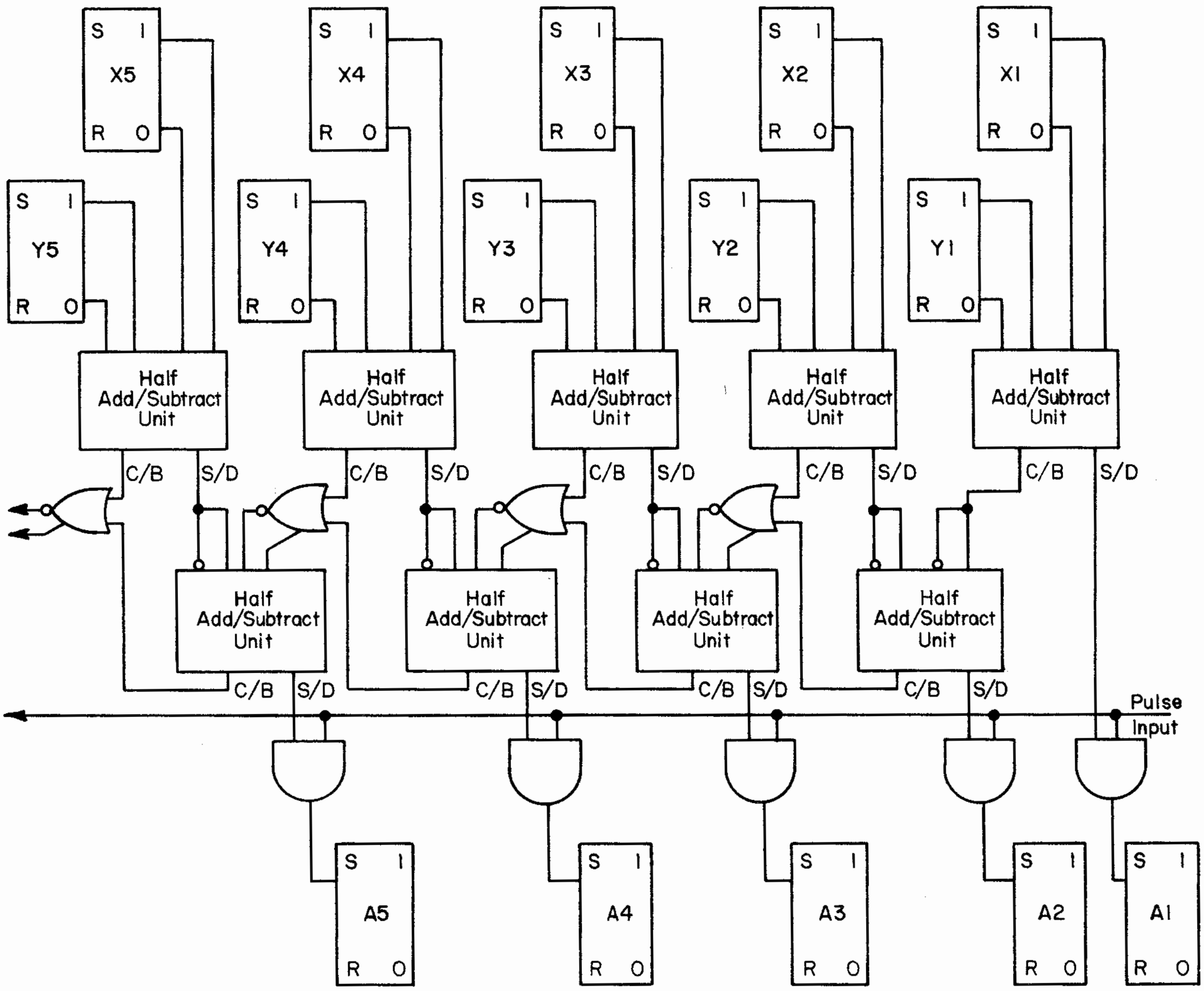


Figure 3-587. Full Parallel Add-Subtract Unit

present at the necessary places, an NRZ sum is present at the other inputs to these gates and the sum is transferred to the "A" (accumulator) register. The "Accumulator" accumulates binary configurations such as sums, products, differences, quotients and results of logical or transfer operations.

3-18.9.6 Incremental Adders in Special Purpose Circuits

At a specified time, an increment is added to an already existing binary word which will modify that binary word. The increment must be a single power of two because of the limits of the carry circuit. Any power of two may be added to the contents of the register containing the binary word. Figure 3-588 illustrates the use of a half-adder as a special purpose incremental adder.

3-18.10 STORAGE DEVICES

Memory is one of the more important units in a digital computer. Although memory or storage units differ from one computer to another, they all perform the same basic functions. They must be able to receive and retain information in terms of binary digits and deliver them to the computer upon demand. Usually when the computer commands the memory to supply a word it is desirable that the

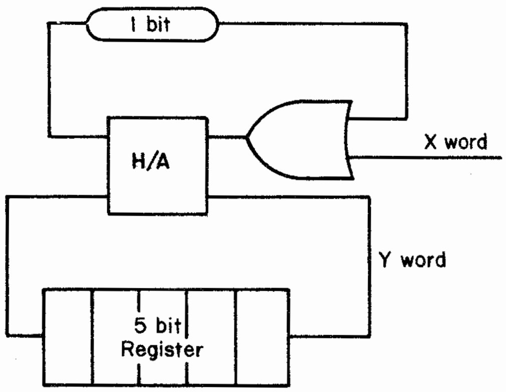


Figure 3-588. Special Purpose Incremental Adder

memory unit still retain the word, so that the computer can use it again later if the need arises. In the type of computer where orders describing computer operations are stored in the memory unit, the memory not only stores words whose binary configuration is a definite binary quantity (called "data"), but it also holds words whose binary configuration is a code (rather than a quantity) to tell the computer what to do. This is called a "stored program" type computer. Almost all general purpose computers are of this type. In the following discussion all words, whether they are "data" words or "instruction" words, are of the same length. Thus, in a twenty-bit computer, all words are twenty-bit words and the memory unit is capable of storing twenty-bit words only.

3-18.10.1 Types of Storage Devices

Four major types of storage systems have been developed to date. The first two listed below are of the non-magnetic type. The last two are magnetic and are the two most widely used in the digital field today. Only the latter two are detailed in this section.

1. Electrostatic Storage. The electrostatic charge on the surface of a cathode ray tube (CRT) is employed as the storage, and the deflection of the electron beam is the access control.

2. Delay Line Storage. A pulse or the absence of a pulse traveling along a delay line is used for storage, and time is used for access control.

3. Magnetic Surface Storage. Magnetic spots or areas on a magnetic surface are utilized for storage, and access to the individual spots is gained by physical positioning of the surface. For example, in audio magnetic tape recorders a particular part of a song, for example, is stored at a particular place on the tape.

4. Magnetic Core Storage. A toroid-shaped piece of saturable magnetic material serves as the storage element, and a coincident current scheme provides access.

3-18.10.2 Definition of Terms

Several terms must be understood which have specific meanings when used with respect to storage devices. It is necessary that these terms be familiar in order to comprehend much of the following material.

BIT SPACE or MEMORY CELL	A digital storage system contains elementary locations, each of which is capable of storing a single digit. Bit is a contraction of "binary digit" and is popularly used in computers to describe 1s or 0s.
WORD	A group of bits make up a word. It is usually a representation of a definite number. The number of bits per word is fixed and uniform in this discussion. Note that a computer is usually involved in the transfer of discrete words from one point to another, no matter what actual function it is performing at any one time.
ADDRESS	Each word in a memory unit must be identified as to its exact location in the unit so that the computer has a means of reading-out from memory the particular word it wants.
WRITING	The method of recording information (words) at the desired position or address in the memory.
READING	The method of extracting information from the desired position or address in the memory.
TIMING	The means of control over the movement of the words (ones and zeros) throughout the computer and between the computer and memory so that these words appear bit by bit at the proper point at the proper time for logical operation.

MODES OF ACCESS	Divided into parallel or serial, this refers to the method used to read or write a word. Serial relates to reading or writing one bit of the particular word at a time. Parallel relates to reading or writing all bits of the word simultaneously.
ACCESS TIME	Simply stated, this is the time required to obtain a word from storage to the computer, starting at the time the computer is ready for the word and ending at the time it produces the word. In large capacity storage units this may or may not constitute a problem, subject to the type of memory unit. Much thought has gone into ways of decreasing the access time on those large capacity devices where it has constituted a definite problem.
VOLATILITY	A "volatile" system is one where the information stored is lost when the power is turned off. In a non-volatile system the information is not lost.
PERMANENCE	The ability of a storage device to retain information.
RANDOM ACCESS	For this discussion, random access refers to a device in which it takes just as long to obtain access to one address as it does to obtain access to any other address.

3-18.10.3 Delay Line Storage

While many types of delay lines can be, and are used, the mercury delay line is the most popular. A mercury delay line is basically a relatively simple structure, as illustrated in Figure 3-589. Delay lines vary considerably in construction, but are basically a tube (glass or steel) to which a quartz crystal transducer has been attached at each end. High frequency pulses are applied to the electrodes on each face of the crystal at one end, and the crystal, as a result of the piezoelectric effect, vibrates. The vibrations pass down the mercury-filled tube to the crystal at the far end, where they induce a voltage across the electrodes of that crystal, again because of the piezoelectric phenomenon. The amount of delay

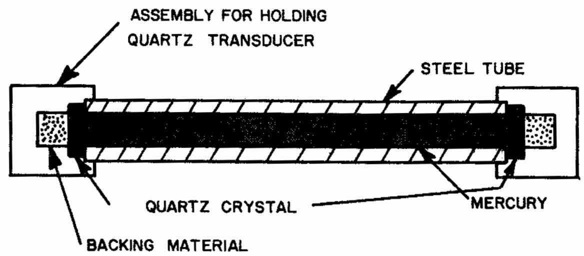


Figure 3-589. Mercury Delay Line

depends upon the length of the mercury column and the velocity of "sound" in mercury, which is approximately 57 inches per millisecond. If information is to be stored for a longer period of time, it must be recirculated through the delay line until it is needed. Delay line storage utilizes simple circuitry. It is a relatively inexpensive type of storage, but has several disadvantages. Delay lines have a low capacity storage, they are volatile, and the information must be recirculated if it is to be stored for any length of time. External noises and vibrations also affect operation.

3-18.10.4 Electrostatic Storage

Bits are stored by an electrostatic charge on the face of a special type of cathode ray tube. The storage is done by an electron beam which strikes the phosphorized surface of the tube. The beam is also used as the access control. This type of storage was widely used by early computers, but has generally been replaced by magnetic core storage. The electrostatic storage has the advantages of rapid access time, and random access; however, the permanence of this type of storage device is relatively low, since it is dependent upon the persistancy of the CRT phosphor. Information stored must be continuously rewritten. Other disadvantages are that this storage device is volatile and extremely high voltages are required.

3-18.10.5 Magnetic Spot Storage

The storage medium used in magnetic spot storage is a surface on which a thin layer of magnetic material has been deposited. Small areas, or spots, on this surface are magnetized to store bits. The lines of flux in this spot will point in a certain direction to represent a "1", and in the opposite direction to represent a "0". The spots are magnetized by a magnetic head similar to those used in audio tape recorders. The sensing, or reading, of stored bits is accomplished by means of relative motion between the magnetized surface and the same or a similar head. The changing flux lines, passing through the head as a result of this motion, will induce a voltage in an output winding. The polarity of this voltage will indicate whether a "1" or "0" is being read. There are many types of devices which use magnetic spot storage tapes, drums, and discs being the most common), but the basic principles of writing and reading are the same in each case. The bits are usually positioned in tracks or channels on the surface so that the bits in any one track will pass sequentially past the magnetic head. The major advantage of magnetic spot storage devices is a very large storage capacity since a great number of bits can be stored in a very small area.

3-18.10.5.1 Basic Principles

Figure 3-590 shows a writing and reading head and the magnetic surface. The read/write head is a piece of ferromagnetic material with a gap at the point nearest the magnetic surface. When current flows in the writing coil, the magnetic field set up by the coil will cause the head to become magnetized. The lines of flux at the gap will tend to pass through the magnetic surface rather than through the air between the tips, since the air offers a higher reluctance to the lines of flux than the magnetic surface. (Refer to B in Figure 3-590.) If a shim made of diamagnetic material (such as brass or silver) is placed between the tips (A in Figure 3-590), even fewer lines will pass between the tips and more will pass through the magnetic surface. Thus, the effect of the shim is to increase the flux density in the magnetic surface. The gap between the pole tips is usually .001-inch or less, and the head is separated .001-inch or less from the surface, so the spot being magnetized is extremely small. The strength of the magnetomotive force is great enough to cause this spot on the surface to be magnetized to saturation in the desired direction. After the write operation is completed the spot will remain magnetized indefinitely since the magnetic material on the surface has high permanence. This spot can be thought of as a small bar magnet with the lines of flux returning through the air from one end of the bar to the other (Figure 3-591). The polarity of this tiny bar magnetic is dependent on the direction of current flow in the write winding. In A of Figure 3-592, with current flowing in the direction of the arrows, the lines of flux in the head will run counter-clockwise and the spot will be magnetized as shown in the figure. This condition might represent a binary "1". In B of Figure 3-592, current is reversed, the lines of flux in the head run clockwise, and the spot is magnetized as shown. This condition would represent a binary "0". Note that the magnetic surface is moving past the magnetic head during this write operation, but the pulse applied to the write winding is of a very short time duration. The magnetized spot on the surface will therefore be very small. Previous information need not be erased in order to write new input. The mmf exerted on the surface is strong enough to change the polarity of the spot of the new condition, when writing either a "1" or a "0", regardless of the previous condition of that area. To read stored information, the magnetic surface is passed underneath a magnetic head similar to the one used to write, or under the same head (A in Figure 3-593). As the magnetized spot passes beneath the read head the lines of flux from the spot will pass through the magnetic head because it has a lower reluctance from the air (B in Figure 3-593). This change in flux in the read head will induce a voltage in the read coil. The polarity of this voltage will be dependent upon the direction of the lines of flux and will be maximum at the time of greatest flux change. As the spot passes by the read head, the lines

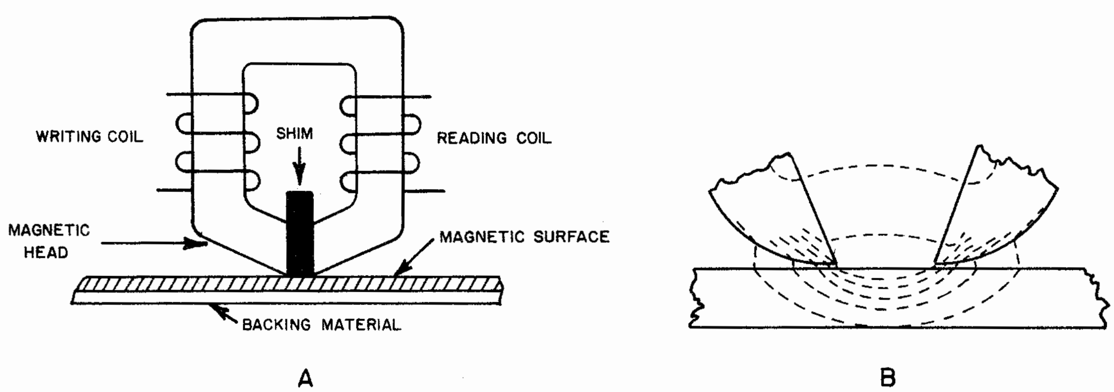


Figure 3-590. Magnetic Read/Write Heads

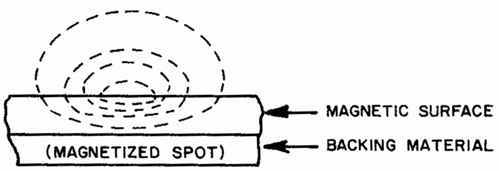


Figure 3-591. Bar Magnet Effect

of flux passing through the head will decrease to zero since the magnetic material in the head has extremely low permanence. This change in flux will induce a voltage of the opposite polarity so that the output for this entire operation would be as shown in C of Figure 3-593. If the bit being read in Figure 3-593 was a "0", the output for reading a "1" would be 180° out of phase with the output shown, since the polarity of the voltage induced in the read coil is dependent upon the direction of the lines of flux.

3-18.10.5.2 RZ Recording

A pulse of current is applied to the write winding in one direction to write a "1" and a pulse in the opposite direction is applied to write a "0". This method is called return-to-zero because the current returns to zero between bits of information. This type of storage has one major disadvantage. When several bits of the same type (several "1's" or several "0's") are stored in sequence, the lines of flux on the spots will tend to merge if the bits are stored close together. This factor limits the storage density possible with RZ recording. Consequently, RZ storage is sometimes used for clock pulses only. Clock pulses must be distinct, individual pulses, and RZ recording will meet this requirement.

3-18.10.5.3 NRZ Recording

When the individual writing pulses are spread out in time so that they occupy a full bit-time, the recording method is called non-return-to-zero or NRZ. The reason for the name is that the pulses lose their individualities and the current does not return to zero between successive "0s" or successive "1s".

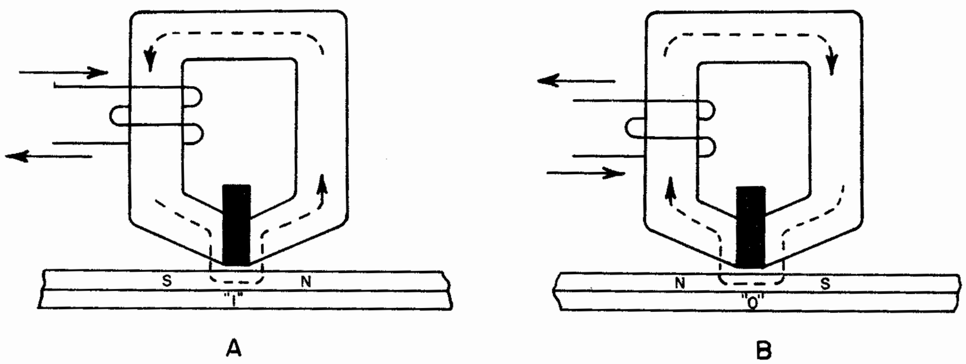


Figure 3-592. Binary Writing of 1 and 0 on Magnetic Material

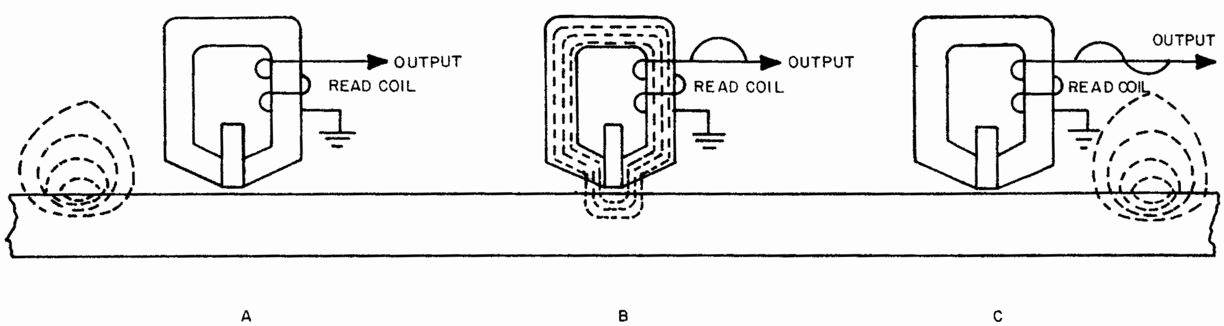


Figure 3-593. Reading 1s and 0s

Instead, the storage surface is continuously magnetized in one direction or the other. The NRZ method of storage is illustrated in Figure 3-594. A plot of the writing current used to store a sample sequence of binary digits is A of Figure 3-594. Note that the first three digits are "1s" and that current is constant for these three bit times. Current changes only when a "1" is followed by a "0" or when "0" is followed by a "1". Part B in Figure 3-594 is a plot of the flux on the surface, which was produced by the writing current. Note that, instead of individual spots for successive "1s" or successive "0s", they are represented by a spot, magnetized in the opposite direction and occupying two memory cells. A single "1" or "0" would be represented by a spot occupying one memory cell. Part C in Figure 3-594 is a plot of the voltage which would be induced in the read winding by the memory cells in B of Figure 3-594 as they passed beneath the read head. Since only a change in flux will produce a voltage in the read winding only a change from "1" to "0", or "0" to "1", will result in an output. The first bit of the group of these "1s" will produce an output, and the changing flux between coils 3 and 4 will result in an output of the opposite polarity. As can be seen from the diagram, a change from "0" to "1" will produce a negative pulse, and a change from "1" to "0" will produce a positive output (negative logic).

3-18.10.5.4 NRZ Read-Write Circuits

Figure 3-595 shows a single magnetic head being used to both read and write. To write, a write command or level is applied to AND gate #1 along with a control level. This control level would usually be one WT in length and is used to control the time that the writing takes place. When AND gate #1 is qualified its output will supply forward bias to Q3 which in turn will allow current to flow through either Q1 or Q2. Q1 and Q2 provide the push-pull action necessary in recording. When the data input is a "1", Q1 will conduct, current will flow in the write winding, and a "1" will be written into the surface. When the data in is a "0", $\overline{\text{data}}$ will be a "1" and Q2 will conduct causing current to flow in the opposite direction in the write coil and a "0" will be written into the surface. When reading, the voltage produced in the "read" winding is applied to an amplifier with no phase inversion and to an amplifier which does have a phase inversion. These amplifiers are biased so that zero volts at the input will result in zero volts at the outputs. A voltage from the read winding which represents a "1" will produce a "1" out of the amplifier to O/S #1. The one-shot is used to shape this pulse. The output from O/S #1 will set the flip-flop and produce a "1" from AND gate #2 when the READ command and the control level are "1s". The control level would usually be one WT in length and would be used to control the time that the reading takes place. A voltage from the read winding which represents a "0" would produce a "1" out of the inverter to trigger O/S #2 and reset the flip-flop. This will result in a "0" output until the next change from "0" to "1".

3-18.10.5.5 Types of Magnetic Spot Storage Devices

There are several types of magnetic spot storage devices, but the writing and reading principles for all types are the same.

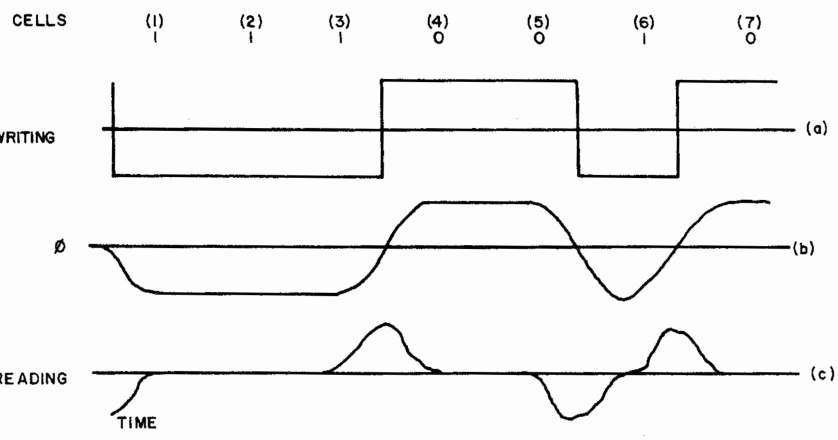


Figure 3-594. Waveforms

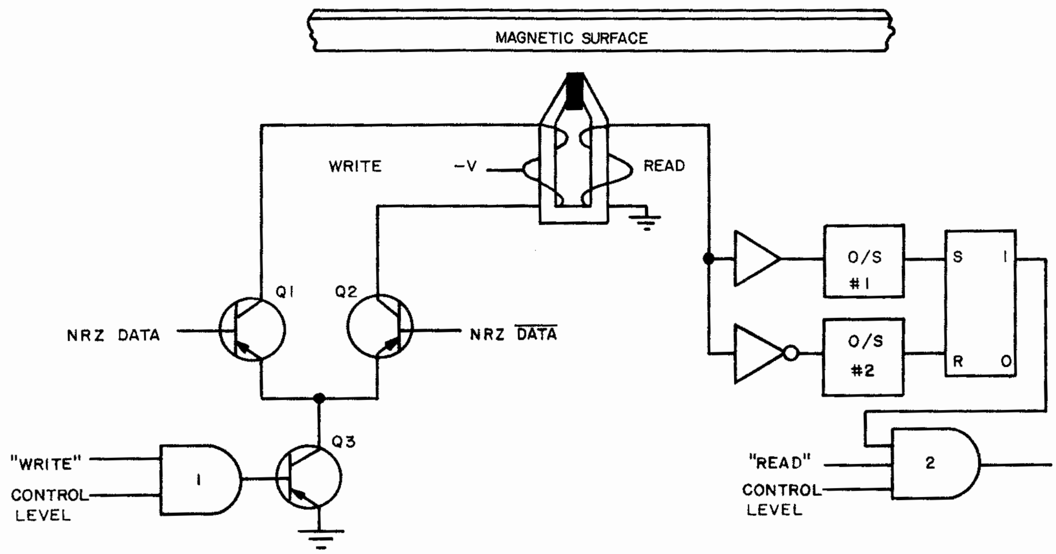


Figure 3-595. Typical Read/Write Head

3-18.10.5.5.1 Magnetic Drum

Magnetic drums are used by many computers as the main working memory, and by many others as a back-up auxiliary storage unit where magnetic cores are used as the working memories. The drum is a cylinder with a magnetic coating on its surface (not on the ends) (Figure 3-596). The drum rotates around the axis of the cylinder, and magnetic read-write heads are placed close to the magnetic surface. Each head has access to one channel or line of bit-cells around the circumference of the drum. All information on the drum will pass beneath the magnetic heads, once at every drum revolution. This is called "cyclic access." The maximum access time of a drum memory is thus the time of one drum revolution. This time would be in milliseconds, depending upon the speed of the particular drum.

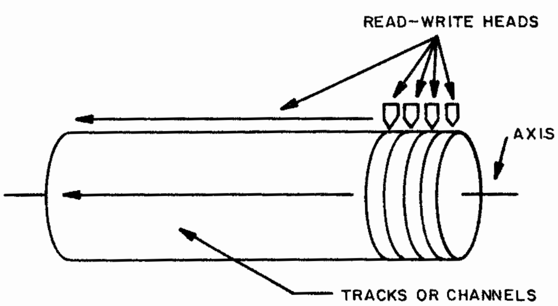


Figure 3-596. Typical Magnetic Drum

3-18.10.5.5.2 Magnetic Tapes

Tape is a commonly used form of magnetic storage surface. It offers large storage capacity but is much slower than the drum method. For this reason, tape storage is usuallly used only as auxiliary or back-up storage, rather than as the working memory. The tape consists of a long strip of flexible base material having a coating of magnetic material. As with the drum method, information is stored in channels or tracks, with a read-write head for each track. The tape is stored on reels and moves past the read-write heads in the same manner as that employed in audio tape recorders. This technique is called "sequential access."

3-18.10.5.5.3 Magnetic Disks

Magnetic disks offer high storage capacity and relatively rapid access time. The magnetic coating is applied to both sides of the disk so the total storage area is equal to twice the area of the disk. The channels are concentric or spiral as shown in Figure 3-597. A separate read-write head may be used for each channel on concentric systems or a single moveable head may be used to provide access to all channels. The disk rotates past the heads to provide cyclic access. Access is attained in the spiral storage disk by using the phonograph principle.

3-18.10.6 Magnetic Core Storage

The magnetic core is in the form of a small ring of easily magnetized material. This material must be capable of being saturated and must have

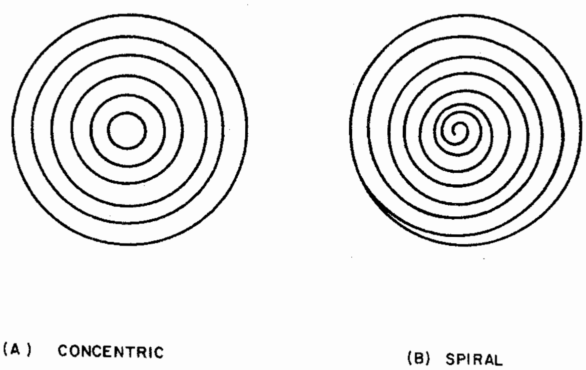


Figure 3-597. Magnetic Disk Channel Layout

high retentivity so that once it is magnetized it will retain its magnetized state indefinitely. This means it must have low reluctance, high permeability, and high retentivity. A core with these characteristics would exhibit a hysterisis loop as shown in Figure 3-598. Note that as the force applied (Hc) drops back to zero, the core retains its magnetization. Also that if the force is reversed, it takes a definite quantity of reverse force before the core reverses its state. Once this force is sufficient to reverse the core's state, it does so rapidly. The equivalent of a binary 1 is stored in a core by magnetizing it in a known direction; the equivalent of a binary 0 is stored by magnetizing the core in the opposite direction. The cores are mounted in frames or planes. Each core in the frame represents one bit of a binary word. Thus, if the memory contains 1024 words, each frame will have 1024 cores. The number of bits in each word will determine how many frames are used. If the computer words are 20 bits in length, 20 frames will therefore be needed. When any word is written in (stored in) the memory, the information will be written into only one core of each frame. The problem which now arises is: how to select the proper core in each frame for writing and reading. Part A in Figure 3-598 shows a portion of one frame. Since only one core in this frame is to be used at any one time, that core must be controlled without affecting any of the others. The selection principle used for both reading and writing is called "coincident current" selection, and is based on the rectangular nature of the hysteresis loop. As indicated in B of Figure 3-598, a magnetomotive force (mmf) of Hc is sufficient to set the core to the "1" state, but a field of half this value (Hc/2) is not sufficient.

3-18.10:6.1 Reading

To select core 23 (Part A of Figure 3-598), and read the bit stored there, a current of the proper direction and magnitude to produce an mmf of -Hc/2 is passed through each of the drive windings, X2 and Y3. The current in X2 will produce an mmf of -Hc/2 in cores 21, 22, and 23. The current in Y3 will produce an mmf of -Hc/2 in cores 13, 23, and 33. The mmf produced by the current in either X2 or Y3 is not enough to change the state of any core, but note that both X2 and Y3 are common to core 23. The mmf produced by both X2 and Y3 therefore combines here to cause core 23 to be magnetized in the "0" state. ((-Hc/2) + (-Hc/2) = -Hc.) If the core was previously in the "1" state, a change in flux resulted from the mmf produced by the current in X2 and Y3. This change in flux will induce a voltage in the sense winding. Since only one core in each frame is changed at any one time, a single sense winding can be used for the frame. This winding is common to all cores in the frame, as shown in Figure 3-598. The voltage induced in the sense winding is amplified by a sense amplifier and the

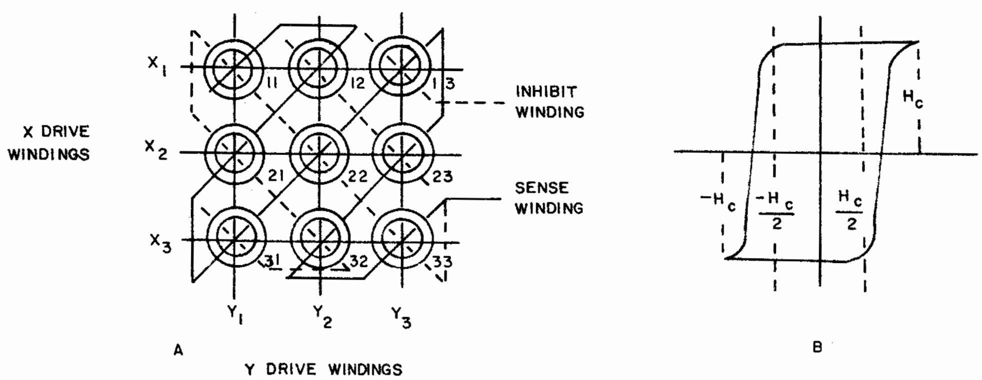


Figure 3-598. Core Memory Matrix

output of the amplifier, which represents a binary 1, will be sent to the computer. If the core was in the "0" state prior to reading the information, no voltage will be induced in the sense winding and there will be no output from the sense amplifier. The absence of an output at this time represents a binary 0. Note that regardless of the state of the core prior to reading the information, the core will always be in the "0" state after being read. This type of memory device has a destructive read-out. Therefore, the core must now be returned to its original state. To do this, the read operation is followed by an automatic write operation. During the write operation, a current opposite to that used in the read operation would be applied to the selected X and Y windings (in this case X2 and Y3). This current is of sufficient magnitude to produce an mmf of +Hc/2. Because X2 and Y3 intersect at core 23 the combined mmf produced by these two windings would be great enough to switch core 23 to the "1" condition. A fourth winding, called the inhibit winding, is represented by the dotted line in Part A of Figure 3-598. Note that, like the sense winding, the inhibit winding is common to all cores in this one frame. If the sense winding output was a "0" during the read operation, a current which will produce an mmf of -Hc/2 in all cores will be applied to the inhibit winding during the write operation. The forces acting on core 23 would thus be (+Hc/2) + (+Hc/2) + (-Hc/2) = +Hc/2. The total force +Hc/2, is not great enough to switch core 23, and it will therefore remain in the "0" state: i.e., the condition it was in prior to the read operation. If the sense winding output was a "1" during the read operation, no current is applied to the inhibit winding during the write operation. With no current in the inhibit winding, the forces acting on the selected core would be (+Hc/2) + (-Hc/2) = +Hc. A force of +Hc will switch the core back to the "1" state, its original condition.

3-18.10.6.2 Writing

To inject new information into the memory, the same operations, read, inhibit, and write, are performed during the memory cycle. During the read operation, all selected cores (one from each frame) are switched to "0", and the outputs from the sense windings are not used. The sense amplifiers are disabled, therefore all the read step accomplishes is to switch all cores at the selected address to "0". During the write operation, however, the new information to be stored will determine which inhibit windings will be pulsed. If a "0" is to be written in the selected core of a particular frame, then that inhibit winding would be pulsed. No current would be applied to the inhibit windings of those frames where a "1" is to be written in to the selected cores. Magnetic core memories are considered the best type of working memory because of their extremely low access time (as low as 0.2 microseconds). Also, since the cores are selected electronically with no mechanical movement involved, they are random access devices.

3-18.11 COMPARATORS

In order to perform a variety of functions, digital computers must be able to compare two binary words and detect any differences. Several different comparisons can be made, but all fall into two categories: comparing for identity, and comparing for magnitude. There are many variations of each type of comparator, and while both types will not be described, no attempt will be made to cover all designs.

3-18.11.1 Exclusive "OR"

The comparator used to compare two words and produce a "1" output when the two words are not identical is called an "Exclusive "OR" circuit." To determine whether the two words being compared are identical, it is necessary to compare all bits in each word bit by bit. For example, in comparing the two words shown in Figure 3-599, the LSD bits of each

0/0110101	X WORD
0/0110101	Y WORD

Figure 3-599. Word Comparison

word are compared first. Successive bits are then checked, with the MSD digit being the final bit compared. A "1" output from the Exclusive "OR" circuit should occur under two conditions only: when the X digit is a "1" and the Y digit a "0", or when the X digit is a "0" and the Y digit a "1" (Figure 3-600).

0/0000011	X WORD
0/0100010	Y WORD

Figure 3-600. Word Comparison

The truth table for the exclusive "OR" circuit is shown in Figure 3-601. The Boolean expression derived from this table is $\overline{X}Y + X\overline{Y}$. One circuit configuration of the Exclusive "OR" is shown in Figure 3-602. The X and Y words are shifted into this circuit simultaneously and serially. A "1" output from the "AND" gate anytime during the comparison stage indicates that the two

X	Y	OUTPUT
0	0	0
0	1	1
1	0	1
1	1	0

Figure 3-601. Word Comparison

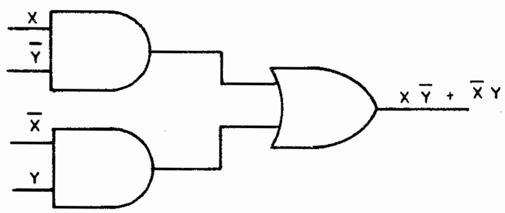


Figure 3-602. Comparator Circuit

words were not identical. The absence of a "1" output during the comparison stage indicates that the two words were identical. The military standard symbol for the circuit presented in Figure 3-602 is shown in Figure 3-603.

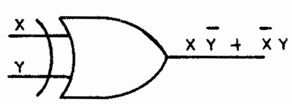


Figure 3-603. Exclusive OR Gate Comparator

3-18.11.2 Algebraic Comparator

The algebraic comparator is a serial binary comparator. It will compare two numbers and, if a difference does exist, will generate an output indicating which number has the largest algebraic value. The algebraic comparator will compare two positive words, two negative words, or a positive and a negative word. In all comparisons, the number with the largest algebraic value will appear the larger; a small negative number will appear larger than a large negative number. In the comparator shown in Figure 3-604, the two words, X and Y, are shifted serially, with the LSD bit shifted first. Negative numbers will be in the "two's complement" form. The Tn signal is at a non-return-to-zero (NRZ) level which will be a "1" during the time that the sign bits are being compared. Since both output "AND" gates (11 and 12) are qualified with Tn, an output can occur only at this time. For example, assume that the comparator is to be employed to compare the following two positive numbers:

X word 0/10110 Y word 0/01100

Flip-flops A and B are reset to To. At TP1 time, both LSDs are "0s," thus no gates are qualified. At TP2 time, the X bit is a "1" and the Y bit is a "0". "AND" gates 1 and 4 will be qualified to set F/F A, and a pulse will be applied to the reset input of F/F B, since X is a "1" and Y is a "1". "AND" gates 5, 6, 7, 8, 9, and 10 do not need to be considered until Tn time since the output gates, 11 and 12 cannot be qualifed until that time. F/F A in the set condition indicates that, up to this point, the X word is larger than the Y word. At TP3 time both bits being compared are "1s" and no gates will be qualified. At TP4 time the X word is a "0" and the Y word is a "1". "AND" gates 2 and 3 will be qualified to SET F/F B and RESET F/F A. F/F B in the SET condition indicates that up to this point, Y is larger than X. Successive comparisons indicate that the X word is larger. When comparing two negative numbers, the operation of the comparator is the same as described above. For example, assume the input are:

1/00110 X word 1/10110 Y word

During TP1 through TP4, both inputs are identical and no gates are qualified and at TP5 F/F B will be SET. At Tn time, X and Y are both "1s," F/F B is SET so "AND" gate 7 will be qualified. Its output passes through "OR" gate 4 to qualify "AND" gate 12. An output from "AND" gate 12 indicates that Y is larger than X. When comparing a positive number with a negative number, all bits are compared. However, only the sign bit has significance, since any positive number is considered to be larger than any negative number. If X is positive and Y is negative, the X sign bit will be a "0" and the Y sign bit will be a "1". At Tn time, "AND" gate 9 would be qualified and its output would pass through "OR" gate 3 and "AND" gate 11 to indicate that X is the larger word. If X is negative and Y is positive, the X sign bit will be a "1" and the Y sign bit will be a "0". At Tn time, "AND" gate 10 would be qualified and its output would pass through "OR" gate 4 and "AND" gate 12 to indicate that Y is the larger word. One more factor must be considered. When comparing two positive numbers or two negative numbers, the output is dependant upon which flip-flop is the "set" condition. This is not true when comparing a negative number with a positive number. If X is positive and Y is negative, it is possible

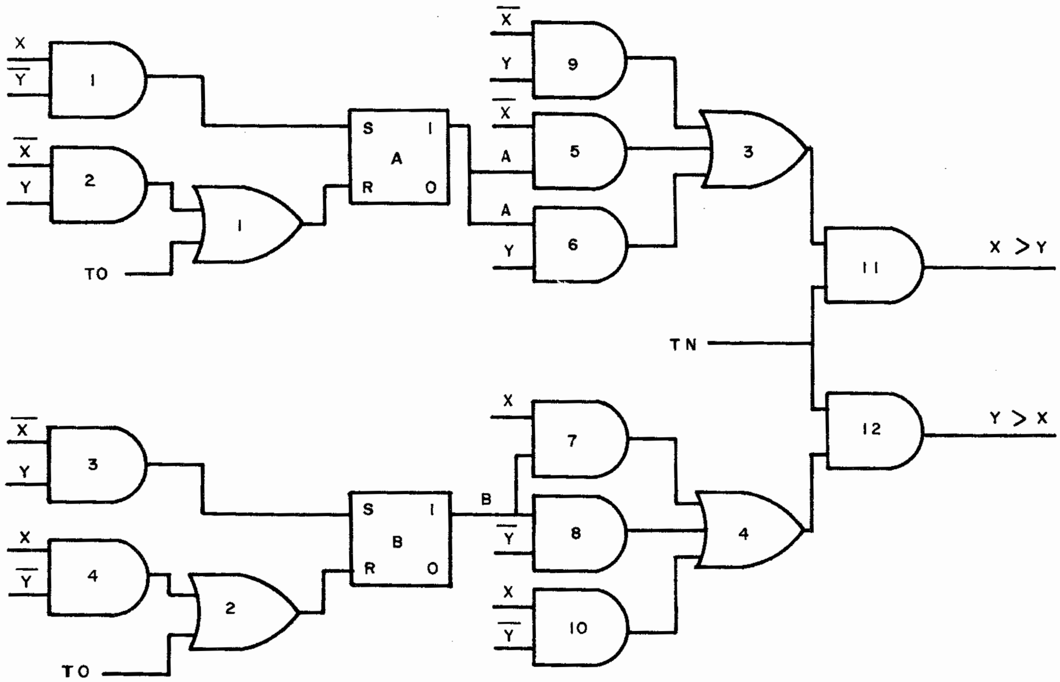


Figure 3-604. Typical Comparator Circuit

the F/F B will be in "set" condition at Tn time. If the output of F/F B were allowed to pass through to the output, this would give erroneous indications. Therefore, if X is a "1" and Y is a "1", the output of F/F B must not pass through "OR" gate 4. This can be expressed by the Boolean expression $B(X + \overline{Y})$ or $BX + B\overline{Y}$. This expression, along with the output of "AND" gate 10 which is qualified when X is negative and Y is positive forms the full output expression of "OR" gate 4: namely, $BX + B\overline{Y} + X\overline{Y}$. If X is negative and Y is positive, it is possible that F/F A will be set at Tn time. If X is a "1" and Y is a "0" at Tn time, the output of F/F A must not pass through "OR" gate 3. This condition can be expressed as $A(\overline{X} + Y)$ or $A\overline{X} + AY$. This expression, along with the output of "AND" gate 9 (which is qualified when X is positive and Y is negative) forms the full output expression of "OR" gate 3: namely, $A\overline{X} + AY + \overline{X}Y$. No output will result from this circuit when two words being compared are identical.

3-18.12 ALIGNMENT

Circuit adjustments may be necessary when input or output units are either connected or disconnected from a computer. These adjustments may vary the gain of amplifier stages, the frequency of a timing circuit, or the voltage furnished by a power supply. The need for other adjustments will be indicated by preventive-maintenance checks. Power-supply outputs must be adjusted to furnish the correct current to magnetic cores and the correct voltage to amplifier and oscillator stages. The speed of magnetic drum and tape devices must be adjusted to provide computer signals having the correct pulse width and timing.

3-18.12.1 Core Memories

Core memory circuits may require realignment because of the following reasons:

1. Changes in operating currents due to the aging parts.
2. Replacement of plug-in units, such as digit plane drivers, memory gate generators, or timing stages.
3. Accidental changing of the original potentiometer settings.

The procedures for adjusting a core memory require checks of the memory timing cycle, the digit-plan-driver current, the read current, and the write current. The memory timing cycle can be measured with an oscilloscope calibrated to measure time. The time between the clear-memory, start-memory, set-read, clear-read, set-inhibit, set-write, clear-write, and clear-inhibit pulses must be checked. If a discrepancy exists in the timing of any of these pulses, an adjustment must be made. In some memory units this is done by changing the taps on the memory clock delay line. An oscilloscope calibrated to indicate voltage can be used to determine the digit-plane-driver, read, and write currents. Maintenance programs can be used to inject the proper computer signals into the memory-coil arrays. You can then calculate the current through each drive line after measuring the voltage drop across the terminating resistor for each set of windings. Potentiometers located in the memory units are used to set the current to the correct amplitude. Recheck and, if necessary, readjust the read and write currents after the first adjustments are completed. This readjustment may be necessary because of interaction between the two circuits. After the read and write currents have been adjusted, you can check the balance by means of an oscilloscope and, if necessary, make a fine adjustment. Connect the oscilloscope probe to a read-write drive line. Run a maintenance program to apply synchronized bursts of read and write pulses to the drive line. If the read and write currents are equal, the oscilloscope display should be similar to that shown in part A of Figure 3-605. If any unbalance is present, the oscilloscope display will resemble the display shown in part B or C of the figure. Either the read or write current potentiometer should be adjusted slightly to provide a balance.

3-18.12.2 Magnetic Drums

Magnetic-drum read and write head adjustments are always required in the initial stages of installation and after replacement. When the read and write functions are accomplished by two drum heads, the heads must be adjusted with respect to both signal amplitude and timing. Since these two adjustments are interacting, they must be made concurrently. The interaction of these adjustments results from the construction and location of the heads, head bars, and rotor drum. When writing and reading are accomplished by a single drum head, the timing will always be correct and only the amplitude adjustment must be made. The use of two oscilloscopes is recommended for drum head adjustments. One oscilloscope is connected to the input of the drum read amplifier to monitor the head amplitude, and the other is connected to the output of the drum read amplifier to provide a check of timing. Typical waveforms that may be obtained are shown in Figure 3-606.

3-18.12.3 Timing Adjustment

For a coarse adjustment, loosen the head-retaining screws and move the head the desired

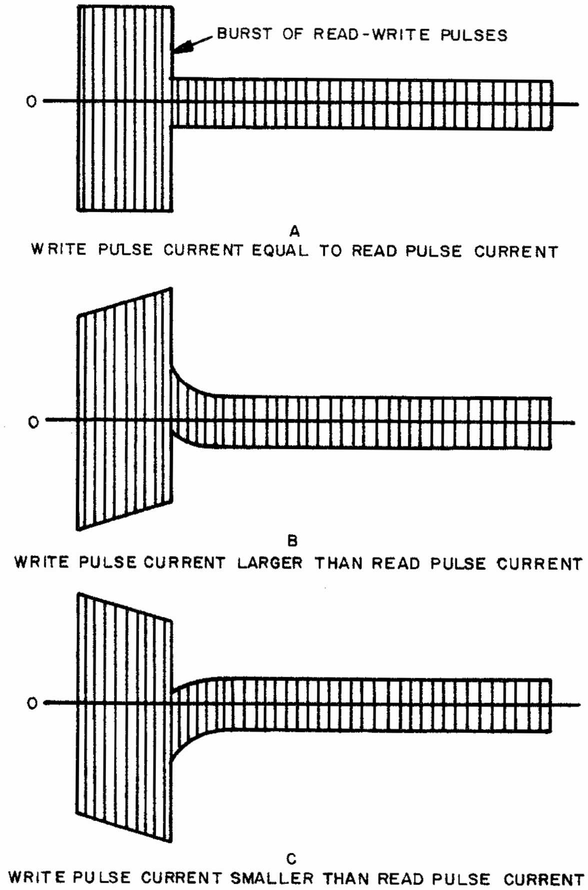


Figure 3-605. Magnetic Core Read and Write Current Balance Check Waveforms

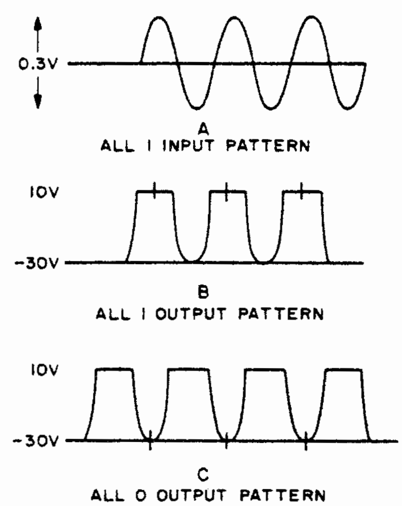


Figure 3-606. Typical Drum Read Amplifier Test Waveforms

amount. Then tighten the screws until they are just snug enough to retain this position. Observe the timing pattern on the oscilloscope, and use the retaining screws to rock the head for a fine adjustment. One screw should be tightened and the other loosened to tilt the head to the position which provides accurate timing. Movement of the screws in either direction should be slight. Excessive tightening of a screw may cause the head to be cocked an abnormal amount and result in a distorted input signal to the read amplifier.

3-18.12.4 Signal Amplitude Adjustment

The output signal amplitude of a read head is adjusted by varying the air gap between the drum surface and the read head core. On some magnetic drums, the read heads are adjusted to provide a signal output which is 75 percent of the amplitude that is measured when the head is in contact with the drum surface. To make this adjustment, proceed as follows:

1. Connect an oscilloscope to the output of the read head.
2. Loosen the lock nut on the amplitude screw just enough to permit a snug fit.
3. Each time an adjustment is made, tap the amplitude adjustment screw lightly to overcome static friction in the associated mechanism.
4. Turn the amplitude adjustment screw until the head core just makes contact with the drum surface. The head core should remain in this position only long enough to measure the head peak-to-peak voltage output with the oscilloscope; otherwise the head may be damaged.
5. Turn the amplitude adjustment screw until the amplitude of the head output waveform is reduced to 75 percent of the value measured in step 4. For example, assuming that the contact amplitude is 300 millivolts, you should adjust the head to provide an output equal to 75 percent of 300 millivolts, which is 255 millivolts.
6. The acceptable limits for the output of a drum head are usually listed in the equipment technical manual. For the example above, the final amplitude setting must fall within the limits of 125 to 300 millivolts; that is, if the 75-percent value is higher than 300 millivolts, it must be reduced to this maximum value. If the resultant is lower than 125 millivolts, the head is probably defective and may have to be replaced.
7. If noise spikes appear at the output of the drum head at the 75 percent setting of the amplitude adjustment, they should be minimized by lowering the amplitude below this value.
8. When the desired amplitude is obtained, tighten the lock nut, taking care not to disturb the adjustment.

3-18.12.5 Timing and Amplitude Relationship

A definite timing and amplitude relationship exists between corresponding read and write heads, and this relationship must be taken into account when adjustments are deemed necessary. If a timing error is detected in a read head and you cannot move the head far enough to obtain correct timing, the following procedure is recommended:

1. Move the corresponding write head in a direction to correct the timing error. Write a new test signal with the head in the new position.
2. Adjust the write-head amplitude. This is accomplished by removing the write-head plug from the write-head-amplifier cable connector and applying the output of the write head to the vertical input of an oscilloscope. Then using the write head as a receiver, adjust the gap between the write-head core and the drum surface to provide an amplitude equal to 75 percent of the contact amplitude. This must be within the limits specified in the equipment technical manual. The limit for the write head may be narrower than for the receive head; for this example assume the limits to be 175 and 200 millivolts.
3. Reconnect the write head to the write-head amplifier, and write a new test pattern. Readjust the read head for correct timing and amplitude. If the minimum limit remains unattainable, replace the head.

3-18.13 PREVENTIVE MAINTENANCE

Preventive maintenance for computer equipment is basically the same as for other electrical equipment; it consists of inspection, cleaning, preservation, lubrication, and performance checks. However, a unique feature of computer preventive maintenance is that the reliability programs can be used to quickly determine whether a computer is operating properly. In some computers marginal checks can be controlled by the reliability program, or performed manually to detect deterioration of computer parts before a failure occurs. Because of the extensive amount of wiring in large computers, wiring and cabling should be carefully inspected. Check the wires for loose or broken lacing, and frayed insulation. Check the cables and connectors for improper placement that might subject them to strains or kinks. Inspect the connectors and wires for burned or charred parts, dirt, cracks, and breaks.

Inspect switches for loose mountings and connections. Examine the contacts for dirt, pitting, and corrosion. Test the action of the switches and see that they operate without binding. In gang and wafer switches, see that the moveable blade makes good contact with the stationary member. Make sure that the stationary contact leaves spread as a movable blade slides into them. Some switches have contacts that are impossible to reach without damaging the switch assembly. Check these switches for defective mechanical action and looseness of mountings and connections.

3-18.13.1 Magnetic Drum Units

To eliminate noise which may be generated by the rotating drum, periodically inspect the static-grounding brush that is mounted at the end of the drum assembly. Replace the brush when it wears to less than half its original length. After a brush has been replaced, check to make sure that there is low resistance between the end plate of the drum unit. This check will assure that the brush is seated properly in the holder and is making good electrical contact. The drum rotor, drive motor, bolt, and pulleys are all subject to mechanical wear during normal operation; therefore, they must be regularly inspected and maintained. Check the drum drive motor pulley for tightness. An indication of a loose pulley is an accumulation of belt rubber on the pulley and pulley guard. If the pulley is allowed to remain loose, the belt will be chafed and its life shortened.

3-18.13.2 Tape Drive Units

Many digital computers include tape handling equipment. This equipment must be cleaned, lubricated, and periodically checked to ensure proper operation. Daily maintenance includes cleaning the head assembly at the beginning of each day's operation and running a short reliability maintenance program, if the computer has this feature. Weekly maintenance may include the cleaning of items such as the friction drive clutch and the air filters. Check the machine's forward and reverse transfer time, the high-speed rewind, creeping of the tape reel, and the tape break circuitry. Check the ability of the equipment to reproduce test pulses, to measure the tape speed, and to measure the moving coil and erase coil currents. A more comprehensive reliability maintenance program than is used for the daily check can then be run to determine whether error-free operation can be achieved for a period of 15 minutes.

3-18.13.3 Magnetic Tape

Dust and dirt can reduce the intensity of the reading and recording pulses by increasing the gap between the tape and the head. Therefore, take care of tape in the following manner:

1. Keep tape in a dust-proof container whenever it is not in use on a tape unit.
2. While the tape is on the machine, keep the container closed and store it where it is not exposed to dust or dirt.
3. Store tapes in some type of cabinet that is elevated from the floor or deck and away from sources of paper or card dust. This should minimize the transfer of dust from the outside of the containers to the tape reel during loading and unloading operations.
4. Do not use the top of tape units as a working area. Placing materials on top of the units may expose the tape to heat and dust from the blowers in the unit.
5. When identifying tape reels, use a material that can be removed without leaving a residue. Adhesive stickers, that can be easily applied and removed, are satisfactory. Never alter a label identification by means of an eraser, since particles from the eraser may come in contact with the tape.
6. Inspect tape containers periodically and remove any dust by washing with a regular household detergent.
7. When necessary to clean tape, gently wipe it with a clean, lint-free cloth moistened with an approved cleaning fluid, such as trichlorethylene.

Recorded information normally comes very close to the edge of a tape. Therefore, for proper operation, the edges of the tape must be free from nicks and kinks. To accomplish this, observe the following precautions:

1. When removing a tape reel from a recorder, handle the reel near the hub whenever possible. If there is resistance in removing the reel, press it from the rear as near to the hub as possible. Under no circumstances should the reel be rocked by grasping the outer edge.
2. Avoid throwing or dropping reels, and do not make contact with the exposed edge of the tape. Dropping a reel can easily damage both the reel and the tape. The use of a reel and tape after the reel has been dropped is usually unsatisfactory. Therefore, never throw or mishandle reels, even while they are in their containers.
3. When mounting reels on the recorder, push them firmly against the stop on the mounting hub to ensure good alignment.
4. When placing the tape on the take-up reel, carefully align the tape to prevent damaging the edge on the first few turns.

5. If a tape break occurs, wind the resulting pieces onto two small reels. Splicing is not recommended unless it is necessary to make a temporary splice to recover information.

Magnetic tape is sensitive to changes in humidity and temperature. Recommendations for tape storage are as follows:

1. If possible, store the tape in the computer room where it is to be used. Location of tape storage near the tape drives reduces both handling and variations in atmospheric conditions.
2. The storage area should be kept at a temperature of 18.3 °C to 26.7 °C and at a relative humidity of 40 to 60 percent.
3. If the tape must be removed from the computer room, hermetically seal it in a plastic bag to reduce the effect of temperature and humidity changes on the tape's physical dimensions. If the tape is not hermetically sealed, before it is reused it should be allowed to remain in the computer room for a length of time equal to the time it was out of the computer room. If the tape is out of the computer room for a period longer than 24 hours, it should be conditioned to the computer room for 24 hours before being reused.
4. For long-term storage, enclose the reel and container in a hermetically sealed plastic bag. Store in an area of constant temperature. Either cold or hot temperature can harm tape. A temperature between 4.4 °C and 49 °C is satisfactory.

3-18.14 TROUBLESHOOTING

Computer circuits such as multivibrators, clock oscillators, and gates can be tested in the same manner as similar circuits in radar, radio, and multiplex carrier equipment. Test equipment such as oscilloscopes, voltmeters, ohmmeters, frequency meters, tube testers, and transistor testers can be used to determine which parts are actually defective and, if possible, what has caused the failure. However, large computers have such enormous numbers of circuits that efficient troubleshooting to locate defective circuits can be accomplished only with the aid of maintenance programs, built-in test equipment, and special test sets designed to be used with specific computers. In computers which feature maintenance programs, reliability checks may be used to determine the area of a computer in which a failure has occurred. Diagnostic programs can be run to localize the trouble to a group of circuits or plug-in units. For example, if trouble appears to be in a plug-in matrix assembly, the suspected matrix assembly can be replaced by a unit known to be operating correctly. If this results in proper computer operation, test the defective matrix assembly with the aid of a matrix test set to locate the defective parts. If a special matrix test set is not furnished for the computer, the assembly must be tested with general-purpose test equipment to find the defective parts. Test equipment such as test-pattern generators, trouble indicators, and memory units are furnished with some computers as built-in equipment. Test-pattern generators can be used for signal-tracing applications or to provide a simulated signal for maintenance purposes. Visual and audible alarms are provided to indicate errors in calculations or the existence of conditions that might cause damage to the computer. LED indicators are often connected to flip-flop stages to indicate the status of registers. By observing the indicators and alarms, the technician can often decide what action should be taken to correct a malfunction.

3-18.14.1 Memory Units

Trouble within a memory unit can usually be located by means of diagnostic maintenance programs. An example of a troubleshooting procedure is as follows:

1. Apply power to the computer and prepare the computer for test mode operation.
2. Check the supply voltages to the memory units; if they are not set properly, adjust them to their correct values.
3. Determine which memory unit is faulty by analyzing the computer error indication, program error, or error printout.
4. Run a diagnostic maintenance program. When an error printout occurs, check the computer condition indicators to determine in which routine the failure occurred.
5. Set the maintenance memory unit to repeat the routine which produced the first error. Run this routine to make sure that the error will occur when the routine is run separately. If the test routine does not produce the error when run separately, set the maintenance program to the previous routing so that program operation will switch to the first error printout.
6. If the trouble is intermittent, use the marginal-check version of the program to apply prescribed margins to the memory unit. If the original memory failure is detected under marginal-check conditions, verify the failure margin by applying a manually controlled voltage excursion to the tested marginal unit while running the routine.

7. Check the memory unit for the remaining errors that were detected by the maintenance program. This is accomplished in the same manner as in steps 5 and 6.

8. Analyze the error printouts to determine whether the error is a bit failure or a selection failure. If the error is a bit failure, determine which bit is causing the failure, and check the associated sense amplifier or digit plane driver. If the error is a selection failure, further analyze the printout to determine which selection line or lines are causing the failure, and check the associated driver stage.

9. Refer to the equipment technical manual troubleshooting guides to determine which circuits should be tested. Rerun the maintenance routine and observe the action of the suspected memory circuits.

3-18.14.2 Magnetic Drum Units

The general procedure for troubleshooting magnetic drum units of a computer is to verify the existence of a failure and determine its source and cause. Maintenance programs are the chief means of determining the particular areas in which malfunctions have occurred. However, in equipment where maintenance programs are not effective or not available, manual testing must be performed at the computer maintenance console or directly at the magnetic drum units. Marginal-checking procedures are applicable for troubleshooting intermittent malfunctions.

3-18.14.3 Maintenance Programs

If the computer has this feature, the technicians can run a diagnostic program whenever a drum unit is not functioning correctly. A drum unit maintenance program can be used to check the fields, status controls, disconnect counter, field switch, and information circuitry. To make a test, place the drum unit in the computer-test mode and initiate the maintenance program which is stored on cards or magnetic tape. The program first clears the drum fields, then writes a test pattern. The computer reads and checks information transfer. Errors are indicated if the program halts before the program is completed or if incorrect information is printed out. Maintenance tables are normally furnished with the equipment technical manuals as aids to determine the source of trouble from the results of a maintenance check.

3-18.14.4 Write-Read Check

The bits of information that are recorded on a magnetic drum by a test pattern may be displayed by lamps on a check register in the drum unit. These indicators are used in conjunction with tables in the equipment technical manual to help the technician to analyze the cause and location of a malfunction. Such tables are arranged according to the types of bit errors which may be encountered: i.e., missing bits common to two or more fields on one drum, a missing bit in one field, or two adjacent bits in error in one field. When one of these conditions is present, the corresponding table should be used to locate the trouble within a defective circuit or device. An arrangement of computer circuits for testing a magnetic drum is shown in Figure 3-607. Several test patterns should be used individually or in combination to eliminate the possibility that an error may go undetected or be misinterpreted. Examples of such test patterns are as follows:

1. Straight "0" bit pattern.
2. Straight "1" bit pattern.
3. "0" and "1" bits alternating in the first register and complementing in the following register: e.g., first register, 010101, second register, 101010, etc.
4. All "1" bits in the first register, all "0" bits in the second register, etc.

A test pattern applies a "1" or "0" bit to each write head for the purpose of receiving identical information from corresponding read head. The use of one test pattern may not prove sufficient in all cases to thoroughly identify a trouble. One pattern or a combination of two or more patterns may be required to locate the trouble which causes bit errors. For example, a straight "0" bit pattern, which is applied to the drum write heads, may appear as all "0" bits on the check register, and yet one of the read heads may be defective and produce only "0" bits. The use of a straight "1" bit pattern, in this case, will reveal the particular read head which is at fault. Further checks must then be made to determine whether the failure is caused by a defective head, a misadjusted head, or the read amplifier stages that are connected to the head. If the read head and corresponding amplifier stages are operating correctly, test the corresponding write head and its associated circuitry.

3-18.14.5 Runout Test

Excessive bearing wear, unbalance, and ovalness of a drum rotor can be checked by means of a runout test. To perform this test, connect an oscilloscope to a read head as shown in Figure 3-608. Operate the drum motor at its normal speed, and apply an all "1" bit test pattern to the write head. For the drum to be considered satisfactory, any change in the amplitude of the test pattern bits, V_{MAX}

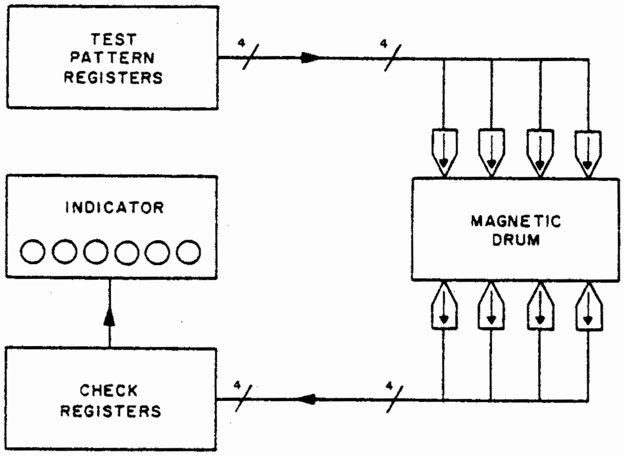


Figure 3-607. Magnetic Drum Write-Read Test, Block Diagram

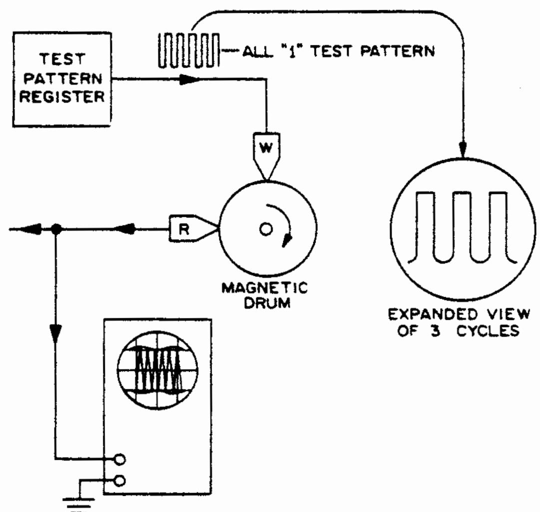


Figure 3-608. Magnetic Drum Runout Test, Block Diagram

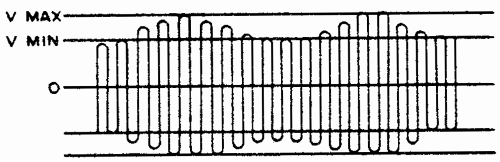


Figure 3-609. Runout Test Waveform

minus V_{MIN}, must be less than 20 percent of the maximum bit amplitude. This measurement should be made from read heads located at each end of the drum, and also from a read head at the approximate center of the drum. The read head waveform should resemble runout test waveform Figure 3-609.

3-18.14.6 Noise Test

Noise being generated by a magnetic drum can be determined by using an oscilloscope in the same manner as in the runout test. For this test, disconnect the test-pattern generator from the write head, and erase all signals from the drum. Then start the drum rotor and observe the display on the oscilloscope. The oscilloscope display may resemble the waveform shown in Figure 3-610. Any voltage present is a result of residual electrical noise or spurious magnetization of the drum rotor. This test must be performed for each recording track on the drum rotor. Refer to the equipment technical manual to determine the noise spike amplitude that can be tolerated and the number of tracks that can be noisy before the drum rotor must be replaced. For the example shown in Figure 3-610, 5 millivolts is considered to be the maximum noise pulse amplitude that can be permitted.

3-18.14.7 Crosstalk Test

A crosstalk test determines whether each channel processes data free of interference from other channels. To perform this test, erase all information from the track to be checked, and connect the read-head output to the oscilloscope. Then operate the drum and apply test patterns to all channels except the channel being tested. Use a calibrated oscilloscope to measure the output noise level due to crosstalk.

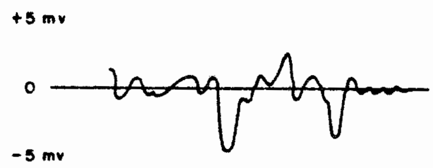


Figure 3-610. Noise Test Waveform

This test must also be performed for each track on the drum. The noise level and number of noisy channels that can be tolerated will be listed in the equipment technical manual.

3-18.14.8 Magnetic Tape Units

Troubleshooting of computer magnetic tape units can also be accomplished by means of diagnostic maintenance programs. Such a program may use the control units, memory units, line printer, and marginal check equipment to test the tape units. Such a program may be composed of three test routines, as follows:

1. Routine 1 uses four different test patterns to check the timing of the write, read, backspace, and rewind instructions, as well as their operation. This routine also checks the write and read end-of-file circuits, the read zero word instruction, the disconnect circuits for the conditions either read more or read less than the number of words contained in a record, and all information transfer paths involved.
2. Routine 2 uses a random number pattern to check the timing of the read, write, and rewind instructions and their operation. This test is accomplished by checking the rewind operation, the stepping of the tape clock, the tape speed, and the effects of a number of short random-number records.
3. Routine 3 is used to check the file-protect-on circuitry. The routines to be run will depend on the error indication. These three maintenance program routines can be run in five different modes, as follows:

a. Mode 1 uses routine 1 only
b. Mode 2 uses routine 2 only
c. Mode 3 automatically applies excursions for prescribed margins or margins to failure, using either routine 1 or routine 2 depending on which marginal-check word is being used.
d. Mode 4 selects the tape drives available and uses routines 1 and 2 on each tape drive automatically. This mode is normally used for preventive maintenance and not for troubleshooting.
e. Mode 5 uses routine 3 only.

When these routines are run, all errors will be indicated by a printout, which will include the type of error, the trouble location, and possible plug-in units that caused the failure. In modes 1, 2, and 5, errors will also be indicated by a programmed halt, with the computer accumulator designating when the error occurred. In mode 3, the marginal check words will also be printed, designating the selected line and excursion that caused the failure. In mode 4, the program will print out errors, but will halt only after all tape drives have been tested.

3-18.15 REPAIR PROCEDURES

When the cause of a computer failure has been found by troubleshooting procedures, corrective action, such as adjustment, alignment, tuning, cleaning, and replacement of parts will normally be performed. The replacement of computer parts involves the same basic procedures as are used for radio, radar, and carrier equipment. When replacing parts with many wire connections, such as switches, relays, connectors, and transformers, each wire should be tagged and identified. The wires may then be detached from the defective part and the replacement made without inducing an error in the wiring. Many computer circuits are contained in modules, which may be either plug-in units or soldered to the equipment. The extent of repair for module units will depend on what is considered to be economically practicable. The units may be repaired in filed maintenance shops or sent to depots. Appropriate soldering techniques must be used when repairing computers that use semiconductors, miniature circuits parts, and printed circuits to avoid overheating the parts. The voltages used in ohmmeter measurements of semiconductor circuits must be limited to safe values.