SECTION 5

ANTENNA AND TRANSMISSION LINE MEASUREMENTS

5-1 DIRECT RF POWER MEASUREMENT

Direct measurement of the effects of RF power is possible up to a frequency range of 60 megahertz. An example is the measurement of the load voltage and current in dc applications. The basic procedure can be modified by using a thermocouple ammeter at lower radio frequencies. Figure 5-1 illustrates the circuit arrangement required to use the unity power factor of the resonant antenna in conjunction with the formula $P = I^2R$ to obtain a direct power measurement. The meter should be calibrated to read the square of the RF antenna current. However, an accurate determination of the effective antenna resistance by the variation, substitution, or bridge method must precede the actual current measurement.

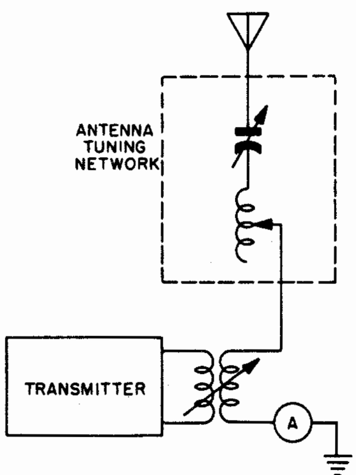

Figure 5-1. Circuit for Measurement of Antenna Input Power

5-1.1 BASIC VARIATION METHOD

The basic variation method of measuring the antenna resistance requires an RF generator covering the desired frequency range, with an output power rating of approximately 50 watts; a wavemeter accurate to 0.25 percent, as required by the FCC; an alternating type of ammeter to 2.0 percent; the proper range value of tunable capacitor and inductor; and a decade-type resistor. Figure 5-2 illustrates the circuit arrangement required to perform a series of

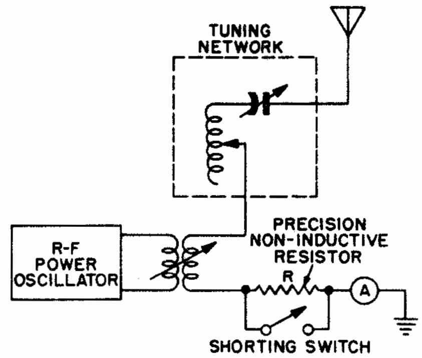

Figure 5-2. Variation Method of Measuring Antenna Resistance

measurements at different frequencies, centered on the stipulated transmission frequency. A graph of resistance versus frequency should be plotted, with the frequency on the X axis. The resistance obtained at the assigned frequency is then the value to be used in the power calculation. The tuning inductor and capacitor are used to vary the resonant frequency about the fundamental frequency, and the wavemeter indications are plotted. The first resistance measurement should be made at the natural frequency of the antenna system by connecting the antenna to ground or to a counterpoise through the coupling coil introducing the applied signal. The RF oscillator is then tuned until the deflection of the antenna milliammeter is at maximum. The deflection should occur gradually as resonance is approached from either direction. An abrupt dip, followed by a quick return to the original position, indicates overcoupling between the driver and antenna stage. Record the current reading at resonance, then, after connecting a known resistance into the circuit, again note the current reading. The following formula will then give the antenna resistance in ohms.

$$R_a = \frac{R_s I_s}{(I_a - I_s)}$$

where:

R_a = antenna resistance
I_s = current measured with standard resistance in circuit
I_a = current measured without standard resistance in circuit
R_s = value of standard resistance

The procedure described above permits evaluating the resistance at the fundamental frequency. The shielded tuning network is then to be connected and the antenna resistance at the frequency of transmission is determined. The tuning network is brought into resonance and the antenna current is again read. With the standard resistor in the circuit, this procedure is to be repeated for many additional frequencies, and a resistance-versus-frequency chart is plotted the same as before. The value of the measured antenna resistance should agree with the calculated value.

5-1.2 TWO-RESISTOR VARIATION METHOD

Figure 5-3 illustrates a method used to disclose the existence of stray capacitive paths to ground. This particular circuit makes use of an L-C tuning network, whereas a calibrated capacitor is normally used alone. With both standard resistors shorted out, the tuning network of the antenna network should be adjusted to resonance, as indicated by maximum antenna current. The oscillator output is then adjusted for a convenient meter deflection and left untouched for the remainder of the procedure. The antenna current is recorded with the resistors out of the circuit. After inserting a known value of R_1 into the circuit, the antenna current is again measured and antenna resistance is computed using the formula provided in the previous paragraph. R_1 resistance is then shorted out and a known value of R_2 is inserted. Another resistance measurement is made and compared with the other method of obtaining antenna resistance.

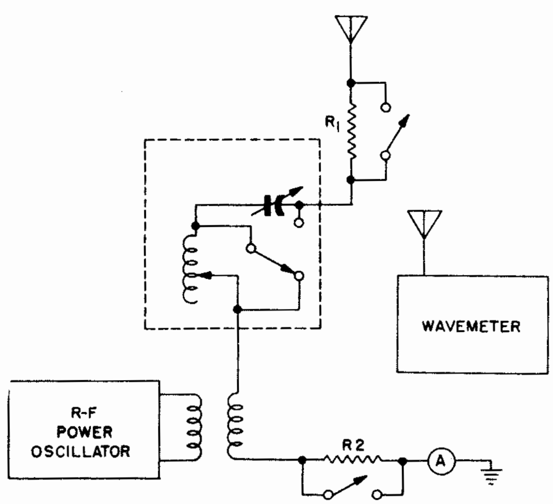

Figure 5-3. Two-Resistor Variation Method of Measuring Antenna Resistance

The two values will agree only if no stray capacitance exists between the measuring circuit and ground. Of the two values obtained, the resistance measured with the use of R_1 is the more accurate, and should therefore be used in the final calculation.

5-1.3 SUBSTITUTION METHOD

Replacement of an antenna by its electrical equivalent is called the substitution method of determining antenna resistance. Resistors, coils, and/or capacitors may, either singly or in combination, provide the proper load. Figure 5-4 illustrates a substitution method that uses capacitance. With the resonant frequency known, the switch is set to position 1 and the antenna circuit current is measured and the reading is recorded. The switch is then set to position 2. If an inductor (such as L in Figure 5-4) was used to resonate the antenna, a resistor and a tuning capacitor should be used, shown as R and C, respectively. However, if a capacitor was used to obtain antenna resonance, then a precision inductor with a resistor must be used. Connect the antenna tuning element to points a and b, as indicated in Figure 5-4, and tune capacitor C (or inductor L, if used) until this second circuit is in resonance, as indicated by maximum milliammeter deflection. Resistance R is then used until the meter reading is the same as when the antenna was connected. The resistance of R is now equal to the antenna resistance, and the reactance of the tuning capacitor is equal to the antenna circuit reactance at the resonant frequency. Using the wavemeter, the oscillator is then adjusted to various other frequencies about the resonance point and the foregoing steps are repeated at each frequency.

5-1.4 RF BRIDGE METHOD

The most rapid and accurate method of determining antenna impedance employs an RF bridge. Figure 5-5 illustrates a type of RF bridge, with a well-shielded signal generator and a receiver serving as an indicating device. In Figure 5-5, capacitor C_1 is attached to a calibrated dial that indicates the unknown resistance. Shunting C_1 is C_2, which is adjusted for initial balance when the terminals to which the unknown impedance is connected are shorted. (These terminals will be referred to in the discussion as the "unknown" terminals.) Because of the inclusion of C_2, the initial setting of C_1 can be zero, and its dial will specify the unknown resistance directly in ohms. Capacitor C_3 specifies the reactance of the unknown impedance. Its capacitance is decreased when measuring inductive reactance, and increased when measuring capacitive reactance. However, its dial is calibrated so

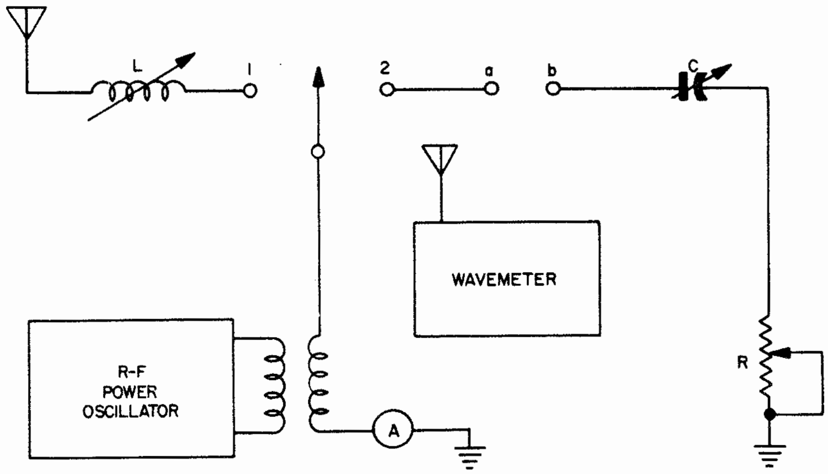

Figure 5-4. Substitution Method of Measuring Antenna Resistance

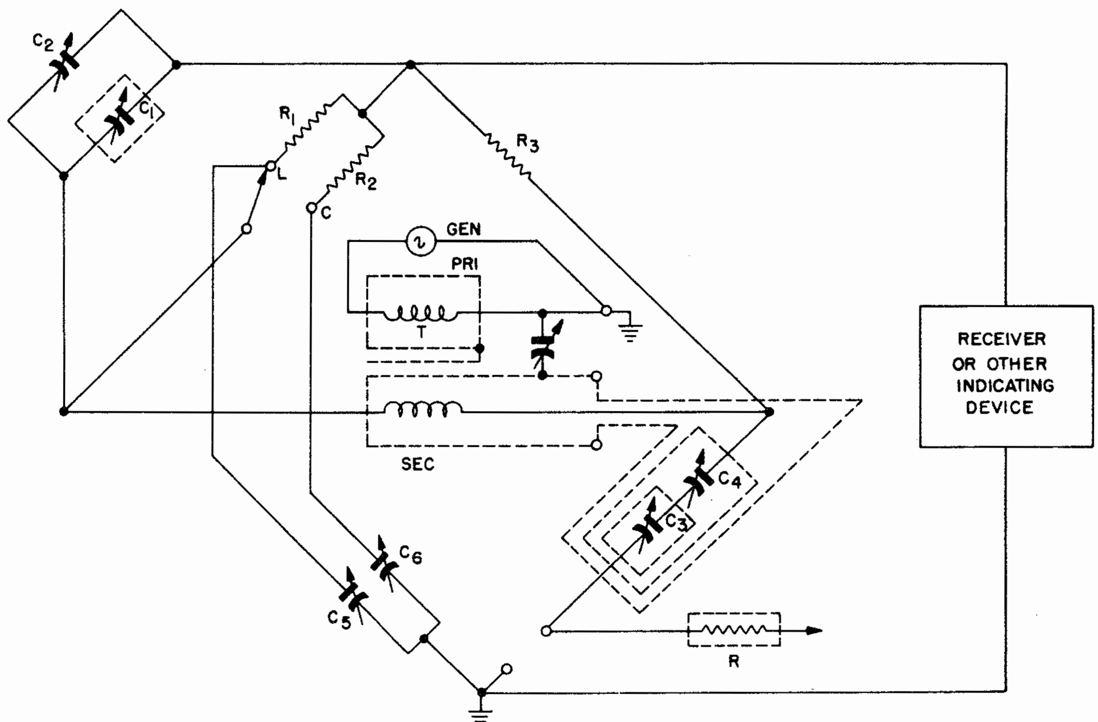

Figure 5-5. Circuit of Typical RF Bridge

that the reading increases (from its original balance setting) with a measurement of inductive reactance, and decreases with a measurement of capacitive reactance. When measuring inductive reactance, the switch in the upper left-hand arm is switched to the L position, and bridge balance is accomplished with C_3 set at zero. A second balance is obtained after the connection of the unknown impedance. The dial setting, divided by the frequency in megahertz, gives the inductive reactance directly in ohms. When a capacitive reactance is being measured, set the above switch to C and set C_3 to maximum. The final dial reading is then subtracted from the maximum dial indication and divided by the frequency in megahertz to obtain ohms directly. As shown in Figure 5-5, extensive shielding is important, as it protects the inherent balancing property of the bridge. An additional resistance, R, placed in a special test lead outside the bridge because of capacitance effects, is connected to the unknown impedance arm, permitting C_1 to be set at zero ohms during preliminary balancing.

5-1.4.1 RF Balanced Bridge

The TS 147 Radar Test Set represents an RF balanced bridge method for measuring RF power. A simplified drawing of such a bridge is shown in Figure 5-6. If a 1 milliwatt current is applied to input 'A' and a similar current is applied to input 'B', the meter will then read zero. Since a meter with a mid-scale zero indication is used, an off-balance to either side can therefore be noted. The variable attenuator is adjusted to achieve a meter reading of zero. The attenuator dial reading is thus indicating 'A' input in +dB above 1 milliwatt.

5-1.5 RF POWER OUTPUT METER METHOD

An RF power output meter is a compact device that may be used to measure power if only approximate accuracy is permissible. On low-power measurements, the technician can dissipate the total power within a resistor and use a percentage of the resulting voltage drop to operate a meter calibrated in watts. Figure 5-7 illustrates an RF power meter circuit, with the load resistor connected across the input terminals to dissipate the applied load. Carbon piles, consisting of carbon disks stacked contiguously on a suitable rod, are sometimes used for this resistor. Low power can be applied continuously to the meter, whereas higher power should be applied only briefly. The upper frequency limit is determined by the capacitive voltage divider shown in the figure. This is because a frequency is reached where the reactance across the loading resistor causes a mismatch in the termination of the transmission line. The standing-wave ratio is then so great that the meter readings are inaccurate. For a further explanation of the effects of standing waves consult the following paragraph of 5-5.

5-1.5.1 Energy Source

For general purpose use it is desirable that the output meter be capable of supplying a wide range of load impedances and of measuring a wide range of power. Figure 5-8 illustrates how the impedance which the output meter offers to the source of energy is controlled by varying the transformer taps, while the power range is varied by the attenuator placed between the rectifier instrument and the transformer secondary. To compensate for transformer loss variations at different taps, series resistors are inserted to provide a percentage of the total energy to the attenuator.

5-1.5.2 Dummy Loads

In the measurement of the output of oscillators and transformers, a dummy load of the same impedance and of equivalent power dissipating

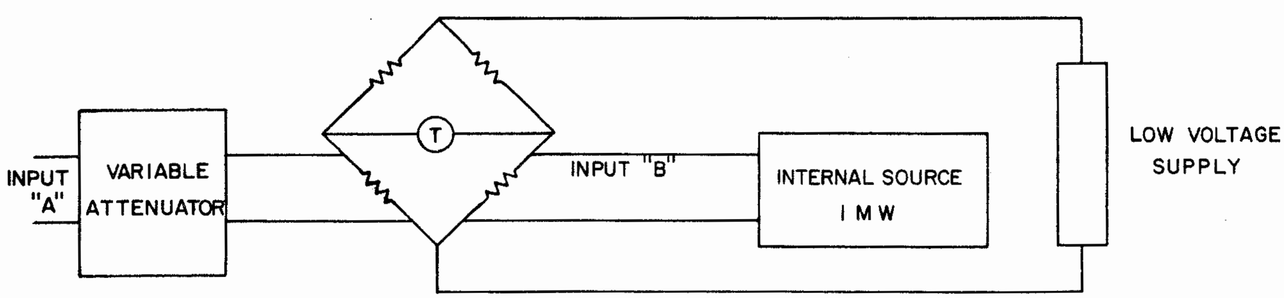

Figure 5-6. Typical RF Balanced Bridge

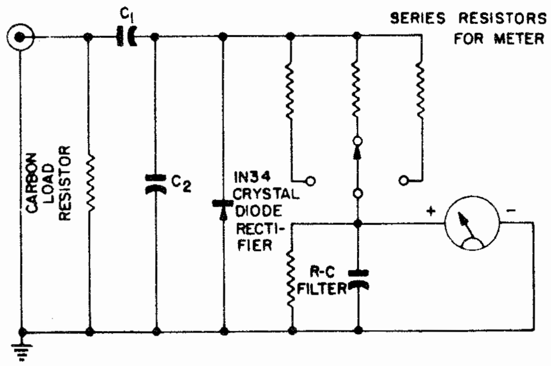

Figure 5-7. Circuit of Typical RF Power Meter

capability may be substituted as the load to be used in the circuit. If the resistance of the dummy load can be measured directly, then the determination of power is simple. Figure 5-9 shows the dummy load measuring method being actually applied. An approximate method of measuring the available power of an RF transmitter involves converting the power to a dc voltage. This is done by means of a measuring circuit consisting of a rectifier-filter network connected in tandem with a load resistance. The latter is large as compared with the internal resistance of the rectifier. This dc power may be measured by conventional means. An adjustable transformer may be used to couple the RF source to the measuring circuit for maximum power output. This method assumes that both the internal impedance of the source and the input impedance of the measuring circuit are resistive. Losses in the transformer and rectifier must also be considered.

5-1.5.3 Dummy Load Limitations

The dummy load method of RF power measurement cannot be used successfully on pulsed RF oscillators or pulsed transformers. Additional losses of connecting cables and connectors must also be considered, as they attenuate the RF before it is applied to the Power Measuring Device. Some power meters are designed to measure pulsed power, similar to those in the circuits previously described. They are narrow-band types, however, and can only be used with the equipment for which they were specifically

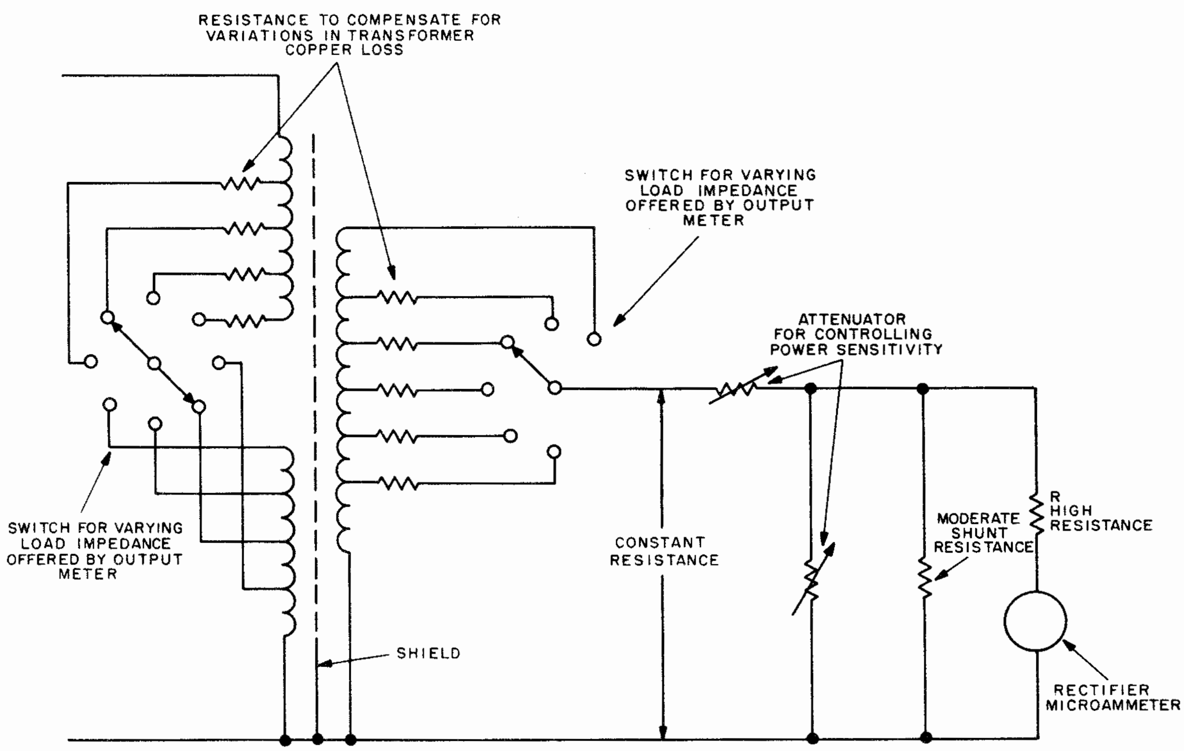

Figure 5-8. Circuit Diagram of Power-Level Meter Capable of Covering a Wide Power Range, and of Offering a Variety of Load Impedances

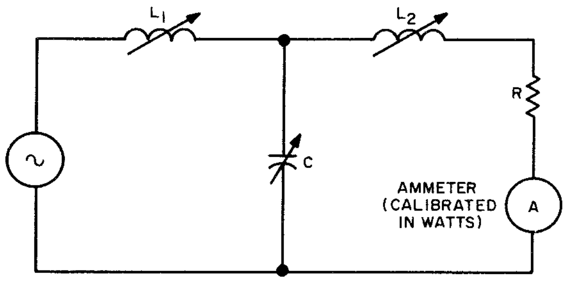

Figure 5-9. Use of a Variable Dummy Load to Measure Transmitter Power by Dissipation in a Fixed Resistance

designed and constructed. Accuracy is thus greatly reduced and considerable driving power is necessary. These RF power metering systems are therefore used more for relative indications than precise measurement of the actual power.

5-2. INDIRECT RF POWER MEASUREMENT

5-2.1 GENERAL

As the frequency of RF energy becomes high, direct methods of measuring RF power are not effective. An indirect method must therefore be used, capable of converting the RF power under test to some other form of energy (such as heat or light) for ease of evaluation. The secondary energy produced must be related to the RF energy, and the associated time interval to determine the RF power must also be taken into account.

5-2.2 LAMP LOAD METHOD

This method of measuring RF power involves the use of a bank of incandescent lamps as a noninductive dummy antenna for the transmitter. A photographic illumination meter is arranged to measure the resulting brilliance of the lamp bank. The antenna connections are then removed from the lamp bank and a variable dc (or ac) power source is used to light the bank of lamps to the same brilliance. An ammeter-voltmeter (or wattmeter) is then connected to determine the required power necessary to bring the dummy antenna indication to the same brilliance. This power is equal to the transmitter output. The lamp bank impedance should match the transmitter output impedance. As an example, consider the termination requirement of a 50-ohm line and conventional 28-volt, 4-watt lamps for the load. Three lamps could be connected in parallel to provide 65 ohms at 12 watts for a termination (the hot resistance of each lamp is 196 ohms). The graph shown in Figure 5-10 illustrates that the light intensity from incandescent lamps can be correlated with power. If the eye could accurately distinguish between various light intensities, this graph would provide a direct means of power evaluation. Test instruments could also be employed to directly measure the light intensity, and the graph consulted to obtain the equivalent power. However, this method is not normally employed.

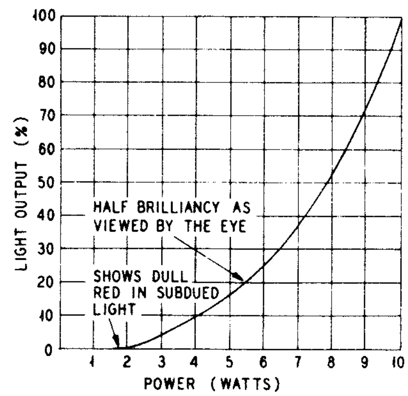

Figure 5-10. Typical Graph of Light Output Versus Power for an Incandescent Lamp

5-2.2.1 Alternate Methods

As an alternate method, a pair of identical lamp combinations, as shown in Figure 5-11, are used. One lamp load is energized by the RF source, and the other by a 28-volt, dc power source. Two adjacent lamp banks, adjusted by a potentiometer across the dc source, will permit judging when the intensities

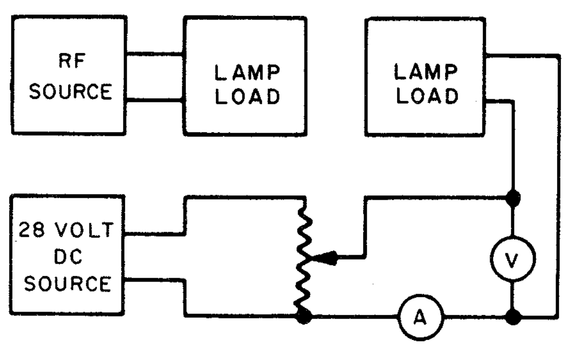

Figure 5-11. Test Setup for Lamp-Load Method of Indirect Power Measurement

are equal. When the banks are equal in brilliance, ascertain the power by reading the voltmeter and ammeter. Figure 5-12 shows a method capable of greater accuracy. In this method, the lamp load is matched to an RF transmission line, with a tunable matching network to obtain maximum lamp brilliance. As shown by the illustration, the light is directed upon a photoelectric cell which provides a current to a meter calibrated directly in watts. If multiple ranges are required, lamp banks of suitable ratings must be provided. The two primary considerations of the lamp-load method of power measurement is that the wattage rating of the lamp load must be greater than the power being measured, and the net load resistance should match the RF source impedance.

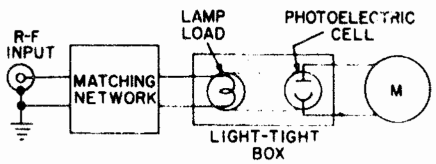

Figure 5-12. Diagrammatic Setup for Photoelectric Type of Lamp-Load Power Test

5-2.3 BOLOMETER METHODS

Since none of the methods described previously are effective in the UHF range of the frequency spectrum, a test equipment employing a bolometer has become the standard instrument for power measurements in this range. At lower frequencies simpler methods are used, even though there is no frequency restriction on the use of bolometers. There are two primary types of bolometers: the barretter, and the thermistor. Both are loading devices which undergo resistance changes when variations in dissipated power occur. The barretter increases in resistance with power increases, while the thermistor decreases in resistance as the power increases. With either device, the resistance must be measured both before and after the application of power. The bolometer is initially balanced with either low-frequency power or dc bias power. Then, when RF power is applied, the low-frequency power is reduced until the bridge is again balanced. The RF power being measured is then equated to the bias power that was reduced, if the resistance of other circuit elements was unaffected by the power dissipated in them. Another method involves calibrating the bridge imbalance with known power. An RF power source can be used whose value is separately determined by substitution of known low-frequency power in a balanced bridge circuit, or (alternatively) known low-frequency power of a different type from that used to operate the bridge. The imbalance produced by the unknown RF power is then converted into a power reading with the aid of the previously obtained calibration. In other words, there is a condition of balance when no RF power is applied; but when power is applied, there is a condition of imbalance owing to the bolometer resistance change. It is this change of resistance that is to be calibrated into power units. Basically, the barretter must be matched to the RF line after the application of power. At low-power levels, the resistance-versus-power curve of a barretter is characterized by a square-law relationship.

5-2.3.1 Thermistors

The thermistor must be matched in the same manner as the barretter. However, the thermistor may be operated at higher temperatures than the barretter; it is more rugged mechanically and electrically; it can withstand larger pulse energies; and it exhibits negligible pulsed-power measurement errors. The disadvantage of a thermistor is that excessive power applied changes the thermistor resistance enough that an RF mismatch is produced and could cause a burnout. The bolometer, in turn, is limited in its use as a measuring device because of the low power it is capable of dissipating. However, with the addition of attenuators, the power-measuring capability will increase, with the power capabilities of the attenuators then being the limiting factor.

5-3 DIRECTIONAL COUPLERS

A device which extracts a fraction of the power associated with only one of the traveling waves in a line, while being unaffected by the wave traveling in the opposite direction, is called a directional coupler. This type of device can be employed to measure the incident or reflected line power, display the reflection coefficient of a load, monitor or adjust the load, compare power meters of differing power ranges, and measure net power delivered to an arbitrary load. The directional coupler provides a much more accurate power indication of a mismatched load than will any other nondirectional voltage-or current-actuated device. Other power-measuring devices respond to the vector sum of the incident and reflected power, whereas the directional coupler responds only to the selected component. A directional coupler consists of a main or primary transmission line, carrying the power to be sampled, coupled to an auxiliary line through a suitable structure (such as a loop, probe, aperture, or slot). A fraction of the power associated with a wave traveling in the positive direction in the main line is coupled to the auxiliary line, whence it proceeds to the output terminals. A similar fraction of the power associated with the nonpreferred wave in the main line is also coupled into the auxiliary line, and proceeds to the matched termination at the opposite end for absorption. The auxiliary line termination must be well matched, otherwise the amplitude of the reflected wave induced in the main line by the mismatched coupler may be larger than that of the original reflected wave. When coupled, it may reflect from the auxiliary line termination at an amplitude comparable with the desired component amplitude at the output, causing a large measurement error. Figure 5-13 is a schematic of a waveguide directional coupler. Two of these couplers can be connected to measure net power, provided one coupler is connected to sample incident power and the other to termination is well matched. This is necessary because the amplitude of the reflected wave induced in the main line by the mismatched coupler may be larger than that of the original reflected wave. When coupled, it may reflect from the auxiliary line termination at an amplitude comparable with the desired component amplitude at the output, causing a large measurement error. The difference in the power outputs of the two couplers gives a measure of the load power. Directional coupling can also be achieved by using an array of two or more coupling elements, related such that the induced voltage waves in the auxiliary line interfere constructively for the desired direction of propagation, and destructively for the undesired direction. Figure 5-14 illustrates two rectangular waveguides sharing a common narrow wall, with two coupling holes a quarter-guide-wavelength apart. The forward-traveling wave (toward load A in the main line) induces a forward and a backward wave in the auxiliary line at each of the two coupling holes. Corresponding induced waves at these holes have nearly equal amplitudes. The induced forward-waves are in phase and reinforce each other at the output of the auxiliary line at B. The induced backward waves, however, are 180 degrees out of phase and therefore will cancel. Similarly, a backward-traveling wave in the main line, caused by a reflection from load A, induces backward waves in the auxiliary line, which add in phase and are absorbed in the dummy load. However, it induces forward waves 180 degrees out of phase, which therefore cancel before reaching the output. As the directivity of such a coupler is frequency-sensitive, wideband directivity and flat coupling can be achieved. Directive properties can also be achieved with a single aperture or element, provided it is suitably placed in the common wall of two waveguides or a transmission line. A simple type of directional coupler, together with equivalent circuits showing the action of the electric and magnetic induction, is illustrated in Figure 5-15. If both the A and B ends are terminated in well-matched thermocouple elements, the opposed output can be read with a microammeter, or a wide-band type of wattmeter can be employed to measure power in an arbitrary load.

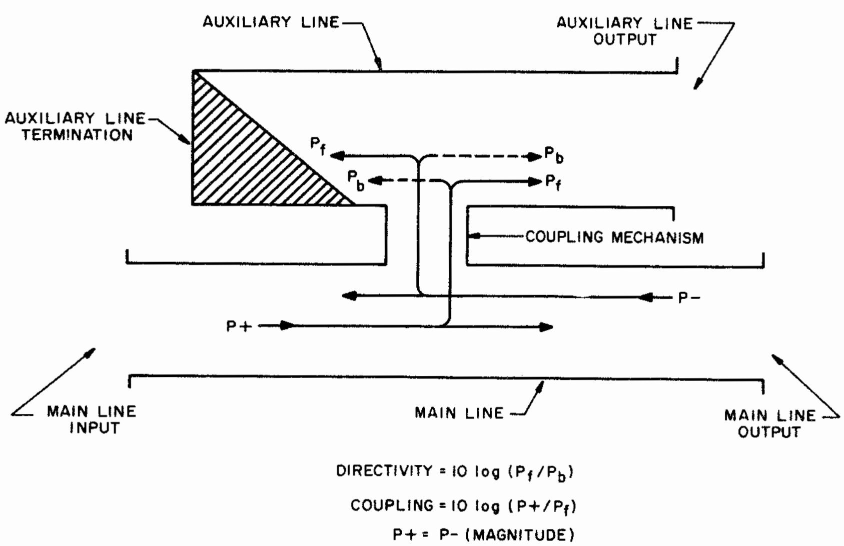

Figure 5-13. Schematic of Waveguide Directional Coupler

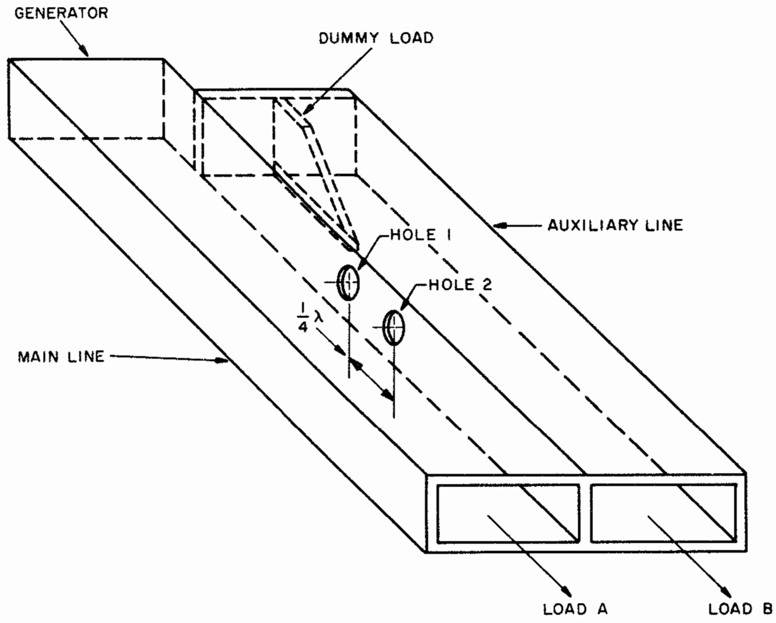

Figure 5-14. Two-Hole Waveguide Directional Coupler

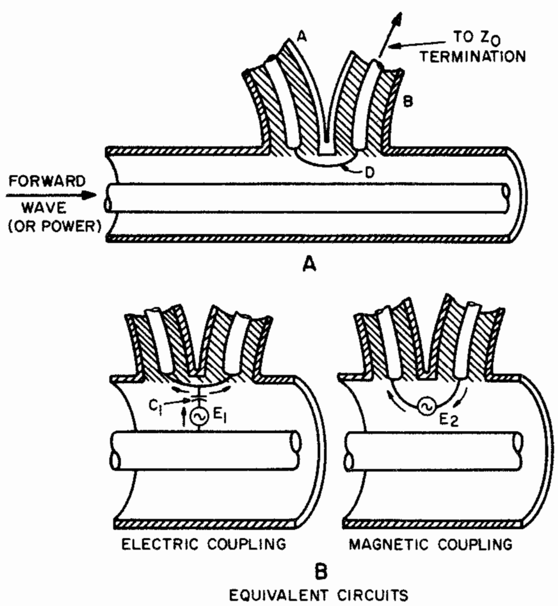

Figure 5-15. A Simple Type of Directional Coupler Together With Equivalent Circuits Showing the Action of the Electric and Magnetic Induction Coupling

5-4 PEAK POWER MEASUREMENT

5-4.1 NOTCH WATTMETER METHOD

Figure 5-16 illustrates the method used to produce a "notch" by interrupting a CW oscillator in time to coincide with the RF pulses being measured. Both the notch and the RF pulse envelope appearing within the notch may be displayed if crystal detection is employed, followed by application to an oscilloscope for viewing purposes. The CW oscillator power level must be adjusted until the notch depth equals the

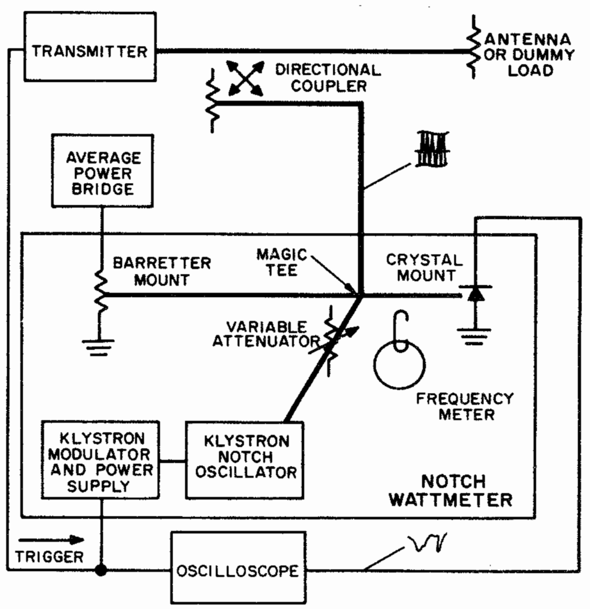

Figure 5-16. Notch Measurement Technique

maximum height of the RF pulse, and the bolometer bridge is then used to obtain the CW measurement. The CW power measurement is thus equal to the peak pulse power. The oscillator and pulse frequencies should be the same in order to avoid errors due to the frequency sensitivity of the crystal. The "magic tee" should be well-matched at all arms to ensure isolation of the CW and pulsed power channels and an equal power split between the crystal and the barretter mount.

5-4.1.1 Alternate Method

Another method of measuring peak power in terms of CW involves the simultaneous mixing of two powers at a microwave receiver or crystal detector input, followed by a wideband oscilloscope for viewing RF pulses. Figure 5-17 shows a block diagram of the necessary connections required to measure peak power by mixing CW and pulsed signals. When only CW is applied to the receiver, no change occurs in the output level, and the oscilloscope baseline at the receiver output terminals remains unchanged. However, if simultaneously both CW and pulse power are applied at the same carrier frequency to the receiver mixer, the video pulse appearing on the oscilloscope may be positive, negative, or zero, depending on the relative phase and amplitude of the two signals. In practice, since the two signals cannot remain coherent, the video pulse on the oscilloscope screen will be composed of a series of horizontal lines representing arbitrary phase combinations of the signals. Figure 5-18(A) illustrates the video pulse when both the CW and pulse voltages are equal; Figure 5-18(B) shows the video pulse when the CW power level is adjusted such that the peak voltage of the pulse is twice that of the CW signal, and the peak pulse power is four times the CW power. This last adjustment can be made by bringing the bottom of the video pulse into coincidence with the oscilloscope baseline without measuring the pulse heights.

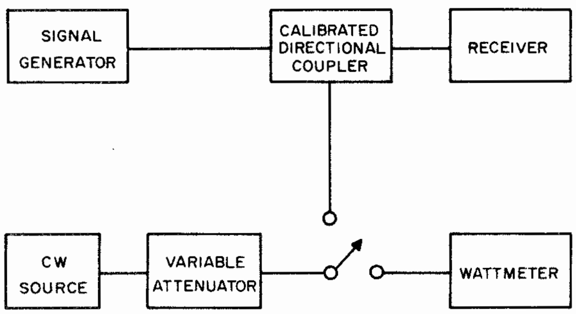

Figure 5-17. Block Diagram of Peak Power Measurement by Mixing CW and Pulsed Signals

5-4.2 HETERODYNE METHOD

The CW signal can be heterodyned with the pulse signal to produce a beat frequency of 1 or 2 megahertz for display within the RF pulse envelope. Figure 5-19 shows how the CW level is adjusted to cause the beat voltage minima to coincide with the oscilloscope base line. The peak pulse power is again four times the CW power.

5-4.3 INTEGRATION-DIFFERENTIATION METHOD

As shown in Figures 5-20 and 5-21, a barretter can be used in an integration-differentiation

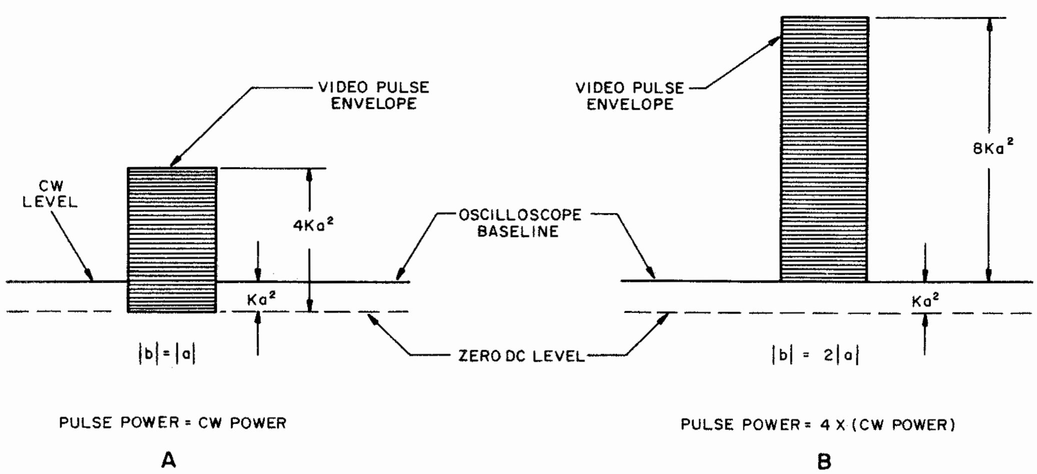

Figure 5-18. Effect of Mixing CW Pulsed RF Signals of the Same Frequency

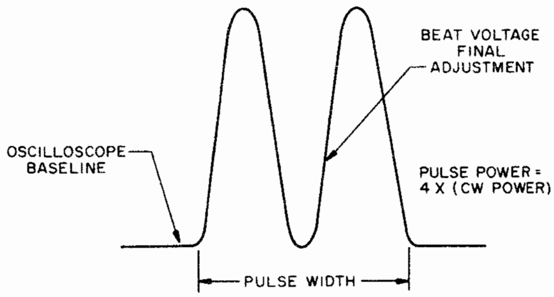

Figure 5-19. Effect of Heterodyning CW and Pulsed RF Signals

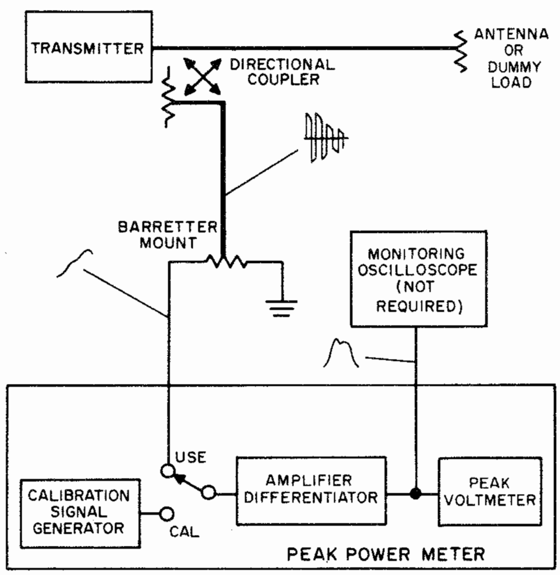

Figure 5-20. Barretter Integration-Differentiation Technique

technique to obtain the pulse envelope necessary for the measurement of peak RF power. This method relies on the reproducibility of the characteristics of the barretter, and incorporates the pulse energy sensitivity and thermal time-constant into the calibrating equipment used to standardize the output meter. The pulse signal must be integrated by the barretter, and then amplified and differentiated to recover the original pulse shape. The peak amplitude of the restored pulse can be measured with a peak voltmeter, which could be made into a direct-reading instrument with the aid of a calibration signal generator. The signal generator provides standard pulses of known level, and applies them to an integrating network divider which provides an input to the amplifier-differentiator peak-voltmeter circuit equivalent to that furnished by the barretter

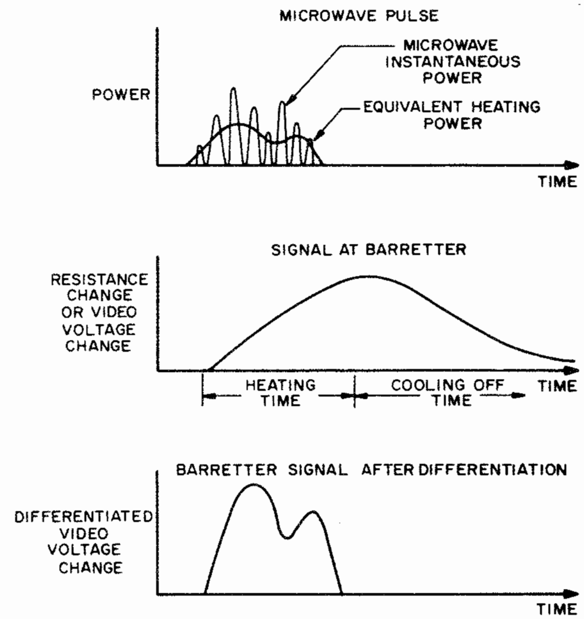

Figure 5-21. Sequence and Appearance of Signals in the Barretter Integration-Differentiation Technique

when subjected to an RF pulse of standard peak power. Figure 5-21 illustrates the signal transformation between input-to-barretter and input-to-peak voltmeter. A broadband amplifier with a response time of 0.03-microsecond and about 90-dB gain must be utilized to deal with the fast rise-time pulses and minimum peak power of 10 milliwatts. The amplifier must also have a range of 100 to 500 milliwatts, depending on the pulse width.

5-4.3.1 Accuracy

The accuracy of the above method is about the same as the accuracy of the previously discussed methods. However, it is substantially faster once the equipment has been set up. The estimated accuracy is about 8 percent. There is a burnout limitation of 30 watts maximum power with a 0.1-microsecond pulse width, and 300 milliwatts with a 10-microsecond pulse width. The pulse duration must not exceed about 15 microseconds, because the short time-constant of the barretter does not permit accurate integration of wider pulses.

5-4.4 AVERAGE TO PEAK POWER CONVERSION

An easy and relatively accurate method of measuring peak power of a radar with certain factors known (PRF, pulse width) is described in Section 3 under "Peak Power Measurement". The method employs an average-power measuring device, such as the AN/USM-177 power meter.

5-4.5 LOW PEAK POWER MEASUREMENT

In many cases, low-power pulsed RF can be measured with the aid of an RF detector, a suitable dummy load, and an oscilloscope. Suitable detectors are readily available through the supply system, or can be constructed by following the schematics of the equipment's monitoring system. The construction of such a detector requires knowledge of diode junction characteristics to ensure that a diode will be selected that will function in the RF range to be measured. A detector such as illustrated in Figure 5-22 is connected to a dummy load common to the RF source whose power output is to be measured. The display on the oscilloscope then represents the peak voltage appearing across the dummy load. When used in conjunction with a thermistor-type power meter, the detector can be conveniently calibrated for later use.

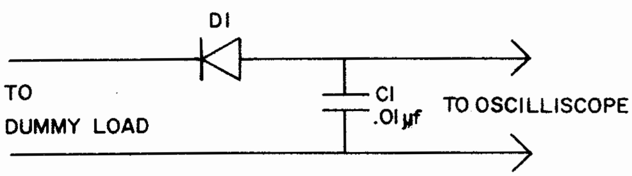

Figure 5-22. Detector for Measuring Low Peak RF Power

5-5 STANDING WAVE MEASUREMENTS

5-5.1 GENERAL

A line used to transfer power is termed a transmission line, with resistances and inductances uniformly distributed along its length and with capacitances distributed across the line's conductors. In many modern techniques, particularly in microwave techniques, use is made of transmission lines having (as nearly as possible) no losses. Such lines are often equipped with probes which slide axially along the line and pick up signals proportional to the voltages across these lines. A plot of this voltage with respect to distance gives a standing-wave plot, and this information may be used for calculating the terminating impedances on the line. This is the principal method of impedance measurement in the microwave and VHF range. The characteristic impedance of any line is defined here as that value of input impedance which the line would display if its length was indefinite. Any line terminated with a resistive load equal to its characteristic impedance will completely dissipate all of the transmitted power applied to its load end, and will not reflect any energy back toward the input power source. Any other termination will cause reflection of energy toward the source, resulting in standing waves along the line.

5-5.2 STANDING WAVES

A standing wave is so-called because if a proper measuring device is used to measure voltage or current along the line, a sinusoidal variation of voltage or current can be measured to determine its amplitude, phase, or frequency. An RF transmission line which is open-circuited at its load end will have infinite impedance across this open circuit, whereas one which is short-circuited at its load end will measure zero impedance across the short circuit. When a generator is connected to either type of line, the voltage and current waves travel along the line until they reach the load. An infinite impedance load will not permit such a flow of current, and a zero impedance load will not permit the drop of any voltage. Therefore, as the energy cannot remain at the load, cannot disappear, and cannot be absorbed, it must reverse its direction and return to its source. Since the reversal of direction is somewhat similar to light wave reflection from a mirror, the term "reflection" is used to describe the effect that takes place at the load end of an open or shorted transmission line. An RF transmission line which is terminated in a capacitive reactance equal to its characteristic impedance will have its standing waves distributed essentially the same as if the line were open. This is because a current peak will be closer to the termination than will a voltage peak. If an inductive reactance is used for this same termination, the standing waves will then be distributed as if the line were shorted, and a voltage peak, rather than a current peak, will be closer to the terminated end.

5-5.2.1 Standing Wave Characteristics

A standing wave on a transmission line can be mentally visualized if a sine wave is viewed as traveling down a line and continually changing from maximum, through zero, to minimum, and back again. Since this sine wave of voltage or current passes a given point at a particular velocity, as determined by the frequency, the amplitude of the wave at any instant can be represented by the vertical position of a point, P, with respect to the horizontal zero axis (refer to Figure

5-23 below). Part (A) of Figure 5-23 shows that the polarity and amplitude are the same at any two points separated by a whole number of wavelengths. Part (B) of the figure shows the position of the incident wave, I, with respect to the reflected wave, R, on a line with an open-circuit termination. If the line were shorted, the incident wave and reflected wave would reverse positions, and the resultant standing wave would reverse its maximum and minimum points from those shown in the illustration.

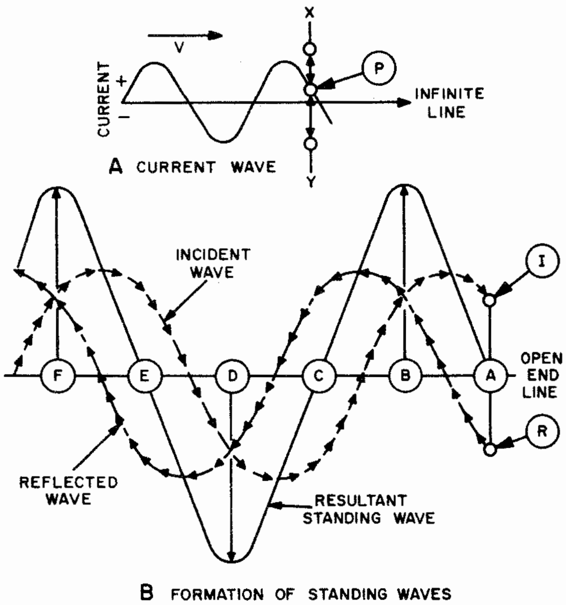

Figure 5-23. Development of Standing Waves

5-5.2.2 Ratios

The ratio of maximum voltage to minimum voltage is called the "voltage standing wave ratio" (VSWR). A similar ratio of currents will have the same value. This voltage or current ratio depends on the ratio of load resistance, at the output or line termination, to the characteristic impedance of the line itself. If a transmission line has standing waves, it is resonant to some frequency. Because of this fact, a given impedance can be transformed into another value by terminating the line at a desired impedance point. A low standing wave ratio indicates a good impedance match, with consequent maximum power transfer. VSWR measurements have proven extremely useful in repair, preventive maintenance, checking, and making adjustments. This is true because the VSWR provides a direct indication of the degree of mismatch measurement. If the tuning device is located between the transmitter and transmission line (Figure 5-24A), the transmission line affects the antenna (load) impedance. In part B of

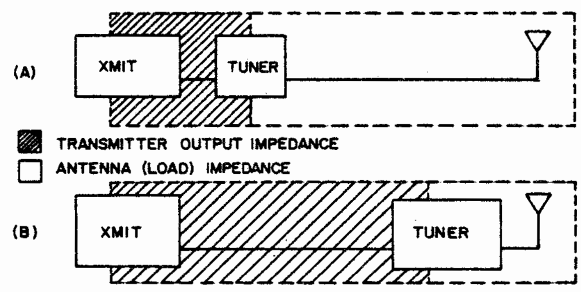

Figure 5-24. Transmission Line Considerations

Figure 5-24, if the tuning device is located at the input to the antenna (load), then the transmission line becomes part of the characteristic output impedance of the transmitter. Adjustments to the output coupling of the transmitter and antenna are performed only when components of the transmitter have been replaced or the transmission line impedance has changed due to age, alteration of line, or physical damage to the line.

5-5.2.3 Measurement

A variety of methods and test equipments may be used to measure the voltage or current distribution along a transmission line. An open transmission line is accessible for coupling to many types of voltage-measuring devices, such as a wavemeter or a grid dip meter. However, at higher frequencies employing coaxial cable or waveguides to minimize skin effect losses, access to the interior of the waveguide or center conductor of the coaxial cable must be gained by using a unidirectional or bidirectional coupler inserted into the transmission line. The coupler contains a slot into which an RF probe is inserted and positioned with respect to directivity. The transmission line length will affect the standing wave ratio, affecting the input and output termination impedance. If the voltage is high at the input terminals and the current is low, then the input impedance is larger than the characteristic impedance; if the current is higher than the voltage at this point, the input impedance is then smaller than the characteristic impedance.

5-5.3 LECHER-LINE METHODS

At frequencies ranging from a few megahertz up to 3000 megahertz, a two-wire transmission-line arrangement, as illustrated in Figure 5-25, may be used for frequency measurement, with accuracies

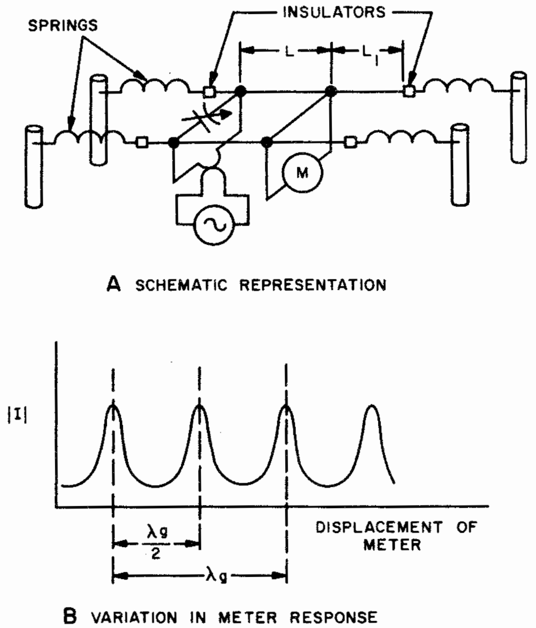

Figure 5-25. Lecher-Wire System of Frequency Measurement

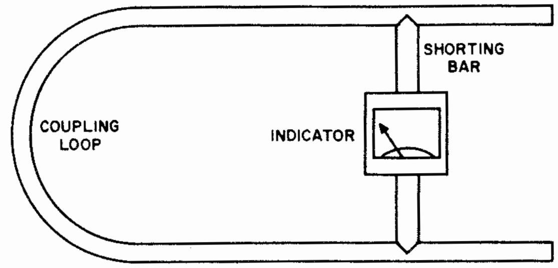

Figure 5-26. Lecher-Wire System for Ultra-High-Frequency Measurement

approaching 0.10 percent. This method comprises a parallel-wire transmission line, usually open-circuited at both ends. The two wires are spring-loaded to prevent sag. The source frequency is loosely coupled to the line, which transmits in the TEM mode. The standing wave patterns can be shifted by using the variable capacitor connected across the input. Measurement is achieved by use of a short-circuiting bridge, in series with a thermocouple meter, paralleled with a low-resistance shunt to decrease the resistance and increase the sharpness of resonance. This bridge arrangement is moved along the line, and the amplitude of current or voltage, maximum or minimum, is recorded. The measured distance between maximum points of current amplitudes is exactly 1/2 wavelength. Figure 5-26 illustrates the proper method of injecting such a meter for UHF standing wave measurement. Instead of a sharp current maximum, a broad or double-humped maximum may be obtained because of interaction between the main section of line and the balance of the line beyond the short circuit. The technician can avoid or minimize this difficulty if he keeps the distance from the far end of the line to the first current minimum different from a quarter wavelength or odd multiple thereof (refer to Figure 5-25). The line must also be shielded beyond the short circuit, and shorting-bars placed across the parallel-wire line in the L_1 section.

5-5.4 PROBES

A magnetic or electric probe can be used to observe the standing wave on a short-circuited terminated line. The wavelength is obtained by measuring the distance between alternate maximum or minimum current points along the line. A typical setup operating at 300 megahertz might use two 10-foot lengths of number 14 phosphor-bronze wires, spaced 1 inch apart and supported parallel to a set of probe guide rails. The line should be partially matched to the source generator by means of a parallel-wire shorting stub connected in multiple with the transmission line and the oscillator output line. Figure 5-27 illustrates the type of probes required for this method of measurement.

5-5.5 NEON-LAMP METHOD

In this method of measurement, a neon bulb or milliammeter is moved along the two-wire parallel transmission line. Points of maximum voltage for the lamp, or points of maximum current with the indicator, will have maximum brilliance or indication, respectively, to place the exact mode or antimode.

5-5.6 SHORTING-BAR METHOD

Figure 5-28 illustrates a method of standing wave or frequency measurement. The current loops can be located with a shorting-bar or knife-edge moved along the length of the Lecher wire. The Lecher line, source, and indicating device must all be coupled very loosely together. As the shorting-bar moves down the line, each resonance point will produce a pronounced dip in the meter (or will glow in a lamp-type indicating device). If the shorting-bar is left in this position, the line will be tuned to the source frequency and will absorb maximum energy from it.

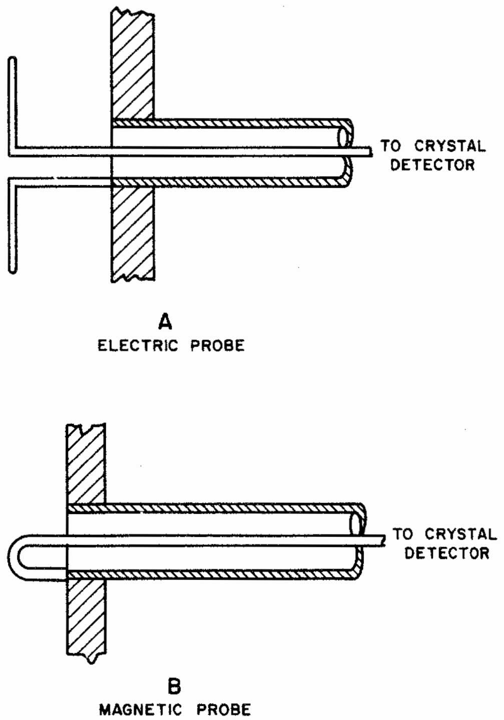

Figure 5-27. Typical Electromagnetic Probes

5-5.7 BRIDGE METHODS

The bridge method permits measuring of the standing wave ratio without actually measuring the actual standing waves. The bridge principle is applicable because the input impedance of a line terminated in its characteristic impedance is a pure resistance equal to the characteristic impedance. A line thus terminated can be used as the unknown resistance in a bridge circuit, and a null will be obtained in the indicating device when the other resistance arms are properly adjusted. Many types of bridges are practicable. For example, any ac bridge which is independent of the applied frequency can be used. The bridge will become unbalanced when the line is no longer properly terminated. Improper termination will produce a reactive component, as well as a resistive component, in the input impedance of the line. As the reading of the indicating device depends on the degree of imbalance, which becomes more severe as the mismatch caused by the termination becomes worse, the indicating device can be calibrated directly to indicate the standing wave ratio. The most common indicator consists of a crystal rectifier, a filtering circuit, and a sensitive dc meter-movement in series with a high resistance.

5-5.8 RESISTANCE-CAPACITANCE BRIDGE

Figure 5-29 shows a resistance-capacitance bridge circuit, with the equation for balance provided. The bridge is theoretically independent of the applied frequency, but the applied frequency must be low enough to avoid skin effect and stray inductance, capacitance, and coupling between circuit elements and wiring. The leads must be kept short to eliminate stray reactance, which causes bridge imbalance. The rectifier circuit wiring must be isolated from other bridge component fields so that "residual" voltmeter readings at the balance point are minimized. Finally, the technician should use only resistors having negligible capacitance and inductance.

5-5.8.1 Calibration

Before calibrating a newly-constructed bridge, the following procedure must be followed if residual readings caused by stray effects are to be held to a minimum:

1. Connect a noninductive resistor which is equal to the characteristic impedance of the line to the output terminals of the bridge.

2. Apply an RF voltage to the input terminals and adjust the variable capacitor for a minimum reading on the meter.

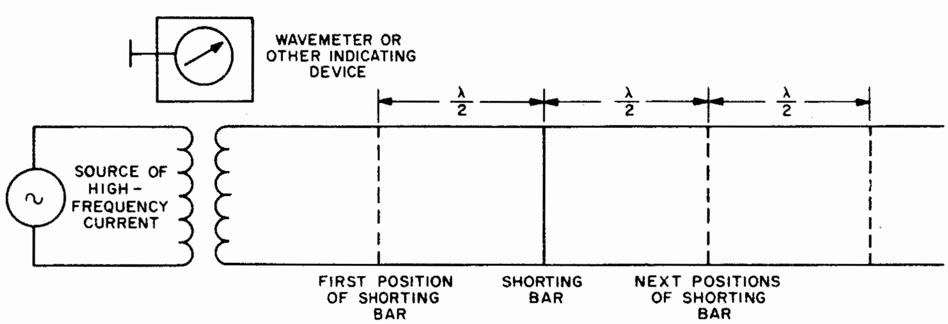

Figure 5-28. Frequency Measurement with Lecher Wires, Shorting-Bar Method

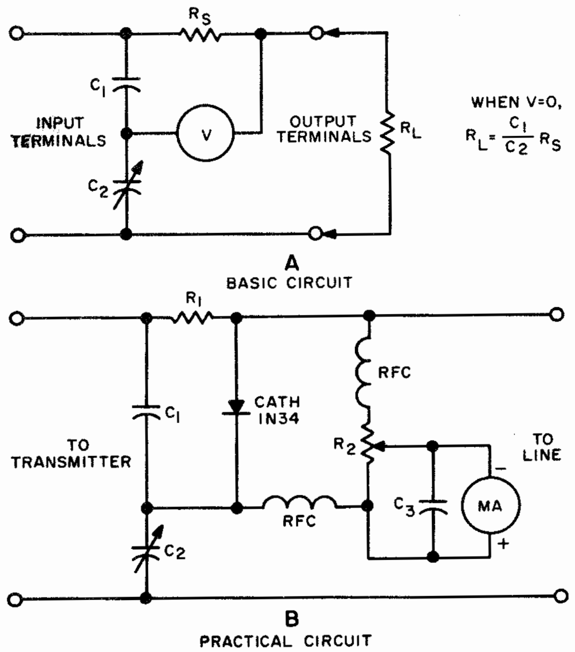

Figure 5-29. Resistance-Capacitance Bridge Circuit for Measuring Standing Wave Ratio

3. Reconnect the resistor to the input terminals and connect the RF power source to the output terminals.

4. Adjust the RF voltage amplitude applied to the meter until a full-scale reading is obtained.

5. Reconnect the bridge in the normal manner (resistor to the output terminals, etc.). If the meter reading is now more than 1 or 2 percent of the full-scale reading, different arrangements (lead dress) of the internal wiring must be tried until the null is reduced to zero or as close as possible to the zero point.

The bridge can be calibrated after completion of the preceding check. Connect the transmission line under investigation to the output terminals of the meter, and connect a succession of noninductive resistors to the load end of the transmission line. Assuming that the bridge was originally balanced for the characteristic impedance of the line, the standing wave ratio can be computed from the equation:

$$SWR = \frac{RL}{Ro} \text{ or } \frac{Ro}{RL}$$

where:

Ro = line impedance
RL = load resistance

Select the formula that yields a ratio greater than unity. The SWR calibration can be recorded on the meter scale directly, recorded on a chart in terms of the meter deflection, or plotted on a graph against the meter deflection. The variable capacitor, in turn, can be calibrated for various characteristic impedances. This is accomplished by applying suitable resistors to the meter and noting the capacitor settings at the respective balance points. A range of 50 to 300 ohms should prove attainable.

5-5.9 ACCURACY OF BRIDGE MEASUREMENTS

To assure accurate measurements, the RF signal applied to the bridge must be properly adjusted each time a calibrated instrument is used. Essentially, this adjustment is a repetition of the previously described reversed-bridge procedure. The following steps are to be performed:

1. Connect the line to the input terminals of the bridge and connect the transmitter to the output terminals.

2. Adjust the transmitter coupling until full-scale deflection is obtained. From this point on, the coupling must be left untouched.

3. Reconnect the bridge in the usual way and proceed with the measurement.

5-5.10 POWER OUT vs. IMPEDANCE MATCHING

For maximum transfer of power out of an RF source, with minimum heating from reflected power, the total output impedance sensed by the RF source must be equal to the internal impedance of that source. That such a state exists is indicated by a VSWR or SWR ratio of about 1.1 to 1. Figure 5-30 shows an RF source, output tuner, VSWR bridge (A), transmission line, and VSWR bridge (B).

5-5.10.1 Tuning Procedure

The industrial/military standard RF impedance is 50 ohms for optimum output of the above system. The tuning procedure for the set-up shown is as follows:

1. Tune the RF source for maximum output, with a 50-ohm load interrupting the line to the tuner.

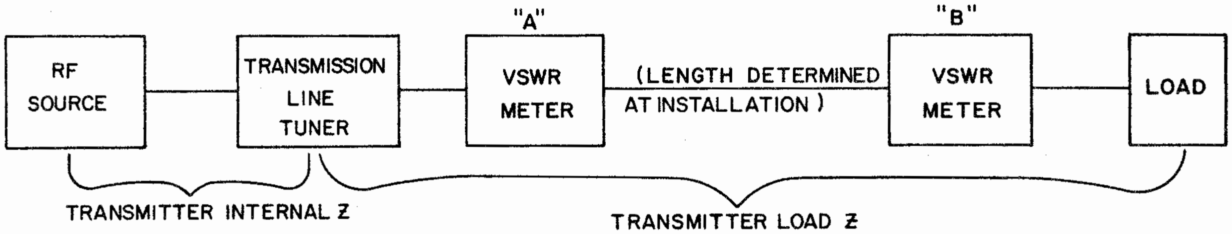

Figure 5-30. Transmitter and Load Test Circuit

2. Place the 50-ohm load so as to interrupt the line between VSWR meter "B" and load; then tune transmission line for lowest VSWR reading on meter "A".

3. Connect load and tune the load for minimum VSWR indication on meter "B".

Most systems only have provisions for tuning the transmitter into a dummy load and then tuning the load for actual use. The VSWR meter, in such instances, is placed in the VSWR meter "A" configuration, rather than the "B" configuration. It is emphasized at this point that the reflected power, as indicated at meter point "A", is not the true reflected power. Assuming that the frequency of operation is such that a loss of 3dB occurs from the output of the transmission line tuner to the load from inherent cable losses, all reflected power from the load would then be attenuated by 3dB before it arrived at VSWR meter "A". The forward power indicated would reflect such power prior to its attenuation. Thus, assuming 100 watts of forward power is being indicated by meter "A", and 25 watts of reflected power is indicated on meter "B", the power indications at the load (SWR meter B) would show 50 watts forward and 50 watts reflected. (The forward power becomes dropped by 3dB, and the reflected power is raised by 3dB.) In a system having a non-tunable load, the transmission line tuner would vary the collective impedance of the transmission line and load to match the output of the transmitter. Such systems have a range of about 45 to 70 ohms in impedance-matching capability.

5-5.11 WAVEMETER METHODS

Wavemeters are of two basic types: reaction and absorption. Both types will absorb a portion of the output power of the device whose frequency is to be measured. The reaction wavemeter should be employed for low-power frequency measurements because it absorbs very little energy. For greater accuracy on high-power devices, the absorption meter should be used, even though it absorbs more power from the device being tested. Because of the power absorbed, its use is therefore restricted to high-power devices. In use, the resonance indicator (a lamp or ammeter) is connected into the tank circuit of the wavemeter. Wavemeters cannot be relied on for a great amount of accuracy because they detune self-excited oscillator circuits to which they are coupled. The brightest indication or maximum deflection will be evidence of fundamental frequency. Therefore, this type of meter is useful only for checking a transmitter's oscillator to determine whether it is operating at the correct fundamental frequency.

5-5.12 REACTION-TYPE WAVEMETER

Figure 5-31 illustrates the reactance wavemeter, containing a coil and a variable capacitor. The wavemeter resonant frequency must be made equal to the frequency being measured by varying the capacitor while the external coil is loosely coupled to the output of the circuit under test. The indicating device will adjust to either maximum or minimum, depending on its location. The scale or dial reading is then used to determine the wavelength by reference to an associated calibration curve or chart. When the wavemeter reaches resonance, the coupling must be reduced for a barely usable indication, or a sharp indication will not be obtained and an error will thus be introduced.

5-5.13 ABSORPTION-TYPE WAVEMETER

This type of wavemeter (Figure 5-32) is basically the same as the reactance type. An indicating device (lamp) provides a self-contained indicating device. The fixed capacitor across the lamp has a greater capacitance than the variable capacitor. Enough

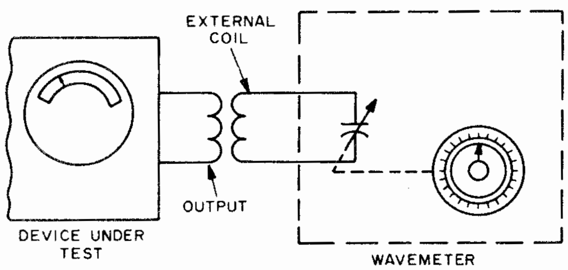

Figure 5-31. Circuit of Reaction Wavemeter

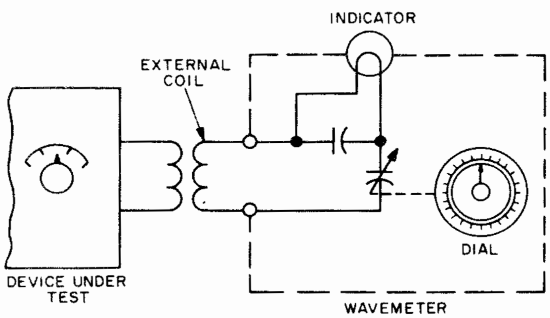

Figure 5-32. Circuit of Absorption Wavemeter

voltage is thus developed at resonance to light the lamp, and at the same time has only a negligible effect on the resonant circuit because it has a relatively low reactance. When this type of meter is coupled into the circuit, resonance must be approached very slowly to prevent the lamp from burning out. This requires varying the dial very slowly and reducing the coupling as resonance is approached. For best accuracy, only a faint glow representing maximum brilliance at resonance should be present.

5-5.14 COAXIAL-LINE WAVEMETERS

When making wavelength measurements with a (coaxial) slotted-line section, the correct procedure requires that it be terminated in a high SWR termination, usually a short circuit, while the moving probe searches the guide section for standing wave minima. If the probe is replaced by a moving short circuit, and provisions are made for coupling energy into the transmission line now short-circuited on both ends, the coaxial cavity will then be resonant each time the movable shorting-plunger is at the position of a standing wave minimum. This position occurs when the cavity is an integral number of half-wavelengths long. Figure 5-33 illustrates some typical wavemeter measurement arrangements that can be used. In part (A) of the figure, the wavemeter is used as a transmission device because it is provided with both input and output coupling loops, and transmits power only when resonant. As a result, the technician can determine the wavelength by tuning the meter to give maximum power indication. In part (B), the meter is used as a reactance or absorption meter, parallel with the main line. At resonance, it causes the power meter in the main line to dip sharply. In part (C), the meter is used for frequency monitoring.

5-5.14.1 Procedure

The procedure for measuring the wavelength with meters like the one shown in part (A) of Figure 5-33, where the plunger displacement is linearly related to the wavelength, involves the determination of the distance between successive resonances. The other wavemeter types, shown in parts (B) and (C), do not have a cavity length or tuning response which is linear with frequency. Therefore, calibration is required. To perform measurements, a meter and one of the following is required: a dc microammeter and RF power-indicating device, or a search meter. Figure 5-34 illustrates a method of testing the individual parts of a transmission line to determine the cause of a high standing wave ratio. The RF source is a pulsed signal generator. An attenuator pad is used in series with the generator output to provide attenuation and to isolate the generator from the rest of the equipment. With this setup, the technician can check the SWR of the part being tested at any frequency within the range of the generator, using the adjustments provided to set to any specified frequency.

5-5.15 COAXIAL CAVITIES

Coaxial cavities are generally used at the lower microwave frequencies because they can be made smaller than cylindrical cavities. Their upper frequencies are limited by the radial dimensions of the coaxial line and by the reduced plunger displacement required between successive resonances. The lower frequency limit is determined by the increased linear displacement required of the shorting plunger. Hence, at low frequencies, cavity size and weight are the limiting factors.

5-5.16 COUPLING

Coupling to these hollow cylindrical cavities can be accomplished by cutting holes in the common wall between the adjacent waveguide and the cavity, as shown in Figure 5-35, where the wavemeter is used as a reaction device. The hole in the waveguide must be positioned to couple to a field component of the desired mode. Figure 5-36A shows an alternate

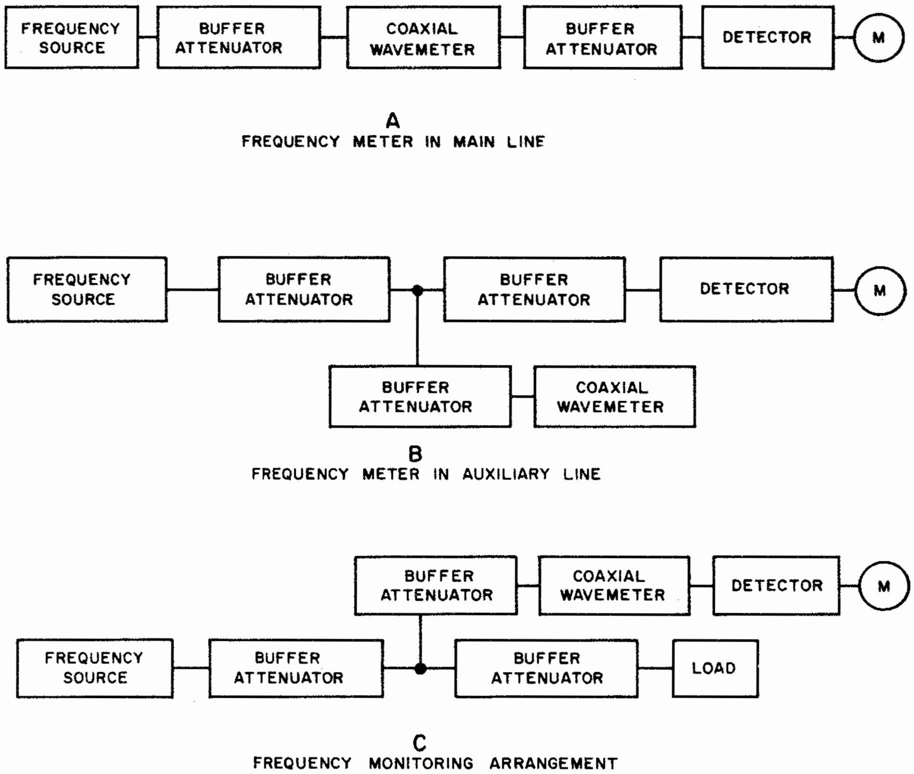

Figure 5-33. Typical Measurement Arrangements Using the Coaxial Wavemeter

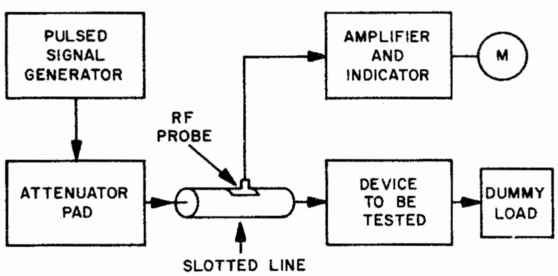

Figure 5-34. SWR Test Setup for Checking Transmission Line Components

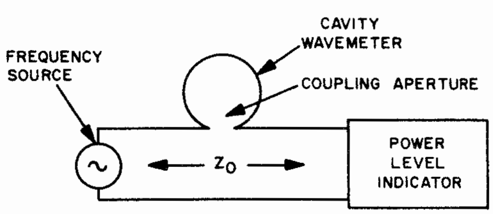

Figure 5-35. Iris-Coupled Wavemeter

method where the wavemeter is coupled to the main guide through an E-plane waveguide-tee arrangement. The shunt line length, L, is chosen so that it matches at frequencies far off (cavity) resonance. The power delivered to the load is maximum. As the frequency is varied or the wavemeter is tuned to resonance, the power delivered to the indicator-load approaches a minimum, as shown in Figure 5-36B. A different fixed coaxial line arrangement is shown in Figure 5-37, in which a loop-excited cavity wavemeter is coupled to the main line through a quarter-wave coaxial stub. Thus, the stub is effectively a quarter-wavelength at frequencies far off resonance, where the shunting impedance is a maximum. However, the shunting impedance is simply the resonant impedance of the cavity, and the power delivered to the matched load (power-detecting element) is a minimum. Figure 5-38 shows two simple methods of making a standing wave or frequency determination. In part (A), the cavity is used as a transmission device and the frequency is determined by varying the length of the calibrated cavity for maximum power. In part (B), a single aperture cavity wavemeter is used as a reaction termination. The line length, L, is

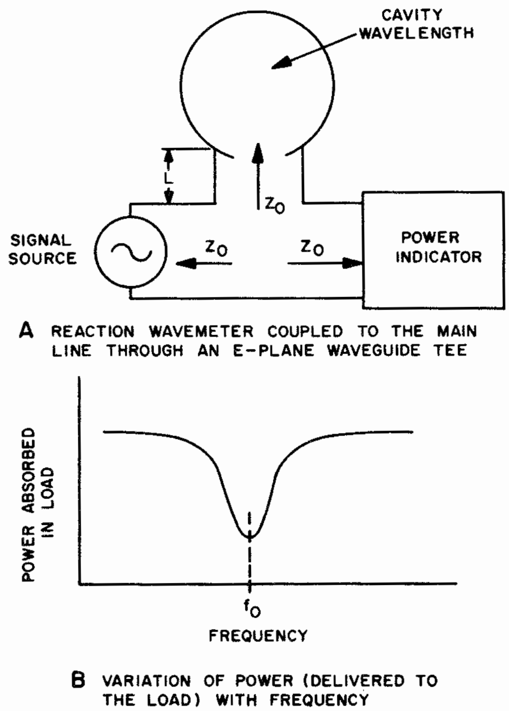

Figure 5-36. Alternative Cavity Wavemeter

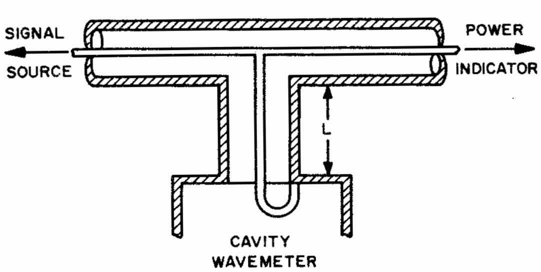

Figure 5-37. Cylindrical Reaction Wavemeter Coupled to Main (Coaxial) Line Through a Quarter-Wave Stub

selected so that at frequencies far away from resonance, the power delivered is maximum at the power detector.

5-5.17 ACCURACY OF CAVITY WAVEMETERS

High-Q, precision, partial-coaxial cavity wavemeters normally use frequency-sensitive, noncontracting shorting plungers to avoid the erratic operation and high contact resistance that frequently occur with spring-finger contacts. Thus, the tuning

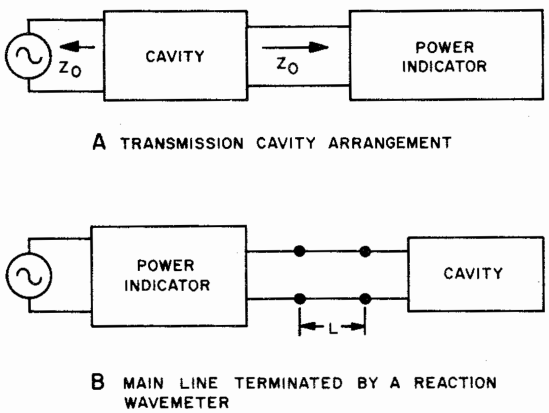

Figure 5-38. Cavity Wavemeters in Main Transmission Line

range of the wavemeter is limited to the frequency range of the noncontacting short circuit. The upper frequency limit is determined by the dimensions of the cavity modes at resonance. Where noncontacting shorting plungers are used, a loss results because of the power leakage past the shorting-plungers. Another loss is due to the inherent backlash in the drive mechanisms of the tuning plungers. This error can be reduced by taking readings of the successive resonant points as the plunger is displaced in one direction. Compensation must also be made for errors due to water vapor inside the cavity structure. The presence of moisture within the air line will cause the dielectric constant of the medium to change and will thus introduce frequency errors amounting to several parts in ten thousand. The effects of humidity can be neutralized by completely sealing the cavity and either evacuating it or filling it with a dry, inert gas. Broad-band cavity-type frequency meters vary in accuracy between 0.01 and 1.0 percent. For very small ranges, absolute accuracies as high as 0.005 percent may be obtained. Other factors that limit the overall accuracy attainable are changes in frequency indications caused by variations in temperature and humidity, detuning due to external reactive loading, and the resolution of reading as affected by the Q-factor.

5-5.18 GRID DIP METER METHOD

A grid dip meter is a multiple-range oscillator which includes a meter in its grid circuit to indicate current flow. For example, a grid dip meter circuit using a Colpitts oscillator with center-tapped low-frequency plug-in coils to maintain the inductive balance regardless of tuning may be used. This meter can be operated as an absorption-type standing wave meter if plate potential is not applied and if the grid and cathode are allowed to act as a diode. The meter reading will increase when the instrument is coupled closely to a source of RF energy.

5-5.19 CRYSTAL VOLTMETER METHOD

In the centimeter region, crystals are used as both detectors and mixers in standing wave measurements. However, their use for direct voltage measurements is limited by the instability and variability of their characteristics, and by the constant need for frequency calibration. A gold-leaf electroscope may be employed to calibrate a crystal voltmeter in the gigacycle range. For example, in a 10-cm calibrator, use a gold leaf 7 mm long, mounted in a slit in the wall of a vertical section of coaxial line. The leaf will deflect toward the center conductor when a field is present. The amount of deflection can be observed by using an optical instrument. A probe of known coupling properties withdraws a portion of the signal to the crystal and indicating device. Tuning stubs to provide a voltage maximum at the left must be installed when applying the 10-cm signal. Calibration is achieved by assuming equality between a known low-frequency signal producing a given leaf deflection, and a 10-cm signal producing the same deflection. When actually making the standing wave measurement, the calibrated crystal in a coaxial line must be used with a probe of known coupling value.

5-5.20 THERMAL METHODS

The largest group of ammeters and voltmeters used for measurements within the VHF and UHF range are composed of thermal instruments that measure power consumed in the measuring device, and are therefore current-squared indicators. The voltage or current being measured is normally impressed on a fine wire, producing heat which can be measured by any thermoelectric technique. The thermoelectric effect or the increase in resistance characteristic can be used to indicate maximum or minimum voltage or current points.

5-5.21 PHOTOAMMETER METHOD

In this method, the unknown current causes the wire to incandesce. The resulting light is then measured with a photocell. This is a high-precision transfer technique whereby a dc standard serves as the basis for the high-frequency measurement. It requires the use of a temperature-sensitive resistance, or bolometer, in one arm of a Wheatstone bridge. This method calls for bridge balancing with a known dc source voltage applied, and then varying the dc to rebalance the bridge when the unknown high-frequency voltage is placed across the bolometer, thus causing resistance change. From the change in applied dc and the circuit constants, the unknown voltage can be computed. (It is assumed that the bolometer resistance is not dependent on frequency, and that equal amounts of dc and ac power will produce identical heating effects.)

5-6 FIELD STRENGTH MEASUREMENTS

5-6.1 GENERAL

The amplitude of an electric field of a radio wave at any given point is termed the "field intensity" or "strength" of that wave, and is usually measured in terms of millivolts or microvolts per meter. The measurement of field strength involves finding the response of an antenna to the incident field. The field intensity of a radio wave can be determined by measuring the RF voltage induced in a receiving antenna. However, before translating the antenna response into actual electric and magnetic field intensities, a two-fold analysis of the behavior of both the antenna and of the field itself must be performed. The measured field will induce a voltage at each point of the receiving antenna. This voltage will have a magnitude and phase that depend upon the antenna shape, as well as the angle of field incidence. In addition, the voltage varies from point to point on the antenna itself. Thus, the total response of an antenna used for field measurements represents a spatial average of the field over the entire antenna. The field being averaged is not a simple propagating wave, since a free-space wave alone cannot exist. Even in line-of-sight transmissions, the combined effects of the free-space wave component and the ground-reflected wave component are considered. Beyond the horizon, the receiving antenna measures the combined effects of the diffracted wave and the ionospheric or troposperic-reflected wave. The measuring antenna yields the average field strength over a finite area; this field strength is a function of time. The time variation of the measured field strength depends on the predominant wave component. Therefore, the free-space wave is essentially constant with time if the transmitter output remains constant, while the sky-wave components are dependent on time, frequency, and the propagation medium properties.

5-6.2 VARIABLES

The accuracy of all field measurements depends on many factors, and the information content of these measurements is a function of the transmission variables. On most occasions, the density of the electromagnetic energy incident at the measured point is actually being measured, when what is desired is the field intensity measurement. For example, a voltage measurement may be either under open-circuited conditions or when voltage is delivered to a matched load. The measurement of voltage delivered to a matched load is really a power measurement. A measurement made at the center of a dipole is a true voltage measurement. The same measurement can be made with a matched dipole if measured across a resistor or transmission line connected to the center of the dipole. However, the open-circuit and the matched voltages differ by a factor of 2. When a thermocouple or bolometer at the center of the dipole is employed, the measurement may actually be that of the thermocouple or bolometer. This is again a power measurement. Therefore, differentiation must be made between a measurement in which the detector has a very low impedance and a measurement of current in a matched load, because these two considerations also differ by a factor of 2. The electric field intensities being measured range from microvolts to several millivolts per meter. The magnetic field intensities are smaller by a factor of 120 π, and are expressed in terms of milliamperes or microamperes per meter. Thus, a plane wave with an electric field intensity of 120 microvolts per meter will have a magnetic field intensity of 1 microampere per meter.

5-6.3 TEST EQUIPMENT

Many different types of field intensity measuring equipments are available in different ranges of frequency coverage. These radio test sets, field strength meters, and radio "noise" meters can be used to measure either relative or absolute magnitudes of field intensities as produced by an excited transmitter antenna. These test equipments can be employed to measure a part of the radiated field, to determine the total amount of energy being radiated. This information can then be used to determine antenna efficiency, directivity characteristics, and signal coverage; to check spurious harmonic radiation; and to make surveys of field intensities. The technician can also detect and locate radiated or conducted interference by using radio test sets. When using field intensity measuring equipment, objects or persons near either the radiating source or the meter must be removed, because they will cause reflections that will cause erratic meter readings. The test equipment antenna must also be extended to its full length to obtain proper operation of the antenna tuning circuit and to provide normal meter indications.

5-6.4 RELATIVE FIELD STRENGTH MEASUREMENT

Simple test equipment, such as a grid dip meter, will suffice to measure relative field strength. To use the grid dip meter for field strength measurement, turn off the plate voltage, connect a loop antenna to the coil terminals of the instrument, insert an appropriate plug-in coil, and tune the meter to the transmission. The received signal will cause grid-cathode current flow, and the relative current magnitude will be indicated by meter deflection. If employing a pickup antenna and a diode or crystal in conjunction with a microammeter, the meter reading will indicate the relative strength of the field acting on the antenna. Because of the nonlinearity of the crystal, the meter reading will not be directly proportional to the field intensity being measured. The sensitivity may be increased by using a microammeter as the indicating device.

5-6.5 ABSOLUTE FIELD STRENGTH MEASUREMENT

Because the voltage induced in a pickup antenna must be compared with a voltage generated by a calibrated oscillator, more elaborate test equipment must be used for the measurement of absolute field intensity than is required for relative field strength measurements. One method is to apply the antenna voltage to a sensitive receiver which usually incorporates two calibrated attenuators: one between the antenna and mixer stage; and the other in the first I-F amplifier. The test equipment contains a meter to indicate the second-detector diode current. This type of test equipment will indicate absolute field intensity in terms of microvolts per meter. Another technique employs a specially designed receiver which has, at the I-F amplifier input, an attenuator calibrated in dB. The mixer-stage output must be exactly proportional to the RF signal voltage present in the grid circuit, and must remain so for input amplitudes up to 1 volt. An auxiliary RF-calibrated oscillator and an electronic voltmeter capable of reading about 1 volt must also be employed. If portability is a factor, then an arrangement using a loop antenna to develop the input RF signal should be used. The auxiliary oscillator output can be coupled to the loop by means of a transformer as shown, or can be applied across a resistor in series with the loop. The local oscillator must be coupled to the input circuit in such a fashion that its output does not become varied as a result of tuning. The input circuit can be checked with an electronic voltmeter, or the mixer plate circuit can be measured with a milliammeter calibrated to indicate a 1-volt grid signal. To avoid saturation of the I-F stages, the I-F amplifier must be preceded by a calibrated attenuator.

5-6.6 ALTERNATE SIGNAL GENERATOR METHODS

Another method of measuring the field strength uses a standard signal generator to introduce a voltage into the antenna circuit, the loop antenna, and the receiver. In this method, the radio receiver is tuned to the signal to be measured, with the loop antenna adjusted for maximum reception. The receiver gain is then set for a convenient deflection of the meter in the output of the detector circuit, and the loop is oriented to eliminate the received signal. The input from the signal generator is now adjusted such that the meter deflection in the detector circuit is the same as before. This amplitude, when divided by the voltage-transfer ratio, is equal to the received signal.

5-6.7 STANDARD ANTENNA METHOD

In this method, an antenna is selected which covers the desired frequency range. This antenna will be used to feed a voltage-measuring device, such as a receiver, with a known adjustable response. The basic arrangement of these equipments is shown in Figure 5-39, and an extended version of the arrangement is illustrated in Figure 5-40. The receiver must be calibrated against a fixed standard voltage. Some means of attenuating this voltage in the receiver must be provided to permit measurements over the entire frequency range of interest. The relationship between the strength of the radio wave and the equivalent lumped voltage that the wave induces in the antenna can then be determined when the antenna-voltage measuring device receives the unknown radiation. The resultant voltage measurement can be used to compute the field strength. The ratio of the voltage induced at the antenna terminals to the electric field intensity which produces this voltage represents the effective length or height of the antenna. The antenna is normally coupled to the voltage-measuring device in such a manner that the measured voltage is not the induced voltage, but bears a fixed ratio. This is called the "voltage transfer" ratio. The effective height of the antenna, the voltage transfer ratio, the calibrating

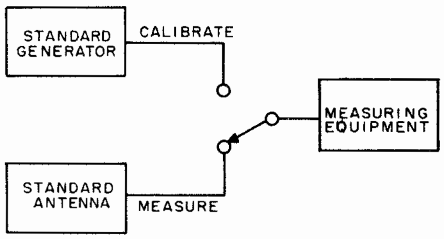

Figure 5-39. Standard Antenna Method of Measurement

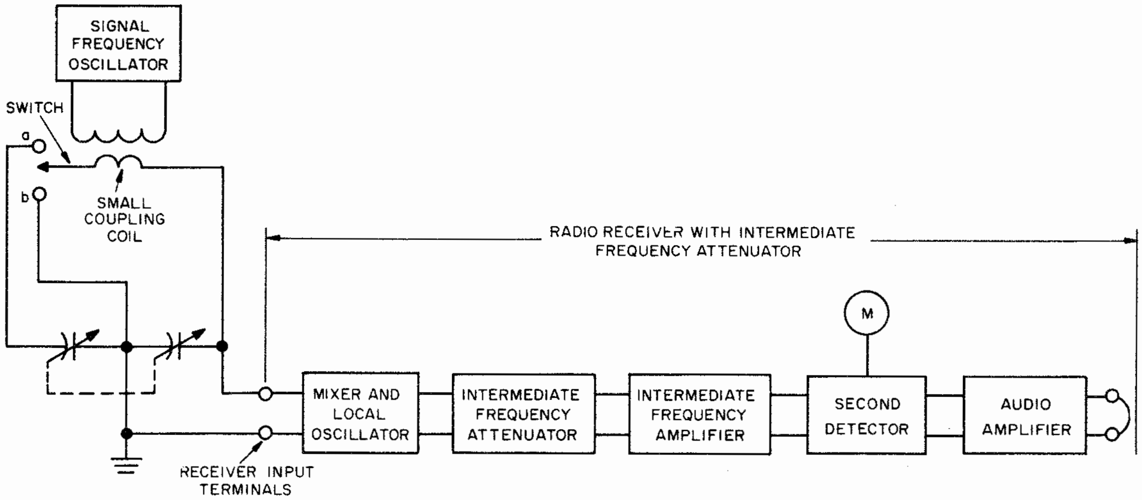

Figure 5-40. Diagram of Field-Strength Measuring Equipment of the I-F Attenuator Type

voltage, and the voltage-attenuator ratio are all factors which the actual computation must consider. The effective antenna height can be measured by placing the antenna in a standard field. The terminal voltage then measured is a function of the induced voltage, and a product factor of the effective length and the voltage transfer ratio. The reciprocal of this value is called the "antenna coefficient," and this coefficient includes the gain factor of the receiver.

5-6.8 STANDARD FIELD METHOD

A standard generator is used in this method to produce the field, as explained previously. The field itself is calculated from the antenna dimensions and from the value of current or power supplied to that antenna. A receiver or similar device is used to measure the equivalent voltage induced by the field at the measurement point. The voltage induced by the unknown field is then measured, and a comparison of the voltage induced by both the known and the unknown fields permits a computation of the field strength of the unknown radiation. The field at the receiving antenna can be computed from the dimensions of the transmitting antenna, its current and voltage distribution, the receiving and transmitting antenna elevations and distance between them, and the effects of ground. At lower frequencies, accurate calculations of the value of the standard field at the receiver are difficult to achieve because of the effect of ground. Therefore, unless the value of the ground constants are actually known, the distance between the receiving and transmitting antennas must be made very short. As this method of measurement requires a different value of standard field for each value of unknown, it is more convenient to measure the antenna voltage transfer ratio of the measuring system in terms of the standard field at the frequencies involved. Subsequent measurements are then made, using the standard antenna measurement method explained previously. Two basic techniques are used in the standard field method: the induction field technique for lower frequencies; and the radiation field technique for higher frequencies.

5-6.9 INDUCTION FIELD TECHNIQUE

Figure 5-41 illustrates the induction field technique to be used at frequencies below 30 mc. A loop antenna is employed at both the transmitter and receiver, with a relatively close distance between them. Although the magnetic field intensity (H) is actually measured, the results are expressed in terms of the equivalent free-space plane-wave electric field intensity (E), and is represented in volts per meter. The impedance (Z) of free space (377 ohms) will then determine the value of E with the following relationship:

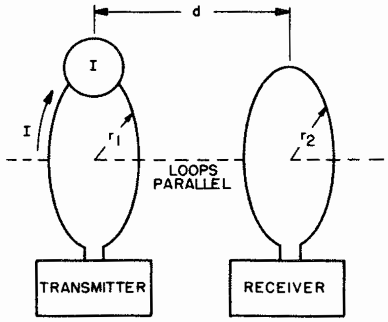

Figure 5-41. Induction Field Technique

$$E = ZH$$

For the loop antennas to possess the uniform current distribution, the circumference of a single-turn balanced loop must be less than one-eighth wavelength ($\lambda/8$). This rule must be applied whether the loop is circular, square, or rectangular.

5-6.10 RADIATION FIELD TECHNIQUE

The close spacing required between receiver and transmitter antennas, and the small antenna loop radius required to ensure a uniform current, make the induction field technique impractical above 30 MHz. A doublet antenna, shown in Figure 5-42, is used for both receiver and antenna. The distances of separation are such that only the radiation

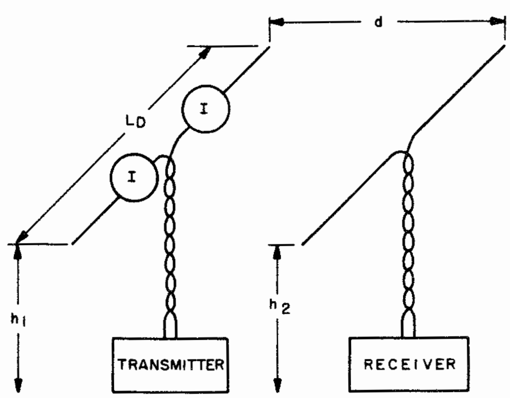

Figure 5-42. Radiation Field Technique

field is significant. If frequencies between 3 MHz and 300 MHz are used, if the polarization is horizontal, and if the distance separation is much greater than the sum of the heights of the two antennas above ground, then the experimental procedure and calculations are facilitated, provided that only the direct and reflected ground waves are involved. To employ this method, select a flat, cleared site where the distance to the nearest reflecting object is great in comparison to the separation d, shown in Figure 5-42, and where the reflection coefficient of the ground is approximately -1. If the use of a doublet antenna is not convenient, a single-turn transmitting loop antenna can be employed, provided that its diameter is very small compared to the wavelength. The effective loop length is calculated with the aid of the following formula:

$$D = \frac{2\pi S}{\lambda} \text{ meters}$$

where:

D = length
S = antenna loop area in square meters and the loop is assumed to have a uniform current distribution
λ = wavelength
π = 3.14

The antenna coefficient of the receiver can be determined for a horizontally polarized standard field, and the receiver can then be used to measure vertically polarized fields with the same accuracy.

5-6.11 FIELD STRENGTH MEASURING DEVICES

5-6.11.1 Antenna

Many basic types of antennas can be employed to measure field strength. However, the choice must be dictated by the polarization of the field component being measured, its intensity, frequency, physical size, and other similar factors. The normal types of antennas used in field strength measurement are the loop, a horizontal or vertical dipole, a simple wire above ground, an electromagnetic horn at the higher frequencies, and others of a more complex nature, depending on the requirements. Each antenna-type has specific uses, and each must be evaluated for the intended field-strength measurement application.

5-6.11.2 Effective Antenna Length

The concept of effective length is used because it simplifies many field-strength computations. An antenna of actual length (L_a) and an arbitrary current distribution must be replaced with a hypothetical antenna of effective length (L) and having a uniform current distribution.

5-6.11.3 Antenna Radiation Resistance

The impedance characteristics of an antenna are similar to that of a transmission line with a complex load, because the antenna is actually a circuit with distributed constants. The reactive component, which is either inductive or capacitive, is not important in this case, because only the resistive component is used in this measurement of field strength. This resistive component takes into account the energy loss of the antenna as a direct result of radiation, ground losses, antenna wire resistance, dielectric losses, etc. For maximum antenna efficiency, all losses must be small in comparison to the energy lost by radiation. This radiation resistance is actually a hypothetical quantity. However, the antenna acts as though this resistance were actually present, because the energy lost by radiation is equivalent to the same amount of energy dissipated via resistance. The radiation resistance referenced to a specific point in the antenna system is that resistance which should dissipate the same energy actually radiated from the antenna system. This resistance can be calculated with the aid of the following formula:

$$\text{Radiation Resistance} = \frac{\text{Radiated Power}}{I_o^2}$$

where: I_o = maximum effective current flowing in the transmitting antenna.

The radiation resistance must be related to a particular point in the antenna system, since the resistance must be such that the current squared, times the radiation resistance, equals the radiated power. The current is different at different points in the antenna, therefore the reference point is normally the point of maximum current, except in the case of a vertical antenna with its lower end grounded. For the vertical antenna, the minimum resistive point (represented by ground) constitutes the reference point.

5-6.12 TEST TECHNIQUES

5-6.12.1 Antenna Directivity Patterns

The directional characteristics of all physical antennas favor radiation in certain directions at the expense of energy in other directions. These directional characteristics can be computed easily for simple elements such as loops or dipoles. However, more complicated structures require more precise measurements. To obtain a complete picture of an antenna's directional properties, measurements must be made at all points on the surface of an imaginary sphere enclosing the antenna. For a small antenna tested at close range, the antenna may be rotated in front of a fixed receiving antenna to measure the relative field strength in all directions. For a large antenna, the receiver is transported around the transmitter in a van or other conveyance, or its pattern may be determined from a smaller-size electrical scale model.

5-6.12.2 Gain and Effective Area

A gain function is normally used to express the directive properties of an antenna. The "absolute gain" of an antenna is defined as the greatest factor by which power transmitted in a specific direction can be increased by a particular antenna in comparison to an isotropic radiator, which radiates uniformly in all directions. The relative gain is the ratio of power supplied to a comparison antenna to provide a directional field strength, to the power supplied to a directional antenna to obtain the same field strength in the same direction. The power gain of an antenna can be ascertained by comparison with an antenna of known gain or by direct measurement. In some cases, it may be calculated. A typical determination involves measuring the ratio of the power input to a transmitting antenna (P_1) to the power output of a receiving antenna (P_2) for two identical antennas. Having determined the ratio P_1/P_2 by one of the methods previously described in this section for the direct-field (free-space) condition, the absolute gain of the antenna can be calculated by applying the following formula:

$$g = \frac{4\pi d}{\lambda} \quad \frac{P_2}{P_1}$$

where:

g	=	absolute power gain of one antenna relative to an isotropic radiator
P_1	=	power input to transmitting antenna
P_2	=	power output of receiving antenna
d	=	diameter of antenna
π	=	3.14
λ	=	wavelength in meters

The power gain of a thin cylindrical dipole, which is short as compared with its wavelength, is approximately 1.5 if the heat loss is neglected. The gain of a thin half-wave dipole is 1.64, and this value increases very little as the diameter of the half-wave antenna is increased.

5-6.12.3 Receivers

Field measurements can be made by using a highly sensitive superheterodyne receiver attached to the output terminals of a receiving antenna. A microammeter must be inserted in the second-detector current. A standard voltage source, as previously explained, can be employed to calibrate the microammeter in terms of the receiver RF input voltage. The anntenna is removed from the receiver input terminals and the standard voltage is applied to the receiver antenna terminals for calibration purposes. With this method, the receiver can be calibrated at a fixed level of sensitivity for each calibration voltage level applied. The field strength can then be determined by comparing the known calibrating voltage produced by the field when the antenna is attached to the input terminals of the receiver, rather than the known comparison voltage. The voltage induced in an antenna can be measured more conveniently with the use of a receiver containing an attenuator in the I-F amplifier section, as shown in Figure 5-40. However, the receiver must be operated such that the output of the converter stage is linearly proportional to the RF signal voltage acting on the converter grid up to large signal amplitudes. The local-oscillator output must be high enough to ensure that this condition holds true. To prevent overloading of I-F stages, the I-F attenuator must be located immediately following the converter tube. The input and converter stages must be able to handle the signal levels encountered. The I-F attenuator method of measuring induced voltages has the advantage that accuracy is independent of signal frequency, depending primarily on the calibration accuracy of the I-F attenuator in conjunction with the linearity of the receiver input circuits. This method is very dependable because it operates at a fixed frequency which is considerably lower than the signal frequency.

5-6.12.4 Vacuum-Tube Voltmeters

If the field strength is high enough, a vacuum-tube voltmeter can be used successfully. However, as field-strength measurements tend to be cumulative, the vacuum-tube voltmeter must have an accuracy of 1 percent. Other than this requirement, measurement by VTM is the same as for the receiver and microammeter method.

5-6.12.5 Thermocouples

When a thermocouple is employed for measurement purposes, the procedure for measuring and comparison is the same. However, the RF current to be measured heats a thermal junction, which produces a dc output current that can be read on a sensitive microammeter. The advantages of using a thermocouple include accuracy, stability, and calibration which is not dependent upon frequency (except at very high frequencies).

5-6.12.6 Bolometers

In the ultra-high-frequency range (300 to 3000 MHz), it is more practical to measure power rather than voltage or current. Therefore, bolometers employing fine wire (Wallaston wire) are used instead of voltmeters or thermocouples. The thin platinum wire used changes its resistance with any change in current flow. Therefore, the bolometer may be used as one arm of a bridge circuit that is balanced by direct current flowing through the bolometer. A small high-frequency current may be superimposed on the bolometer, causing the bridge to become unbalanced because of the resistance change. This, in turn, provides galvanometer deflection. The dc bias current passing through the bolometer should be adjusted to provide the maximum output RF sensitivity. Thermistor beads may be substituted for the Wallaston wire if larger amounts of RF power are to be measured.

5-6.12.7 Crystal Rectifiers

The actual voltage induced in a dipole antenna can be measured with the aid of a crystal rectifier. As shown in Figure 5-43, a selected silicon-crystal rectifier is mounted across the gap at the center of the dipole, and a balanced resistance-capacitance network filters the dc output and prevents RF pickup on the dc line to the crystal. The dc line should be balanced, and the dc voltage may then be measured by any standard low-voltage technique. Since the dc

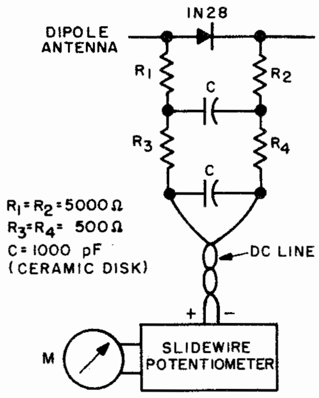

Figure 5-43. Calibration of Dipole Antenna with a Crystal

output resistance of a germanium crystal increases with the RF input resistance, germanium crystals or other types with germanium characteristics cannot be used in such measurements.

5-6.12.8 Standard Oscillators

The oscillator must be adjusted for the correct amplitude modulation and have an output-level control adequate to meet the varying requirements of field-strength comparisons. The unmodulated high-frequency source should have a single-frequency output free from harmonics. When the oscillator is used with instruments of the rectifier-type, such as a vacuum-tube voltmeter, large measurement errors will result if frequencies other than the fundamental are present. The technician must shield and ground the high-frequency source, and the source should not be overly dependent on the load impedance.

5-6.12.9 Frequency Considerations

5-6.12.9.1 LF and MF Radiation Measurement Methods

Radiated power measurement is generally obtained by comparing the measured field strength with the calculated field strength of a theoretically ideal antenna of the same height. For a simple ground-based vertical antenna, the equivalent power for ground-above-field strength may be obtained at varying distances from the antenna (if the antenna is less than one wavelength wave transmission) by measuring the ground strength. The radiated power of field strength at a single point may thus be calculated from the formula:

$$P = \frac{(E_1)^2}{E_2}$$

where:

E_1 = measured inverse-distance field strength at 1 kilometer

E_2 = field strength at 1 kilometer from an ideal ground-based vertical antenna with the same effective length as the actual antenna and a radiated power of 1 watt

5-6.12.9.2 Errors

Unfortunately, in using the above formula it is assumed that perfect measuring conditions exist. However, one measurement taken at a single point will not suffice in practice. Other sources of errors include: inaccuracies in determining the actual conductivity; errors in the receiving point's field strength value, due to directivity of the radiated pattern; uneven terrain and nearness of overhead wires; inaccuracies in specifying the exact distance between the antenna and receiving point; and, finally, errors in the field-strength meter calibration, plus the errors made by the observer in taking the reading. Field-strength readings should therefore be taken on a number of equally distributed radials, extending from about 1/2-mile to about 15 miles from the antenna. Shorter distances may be used for frequencies about 1600 kHz, and longer distances below 500 kHz. The smoothness of the measured data curve depends on the number of radials used and the number of points measured. The field-strength data collected from these measurements should be plotted on log-log coordinate paper as field strength versus distance. When this plot is compared with standard curves (plotted in table or graph form) of ground-wave field-strength versus distance, the equivalent attenuated field strength (E_1) can be determined. The value of E_2 can be obtained from a standard curve for an ideal antenna of the same length. The standard curve is not a plot of the output in millivolts versus the antenna length in wavelengths, but gives the unattenuated field-strength at 1 km from an ideal ground-based vertical antenna radiating energy at the rate of 1 watt.

5-6.12.9.3 Radiation Efficiency

That fraction of the input power actually radiated by an antenna is defined as the "radiation efficiency". The following formula may be used to determine this power fraction:

$$N = 100 \frac{Pe}{Pi}$$

where:

Pi = antenna input power
Pe = equivalent radiated power

5-6.12.10 HF Radiation Measurement Methods

Directional characteristics become factors of measurement considerations at the higher frequencies. At any given direction and distance from the transmitter, the field strength can be determined. The effective radiated power in that direction can then be calculated with the aid of the following formula:

$$Pe = \frac{Eo^2 d^2}{30}$$

where:

Pe = effective radiated power (in watts)
d = distance in meters
Eo = direct wave field strength (free space) in volts per meter

The height, distances, and vertical directivity at higher frequencies may permit measuring the direct field when the ground-reflected wave is small. If this reflected wave cannot be neglected in the computation, the technician can measure and compare the field strength at the receiver as radiated by the measured antenna, and the radiated field strength of a standard antenna at the same height and position. The standard antenna usually consists of a loop of thin dipole operating at frequencies up to 50 MHz. If a thin half-wave dipole is employed as the standard antenna, the power gain (relative to an isotropic radiator) is 1.64 times the power input to the half-wave dipole. In this application, the standard antenna should be mounted at the height of the electrical center of the antenna under test and at a sufficient distance from the antenna to prevent distortion. The standard antenna must be coupled to the transmitter through a coaxial cable, making use of a balance-to-imbalance transformer. This way, one transmitter may serve both antennas, in which case measurements should be made at several points along the line of maximum power to obtain an average value.

5-6.12.11 Microwave Radiation Measurement Methods

Measurements at the higher frequencies are more critical, hence the equipment required is substantially more complex because inductance and capacitance lose their identities within transmission line elements. Microwave receivers usually consist of silicon-crystal detectors used as mixers in conjunction with wide-band I-F amplifiers. Special triodes, klystrons, and magnetrons are used in the RF oscillators. Attenuators frequently are comprised of short sections of waveguide into which a conducting card is inserted to increase the loss. The conducting card is usually a 4-terminal-type network which creates an "insertion loss." Waveguide sections operating below the cutoff frequency may be used as attenuators at intermediate frequencies and at frequencies up to about 10,000 MHz, provided that correction is made for the fact that the operating frequency is nearly equal to the cutoff frequency. In the UHF band (300 to 3000 MHz) and at higher frequencies, it is more practical to measure power rather than current or voltage. In many applications, it may prove convenient to evaluate the incident field in terms of power density (watts per square meter) rather than the electric field strength (volts per meter). The effective radiated power at microwave frequencies can be determined by using the same methods as discussed previously under HF Radiation Measurement Methods (Paragraph 5-6.12.10). This means that the effective power radiated by the standard antenna in free space, in the direction of maximum radiation, is equal to the absolute power gain of the standard antenna times the power input to the standard antenna.

5-6.12.12 Alternate Antenna Method

In this method, the field strength (or power density) is measured in a manner similar to that used at the lower frequencies. Figure 5-44 illustrates a standard antenna input and a receiver input matched to the transmission line impedance. At higher frequencies, a coaxial cable or waveguide must be employed in lieu of the transmission line, and the antenna must consist of a feedhorn, a half-wave dipole, or some other structure having known properties. The computation involves measurement of the received power (Pr) flowing through a point at the end of the transmission line. If the absolute power-gain of the receiving antenna system, including the transmission line losses, is known, then the power density, in watts per square meter, of the field at the receiving antenna can be calculated, using the following formula:

$$\text{Power density} = \frac{4\pi Pr}{Gr\lambda 2}$$

where:

- π = 3.14
- Pr = received power (in watts) into a matched load
- Gr = absolute power gain of receiving antenna system (relative to an antenna radiating uniformly in all directions)
- λ = wavelength in meters

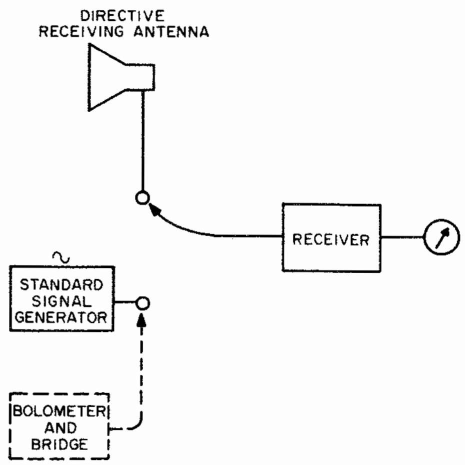

Figure 5-44. Standard Antenna Method at Microwave Frequencies

After computing the power density, calculate the electric field intensity, in terms of volts per meter, with the aid of the following equation:

$$Eo = \frac{6.3 \quad Pr}{\lambda \quad Gr}$$

5-6.12.13 Standard Field Generator

A standard-field generator may be used as a transmitter operating in the microwave region. This standard-field measurement method is essentially the same one that is used at lower frequencies, except for the actual computation. The free-space electric field may be calculated from the following formula:

$$Eo = \frac{30 \text{ Pt Gt}}{D}$$

where:

- Pt = power input (watts) to the transmitting antenna
- Gt = power gain of the transmitting antenna relative to an isotropic radiator (an ideal antenna radiating in all directions)
- D = distance between transmitter and receiver (meters)
- Eo = free-space electric field strength (ms volts per meter)

5-6.12.13.1 Test Site

A level test site, free from reflecting objects, is essential for calibration purposes at microwave frequencies. The distance between the transmitter and receiver must be made small in comparison with the antenna heights above ground, and the distance to the nearest reflecting object must be relatively great. Free-space fields can be established under such conditions. If relatively great distances between antennas are contemplated, the height of both antennas must be measured, and this factor must be included in the computation to determine the electric field strength. The assumed coefficient of reflection of the ground plane must be -1, and the angle of incidence of the ground-reflected wave and the beam width of the transmitting and receiving antennas must be such that there is no discrimination between the direct and ground-reflected waves. If the test site is on uneven ground, the electric field at the receiver location will be distorted, and its measurement accuracy will be questionable. This distortion will be minimized if one or more straight parallel wire fences are installed at right angles to the path of transmission between the antennas. These wire fences should be high enough to shield the receiving or transmitting antenna from direct ground reflections. The computation will then be made on the basis of free-space transmission, using the above formula.

5-6.12.14 Test Equipment Calibration

Test equipment for measuring field strength must first be calibrated in terms of a signal of the same level as the signal to be measured. The standardization of an antenna, for example, is effected by calibrating the voltmeter in terms of known values of the incident field, except in direct substitution methods where only a comparison is required.

5-6.12.15 Loop Antenna Standardization

A loop can be standardized by comparing the induced voltage in the receiving loop antenna with a known RF voltage from a standard signal generator, connected as shown in Figure 5-45. The receiver need not be calibrated because it acts only as a comparison voltmeter. However, the loop must be oriented and tuned such that the received signal produces maximum output. The receiver's attenuator need be set only once to produce the desired arbitrary deflection on the output meter. The loop is then turned at right angles to its previous position so that minimum output is indicated on the meter. The signal generator is then turned on and tuned to zero beat with the residual measured field. The standard voltage is adjusted to produce the same meter deflection as the unknown field. At this point, the signal generator voltage and the induced voltage acting in series with the loop are equal. In the particular case where the natural loop frequency is several times greater than the measured signal frequency, the field strength in volts per meter is determined by the following formula:

$$Eo = \frac{V}{L}$$

where:

- Eo = field strength (volts per meter)
- V = standard signal generator output (volts)
- L = effective length of loop antenna (meters)

The length (L) can be determined by the following equation:

$$L = 0.0296 \times F$$

where:

- F = frequency (megahertz)

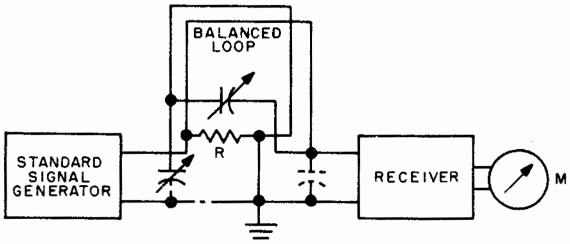

Figure 5-45. Standardization of Loop Antenna

5-6.12.15.1 Limitations

The above calculations for field strength are effective only if the signal generator output impedance is very low so as not to excessively reduce the Q of the loop. The signal generator output is connected to resistor R, which is in series with the loop antenna. This series resistance should be maintained for either the unknown radiation or the standard calibration voltage. Tuning the loop antenna should therefore not affect the generator voltage value appearing across this resistance. If the calibrating voltage is constant, a calibrated attenuator can be used between the mixer and first I-F stage of the receiver. The reading can then be compared with the voltage induced by the field. Interpolation between steps of the voltage attenuator is accomplished by a linear indicator, such as a diode rectifier-and-meter combination. Another calibration method involves applying the calibrating voltage to the receiver input or vacuum-tube voltmeter input and setting it equal to the loop voltage. The loop Q must then be measured to obtain the field voltage induced within the loop. The field strength can be computed from the induced field voltage and the effective antenna length.

5-6.12.16 Dipole Antenna Calibration

For calibration of a dipole antenna, the dipole is removed, and the signal generator is activated to supply a calibrating voltage to the end of the transmission line (see Figure 5-46). The amplitude of this

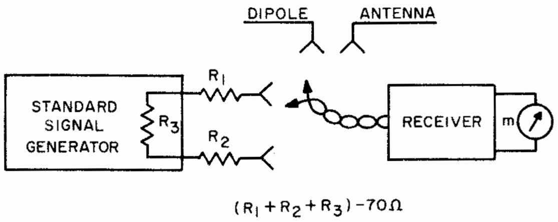

Figure 5-46. Calibration of Dipole Antenna

voltage is then varied to produce the same meter deflection as the unknown field. Therefore, the known voltage of the signal generator is equal to the unknown voltage induced in the antenna by the field. The field strength can then be derived from the following equation:

$$Eo = \frac{Voc}{L_D}$$

where:

Eo = field strength (volts per meter)
Voc = open-circuit voltage at output of the standard generator
L_D = effective length of dipole (in meters)

Shorter than the c wavelength, L_D is about 1/2 of the actual physical antenna length.

NOTE

The radiation impedance of an antenna depends on its height above ground, but is also affected by the presence of other reflecting surfaces. Therefore, the formula given above is in error for heights less than one wavelength.

5-6.12.17 Silicon-Crystal Rectifier Calibration

An additional method of calibrating a dipole, over a limited range of amplitudes, employs direct measurement with a silicon-crystal rectifier of high impedance, as shown in Figure 5-43. Antenna impedance need not be known, and the ground conditions under the antenna will not affect the antenna voltage-transfer ratio. However, for best results, the crystal voltmeter must be recalibrated. Since the antenna is not selective, the silicon-crystal method cannot be used close to any interfering station which produces a field comparable to that of the measured field.

5-6.12.18 Antenna Horn Calibration

Normally, electromagnetic horns must be calibrated for gain by comparison with that obtained from a standard antenna of known gain. The unknown antenna must therefore be connected to a receiver containing a variable attenuator, as shown in Figure 5-47. The receiver is tuned to a distant signal source and the antenna oriented for maximum signal. The sensitivity and attenuation are then adjusted for a convenient reading, and this reading is recorded as P_M. The antenna is then disconnected and the standard generator attached to the receiver terminals previously

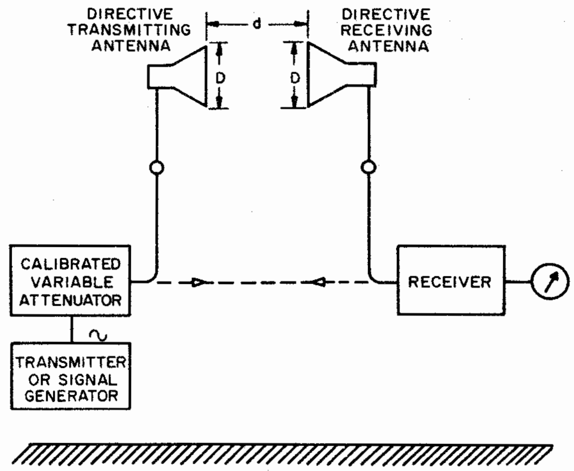

Figure 5-47. Calibration of Horn Antenna

used by the disconnected antenna. Without changing the receiver settings, the receiver horn is then oriented to obtain maximum signal strength. This meter reading is recorded at Ps. The following formula then provides the unknown antenna gain:

$$G = \frac{Pm\ Gs}{Ps}$$

where:

Gs = standard antenna gain

NOTE

If the power is large enough, a bolometer may be used to measure directly, as shown in Figure 5-48.

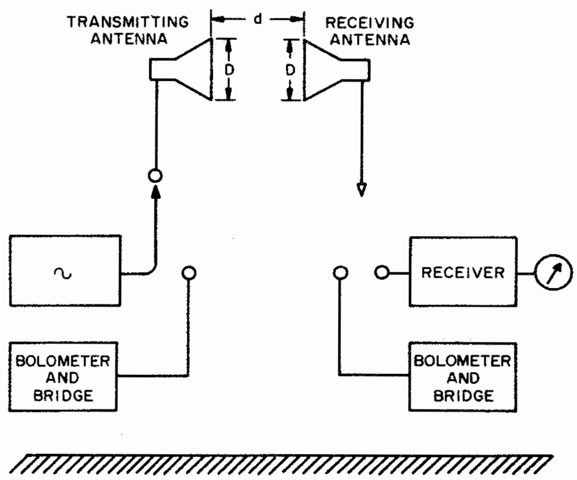

Figure 5-48. Calibration of Standard Horn Antenna

For the above calibration, a flat test-site, free of reflecting objects, must be selected, and the antenna must be located above ground at a sufficient height to make the effect of ground reflections negligible. If necessary, the effect of ground reflections may be reduced by using one or more of the previously discussed parallel diffraction edges placed perpendicular to the transmission path between the antennas.

5-6.12.19 Voltage-Transfer Ratio

The voltage-transfer ratio (antenna coefficient) is the ratio of the measured voltage at the field-intensity receiver to the voltage produced in the antenna by the measured field. This ratio depends primarily on the type of antenna used, the method of coupling the antenna to the transmission line, and the characteristic impedance and attenuation of the transmission line. At frequencies between 10 MHz and 30 MHz, this voltage ratio can be measured by the standard-field method, using a transmitting loop antenna of known current. The receiving loop antenna to be measured is placed a short distance away, and the response of the antenna and indicating receiver is determined. Since the field at the receiving antenna can be computed from the foregoing formulas, and since the receiver response can be determined from the signal-generator measurements, the voltage-transfer ratio can be derived from the measured data. At frequencies from 30 to 300 MHz, dipole antennas are used in conjunction with field-strength meters. The voltage-transfer ratio can be determined by either the standard-field or the standard-antenna method. These measurements are made at distances which are great in comparison with the wavelength so that the radiation field is the predominant component of the total field. To determine the voltage-transfer ratio of either the standard-antenna or standard-field method, first ascertain the field strength of a locally generated field by measuring the induced voltage in a standard dipole. The antenna to be measured is then substituted in the same location as the standard antenna, after which the measurement procedure is performed as previously described.

5-6.12.20 Receiver Indicator Calibration

Calibration of a receiver or other detecting device merely involves standardizing an output indication in terms of a known RF input signal. This is done by introducing an input signal and determining the output reading as a function of the input. The overall receiver gain must be held constant so that a given meter reading corresponds to a unique input signal. If the receiver characteristics are highly linear, a single point is sufficient to calibrate the receiver. However, a more accurate calibration will be achieved if the measurements are accomplished at the actual signal level. At the signal level to be measured, an accurate attenuator is provided in either the I-F or RF receiver section to adjust the reference signal level at the detector. This ensures obtaining the same response as that of the signal to be measured.

5-6.13 ACCURACY OF FIELD-STRENGTH MEASUREMENTS

5-6.13.1 Sources of Error

Field-strength measurement errors fall into two distinct types: errors which occur in all cases; and errors associated only with certain techniques of measurement. Some of the primary sources of error that must be eliminated or minimized are as follows: incorrect calibrating voltage; nonlinearity in attenuators; induction of voltages in the antenna by unshielded receiver components; receiver regeneration; inaccurate introduction of the calibrating voltage due to antenna unbalance; distortion of the field by reflecting objects; distributed-capacitance effects in the loop antenna; inadequate data on local conditions, such as ground constants; spurious frequencies in the output of the voltage generator; and nonlinearity of the detector, mixer, or other output stages adversely affecting the gain.

5-6.13.2 Minimizing Errors of Field-Strength Measurements

The technician must make sure that the voltage-generator output is both sinusoidal and free of spurious frequencies in order to prevent a distorted waveform output. The linearity requirements of the detector mixer, or other output stages must also be closely checked to prevent any nonlinearity in the gain which would produce an output-reading error. An overall facility error could occur because of a faulty attenuator or an improperly matched attenuator. Unshielded nearby equipment may induce voltages into the antenna to influence the field strength. An adequate coupling length must also be provided between the antenna and receiver to prevent extraneous coupling. An error in calibration can exist because of the Q-factor or difference in step-up voltage of a tuned-loop antenna when used in a uniform space field as compared with the condition where a lumped-calibration voltage is inserted at the center or end of a loop. This error, which could amount to 15 percent, is actually due to the distributed capacitance of the loop. The technician can partially compensate for this error by operating the loop antenna in an unbalanced condition with respect to ground. The antenna must not become responsive to unwanted polarization components. An untuned loop may be employed to reduce the distributed capacitance error, although the additional amplification required and attendant noise will result in a signal loss by a factor of 1/Q.

5-6.13.3 Expected Values of Accuracy

The expected accuracy in field-strength measurements is a function of operating conditions and equipment serviceability. However, accuracies of 5 percent at frequencies below 3 MHz and up to 20 percent above 3 MHz may be obtained.

5-6.13.4 Final Error Measurement Data

In view of all the sources of error discussed above, uncorrected field-strength measurements are of little value unless the following information accompanies the field-strength survey data: the field component measured; a physical description of the set-up (topography, distances, heights, reflection sources, etc.); a description of the equipment and procedures; the measurement values employed (rms, peak, average, etc.); and the nature of the modulation, if any.

5-6.14 ATMOSPHERIC EFFECTS

Field-strength or noise measurements are complicated by the variation of field caused by a varying transmission medium, by emission complexities, by ground effect, and by nearby structure reflection. All the field components at the receiving antenna could be measured, however, if a complete measurement of wave fields transmitted via the ionosphere or troposphere is desired. However, in everyday practice, the predominant component of the field is generally selected for measurement and analysis.

5-6.14.1 Ionospheric Measurement

The ionosphere is an ionized region in the atmosphere which begins at approximately 80 kM above the earth's surface, extends upward for a considerable distance, and varies in height as a function of time. It is characterized by a nonuniform electron distribution, with maximum electron density occurring at three distinguishable elevations or layers. Since the properties of the ionosphere directly affect long-range radio-wave propagation, measurements of the elevation and maximum electron density of each layer are of very definite interest in field-strength measuring operations.

5-6.14.2 Ionosphere Measurements Using Oscilloscope

An oscilloscope can be used to display the pulsed signal reflected back from the ionosphere. For normal incidence measurements, the transmitter and receiver are usually located next to each other. The receiver output must be applied to the vertical-deflection plates of the cathode-ray tube, and a timing wave synchronized with the transmitted pulses applied to the horizontal-deflection plates. The time delay between transmitted and received pulses is directly related to virtual height. At the velocity of light, the lapse of 1 millisecond is equivalent to a 300 kM round trip, or a virtual ionospheric height of 150 kM. Actually, this virtual height is slightly greater than the actual height because wave velocity decreases with increasing electron ionospheric density. This difference is only slight, however, except at critical frequencies. When the RF pulse frequency is varied, the height of the layer will appear to jump suddenly at a particular frequency. This type of discontinuity corresponds to the critical frequency, at which the wave is just barely able to penetrate the layer. The maximum electron density of a layer can be determined from the following formula:

$$Ed = Fc^2$$

where:

Ed = maximum electron density of layer in electrons/cc

Fc = critical frequency in kHz

The transmitter referred to above is usually of the pulse-modulated type, with a pulse length of about 100 microseconds and a pulse repetition rate of 60 hertz. The superheterodyne receiver should have a manual volume control and an I-F bandwidth range of 10 to 40 kHz. The frequency of the transmitter and receiver must be varied simultaneously to obtain a multifrequency type of measurement. Therefore, the horizontal deflection of the oscilloscope corresponds to frequency, while the vertical deflection, which is produced by a sawtooth voltage synchronized with the transmitted pulse, is proportional to time and, therefore, to distance. The receiver output must be employed to intensity-modulate the cathode-ray tube such that a spot or trace appears only when a pulse is received. The position of the return along the vertical axis is determined by the time-lapse between the transmitted and received pulses, and is therefore a measure of the virtual height.

5-6.14.3 Atmospheric Noise Measurements

The measurement of fields representing noise or static may be made in the same manner as normal field-strength measurements. However, noise signals can vary from intermittent, random impulses to more continuous radiations which vary in strength daily, hourly, or seasonally. Since this is especially the case with atmospheric noise, accurate noise field-strength measurements are difficult to achieve. An indirect measurement of atmospheric noise can be conducted to determine just how weak a particular type of signal can be made before the noise renders it unintelligible. A direct measurement can be obtained by means of a field-strength receiver and automatic recorder. For both impulse noise and smooth fluctuation noise, the rms voltage output of a receiver is proportional to the square root of the bandwidth. If the noise is composed of sharp, widely separated pulses, the peak amplitude of the noise output is directly proportional to the bandwidth, or to the square root of the bandwidth in the case of a large number of overlapping pulses. The average amplitude differs from both the rms and peak amplitudes, being independent of bandwidth in the former case, and being proportional to the square root of the bandwidth in the latter case. Therefore, it is evident that any atmospheric-noise measurement made by the equipment is a function of both the bandwidth and the time-constants of the equipment. A typical noise-measuring equipment consists of a linear detector whose output connects to an automatic recorder through a dc amplifier which has a charge and discharge time-constant of one minute. The readings are proportional to the average value of the RF envelope, and the recorder is calibrated in terms of the rms value of a standard CW voltage. Because of the nature of the signal, the measurement of noise peaks against time duration is also necessary. Since atmospheric noise measurements are of interest for all broadcasting hours, the percent of time that a number of different field-strength levels are exceeded by the atmospheric noise should be plotted against various times of day.

5-6.15 MOBILE AND AUTOMATIC RECORDING

When measuring the field pattern and strength of transmitting antenna at various azimuth angles and distances, the measuring equipment can be mounted in a vehicle to provide ease of relocation. The antenna should be well clear of the vehicle because the metal body of the vehicle will cause distortion in the surrounding field and consequent measurement errors. The effect of the vehicle on the receiving antenna should be evaluated by making comparison tests with the antenna mounted with and without the vehicle at a common height. If measurements are obtained while the vehicle is in motion, standing waves will be present due to reflections from telegraph lines, power lines, fences, etc. Therefore, the tabulated data provides only an approximate representation of the field if the maxima and minima average are noted. If the received signals are pulse-modulated, the pulse rate must be made high enough so that the receiver detector can distinguish between the pulse peaks and the maxima of the varying standing-wave pattern. The automatic recorder derives its input from the rectified dc output of the I-F amplifier, which is normally used for automatic volume control of the amplifier. The dc input from the AVC detection process will cause a recorder-pen deflection proportional to the dc signal strength. The relation between the recorder deflection and the field strength is substantially logarithmic. The time-constant of the automatic volume control must be adjusted to permit other field-strength measurements in terms of either instantaneous or average values.

5-7 TIME DOMAIN REFLECTOMETRY

5-7.1 GENERAL

Time Domain Reflectometry is a testing and measurement technique which has found rapidly increasing usefulness in the testing of transmission lines, cables, strip lines, connectors and other wideband systems or components. Basically, it is an extension of an earlier technique in which reflections from an electrical pulse were monitored to locate faults and to determine the characteristics of power transmission lines. With the modern development of extremely fast-rise pulse generators and oscilloscopes with equivalent bandwidth, a great increase has occurred in the applications of pulse-reflection measurement methods to high-frequency systems. The arrangement used in time domain reflectometry may be compared to a closed-loop radar system in which the transmitted signal, a very-fast step-pulse, is fed into the system and the reflections resulting from discontinuities or impedance deviations in the system are monitored on an oscilloscope screen. Basic equipment requirements are a very-fast-rise pulse generator, together with a single-channel sampler and a time-base generator. Because the speeds required for time domain reflectometry are much greater than the response of the fastest real-time oscilloscope, signal-sampling techniques are used. In typical commercial reflectometry equipment, a series of 4000 separate samples of the reflected energy are taken every 20 milliseconds. Although each sample appears as a single dot on the oscilloscope screen, the entire sampling combines to form a complete display. The technique utilized in time domain reflectometry consists essentially of feeding a step or impulse of energy into the system and the subsequent observation, also at the point of insertion, of the energy reflected by the system. When the fast-rise input pulse meets with a discontinuity or impedance mismatch, the resultant reflections appearing at the feed-point are compared in phase, time, and amplitude with the original pulse. By analyzing the magnitude, deviation, and shape of the reflected waveform, the nature of the impedance variation in the transmission system can be determined. Also, since distance is related to time and the amplitude of the reflected step is directly related to impedance, the comparison indicates the distance to the fault, as well as an indication of its nature. In addition to showing both the position and the nature (resistive, inductive or capacitive) of each line discontinuity, time domain reflectometry also reveals the characteristic impedance of the line and indicates whether losses are shunt or series losses. A conventional method of evaluating high-frequency transmission systems and components has been the use of standing wave ratio (SWR) measurements to obtain an overall indication of transmission line performance. This method involves feeding a sine-wave signal into the system, and then measuring the maximum and minimum amplitudes of the standing waves which result from the system discontinuities or load mismatches. The ratio between the maximum and minimum values (SWR) is then taken as the system figure of merit. The SWR measurement, however, does not isolate individual discontinuities or mismatches where multiple reflections are present, but only indicates their total effect. Time domain reflectometry (TDR) measurements, on the other hand, isolate the line characteristics in time (location). As a result, multiple reflections resulting from more than one discontinuity or impedance change which are separated in distance on the line are also separated in time at the monitoring point and can be individually analyzed. The combination of modern, very-fast-rise pulse generators and the newer generation of sampling oscilloscopes improves test response to the extent that reflections from discontinuities separated by a small fraction of an inch may be isolated and resolved.

5-7.2 TEST SETUP

Although several suitable arrangements are possible, time domain reflectometry measurements generally utilize an equipment setup similar to that shown in Figure 5-49. The fast step, developed in a pulse generator that is properly matched to a 50-ohm line, is passed through a feed-through sampler and along a short length of coaxial cable to the transmission system or element being tested. A 50-ohm matching impedance is most frequently employed because of its common usage and availability of parts and calibration standards. The short length of coaxial cable serves to move system reflections away from the input step. The oscilloscope is connected to a high-impedance sampling gate which bridges the feed-through section of the 50-ohm line, so that both the incident step and system reflections are displayed on the oscilloscope screen. When the system's terminating load is equal to the characteristic impedance of the system, no reflection will occur. The oscilloscope display will then show only the voltage step presented as the input pulse passes the monitoring point. Also, with a pure resistive load on the output of the tested

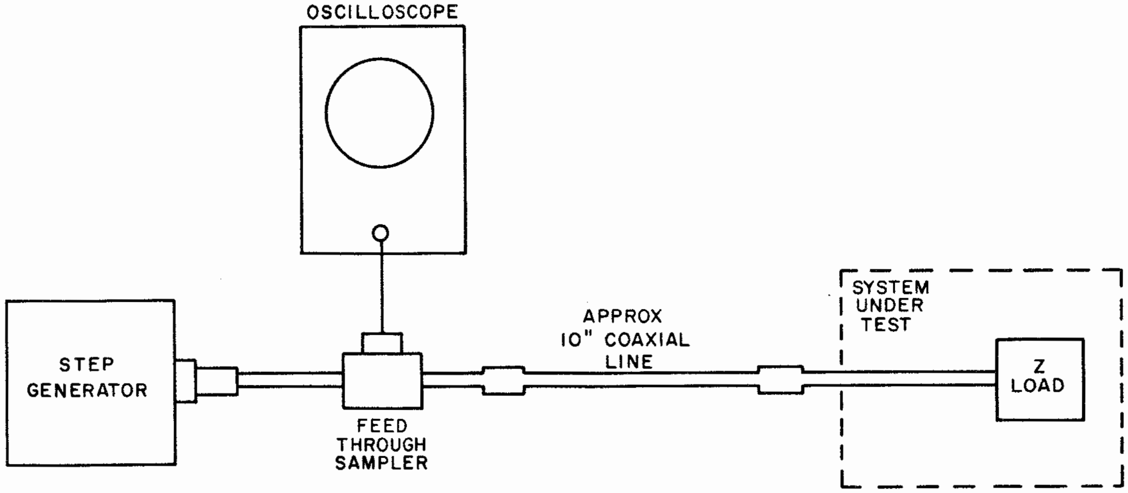

Figure 5-49. Time Domain Reflectometry Basic Equipment Setup

system, the height of the step signal observed on the oscilloscope is a function of the resistance. In the case of reactive loads, the waveform presented will be determined by the time-constant of the load and source impedance. The shape and magnitude of the reflections indicate their nature and the value of the R, L, or C causing the mismatch. A single fault will generally be indicated as a small, positive bump (or negative dip) on an otherwise flat trace. A positive (upward) bump will indicate a rise in the line's characteristic impedance over a short section, or the presence of series inductance at a point in the line. The time required for the reflection to return to the sampling gate will locate the discontinuity. Similarly, a negative (dip) indication shows a lowering of the line impedance at some point, or a shunt capacitance. The three waveforms shown in Figure 5-50 represent idealized oscilloscope presentations of resistance, capacitive, and inductive discontinuities. The distance (x) that each is located from the feed-through sampler

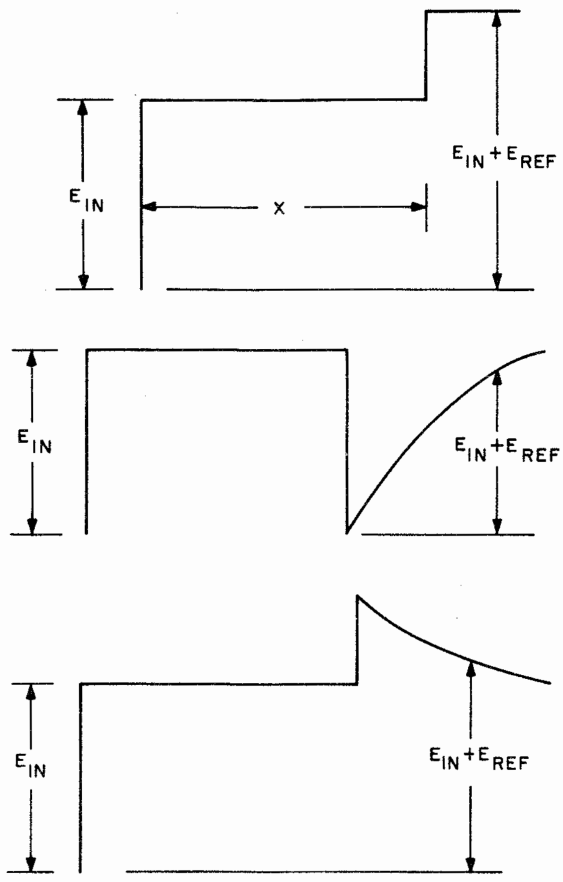

Figure 5-50. Basic Waveforms Representing Resistive, Capacitive, and Inductive Discontinuities

can be measured directly by means of a calibrated time-base on the oscilloscope display. Because discontinuities separated by distance in the system will present reflections separated in time on the oscilloscope display, the time domain reflectometry technique permits measurement of an individual impedance mismatch without interference from others which may be present. The procedure employed in making TDR measurements generally begins by connecting the step-pulse generator by way of the feed-through sampler to the oscilloscope, and then via a short length of 50-ohm coaxial cable to an accurate 50-ohm termination, which simulates the system to be tested. This termination can be a 1 percent deposited-film resistor. Separating the signal input from the system under test by means of at least 10 inches of coaxial cable moves the system reflection(s) away from the leading edge of the input step indication. With the accurately-known value of the terminating resistor in place, the oscilloscope vertical gain is adjusted to provide a calibrated height equivalent to the impedance of the load. Height variations above or below this calibrated value then indicate system impedance variations. With the oscilloscope thus calibrated, the system impedance can be determined from the height of the reflected step. Calibration of the vertical axis in terms of reflection coefficient permits direct interpretation in terms of VSWR measurements. A graph may be constructed for use in translating step-height indications into equivalent impedance. Similarly, calibration of the horizontal oscilloscope deflection against a known cable length allows rapid physical location of system faults or discontinuities. In summary then, time domain reflectometry offers the ability to simultaneously display the transmission characteristics of a transmission system over a very broad band of frequencies and to isolate individual discontinuities so that they may be properly corrected or compensated for. TDR is applicable to the direct measurement of such cable parameters as characteristic impedance, series or shunt losses, and length. It is also useful in the measurement of broad-band reflection coefficients, evaluation of individual components, and many specialized tasks. Increasing use of the technique will result in many additional applications becoming known.

5-8 INSERTION LOSS CONSIDERATIONS

5-8.1 GENERAL

Insertion loss has been defined as the loss in load power due to the insertion of a component or device at some point in a transmission system. Thus, when a network is inserted between a generator and a load, the ratio of the current flowing in the load impedance and that current which would flow in the load impedance with the network removed and the generator connected directly to the load is the insertion loss. It is important to be aware that this loss exists and has the effect of attenuating the load current irrespective of whether image-impedance matching exists. Image operation corresponds to the condition where the power delivered by the generator has the maximum possible value, and the load resistance has the value that will absorb the maximum possible power. In other words, the impedances (assumed to be resistive) at the input to the transmission system are the same when looking backward into the generator, and forward toward the network. Similarly, the impedances (again assumed to be resistive) at the output are the same looking backward toward the network, and forward into the load. Although discussed above in terms of a current (or voltage) ratio, insertion loss is commonly given in terms of the equivalent power ratio and is expressed in decibels. A small modification of the insertion loss occurs when there is mismatching at both the input and output of the transmission system. It does not occur, however, if the generator resistance matches the input image resistance or if the load resistance matches the output image impedance. When a passive four-terminal circuit is introduced between a signal source and a terminating load, and the power (Po) delivered to the load is reduced to P_2, the resulting attenuation is defined as:

$$A = 10 \log \frac{Po}{P_2} \text{ dB}$$

Since the attenuation is caused by the insertion of the four-terminal circuit, it is often referred to as insertion loss. Such attenuations are not only dependent upon the parameters of the inserted circuit, but also upon the load impedance and the impedance of the signal source. As indicated above, they may be entirely dissipative, entirely reflective, or in part dissipative and in part reflective. Consider, for example, the simple circuit shown in Figure 5-51. The source impedance (Z_G) and the load impedance (Z_L) are complex and mismatched. The current delivered is given by:

$$I_o = \frac{V}{Z_G + Z_L}$$

and the corresponding power delivered is:

$$P_o = I_o{}^2 R_L$$

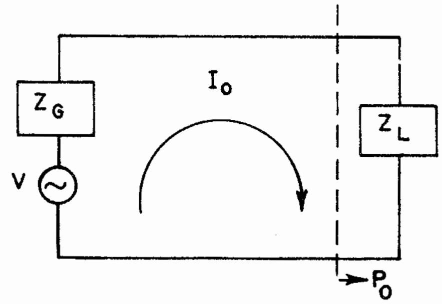

Figure 5-51. Simple Series Circuit of Two Mismatched Impedances

$$P_o = \frac{|V|^2}{|Z_G + Z_L|^2} R_L$$

where R_L represents the real part of the load impedance. Now insert a passive four-terminal circuit into Figure 5-51, as illustrated in Figure 5-52. Its equivalent circuit in terms of the open-circuit impedances of the four-terminal circuit is given in Figure 5-53.

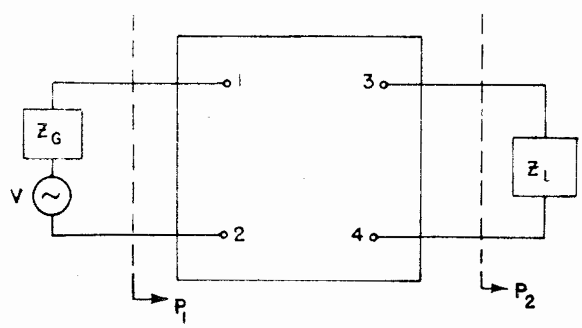

Figure 5-52. Passive Four-Terminal Circuit Inserted Between Two Mismatched Impedances

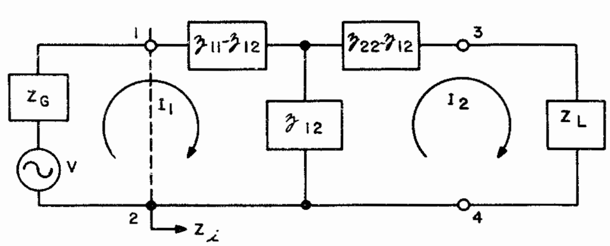

Figure 5-53. Open-Circuit Impedance Equivalence of Passive Four-Terminal Circuit of Figure 5-52

These open-circuit impedances are defined as follows:

Z_{11} = driving impedance of terminals 1-2 with terminals 3-4 opened

Z_{22} = driving impedance of terminals 3-4 with terminals 1-2 opened

Z_{12} = transfer impedance with terminals 3-4 open and terminals 1-2 excited with a driving voltage

From Figure 5-53, the driving-point impedance (Z_i) is computed as:

$$Zi = \frac{Z_{11}X_{22} - Z_{12}^2 + Z_{11}Z_L}{Z_{21} + Z_L}$$

When the passive four-terminal circuit is lossless, its open-circuit impedances are reactive functions. The power output is equal to the power input since no loss of power occurs in the circuit. Thus, from Figure 5-53:

$$P_2 = |I_2|^2 R_L$$

or

$$P_2 = |I_2|^2 R_L = |I_1|^2 Ri$$

where R_1 is the real part of Z_1 contained in the formula above, and the current (I_1) is given by the equation:

$$I_1 = \frac{V}{Z_G + Z_i}$$

Therefore, the output power becomes:

$$P_2 = \frac{|V|^2}{|Z_G + Z_i|^2} Ri$$

Consequently, the insertion loss (A_r) can be found in terms of circuit impedances by substituting the equations for P_o and P_2 in the following equation:

$$A_r = 10 \log \frac{P_o}{P_2}$$

$$A_r = 10 \log \frac{|Z_G + Z_i|^2}{|Z_G + Z_L|^2} \cdot \frac{R_L}{R_i}$$

This last equation represents reflective attenuation, and it may be transformed into a more convenient form by introducing the reflection factor of current (I_1) which is defined by:

$$\rho = \frac{Z_G - Z_i}{Z_G + Z_i}$$

For the usual case, where the source and load impedances are purely resistive (that is, $Z_G = R_G$ and $Z_L = R_L$), the equation for insertion loss reduces to:

$$A_r = 10 \log \frac{4R_GR_L}{|R_G + R_L|^2} \cdot \frac{1}{1 - |\rho|^2}$$

where the reflection coefficient is:

$$\rho = \frac{R_G - Z_i}{Z_G + Z_i}$$

Only under matched conditions, where $R_G = R_L$ and $Z_i = R_G$, does the reflective attenuation or reflective insertion loss reduce to zero. When the four-terminal circuit of Figure 5-53 is lossy, the output power (P_2) can be evaluated by knowledge of the output current. It can be shown that this current is equivalent to:

$$I_2 = \frac{I_1 Z_{12}}{Z_{21} + Z_L}$$

By substituting for I_1, it becomes:

$$I_2 = \frac{Vz_{12}}{(Z_G + Z_i)\ (Z_{22} + Z_L)}$$

The insertion loss attenuation can now be represented by:

$$A = 10 \log \frac{|(Z_G + Z_i)(z_{22} + Z_L)|^2}{|Z_G + Z_L|^2 |z_{12}|^2} \text{ dB}$$

It is of interest to note that the factors influencing attenuation contribute both dissipative and reflective losses; they can be separated into corresponding dissipative and reflective attenuation terms. The dissipative attenuation term is found by matching the impedances. For the usual case, $Z_G = R_G = R_L = Z_L = Z_i$. The equation for the above insertion loss attenuation can be converted into the dissipative attenuation loss as follows:

$$A_{dis} = \left(\frac{z_{22} + R_G}{z_{12}}\right)^2$$

The difference between the insertion loss (A) and the dissipative insertion loss (A_{dis}) represents the reflective attenuation, or:

$$A_r = A - A_{dis} = 10 \log \left(\frac{Z_G + Z_i}{Z_G + Z_L}\right)^2 \left(\frac{z_{22} + Z_L}{z_{22} + R_G}\right)^2$$

A positive insertion loss represents power loss, while a negative insertion loss represents power gain. Power gain can occur in the cases described, since the amount of power delivered is dependent upon the impedance Z_L when the circuit is out, and upon the impedance Z_i when the circuit is in. To assure that an insertion loss is truly a loss, the power delivered to the load before the circuit is inserted must be defined as the maximum power which could be delivered to the load. The general expression for the insertion loss of a component is given by:

$$A = 10 \log \frac{P_o}{P_2} \text{ dB}$$

where:

P_o = delivered output power with the component out of the circuit

P_2 = delivered output power with the component inserted

These conditions were illustrated in Figures 5-51 and 5-52. The insertion loss can be separated into two components by the indroduction of P_i, as shown in Figure 5-52. Since:

$$\frac{P_o}{P_2} = \frac{P_o}{P_1} \cdot \frac{P_1}{P_2}$$

therefore:

$$A = 10 \log \frac{P_o}{P_1} + 10 \log \frac{P_1}{P_2}$$

The first component represents attenuation by reflection, or:

$$A_{ref} = 10 \log \frac{P_o}{P_1}$$

The second component is caused by energy loss, or:

$$A_{dis} = 10 \log \frac{P_1}{P_2}$$