CHAPTER 6

ELECTRONIC AIDS TO NAVIGATION: TACAN, AN/SRN-6 (PART II)

INTRODUCTION

The block diagram analysis of the overall Tacan Radio Set, AN/SRN-6, and the circuit operation of the system receiver was presented in chapter 5. A thorough study of the material in that chapter is recommended before attempting to study the contents of this chapter.

The circuit operation of the coder-indicator, frequency multiplier-oscillator, amplifier-modulator, and antenna system are presented in this chapter. It may be necessary, occasionally, to refer to the radio beacon block diagram (chap. 5, fig. 5-10) and the related discussion to become re-oriented on the overall beacon operation.

CODER-INDICATOR

The coder-indicator of the Radio Beacon, AN/SRN-6 (fig. 6-1) performs the following functions:

1. Generates the 15-cps reference burst each time a 15-cps trigger pulse is generated in the antenna.
2. Generates the 135-cps reference burst each time a 135-cps trigger pulse is generated in the antenna.

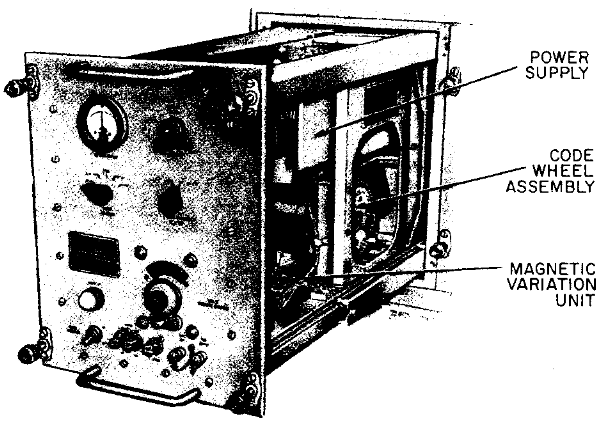

Figure 6-1.—Coder-indicator, overall view.

3. Generates a radio beacon identification call of 1350 cps in International Morse Code keying for transmission at specific intervals.
4. Processes the distance interrogation pulses appearing at the receiver output, and in combination with time delays in the receiver and transmitter, adjusts their net transit time through the beacon to exactly 50 μs.
5. Assigns priorities of transmission to the components of the signal, which consists of the bearing reference bursts, the radio beacon identification call, and the replies to distance interrogations and random noise pulses.
6. Combines the various components of the signal.
7. Encodes the transmission signals to give them the characteristics identifying them as the radio beacon signals.
8. Provides an accurate 1350 cps signal to be used in checking the accuracy of the speed at which the antenna parasitic elements rotate about the central array.
9. Provides a means for adjustment to compensate for variations between magnetic north and true north.

REFERENCE BURST GENERATOR STAGES

The reference burst generator stages (fig. 6-2) produce the reference burst signals which are transmitted by the radio beacon. The 15-cps reference burst generator stages include V601B, V602, and V603. The 135-cps reference burst generator stages comprise V601A, V609, and V610. The circuit operation of the two reference burst generator stages is the same except that the size of the frequency generating components in the separate circuits will differ. The following discussion of the 135-cps reference burst generator stages applies as well to the 15-cps reference burst circuit except where noted.

Pulse amplifier, V601A, receives the trigger pulses from the 135-cps pulsing coil of the antenna system. Since the stage does not employ grid-leak bias and the cathode is grounded, the bias on V601A is zero.

The trigger pulses from the antenna contain both positive and negative excursions. The positive portion of the pulse is generated first. When the input pulse to V601A is positive-going, grid current produces a voltage drop across R658, and the net signal to the V601A grid is held very near to zero volts. When the trigger pulse from the antenna is negative-going, a sharp change in the V601A plate current occurs. The resulting output of V601A is a sharp positive-going pulse, which is coupled by C624 to the grid of the 135-cps gate generator, V609.

V609 is a one-shot keying multivibrator which is used to provide a keying or gate pulse to the pulsed oscillator, V610A. Because of the circuit constants used in V609, the natural period of its output gate pulse permits the generation of six pulses in the 41.7 kc pulsed oscillator, V610A, for each output pulse from V609. These pulses, after amplifying and shaping, represent the 135-cps reference burst pulses.

The circuit operation of the V609 multivibrator is as follows: The B section of V609 is normally conducting. Although the A section grid of V609 is returned to the positive junction of R661 and R660 (with respect to ground), V609A (in the absence of an input trigger pulse) is held beyond cutoff by the voltage developed across the common cathode resistor, R665. This voltage is effective between the grid and cathode of V609A as bias.

The sharp positive pulse from V601A triggers V609A into conduction. The plate voltage decrease of V609A causes C625 to discharge through R663. The resulting negative-going potential developed toward the V609B grid by the drop across R663, decreases the conduction in this tube section. The drop in the V609B current through the common cathode resistor, R665, decreases the grid-cathode potential (bias) on the V609A section, and in conjunction with the trigger pulse, further tends to reduce the bias voltage applied to V609A. Thus, the current rises in V609A, with an accompanying drop in the V609B current, until V609A is conducting saturation current, and V609B is at cutoff.

When the discharge of C625 drops to a point where the negative voltage developed at the grid side of R663 is no longer sufficient to hold V609B beyond cutoff, V609B will again conduct. Since the grid of V609B is tied to the B supply by the divider consisting of R665, the grid-cathode resistance of V609B, and R663, the potential at the V609B grid will again assume a positive value with respect to the cathode. The resulting increase in the V609B current through the common cathode resistor, R665, causes the grid-cathode voltage (bias) on V609A to increase. Thus, the current in V609A will decrease to cutoff, while the V609B current rises to saturation. This condition (V609A cut off and V609B conducting) is the normal condition of the multivibrator which prevails until another positive-going reference trigger pulse is applied to the V609A grid.

The pulse output at the V609B plate is one complete square wave for each input pulse to the grid of V609A. The duration of the output square wave is determined by the C625-R663 time constant. The plate output at V609B pulse is positive-going, and is coupled by C633 through R678 to the control grid of the priority gate stage, V611. The purpose of this input is treated later. A negative-going output square wave from the V609B cathode is coupled by C647 to the control grid (pin 3) of the 41.7 kc pulsed oscillator, V610.

The pulsed oscillator tuned tank, comprising L602, C626, and C650, is connected in the cathode of V610A. The resonant frequency of this tank is 41.7 kc. However, the tank circuit is not active at all times to produce the 41.7 kc output as is the case with conventional oscillators. Note that V610A is cathode biased by L602. Since very little bias voltage will be established between the control grid and cathode of V610A as a result of the current through L602, the V610A current through L602 is large enough to keep the coil saturated. Under this condition, no oscillations can occur in the cathode tank circuit because of the low incremental inductance of L602.

When the negative-going gate pulse from the cathode of V609B arrives at the V610A grid, this tube section immediately cuts off, and the cathode tank circuit is shocked into ringing oscillations at 41.7 kc. The high Q cathode tank will continue to oscillate until the negative gate pulse from V609B is terminated.

The relationship of the oscillator (V610) output to the period of the input gate pulse from V609B permits the oscillator to complete 6 sine wave cycles before the trailing edge of the gate pulse arrives to again cause the saturation of the cathode tank coil, L602. As previously stated, the natural period of the gate pulse is determined by the R-C time constant of C625 and R663 in the V609 multivibrator.

The signal from the ringing circuit is applied through R668 to the control grid of cathode follower power amplifier, V610B. Capacitors C626 and C650 form a capacitive voltage divider which provides feedback to the oscillator tank circuit. Note that when the plate current of V610B is rising (as a result of a positive voltage swing in the tank circuit which is applied to the V610B grid), the charge on the upper plate of C650 is becoming more positive with respect to ground, and the feedback energy from the tube aids the tank oscillations. Conversely, if the grid tank swing is going in the opposite (negative) direction, the V610B plate current will be decreasing as a result of the less positive grid signal (with respect to ground) and C650 will be discharging through R790. This action is also regenerative to the signal in the oscillator tank.

The oscillator tank output is coupled by C627 and CR604 to the grid of the reference burst amplifier, V604A.

The 33.3 kc pulsed oscillator, V603, in the 15-cps reference burst circuit also applies its output through CR605 to the same reference burst amplifier, V604A. A single amplifier can be used for both the north and auxiliary reference burst since these signals do not occur simultaneously.

V604A operates at approximately zero bias and is therefore normally conducting. This stage inverts and amplifies the oscillator (V610 or V603) output to an amplitude which is sufficient to trigger the one-shot multivibrator, V605. Since only the negative portion of the oscillator (V610) output is required at the input to V604A, CR604 and CR605 rectify the input sine waves to a series of negative pulses. These negative pulses are developed at the V604A grid across R616.

The B section of V605 is normally conducting. This multivibrator operates in essentially the same way as described for V609. The positive triggering pulses to V605 may vary in amplitude and waveshape. However, because of the multivibrator action, the V605B output pulses are of uniform amplitude and waveshape. Also, the time spacing of the multivibrator output pulses is an exact duplicate of the input trigger pulses.

Blocking oscillator, V606, is employed to produce sharp voltage spikes capable of driving the encoding delay line, DL601. The V605 multivibrator output is differentiated at the grid of blocking oscillator, V606 (across R626), into sharp positive- and negative-going spikes. Both grids of V606 are held beyond cutoff by negative potentials obtained from the voltage divider comprising R633, R632, and R634. This divider is connected from the -105 volt supply to ground.

When the positive pulse from V605 arrives at the grid of V606A, this tube section will begin to conduct, and plate current will flow through the primary of T602. The current through the primary of T602 causes a voltage to be induced in the secondary, and the V606B grid becomes positive with respect to the cathode. The action causes a V606B plate current, which aids the V606A current through the T602 primary. This, in turn, increases the positive voltage at the V606B grid. The action is cumulative until the plate current of V606B has reached saturation, at which time no further increase in plate current is possible.

Since a positive voltage is induced on the grid of V606B only while current through the T602 primary is increasing, at saturation, the grid potential swings back to a negative voltage, thereby reducing plate current through V606B. The current through the T602 primary is now decreasing or changing in a direction which will induce a negative voltage toward the V606B grid, and the V606B plate current is reduced to zero. V606A is not actually a part of the oscillator, but acts as a trigger amplifier to convert sharp pulses of voltage from V605 into sharp pulses of current in the primary of T602.

R627 and R629 are connected across the primary and secondary, respectively, of T602. These resistors produce a damping effect on the oscillations in T602 so that only one voltage spike is produced for each trigger pulse input.

You will recall that the trigger pulses applied to the blocking oscillator (V606) grid are actually the north reference burst cycles generated in V610A after the pulses have been amplified in V604A and shaped in V605. Thus, for each reference burst trigger pulse from the antenna, V610 produces 6 reference burst cycles and V606 produces 6 pulse spikes.

V606B provides two outputs. The cathode output of V606B consists of positive-going spikes which are applied to the primary of T603. The purpose of this output is discussed presently.

The plate output of V606B consists of 6 negative-going spikes for each reference burst trigger at V610A. These pulses are developed across the divider consisting of C629, R671, R672, and R673. R672 is a synchronization control for the 135-cps gate generator stage, V609. Synchronization takes place in the following manner: The pulses from V606B plate (across R672) appear as negative-going spikes at the V609A grid. You will recall that V609A is conducting during the time of the negative gate pulse to V610A. The negative-going pulses from

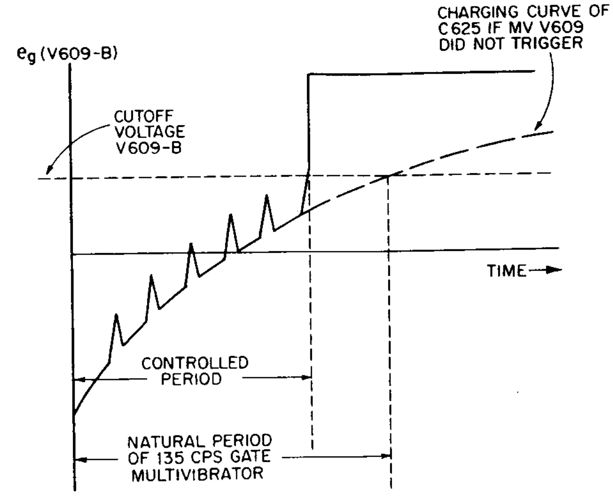

Figure 6-3.—Grid voltage during discharge time of C625.

the R672 arm are amplified in V609A. These pulses "ride-on" or are superimposed on the discharge curve of C625 (fig. 6-3). The amplitude of the feedback pulses from the R672 arm should be adjusted to ensure that the duration of the gate pulse from V609 to V610A causes the production of exactly 6 oscillator cycles from V610 and subsequently from V606B.

The 15-cps reference burst generator circuit, consisting of V601B, V602, and V603, is designed to produce 12 pulses spaced 30 μs apart. The pulsed oscillator, V603, tank circuit, consisting of L601, C608, and C649, operates at 33.3 kilocycles. The V603 oscillator output is gated by the cathode output of V602B. CR604 and CR605 isolate the 15-cps and 135-cps pulsed oscillators. The feedback circuit from the arm of R672 to the V609A grid drives V609A further into cutoff during the period of the 12 input pulses developed by V605 to ensure that the 135-cps gate generator does not operate during this time.

ENCODING DELAY LINE AND OUTPUT AMPLIFIERS

Keeping in mind the circuits discussed, it would be well at this time to bring to mind again the overall purpose of the reference burst circuits. The combined action of the 135-cps reference burst circuits is to produce a group of bearing reference pulses, which consist of 6 pulse-pairs with 12 μs spacing between pulses of a pair, and 24 μs between pulse-pairs. These pulses recur 120 times per second at a 135-cps rate, as determined by the position of the soft iron (auxiliary) slugs on the pulser-plate assembly. (See ch. 5, fig. 5-4, B.) The 15-cps reference burst circuits must produce 12 pulse-pairs with a 12 μs spacing between pulses of a pair, and 30 μs between pulse-pairs. The pulses recur 15 times per second as determined by the north slug on the pulser-plate.

The output at the blocking oscillator V606, cathode (fig. 6-2) is about 166 volts peak-to-peak, and is fed to the primary of autotransformer, T603. This transformer matches the impedance to the encoding delay line, DL601, and has a step-up ratio of 1:3. Thus, the amplitude of the input signal to the delay line is approximately 500 volts peak-to-peak. As mentioned previously in chapter 5 the encoding delay line introduces the major portion of the standard zero distance delay for distance reply pulses.

For each pulse input to the encoding delay line, DL601, the delay characteristics of the line produces 2 positive pulses spaced 12 μs apart. These pulses are delivered to the pulse amplifier, V607A. These pulses are tapped from the 5 and 3 terminals of DL601, and have an amplitude of approximately 15 volts peak-to-peak. When an input pulse is applied to the line, the initial pulse from the line is delayed 32 μs, while the second pulse is delayed 44 μs.

L604 and C614 filter out reflections from the delay line, DL601, to prevent possible interference with the operation of the blocking oscillator, V606.

The output of the delay line, DL601, is developed at the grid of the pulse amplifier, V607A, across R635. CR601 and CR602 block the negative portion of the input cycle. A fixed bias potential of about -21 volts is applied between the grid and cathode of V607A from the R636-R637 junction. V607A amplifies and inverts the positive input pulses.

The negative-going pulses appearing at the V607A plate are coupled by C617 to the grid of the A section of one-shot multivibrator, V615. V615A is essentially zero-biased and is, therefore, the conducting section of the multivibrator in the quiescent condition. V615B is held beyond cutoff by a fixed negative potential (approximately -34 volts) obtained from the junction of R643 and R644. These resistors are connected from the -105 volt supply to ground.

The operation of V615 is similar to the one-shot multivibrators already discussed. The application of negative pulses from the V607A plate to the grid of V615A cuts off this section of the tube to initiate the multivibrator cycle. The duration of the V615B output is determined by the time constant of C618, R787, and R643. C621 provides regenerative feedback to the grid of V615A to decrease the rise-time and fall-time of the output pulse at the cathode of V615B.

The V615B cathode output consists of the shaped reference burst pulses (either north or auxiliary) which are coupled by C619 to cathode follower, V607B. The action of V615 ensures uniformity of the amplitude and width of the reference pulses.

Cathode follower, V607B, functions to isolate the V615 multivibrator from the output, and to match the output circuit to the 50-ohm line which feeds the signal to the transmitter. R649 and R650 are the load resistors for this stage.

A portion of the output of V607B, taken from the junction of R649 and R650 is delivered to test jack, J607, on the front panel of the coder-indicator. This jack is labeled "test output" (fig. 6-1), and enables the operator to examine the output of the coder-indicator.

IDENTIFICATION CODE CIRCUITS

The circuits comprising V612, V613A, V614, V613B, and CR603, make up the 1350-cps tone generator, the tone of which is keyed to produce the identification code characters for the radio beacon. Identification call amplifier stage, V613B, actually produces 2 pulses every 740 μs, or at a rate of 1350 cps.

Reference to the block diagram of figure 5-10 (ch. 5) reveals that the identification tone output of V613B is keyed by V604B and V611, amplified and shaped in V605, V606, DL601, V607A, V615, and V607B before it is fed to the frequency multiplier-oscillator of the radio beacon. You will remember that the encoding delay line, DL601, produces 2 pulses spaced 12 μs apart for each pulse applied at the input to the line. Thus, during a 1 second period of the identification call, 1350 pairs of 100 μs-spaced pulses appear at the input to the encoding delay line, with 2700 pulse-pairs at the DL601 output. The resulting beacon pulse output is maintained at the required rate.

The operation of the identification call circuits is as follows: The oscillator tone (1350 cps) is generated in V612. The oscillator circuit connection is that of a Hartley, with L603, C652, and C653 forming the oscillator resonant circuit. A 135-cps synchronization signal obtained from the B section plate of the 135-cps gate generator stage, V609B (in the 135-cps reference burst circuit) is applied to the control grid of V612A. The signal is amplified in V612A and applied as a positive (regenerative) synchronization input to the oscillator tank circuit. The oscillator tank serves as the plate load of V612A. The tank circuit actually operates at the tenth harmonic of the synchronization input. The 1350 cps output of V612A is the basic identification tone signal.

The L-C tank circuit which acts as the plate load of V612A also serves as the grid input source for V612B. The circuit of V612B is that of a cathode follower. The 1350 cps output of the oscillator appears across R692 and R693, and is coupled through C641 to amplifier, V613A. Limiting at the grid of V613A across R691 when positive signals are applied causes only a very small negative-going plate output. Negative alternations of the 1350 cps signals at the grid of V613A are amplified in this stage, and coupled by C642 to the A section grid of multivibrator, V614.

V614 is connected as a one-shot multivibrator. The purpose of this stage is to produce 2 output pulses for each pulse applied at its input. Thus, the output pulse rate of the identification stages during the transmission of the identification tone is 2700 pulses per second, which are grouped at a 1350 cps rate. These pulses are encoded (double-pulsed) and amplified in the shaping stages (V605, V606 and DL601) so that the output pulse rate from the transmitter during the transmission of identification call is maintained at 2700 pulse-pairs per second.

The V614 circuit is conventional in that it produces one square wave output from each of its plates for each trigger pulse input at its grid. The square wave from each section (occurring simultaneously) is differentiated. The resulting differentiated pulses from each section of V614 are combined to yield pulses at the output of the multivibrator for each pulse input from V613A.

The various waveshapes obtained from the V614 multivibrator are shown in figure 6-4. Frequent reference should be made to this illustration while studying the following discussion of the multivibrator operation.

Waveform 1 (fig. 6-4) shows the voltage change on the plate (pin 6) of V614A as the multivibrator goes through two complete cycles, due to the application of two trigger pulses. In the quiescent state of the multivibrator, V614A (fig. 6-2) is cut off and its plate voltage is at +250 volts (the full value of the B supply voltage). At the instant a trigger pulse is applied, the plate voltage at pin 6 takes a sharp negative swing. This sharp negative swing in voltage is passed through C645 and differentiated across R796 to form a negative spike at the grid (pin 7)

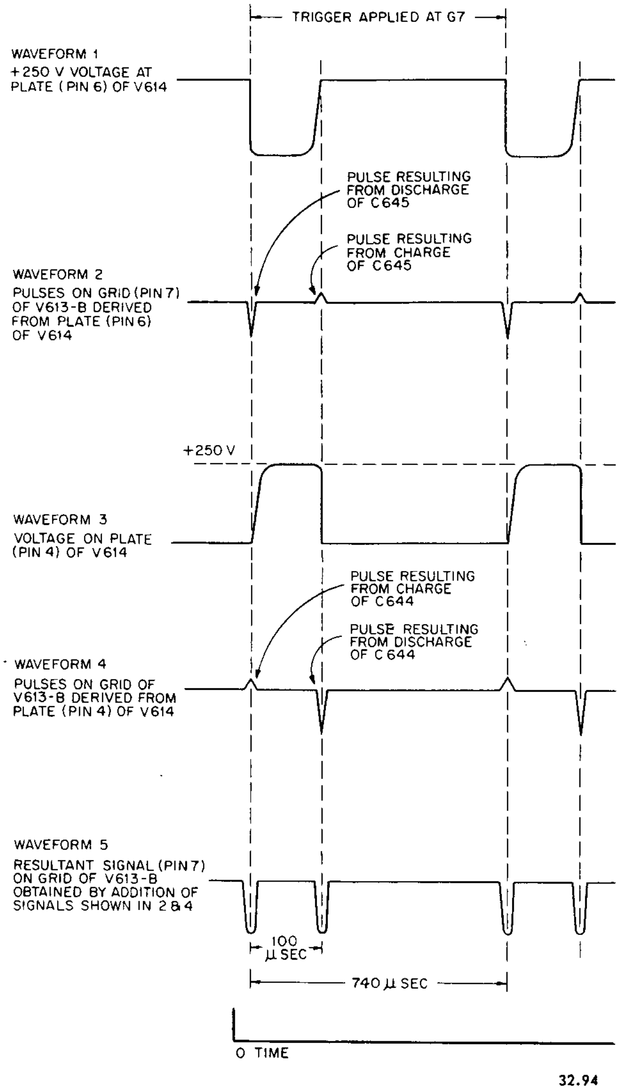

Figure 6-4.—Development of pulse-pairs on grid of V613B.

of V613B. The V613B differentiated input is shown in waveform 2 of figure 6-4.

The negative-going change in the V614A plate voltage (when the trigger pulse arrives) causes C643 to discharge through the path comprising R797, R798, and the conducting resistance of V614A. The voltage drop across R797 and R798, during the discharge of C643, makes the V614B grid negative with respect to the cathode. Thus, V614B, which is normally conducting, will cut off.

The period of the multivibrator output pulse (waveform 1) is determined by the setting of the R798 arm. At the end of the multivibrator period, C643 has discharged to the full value of the plate voltage decrease, and the resulting discharge current through R797 and R798 is reduced. The V614B bias is thus reduced to a point which allows V614B to again assume its normally conducting condition. The plate voltage of V614A begins to rise as a result of the increased grid-cathode bias voltage developed across the common cathode resistor, R699, caused by the return of V614B to conduction. The plate voltage rise of V614A back to the B supply value (waveform 1) is relatively slow as compared to the plate voltage drop since the rise in plate voltage is controlled by the charge of C643 through R699, the cathode-grid conducting resistance of V614B, and R697.

Waveform 3 shows the change in plate voltage of V614B. Note that the V614A and V614B waveforms begin at the same time (upon application of the trigger pulse to V614A). The V614B plate output is positive-going during the period of the multivibrator.

A small positive-going voltage change is developed across the divider comprising R795 and R796, during the charge time of C644 because of the high resistance in this path and the resulting slow rate of charge of C644 (waveform 4). The magnitude of the charging current varies directly with the rate of change of voltage across the capacitor. Therefore, only a small positive-going voltage change will appear at the V613B grid across R796. When V614B again conducts (after the discharge of C643) the tube immediately conducts saturation current. The plate voltage decrease through C644 is much more abrupt, because the C644 discharge path is through R795, R796, R699, and the low conducting resistance of V614B. The faster rate of discharge of C644 as compared to the charging rate causes the amplitude of the negative-going pulse to be developed across R796 (at the V613B grid) during the C644 discharge to be higher than the amplitude of the positive-going pulse developed during the charge of C644 (waveform 4).

Note that the faster change of the C645 or C643 capacitor voltage always occurs at the instant the capacitor is discharging. Because the A section of V614 produces the greater rate of change in capacitor current during the time when the multivibrator is initially triggered, and V614B produces the greater rate of current change through C644 at the end of the multivibrator pulse, the larger amplitude pulses do not appear simultaneously. Rather, the smaller-amplitude positive-going pulse from V614B appears at the grid of V613B coincidently with the higher-amplitude negative-going pulse from V614A. Conversely, the smaller-amplitude positive-going pulse from V614A (at the end of the multivibrator output pulse) appears at the grid of V613B coincidently with the higher-amplitude negative-going pulse from V614B. The resulting voltage across R796 (and to the V613B grid) is the algebraic sum of the two simultaneous voltages applied from V614A and V614B (waveform 5).

When R798 is properly adjusted, the time between peaks of the two pulses from V614 will be exactly 100 μs (fig. 6-4). Each pair of pulses is separated by 740 μs. In this manner, a constant duty cycle of $10^6 \times 2/740$ or 2700 pulse-pairs per second is maintained (after double-coding in DL601) figure 6-2, at the output of the transmitter, while the tone frequency of 1350 cps is preserved in the form of pulse-pairs.

CR603 (fig. 6-2) functions as a clamping diode which presents either a low or high impedance path to ground for the identification call signals depending on the position of the switches (S607 and S604) in the identification call mechanical keyer assembly, A602. When the microswitches (S607 and S604) are closed, ground is applied to the plate side, "P" of CR603, through R687 and the closed contacts of S603. When "P" is grounded, CR603 becomes a low impedance path to ground for the negative pulses which are fed to the grid of V613B. Thus, the tone signal is grounded out through the diode, and does not appear at the V613B grid. When the ground is removed from the diode, by the action of the keyer wheel assembly (discussed later), a negative bias voltage of approximately -50 volts is applied to plate "P" of the diode from the R788-R686 junction. This divider is connected from the -105 volts supply to ground. With -50 volts on the plate of CR603, the diode acts as a high impedance, and the tone signals are developed at the V613B grid across R796. Under this condition, V613B passes the 1350-cps pulse-pairs through C636 to the priority gate stage, V611.

The keyer assembly, A602, comprises a motor driven code keying wheel, a code start timing cam, and two microswitches, S604 and S607. The keyer assembly in conjunction with the gate blocked distance reply amplifier, V604B, controls the selection of either identification tone distance reply pulses as input signals to the priority gate stage, V611. The keying wheel and start code timing cam initiate a beacon identification coded call every 37.5 seconds.

The code keying wheel revolves at a speed of 8 rpm. The periphery of the wheel is divided into segments. The segments can be moved outward from the center of the wheel. By positioning the segments, code characters of long or short duration (dots and dashes) can be set on the wheel.

Timing switch, S604, is controlled by the code start timing cam, which revolves at 1.6 rpm. Assuming that the continuous tone switch, S603, is closed, the timing switch, S604, grounds either the R788-R686 junction through R687, or one side of the keying switch, S607. The timing switch control cam is designed so that the timing switch will ground one side of the keying switch, S607, for one revolution in every five of the code keying wheel.

Recall that the voltage divider comprising R788 and R686 is connected from the -105 volt supply to ground, and that the junction of the resistors is approximately -50 volts with respect to ground. This negative potential is applied to the control grid of the gate blocked distance reply amplifier, V604B, through R685. Note that the distance reply signals from the receiver are applied through C638 to the control grid of V604B.

When the keying switch, S607, and the timing switch, S604, are open simultaneously (corresponding to a tone-on condition), the -50 volts is applied to the V604B grid, and the distance reply pulses are blocked by the V604B cutoff. Likewise, the negative potential is applied to the plate "P" of diode CR603, and the diode does not conduct, and the impedance to ground for the 100 μs spaced pulses from V614 is high. These pulses, which are the identification tone pulses, are developed across R796 at the V613B grid. The tone pulses are amplified in V613B and coupled through C636 to the grid of the B section of the priority gate stage, V611. If reference burst pulses are not being applied to the A section control grid of V611 at the same time that the identification tone pulses are applied, the tone pulses will be passed through the priority gate stage and subsequently to the output of the coder-indicator. The action of the priority gate stage will be considered presently.

When the timing switch, S604, is open and the keying switch, S607, is closed (corresponding to a tone-off condition) the ground path through the microswitches is completed. The potential applied to the control grid of V604B therefore is essentially zero. CR603 acts as a low impedance path to ground for the tone signals, and no resulting amplification of these signals occur, in V613B during this period. Since the bias voltage applied to V604B is reduced, this stage, acting as a cathode follower, develops the distance reply pulses from the receiver across R684 at the V604B cathode. The output signals are coupled through C635 to V611B, and in the absence of reference pulses at the grid of V611A, the distance signals will be passed through V611B and subsequent stages of the coder-indicator to the output jack, J602.

You will note that both the keying switch, S607, and the timing switch, S604, must be open simultaneously to produce a code tone pulse. Therefore, no code tone pulses are possible during the time the timing switch, S604, applies ground to the grid of V604B. This condition exists for the four revolutions of the keying wheel directly following each cycle initiating a beacon identification call.

The priority gate stage, V611, assigns the order of preference for the transmission of the components of the beacon signal. The reference burst signals must be transmitted without regard to the transmission of the beacon identification code pulses or the distance reply pulses. Second in priority are the identification code signals which are transmitted every 37.5 seconds. If the reference burst pulses are being transmitted during the time that identification code is to be transmitted, the priority gate stage, V611, blocks the passage of the identification code only during the time that each reference burst pulse is actually being transmitted. The distance reply pulses are last in the priority grouping. These pulses are passed through the priority gate stage, V611, only during the time that neither reference burst nor identification code pulses are being transmitted.

With neither reference burst gates, identification call pulses, nor distance reply pulses present at V611, V611A will be cathode-biased, and will therefore be conducting. V611B will be cut off with -10 volts applied to its grid from the junction of the voltage divider comprising R682 and R651. Each of the radio beacon signal components is applied to V611 as a positive-going pulse.

If only distance reply or identification call pulses appear at the grid of V611B, the positive input signals override the grid-cathode bias, are amplified, and passed on from the V611B plate to the shaping amplifiers, V605 and V606. However, whenever reference burst pulses are being applied to the V611A grid through C632 (as discussed), the positive gate causes V611A to conduct heavily. This action causes a large voltage drop across R681, which is the common cathode resistor for V611A and V611B. The increase in grid-cathode bias on V611B, in combination with the fixed bias of -10 volts, drives V611B so far into cutoff that the positive-going identification call or distance reply pulses on its grid can no longer override the cutoff bias to bring V611B back into conduction. These signals are now blocked; therefore, the priority of the reference burst pulses over the identification call and distance reply pulses is established.

You will recall that the reference burst signals are applied to the shaping amplifiers from their respective pulsed oscillator stages (V610 or V603). Thus, the reference pulses are not fed through priority gate stage, V611, but are applied to V611A for the purpose of cutting off V611B as described.

The signals permitted to pass through the priority gate stage, V611, are coupled by C648 to the output amplifier circuits via V605. These signals are shaped, double encoded, amplified, and sent to the transmitter as components of a multiplexed signal.

ANTENNA SYNCHRONIZATION OSCILLATOR

A 1350-cps antenna synchronization oscillator, V608, produces an output voltage, which is used in conjunction with a signal voltage from the antenna speed tachometer (not shown), in checking the speed of rotation of the antenna. The tachometer produces a 675 cps output. The two voltages will cause a 2 to 1 Lissajous pattern to be formed on an oscilloscope screen when the antenna speed is correct.

The frequency of the oscillator, V608, is controlled by a tuning fork, Y601. The 1350-cps output of the tuning fork is generated in the grid-cathode circuit of V608A via that portion of the tuning fork coil which is connected between grid and ground of V608A. A portion of the V608A output is used as the 1350-cps reference frequency voltage required in checking the speed of the antenna as mentioned above. This output is available at J609.

The V608A output is also coupled to V608B via C622. V608B functions as a cathode follower, with the input coil of the tuning fork acting as its cathode load. In this manner, sufficient voltage generated within the oscillator circuit is fed back to the tuning fork to sustain the oscillations.

FREQUENCY MULTIPLIER-OSCILLATOR

The frequency multiplier-oscillator chassis (fig. 6-5) of the radio beacon comprises a carrier generating (r-f) chain and a video chassis. The carrier generating chain produces the beacon radio frequency carrier. The circuit

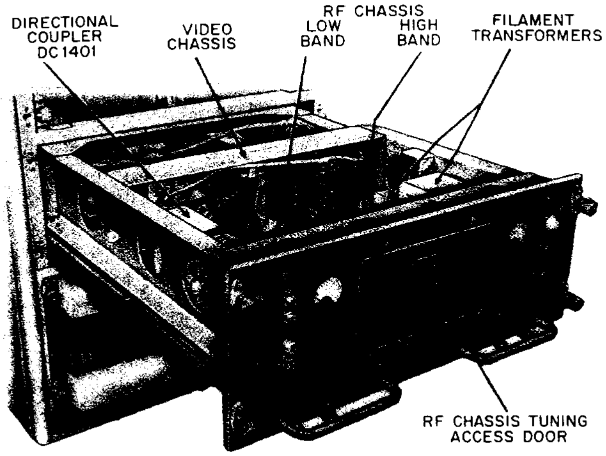

Figure 6-5.—Frequency multiplier-oscillator, overall view.

operation of the stages included in the carrier generating chain is discussed in chapter 5 of this training course. The video chassis produces two shaped pulses of 3.5 and 10 μs, respectively, which are used for keying and modulation of the klystron output.

VIDEO CHASSIS

The positive 1.5 μs pulses from V607B on the coder-indicator chassis (fig. 6-2) are applied via J1401 on the frequency multiplier-oscillator chassis (fig. 6-6) to the one-shot multivibrator, V1402. V1402 ensures uniformity of shape of the input pulses. The positive-going leading edge of the trigger pulse is passed through CR1402 and developed across R1405 at the V1402A grid. This pulse triggers the V1402 multivibrator into operation.

If the input pulse contains a negative alternation, this negative portion of the pulse may tend to switch the V1402 multivibrator action before the normal time of switching. To prevent this false triggering action, CR1402 is inserted in series with the signal path of the trigger pulse to block the negative alternation.

V1402 is connected as a monostable multivibrator circuit, with V1402B conducting during quiescence. The grid of V1402B is returned to the +250 volt B supply through R1403 and R1466. The grid of V1402A is connected to a voltage divider consisting of R1407, R1469, R1446, and R1468. This divider is connected between the +250 and -375 volt supplies.

In the quiescent condition of the multivibrator (V1402B conducting) the heavy V1402B plate current flow through R1446 and R1468 causes the plate voltage of this tube section to be low. The voltage dropped across the divider comprising R1407 and R1469 likewise is relatively low. Since these two resistors are connected from a low positive potential (about +65 volts when V1402B is conducting) to the -375 volt supply, it follows that the potential across R1407 and R1469 will be low when V1402B is conducting as compared to the potential across these resistors when V1402B is cut off. The potential at the V1402A grid when V1402B is conducting is approximately -42 volts. This value of negative (bias) voltage is established between the grid and cathode of V1402A, and is sufficient to hold V1402A beyond cutoff as long as V1402B is conducting.

Figure 6-7, A, shows the 1.5 μs input signal to V1402A from V607B (fig. 6-2). The positive pulse on the grid of V1402A (fig. 6-6) causes a negative pulse to be developed at the plate of V1402A. The negative-going pulse is coupled by C1423 and developed across R1403 and R1466 so that it is negative toward the V1402B grid. As the grid of V1402B goes negative with respect to its cathode, its plate potential rises, and a positive-going pulse is coupled through C1421 to the V1402A grid. This action tends to drive V1402A to saturation while simultaneously driving V1402B to cut off. V1402B will remain cut off until C1423 stops discharging through the series combination of R1403 and R1466.

When plate current again starts to flow in V1402B, the multivibrator returns to its stable condition, with V1402A cut off and V1402B conducting. The time constant of the circuit consisting of C1423, R1403, and R1466, determines the width of the pulse generated. R1466 should be adjusted to provide a square pulse at the output of V1402B approximately 2.4 μs wide (B of fig. 6-7).

Power amplifiers, V1403, V1404, and V1405 (fig. 6-6) are operated in parallel in order to handle the heavy currents required in producing the high-powered pulse necessary to drive the subsequent shaping circuits. Grid resistors, R1410, R1413, and R1414, and plate networks, Z1401, Z1402, and Z1403, are parasitic suppressors. R1409 is the common grid resistor for all of the power amplifier stages. The primary of T1402 acts as the common plate load of the power amplifiers.

The negative potential at the junction of R1409 and R1408 (caused by the connection of these resistors from the -375 volt supply to ground) is applied to the grids of V1403, V1404, and V1405. The fixed bias potential is sufficient to cut off

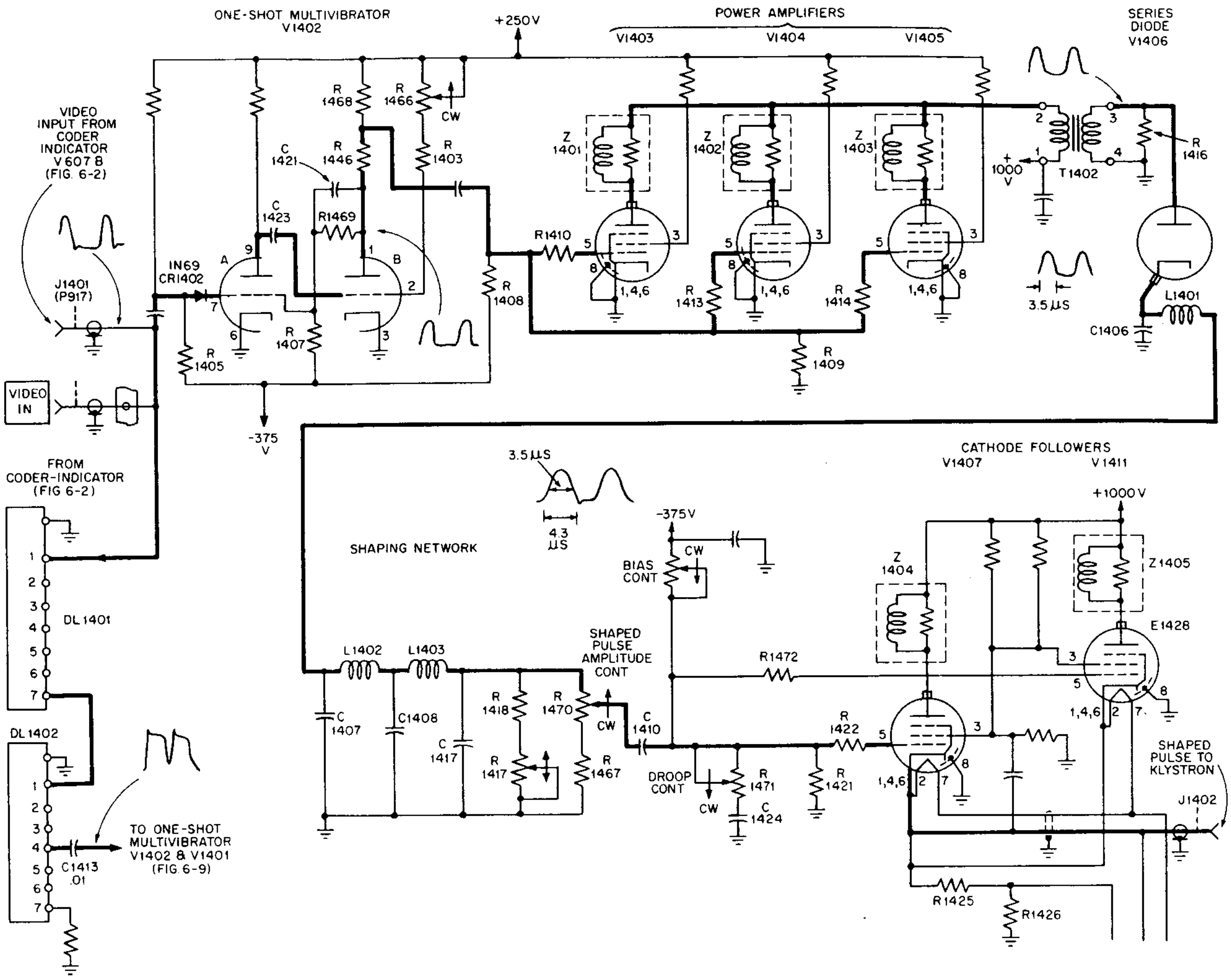

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Figure 6-6.—Frequency multiplier-oscillator, video chassis, schematic diagrams.

the power amplifiers in the absence of a signal on the grids.

When a positive pulse from V1402B is applied to the grids of the power amplifiers, V1403, V1404, and V1405, plate current flows in these stages, and a negative-going rectangular pulse approximately 800 volts in amplitude is developed across the primary of T1402. A positive pulse, approximately 2300 volts in amplitude is induced in the secondary of T1402, and is applied to the shaping network through series diode, V1406.

V1406 prevents negative transient voltages at the secondary of T1402 from being passed to the shaper network. R1416 loads the T1402 secondary, to damp out ringing oscillations which may occur because of the sudden collapse of the inductive field of T1402 when plate current in the power amplifiers is cut off.

The shaping network comprises capacitors, C1406, C1407, C1408, and C1417, and coils, L1401, L1402, and L1403, in a low-pass filter circuit which produces a 3.5 μs pulse. Rectangular pulses 2.4 μs wide (B of fig. 6-7) at the input of the shaper network are reduced in amplitude and shaped into pulses 3.5 μs wide having a gradual rise and decay time (fig. 6-7, C). Control of the pulse shape at this point ensures that a minimum of r-f spectrum will be occupied by the final transmitted r-f output pulse. The delay time of the pulse through the shaper network (fig. 6-6) is approximately 2.8 μs.

R1418 and R1417 are used to properly terminate the shaping network. R1417 provides a method of adjusting the terminating impedance to match the characteristic impedance of the shaper network. This not only prevents

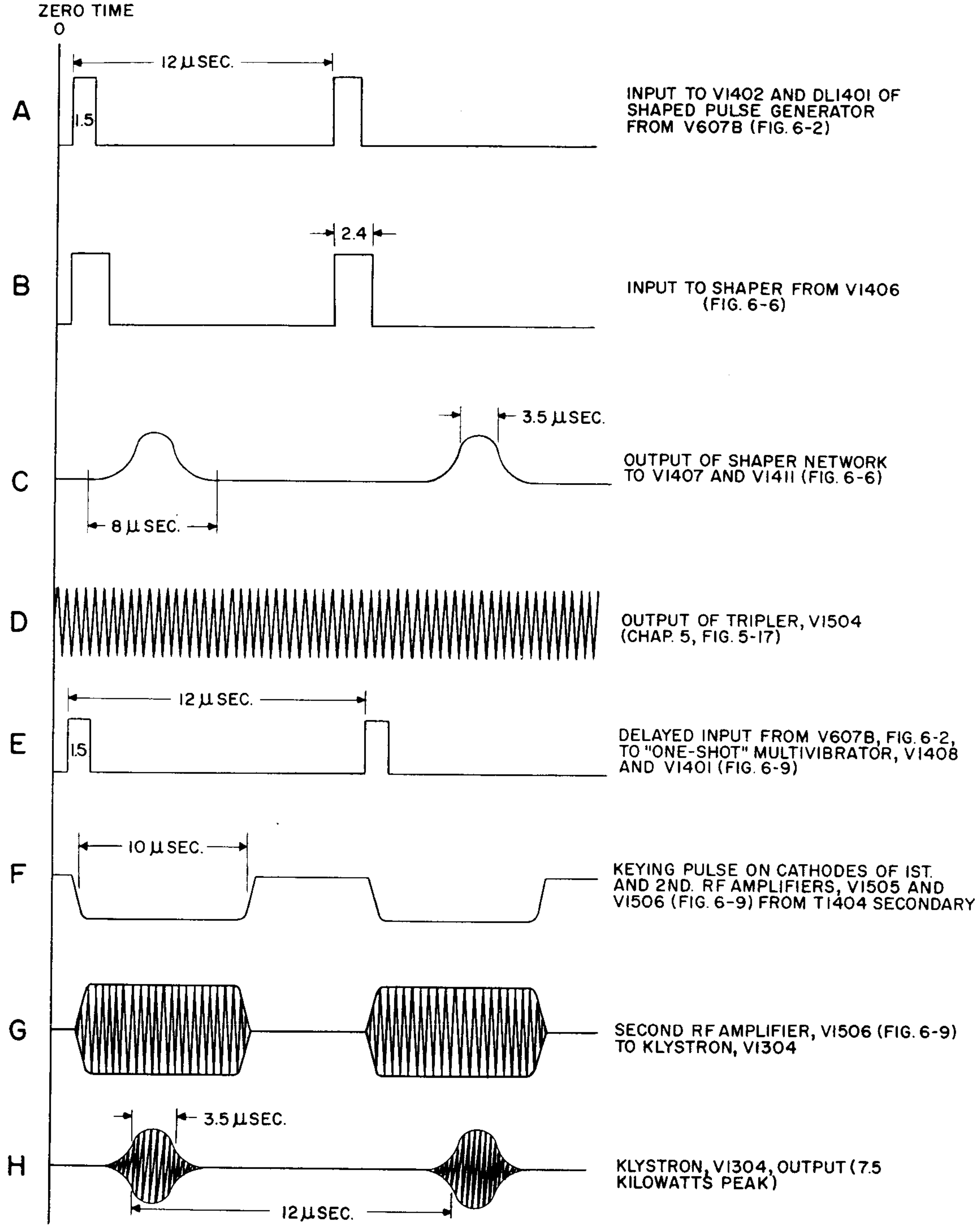

Figure 6-7.—Transmitter pulse sequence.

reflections in the network, it also prevents distortion of the output pulse.

Positive output pulses from the shaper network are coupled through C1410 and parasitic suppressors, R1422 and R1472, to the paralleled grids of the cathode followers, V1407 and V1411, respectively. The cathode follower tubes are operated in parallel to lower the output impedance of the 3.5 μs shaper pulse generator.

R1471 and C1424 form a compensating network in the grid circuit of the cathode follower output stages, V1407 and V1411. When the R1471 arm is moved downward (clockwise) the impedance to ground (particularly for high frequency) is very much decreased so that more of the high frequencies are bypassed to ground by C1424. Moving the control upward (counterclockwise) increases the impedance from the R1471 arm to ground and fewer of the high frequency components are bypassed.

It is required that the north reference burst of 24 pulses shall not deviate from the average amplitude by more than ±2.0 percent. Any tendency toward pulse droop in the output of the radio set may be counteracted by adjusting R1471 to introduce enough pulse boost to equalize the pulse droop.

The output of the cathode followers, V1407 and V1411, is developed across the common load resistors, R1425 and R1426. The output is fed to the amplifier-modulator chassis (fig. 6-8). The signal is applied via C1370, C1371, and T1372 (fig. 6-9) to the klystron amplifier, V1304. You will note that in the frequency multiplier-oscillator, each pulse from the coder-indicator is shaped before it is applied to the klystron (output) amplifier. The space between pulses is maintained at 12 μs.

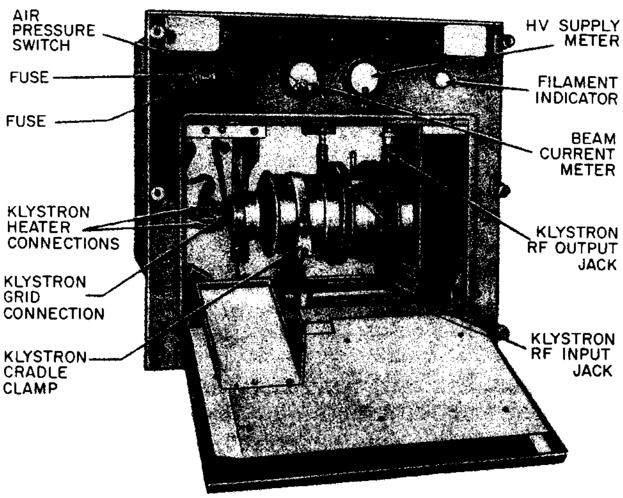

Figure 6-8.—Amplifier-modulator, front view.

In addition to applying the 3.5 μs pulse to the klystron (fig. 6-9), the r-f carrier must also be applied at the same time to produce the required beacon output. The r-f carrier frequency is produced in the carrier generating chain. The circuit operation of the carrier generating chain was discussed in chapter 5 of this training course (ch. 5, fig. 5-18).

The V1504 output, which represents the oscillator, V1501, frequency multiplied 24 times, is applied simultaneously to the hybrid mixer (fig. 5-10) in the receiver, and to the first keyed r-f amplifier, V1505 (fig. 6-9). V1505 and V1506 are keyed r-f amplifiers operated in cascade and employing 2C39A lighthouse tubes in conjunction with a broadband coaxial-type resonant cavity. The r-f output of V1505 and V1506 is keyed by a 10 μs pulse produced in the frequency multiplier-oscillator and supplied via V1409 (fig. 6-9), to allow the passage of the r-f carrier. The 10 μs pulse simultaneously keys the r-f pulse to the klystron, and blanks the receiver of the beacon to prevent the receiver from reamplifying the pulse and sending it again to the transmitter.

Let us consider how the 10 μs pulse required to trigger the first and second r-f amplifiers, V1505 and V1506, is produced. The 1.5 μs pulses from V607B in the coder-indicator (fig. 6-2) are fed through delay lines, DL1401, and DL1402 (fig. 6-9), in series to the one-shot multivibrator stage consisting of V1408 and V1401. The shape of the input pulse is shown in figure 6-7, E. The total input delay through DL1401 and DL1402 is about 2.8 μs. This corresponds with the delay of the shaped 3.5 μs pulse through the shaper network.

The circuit connections and circuit operation of the V1408 and V1401 multivibrator are conventional except that the two halves of the circuit employ duotriode tubes with the sections of each tube connected in parallel. This increases the current carrying capabilities of the multivibrator, and ensures conservative operation of the sections of the tubes.

V1401 of the multivibrator is normally conducting. Positive input pulses from DL1402 trigger V1408 to start the one-shot multivibrator action. CR1401 blocks negative pulses coming from DL1402 which, if permitted to reach the V1408 grid, might cause false triggering of the multivibrator during the time that V1408 is conducting.

The time constant of the multivibrator is determined by C1416, R1402, and R1465. The latter should be adjusted to set the output pulse width at 10 μs.

Pulse amplifier, V1409 (fig. 6-9) consists of a parallel connected twin triode. Pulses from the plate of V1401 are coupled to the grids of V1409 through C1414 and grid blocking resistors, R1435 and R1440, respectively.

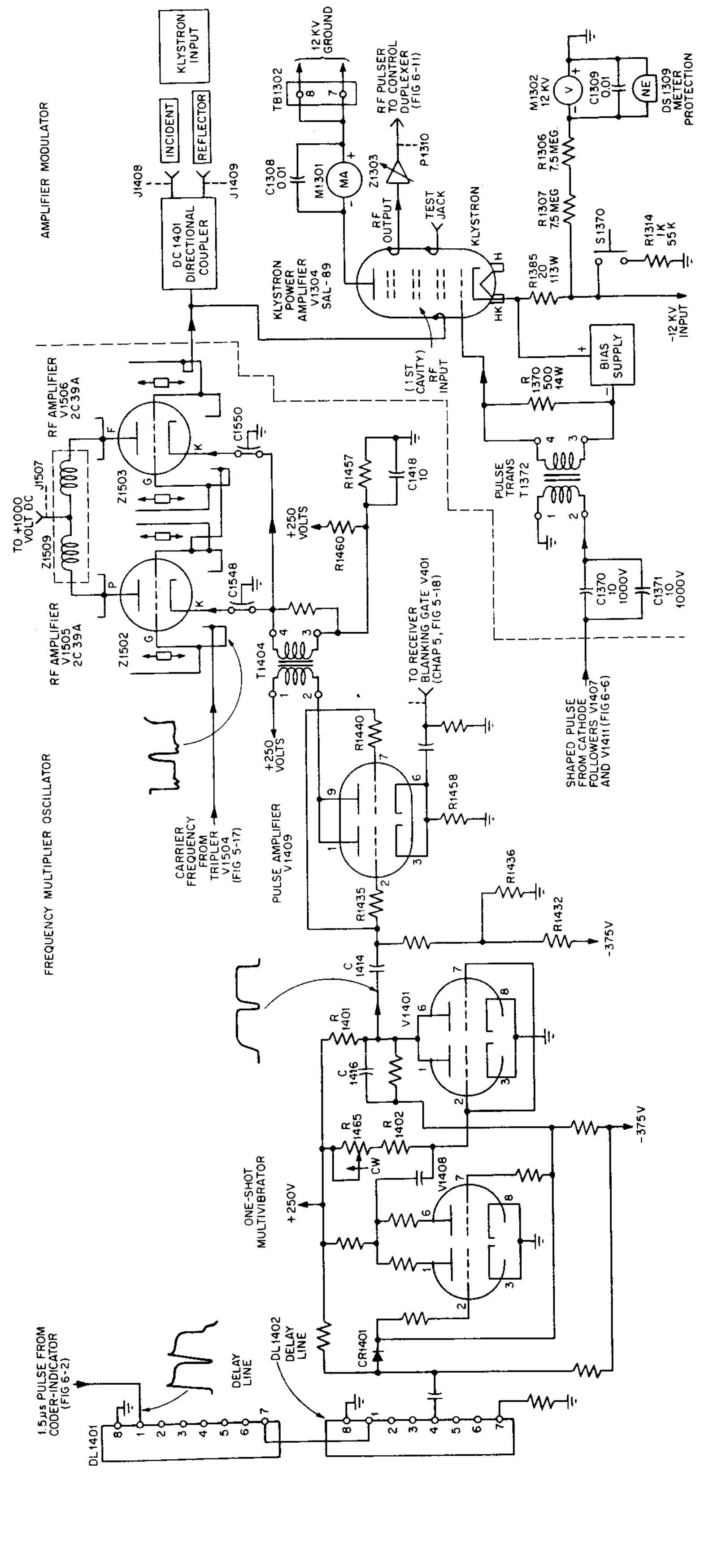

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Figure 6-9.—Amplifier-modulator (transmitter output) schematic diagram.

Approximately 25 volts fixed bias is applied between the control grid and cathode of V1409 from a voltage divider consisting of R1436 and R1432. The positive 10 μs pulse from V1401 permits V1409 to conduct, thereby producing a plate output across the primary of T1404.

T1404 couples the 10 μs output pulse from the plate of V1409 to the cathodes of r-f amplifiers, V1505 and V1506. A second output from V1409 is developed across the unbypassed cathode resistor, R1458, and fed to the receiver blanking gate, V401, via J506 (ch. 5, fig. 5-19), to serve as the 10 μs receiver blanking pulse. This pulse blanks the receiver during the period that the transmitter is operative.

As already stated, the 10 μs output of V1409 (fig. 6-9) which keys the r-f amplifiers, V1505 and V1506, is developed across the primary of T1404. R1457 and R1460 serve as a voltage divider between the +250 volt B supply and ground to develop a positive voltage at the junction of these resistors, which is fed through the secondary of T1404 to the V1505 and V1506 cathodes. This potential (approximately 25 volts) biases these stages beyond cutoff.

The negative 10-μs pulses induced in the T1404 secondary are shown in F of figure 6-7. The pulses reduce the existing grid-cathode bias on V1505 and V1506 (fig. 6-9) to a point which allows these tubes to conduct.

Recall that a constant r-f carrier signal is applied to the grid of V1505 (fig. 6-7, D). The r-f carrier signal is passed through the r-f amplifiers, V1505 and V1506 (fig. 6-9), only during the time that the negative 10-μs keying pulse is present at the cathodes of these stages. The r-f output of V1506, which consists of 10-μs pulses of r-f energy at the carrier frequency (fig. 6-7, G), is fed to the klystron, V1304, in the amplifier-modulator (fig. 6-9).

AMPLIFIER-MODULATOR

The amplifier-modulator (fig. 6-9) consists of the klystron r-f amplifier, V1304, a regulated bias power supply, and the associated control circuits. The r-f power amplifier, V1304, employs a SAL-89-type klystron. The klystron is a three-cavity amplifying tube, which has a control grid used to intensity-modulate (vary beam strength) the klystron beam current, and an r-f input jack connected into the first cavity, which is used to velocity-modulate or bunch the beam electrons.

In the quiescent state, a constant negative potential of -12 kilovolts (with respect to ground) from the high voltage power supply (not shown) is applied to the V1304 cathode. However, the klystron beam current is cut off by a -120 volt potential applied between the control grid and cathode of V1304, through the secondary of T1372. This potential is obtained from a series regulated bias supply which is ungrounded.

The 10 μs r-f pulse envelope from V1506 is applied to the first cavity of V1304. The 3.5 μs-shaped pulse is applied from V1407 and V1411 via T1372 to V1304 between the cathode and grid so that the peak amplitude of the pulse occurs at about the center of the 10 μs r-f pulse envelope. The shaped pulse modulates the r-f to form the output pulse (fig. 6-7, C, G, and H). Thus, the action of the klystron amplifier is like that of a coincidence stage.

Because the 3.5 μs-shaped pulse is delayed 2.8 μs in the shaper network (fig. 6-6) the 10 μs trigger pulse must be delayed the same amount by the delay lines DL1401 and DL1402 (fig. 6-9). This action causes the peak of the 3.5 μs-shaped pulse to coincide with a point midway between the beginning and the end of the 10 μs r-f pulse envelope period (fig. 6-7, C and G).

The negative-going 3.5 μs-shaped pulse from the cathodes of cathode followers, V1407 and V1411, are fed to the primary of T1372. The positive-going pulses from the secondary of T1372 are applied between the control grid and cathode of V1304, and cause the klystron tube to conduct a beam current. The instantaneous value of the klystron beam current is directly proportional to the instantaneous value of the 3.5 μs-shaped pulse.

Actually the width of the 3.5 μs-pulse is greater than 3.5 μs. However, the klystron is gated so that the effective width of the output pulse is about 3.5 μs.

It is desired that the peak of the 3.5 μs intensity-modulating pulse (fig. 6-7, C) should arrive between the control grid and cathode of V1304 at a time near the center of the 10 μs-keyed r-f pulse (fig. 6-7, F) to the first cavity of V1304. Since the 3.5 μs pulse (fig. 6-7, C) is delayed 2.8 μs behind the input pulse (fig. 6-7, A) in the shaper network (fig. 6-6), it is necessary to delay the 10 μs pulse (fig. 6-7, F) by an equal amount in DL1401 and DL1402. Because of the long rise and fall time of the shaper output pulse (fig. 6-7, C), the mid portion of the 3.5 μs pulse occurs at the center portion of the 10 μs keyed r-f pulse.

The output of the klystron, V1304 (fig. 6-9) consists of specially shaped r-f pulses (fig. 6-7, H), which have a minimum peak power of 7.5 kw and a repetition rate identical to the repetition rate of the pulse train at the output of the coder-indicator. The output is coupled to a double-slug tuner, Z1303 (fig. 6-9), which provides a means of matching the output impedance of the klystron with the input circuit impedance of the control-duplexer.

CONTROL-DUPLEXER

The control-duplexer (fig. 6-10) is a passive network which permits both the transmitter and receiver of the radio beacon to be operated from the same antenna. The pulses r-f transmitter output from the klystron r-f amplifier, V1304 (fig. 6-9) is fed through matching slug, Z1303, to a transmission line r-f filter (Z1156 or Z1157 of fig. 6-11). Z1156 is used with low-band frequencies, and Z1157 is used with high-band frequencies. Each of the filters consists of a pair of tunable resonant cavities.

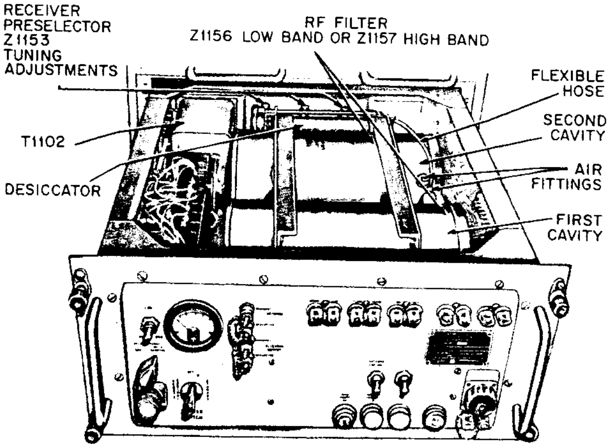

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Figure 6-10.—Control-duplexer, overall view.

The transmission line filter is located in the line between the transmitter and the antenna, with the receiver input connection branching off on the antenna side of the filter. The receiver input path also contains a filter, Z1153. The transmission line filters (Z1156 and Z1157) are tuned 63 megacycles, one above, and the other below the receiver frequency, so that Z1153 appears as an open circuit at the transmitter frequency. Thus, the transmitter output is fed directly to the antenna.

Because of the high Q's associated with the resonant cavities, the control-duplexer exhibits sharp frequency response characteristics. In this way, the transmission of adjacent frequencies is reduced.

The low-band transmission line filter, Z1156, is used at frequencies between 960 and 1025 mc. Z1157, the high-band filter, is used at frequencies between 1150 and 1215 mc. The performance standards for the transmission line filters are as follows:

1. The signal insertion loss at the transmitter frequency is about 1.4 db maximum.
2. Within ±0.2 mc of the transmitter frequency, the filter response is within 3 db of maximum.
3. At ±0.75 mc from the transmitter output frequency, the filter response is a minimum of 20 db down.
4. The transmitter response curve is down sufficiently at the receiver frequency (transmitter frequency ±63 mc) so that the transmitter

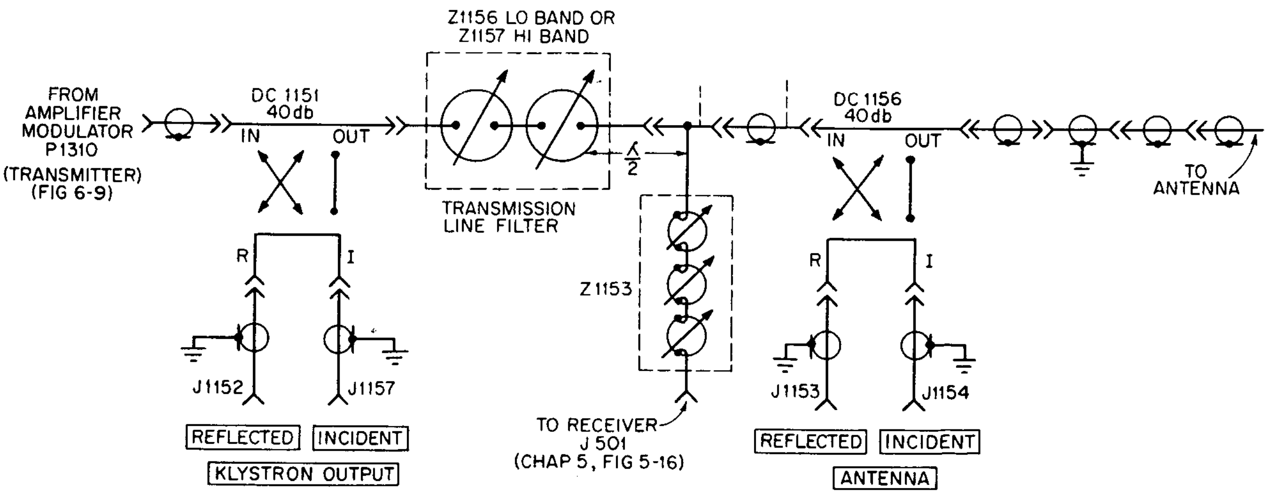

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Figure 6-11.—Control-duplexer, schematic diagram.

frequency is completely rejected at the receiver.

5. The temperature operating range of the filter is from -54 to +65 C.

The preselector cavities for the receiver (A, B, and C, of Z1153) are tuned to the receiver frequency, which is 63 mc above the transmitter frequency for the low-band, and 63 mc below the transmitter frequency for high-band operation. Because of the wide frequency separation, signals from the antenna at the receiver frequency are routed to the radio receiver, and signals at the transmitter frequency are rejected.

Samples of the klystron-incident and klystron-reflected fields and of the antenna-incident and antenna-reflected fields are made available by directional couplers, DC1151 and DC1156, respectively. During tuning, the klystron output may be obtained at J1152 and J1157, while the antenna output can be obtained at J1153 and J1154.

The control-duplexer contains a desiccator (drying agent) that controls the humidity of the air in the two filter cavities (Z1156 or Z1157 and Z1153). Because of temperature changes, the filter cavities tend to breathe (draw in and expel air). This air exchange takes place through the desiccator, with the desiccant absorbing moisture from in the air, thereby maintaining a relatively dry atmosphere in the cavities.

ANTENNA SPEED CONTROL

As previously discussed, the speed of rotation of the antenna cylinders must be maintained at 900 rps, since this speed directly determines the number of times per second that the antenna lobes will pass a given point in space. The 15-cps lobe should pass a given point 15 times per second, resulting in 15-cps modulation of the transmitted energy at that point. The 135-cps lobes should pass a given point at a 135-cps rate, to produce a 135-cps modulation at a given position.

An antenna speed control servo system is employed in the beacon which functions to maintain the beacon antenna rotation at the required speed (900 rpm). The speed control system includes a spin motor, B3203 (chap. 5, fig. 5-12), a speed control tachometer, G3202, a magnetic preamplifier, AR2080, and a power magnetic amplifier, AR2081. A block diagram discussion of the preamplifier and power amplifier sections of the control system is given in chapter 5 of this training course with the aid of the block diagram of chapter 5, figure 5-12.

PHASE SENSITIVE DETECTOR

The voltage developed by the tachometer is fed to the primary of the phase sensitive detector transformer, T2101 (fig. 6-12). T2101 has two secondaries, corresponding to terminals 5-7 and 3-4, respectively. The secondary output from the 5-7 terminals of T2101 is applied through R2104 and R2102, and R2103 and R2101, respectively, to diagonally opposite corners of two associated bridge rectifiers A and B. The 3-4 secondary winding output of T2101 is fed through a series tuned circuit, and is coupled from the secondary (terminals 4-6 of Z2101) to the other set of diagonally opposite corners of bridge rectifiers A and B. The tuned circuit includes the primary of the reference transformer and is adjusted to resonance at 675-cps by C2103.

The rectifier load is connected from the center-tap (terminal 5) of Z2101 to the center-tap (terminal 6) of T2101, and includes C2102 in series with the 10-6 winding of T2101. C2102 is paralleled by the error indicator, and its series current limiting resistor, R2110, in one path, and by the voltage magnetic preamplifier control winding (terminals 7-8 of L2103), which is connected in series with the parallel combination of R2112 and C2101 in the other path.

The frequency of the reference voltage for the phase sensitive detector is stabilized by the flywheel effect of the series resonant circuit comprising, C2103 (fig. 6-12) and the reference transformer primary, terminals 1 and 15 of Z2101. Because this circuit is series resonant at 675 cps, the voltage (IX_L) across terminals 1-15 is 90° out-of-phase with the input voltage terminals 3-4 of T2101, and therefore the secondary voltage across terminals 4-6 of Z2101 is 90° out-of-phase with the tachometer voltage across terminals 5-7 of T2101. The relation is indicated in the sine waveforms and vectors in figure 6-13, A, and represents the condition for zero error signal.

Because of the 90° relation (for zero error) between E_{4-6} and E_{5-7} the polarities of these voltages will reverse, with respect to each other, 4 times during each cycle of applied tachometer voltage.

The relative polarities of the E_{5-7} and E_{4-6} voltages during the first 90° (designated 1) are shown in figure 6-13, A. The current through the load (R_L) during this period is to the left as indicated by the arrow, and follows the path (fig. 6-13, C) from terminal 6 of Z2101, through the lower right B rectifier, through R2103, through the 7-6 terminals of T2101,

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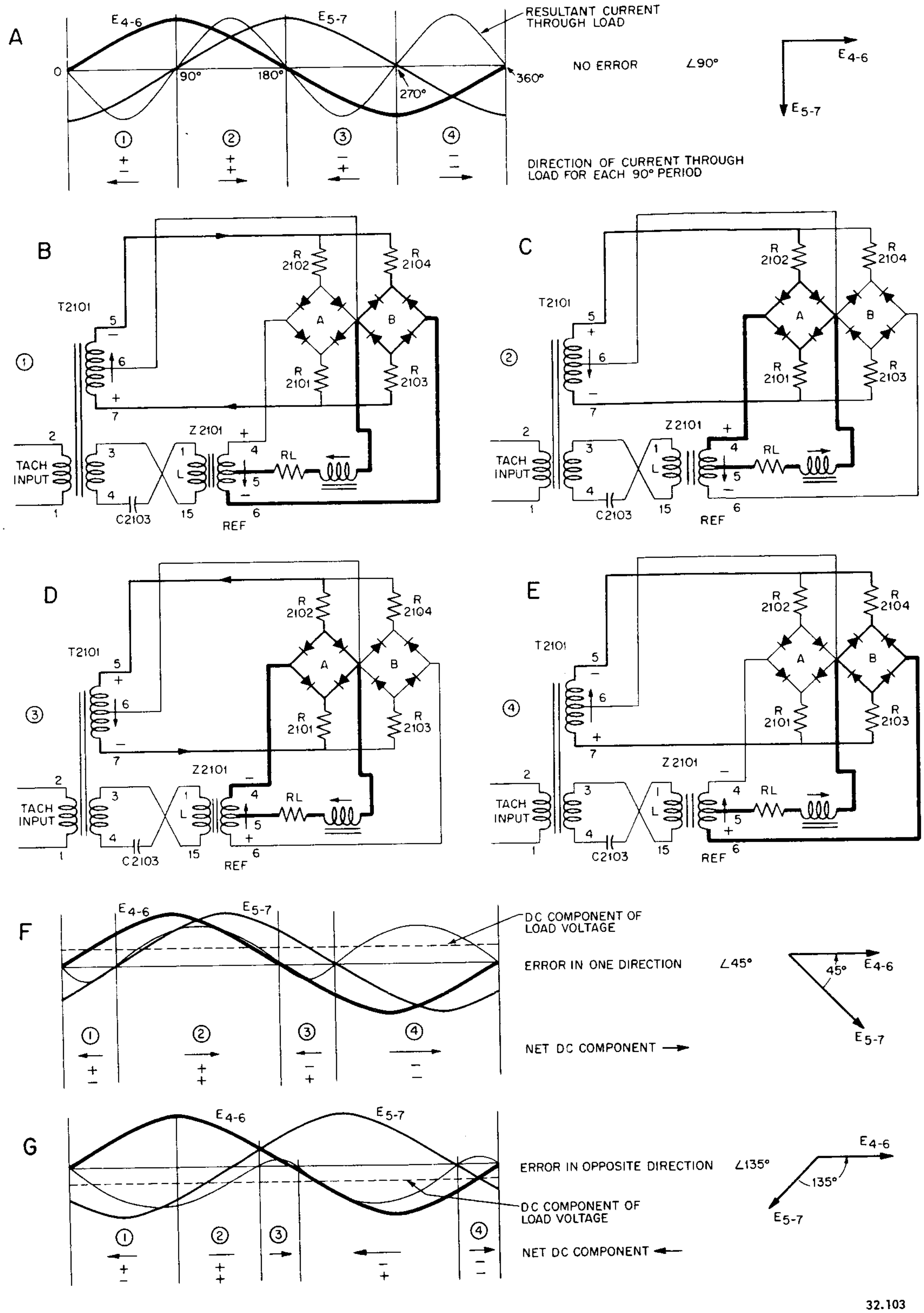

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Figure 6-13.—Simplified diagram showing operation of phase-sensitive detector.

through the load, R_L, to terminal 5 of Z2101, thereby completing the circuit path.

During the second 90° period (designated ② in figure 6-13, A) the current through the load is in the opposite direction. The polarities at T2101 and Z2101 during this period are shown in figure 6-13, C. The current path is from terminal 5 of Z2101, through the load (R_L), through terminals 6-7 of T2101, through R2101, and the lower left A rectifier, returning to terminal 4 of Z2101.

The direction of the load current during the following two 90° periods is as shown in figure 6-13, D, and figure 6-13, E. You will note that the direction of the load current during periods ① and ③ is the same, and that the direction during periods ② and ④ is reversed with respect to ① and ③. The amplitudes of the load currents for alternate quarter cycles is the same and the net d-c component across the load is zero.

The conditions just described represent a condition when no error signal is introduced to the phase sensitive detector from the tachometer. If the speed on the antenna tries to increase, the frequency of the tachometer voltage at the 1-2 primary winding of T2101 also tries to increase. However, because of the phase detector action, the feedback signal prevents any more than a few degrees of phase shift between the two voltages. The phase of the voltage induced in the 5-7 winding of T2101 (with respect to the no-error voltage in the 5-7 winding) will be advanced. When the error signal is received, the flywheel effect of the reference tuned circuit will sustain this circuit at 675 cps for a period dependent on the circuit Q. Thus, the phase angle between the voltage in the 5-7 winding of T2101 and the reference voltage in the 4-6 winding of Z2101 decreases below 90°. This condition is represented by the vector at figure 6-13, F.

Even though the tachometer voltage may vary ±2 cps, the two voltages E_{5-7} and E_{4-6} can never deviate from their 90° relationship more than a fraction of a cycle because they are derived from two windings on a common transformer core, the primary of which supplies the tachometer input voltage.

Note in figure 6-13, F, that the resulting phase shift in the 5-7 voltage with respect to the 4-6 voltage has produced an unbalance in the periods during which current flows in one direction through the load. The resulting polarities at the 5-7 secondary of T2101, and the 4-6 secondary of Z2101 during periods ② and ④ (fig. 6-13, F) will occur over a longer period than the reverse polarities during periods ① and ③. Thus, the load current during periods ② and ④ will remain in the same direction longer than during periods ① and ③. It follows that the load current will not be a pure a-c (as was the case for the no-error condition) but will contain a d-c component which is proportional to the phase shifts in the 5-7 and 4-6 voltages. This d-c component is the error signal and varies with the phase shift from the 90° relation (fig. 6-13, A). Thus, the shift in phase of the two input voltages has produced a d-c component of load current as indicated in figure 6-13, F. Note that the circuit is sensitive to phase variations in the E_{5-7} and E_{4-6} voltages, and is therefore called a phase sensitive detector. The error signal is amplified and applied to the spin motor in a manner to prevent an increase in speed.

If the tachometer input error signal were in the opposite direction as a result of an attempted decrease in antenna speed, the voltage at the 5-7 secondary of T2101 will lag the reference voltage at the 4-6 secondary of Z2101. This produces a phase relationship between the two voltages as shown in figure 6-13, G. Now the current through the load produces a predominant voltage across the load with the d-c component in the opposite direction to that in figure 6-13, F.

The output error voltage from the phase sensitive detector is applied to the voltage preamplifier control winding in series with a network consisting of R2112 and C2101. The large capacitance of C2101 permits this capacitor to charge to the d-c component of the phase detector output. The discharge of C2101 through R2112 in parallel with the 7-8 control winding of L2103 produces a more steady d-c control current to the voltage preamplifier. C2101 also removes noise pulses which might cause false operation of the voltage preamplifier.

VOLTAGE MAGNETIC PREAMPLIFIER

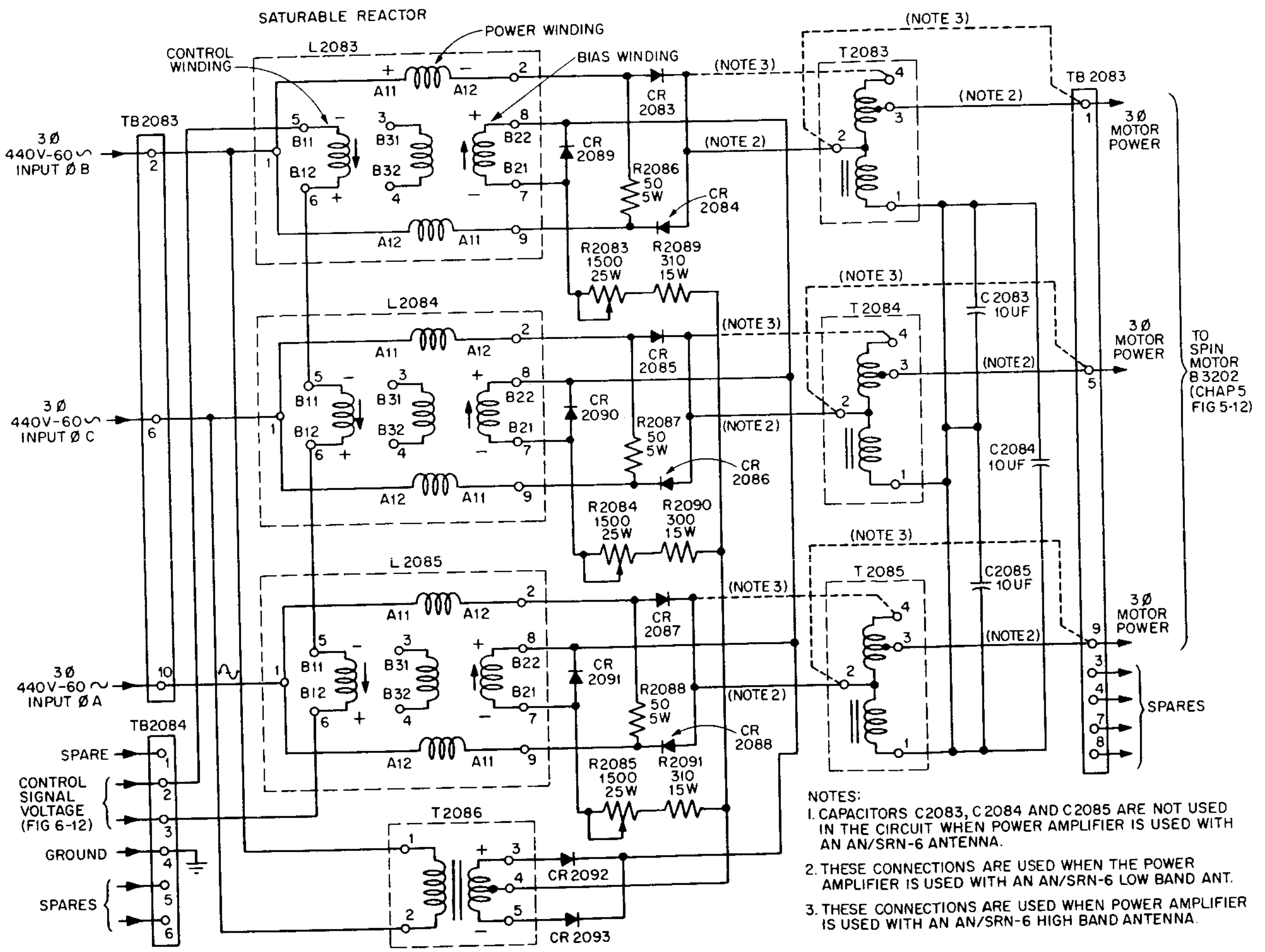
The voltage magnetic preamplifier, L2103, receives the speed error signal from the phase sensitive detector at its 7-8 control winding, amplifies the error signal, and converts it to an equivalent d-c component. This stage supplies the control current to the antenna speed control power amplifier (discussed later). The power amplifier, in turn, controls the speed of the spin motor.

Bias in the 1-2 winding of the voltage preamplifier is established by the bridge rectifier ((C) fig. 6-12). The direction of the bias magnetomotive force (mmf) is represented by the arrow below the 1-2 winding.

The a-c input voltage applied between terminals 13 and 14 of L2103 is half-wave rectified. The conduction path is from terminal 4 of the transformer through CR6, through R2117 (in one path), through CR3, and through the paralleled circuit comprising CR1 and the A winding of L2103 in one branch, and CR2 and the B winding in the other path, returning to terminal 3 on the transformer. R2117 is paralleled by the 9-10 feedback winding of L2103 in series with R2113, and by the power amplifier control winding via terminals 3 and 2 of TB2108.

The direct current in the 9-10 feedback winding of L2103 and the control winding of the preamplifier (fig. 6-14) establishes the quiescent (no-error) mmf's in these windings. Opposing mmf's are established in the A and B windings of L2103 (fig. 6-12), causing these windings essentially to block out the 60 cycle power supply frequency from the control and feedback windings. The A and B windings of L2103 are connected in parallel, and the parallel combination of these windings is connected in series with the control winding of the power amplifier. Thus, as the impedance of either the A or B winding decreases, the current through the power amplifier control winding increases.

If the error signal aids or reinforces the bias mmf (as shown by the solid arrow at the control winding), the A and B winding impedance decreases, and the current through the power amplifier control winding (fig. 6-14) increases. Conversely, if the error signal in the 7-8 control winding (fig. 6-12) opposes the bias mmf, the impedance of the A and B winding increases and the power amplifier control winding mmf decreases.



32.104

Figure 6-14.—Antenna speed control power amplifier.

Positive feedback is obtained from the 9-10 feedback winding of L2103 which tends to aid the action of the amplifier as determined by the mmf of the control winding. The positive feedback increases the gain of the amplifier.

Note that if the impedance of the A and B windings is decreasing (core material becoming saturated) the current through the bridge rectifier and through the 9-10 feedback windings also increase. Thus, the mmf of the feedback winding also tries to increase the saturation of L2103. Conversely, if the A and B windings impedance are increasing, the rectified current is decreasing, and the 9-10 feedback winding mmf also decreases. This action aids the mmf of the A and B windings to further the de-saturation of the core material of L2103.

L2102 is a stabilizing inductor placed in the circuit to compensate for variations in the reactor magnetic field.

POWER AMPLIFIER

The function of the power amplifier (fig. 6-14) is to control the amplitude of the three-phase power input to the antenna spin motor and thereby control the speed of the antenna rotation. The control signal is developed in the phase sensitive detector (fig. 6-12), amplified in the voltage amplifier, and fed directly to the series connected 5-6 control windings of L2083, L2084, and L2085 in the power amplifier (fig. 6-14).

The output saturable reactors, L2083, L2084, and L2085, are each in series with one phase of the 3-phase power line. During the positive half-cycle of the line voltage input at terminal 1 of L2083, L2084, and L2085, a current passes through the 1-2 power windings of each of the reactors through CR2083, CR2085, and CR2087, through the 2-3 windings of T2083, T2084, and T2085, and through the respective windings of the spin motor. During the period of the negative input at terminal 1 of the reactors, the conduction path is through the 1-9 power windings of the reactors and through CR2084, CR2086, and CR2088, respectively.

Since all of the components in the separate windings are series connected, it follows that an impedance increase of either of the components will decrease the motor voltage, and the speed of the motor will decrease. On the other hand, an impedance decrease in either of the series connected components will increase the spin motor voltage, and the motor speed will increase.

You will recall that the control (error) signal in the 5-6 windings of the reactors is applied from the voltage amplifier of figure 6-12. This signal enters the power amplifier via terminals 2 and 3 of TB2084. Since a d-c current passes through the 5-6 control winding of the reactors, the control mmf established in these windings is always in the same direction, increasing or decreasing as the error signal changes direction.

Bias mmf is established in the 7-8 windings of each of the reactors by the action of the transformer, T2086, and the diodes, CR2092 and CR2093. CR2092 conducts when terminal 3 is positive with respect to terminal 5, and CR2093 conducts when terminal 5 of the transformer is positive with respect to terminal 3. The rectifier diodes (CR2089 thru CR2091) connected across each of the respective windings suppress harmonics introduced in the control and bias windings of the reactors. R2083, R2084, and R2085 are used to regulate the current, and therefore the 7-8 winding bias mmf.

In the quiescent condition (no-error signal), the bias and control mmf's are in opposition. This action sets the degree of core saturation to a point which will allow either an increase or decrease in the saturation of the core, depending on the direction of the error signal.

The speed of the 3-phase spin motor is controlled by varying the voltage applied to the stator windings. The cage rotor has relatively high resistance so that any change in voltage applied to the stator (primary) will vary the strength of the revolving field and thus vary the voltage and current induced in the rotor (secondary).

The voltage applied to the stator winding of the spin motor is varied by changing the effective impedance of the series saturable reactors in each line wire to the motor.

Since the action of each of the reactors is the same, the following discussion of L2083 can be applied to the other reactors (L2084 and L2085) as well.

The impedance of the 1-2 and 1-9 power windings of saturable reactor, L2083, is determined by the saturation of the L2083 core material. The control and bias windings affect the core material so that the impedance of the 1-2 and 1-9 power windings is decreased for one error signal in the 5-6 control winding, and increased for an error signal in the opposite direction.

If the error signal causes an increase in the 5-6 control winding mmf, the greater opposing mmf of this winding decreases the 7-8 bias winding mmf. This action decreases the degree of saturation of the core of L2083, and the impedance of the 1-2 and 1-9 windings increases.

The increased impedance of these windings lowers the spin motor voltage, and the motor decelerates.

Conversely, if the error signal in the 5-6 control windings decreases the control mmf, the bias mmf will be less opposed. The degree of saturation of the L2083 core material will be increased, and the effective impedance of the 1-2 and 1-9 power windings is decreased. Thus, the motor voltage is increased and the motor accelerates.

BEARING SERVO SYSTEM

The bearing servo system of the radio beacon (ch. 5, figs. 5-13 and 5-14) provides an electromechanical means of adjusting the pulser-plate of the beacon to ensure that any calibrated AN/ARN-21 receiver will receive the proper magnetic bearing. The pulser-plate positions the 15-cps pulser coil so that the 15-cps reference burst of the radio beacon is in its correct position relative to magnetic north. The 15-cps reference burst signal is pulsed so that the tenth pulse of the train of pulses of the 15-cps burst will occur precisely at the positive zero crossover of the 15-cps fundamental and 135-cps ninth harmonic.

The bearing servo system used with the stabilized shipboard antenna operates as follows: The angular position signals for true north from the ship's gyro compass are continuously transmitted through the ship's bus to the radio beacon. The magnetic variation unit located in the coder-indicator corrects the ship's gyro compass information for the difference between magnetic north and true north. This corrected signal maintains the radio beacon 15-cps reference pulse coil subassembly fixed with respect to corrected (magnetic) north.

The magnetic correction signals are manually set into the 1- and 36-speed differential transmitters, CDX601 and CDX602, of the magnetic variation subassembly. The sum of the gyro compass and magnetic variation signals is forwarded to the bearing 1- and 36-speed control transformers, T3201 and T3202, in the antenna base.

As long as the rotors of the control transformers are at the same effective angular position as those of the differential transmitters after summing up the signals of the ship's gyro compass and magnetic variation signals, no control transformer output signal is sent to the bearing servo amplifier. If the equilibrium should be upset because of displacement of the rotors of CDX601 and CDX602 to new positions by changing the setting of the magnetic variation unit, or if the ship's course were to be changed, an error voltage would be introduced into the differential transmitters and transmitted from the control transformers to the servo amplifier.

The magnitude of the error voltage fed to the balanced bearing magnetic amplifier is a function of the rate of change of the control transformer shaft position. If the error signal from the differential transmitters exceeds 2.5 volts (the equivalent of a shaft position 2.5 degrees away from null), the one speed transformer voltage takes control. For shaft position less than 2.5 degrees away from null (equivalent to voltages below 2.5 volts), the output of the 36-speed control transformer is the controlling signal.

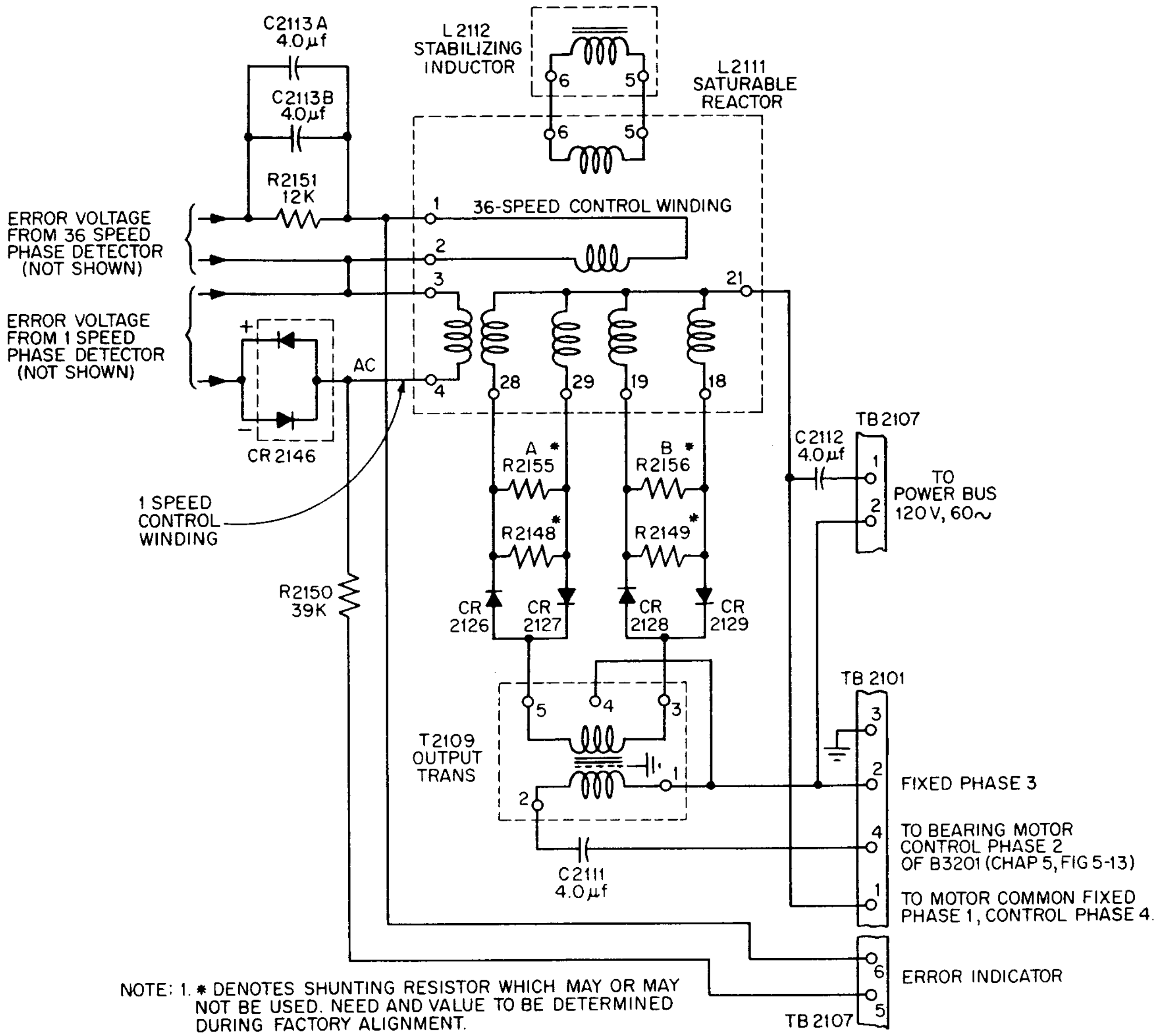
BEARING MAGNETIC AMPLIFIER

The bearing magnetic amplifier utilizes two phase-sensitive detectors; one for the 1-speed error voltage, and the other for the 36-speed error voltage. The phase-sensitive detectors of the bearing magnetic amplifier circuit are not shown but are similar to the phase-sensitive detector presented in the discussion of the antenna speed control system. The output signals of the phase-sensitive detectors are fed to respective control windings of saturable reactor, L2111 (fig. 6-15).

A de-emphasis network, CR2146, is connected in series with the 1-speed control winding. This network uses two rectifiers connected back-to-back in order to obtain a nonlinear bilateral resistance. The resistance of this circuit is high at low levels of voltage, owing to the characteristics of the diodes (called Zener effect), and the resistance is decreased considerably for high levels of voltage. Thus, when the 1-speed transformer error voltage is low, the gain of the 1-speed circuit is decreased. Conversely, when the error voltage is large, the gain is increased and the 1-speed error voltage has the greater control. Under this condition, the lowered resistance of the CR2146 diode pair permits a much higher current to enter the 1-speed control winding.

The series rectifiers (CR2126, CR2127, and CR2128, CR2129), one in each leg of the balanced amplifier circuit, prevent the load current in the leg from reversing. The circuit is said to be balanced, since the center-tap of the output transformer, T2109, is at the electrical center of each branch of the circuit.

The two branches of the amplifier circuit are designated the A and B branches, respectively.



32.105

Figure 6-15.—Balanced bearing magnetic amplifier, schematic diagram.

The flux of branch A is in opposition to the flux of branch B.

The 120-volt 60-cycle input at terminals 1-2 of TB2107 cause a quiescent current alternately through the two legs of the A and B branches of the amplifier. The resulting voltage developed across the output transformer, T2109, for the quiescent condition is always zero.

The consideration of the circuit operation of the magnetic amplifier in the bearing and the roll and pitch servo systems will be limited to a general explanation, pointing out the major operating features of the circuit. For a detailed explanation of a representative magnetic amplifier in the AN/SRN-6, refer to the discussion of the antenna speed control servo system discussed earlier in this chapter.

Parallel resistances R2148, R2155 and R2149, R2156, connected across the load windings, are used for balancing the a-c output across the 5-4 and 4-3 terminals of output transformer, T2109. The value of the resistances is chosen to give zero a-c voltage output for zero d-c current in the control windings. These resistors also set the bias current in the load windings.

The operation of the amplifier is as follows: The mmf of branch A is in opposition to the mmf of branch B. Assume that there is an error voltage which tends to saturate the core of branch A, and desaturate the core of branch B. This condition unbalances the circuit, resulting in a heavy current flow in the load winding of the saturated branch. If the polarity of the error voltage is reversed, branch B becomes saturated, while branch A becomes desaturated.

The branch conducting the greater current will produce a predominant alternating flux across one-half of the primary of the output transformer, T2109. This output is coupled from the secondary of T2109 to the control phase winding of the bearing servo motor, B3201 (ch. 5, fig. 5-12).

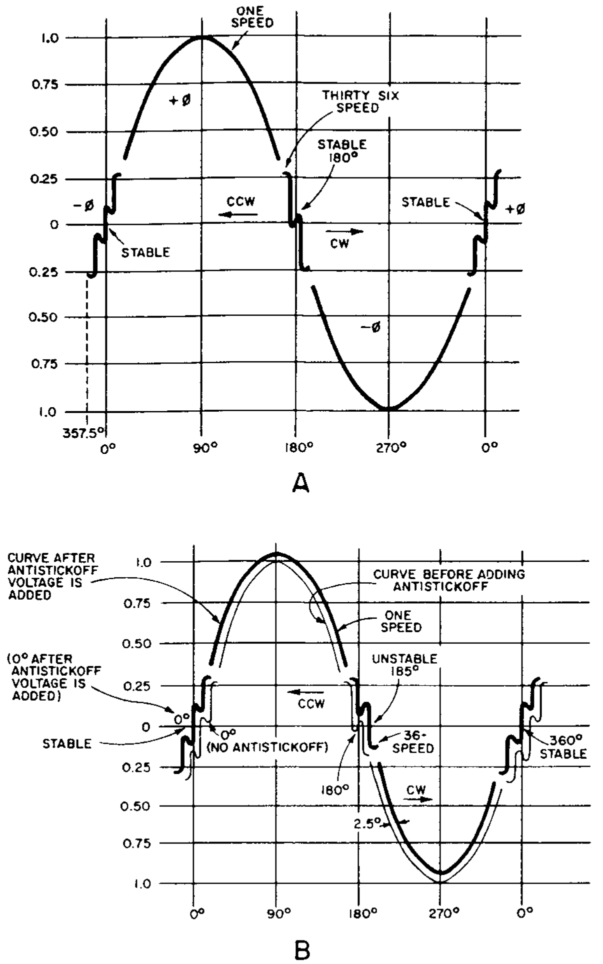
Because of the balanced arrangement of the two branches of the amplifier (fig. 6-15) with respect to the center-tap, the mmf in the 5-4 section of T2109 will be predominant for one error and the mmf of the 3-4 section of T2109 will predominate for the opposite error. Thus, the current through the motor (and the direction of motor rotation) reverses with a reversal of error voltage.

ANTISTICKOFF VOLTAGE

As stated earlier, signals from the magnetic variation subassembly of the coder-indicator are introduced into the 1- and 36-speed differential transmitters in order to cause the pulser plate assembly in the antenna pedestal to be corrected to magnetic north. A single dial (ch. 5, fig. 5-13) calibrated from 0 to 360°, and fastened to the coarse (1-speed) differential transmitter and control transformer (CDX601 and T3201, fig. 5-12) yields only a fair degree of accuracy in providing information to the servo motor, B3201. A second dial, fastened to the fine (36-speed) differential transmitter and control transformer, CDX602 and T3202, is calibrated so that one revolution of the fine control transformer, T3202, equals only 10° of the full coverage of the coarse control transformer, T3201. This provides a method of reading the position of the control shaft with 36 times the accuracy of a system when only the coarse control is provided.

The voltage which controls the bearing servo motor is shown in figure 6-16, A. The curves shown have the appearance of sine waves but actually represent an infinite number of rms voltage values for error signals throughout a 360° rotation of the rotor of the control transformer.

The amplified control transformer output drives the bearing motor. The motor is a 2-phase induction motor having a continuously



32.106

A. Pure 1- and 36-speed synchro output voltages (no antistickoff voltage). B. 1- and 36-speed voltage with antistickoff voltage added.

Figure 6-16.—Addition of antistickoff voltage.

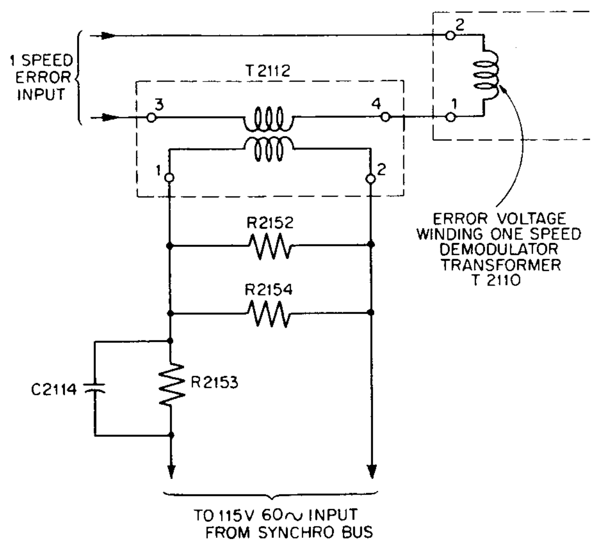
excited fixed phase and a variable control phase. The direction of the motor torque is determined by the instantaneous polarity of the control voltage from the control transformer with respect to that of the fixed phase.

When the error voltage is any value above the zero reference line, the direction of the motor rotation is assumed to be counterclockwise (ccw). Conversely, when the error voltage is below the axis the instantaneous polarity of the control phase of the motor is reversed with respect to the fixed phase and the motor rotates in the opposite direction (clockwise).

If the error voltage is less than 2.5° a deemphasis network which consists of Zener diodes connected as shown in figure 6-15, permits the 36-speed voltage to control the motor. For error signals greater than 2.5° the deemphasis network breaks down and the 1-speed voltage controls the motor. Thus, in figure 6-16, A, the portion of the curve from 357.5° to 2.5° (corresponding to 2.5° on either side of 0°), the 36-speed voltage controls the servo system. For errors larger than 2.5° the 1-speed voltage is shown to be the controlling signal.

The use of an even gear ratio between the fine and coarse control transformers presents a problem in synchronizing the servo system. Note that at 180° (fig. 6-16, A) the 36-speed voltage crosses the zero axis three times in the area from 175° to 185° to produce a negative voltage at 179° and a positive voltage at 181°. When the system is at 179° the error voltage drives the motor clockwise; at 181° the servo motor is driven counterclockwise. This causes the system to lock-in at 180°, so that the system synchronizes at a false null of 180°.

The elimination of the 180° false null is accomplished by the addition of an a-c voltage, called an antistickoff voltage (fig. 6-17) to the 1-speed voltage (fig. 6-16, B). The addition of the antistickoff voltage shifts the curve upward at 180° so that the crossover point occurs at 185°. The 180° false null condition cannot exist, since for an error condition of 185° or less, the error voltage is positive and the servo motor is driven counterclockwise away from the false null to null at 0°. If the error condition exists



32.107

Figure 6-17.—Method of adding antistickoff voltage.

above 185° the servo motor will be driven clockwise away from the false null to null at 360° or zero degrees. Thus, the addition of the antistickoff voltage to the 1-speed control transformer voltage causes only the 0° null position to exist as a stable point in the servo system.

To correct the 1-speed zero at the zero shaft angle position, the 1-speed control transformer is re-zeroed to a point 2.5° from zero. This action cancels the error introduced by the addition of the antistickoff voltage at the zero shaft position, and adds the 5° error to the 180° crossover point.

ROLL AND PITCH SERVO SYSTEM

The roll and pitch servo systems (ch. 5, fig. 5-15, A) are separate but similar circuits which maintain the antenna in the vertical position with respect to the horizontal plane, despite roll and pitch motions of the ship on which the antenna is mounted. The roll servo system can correct up to 25 degrees of roll, while the pitch servo system can correct up to 6 degrees of pitch. Pitch and roll information is obtained from the stable element of the ship, which is gyroscopically controlled to maintain a true horizontal position.

Roll and pitch information is fed from the ship's bus through the stow-stabilized switch, S606, in the STABILIZED position (opposite to the position shown) to the respective control transformer of the roll or pitch control system. The error signal developed by the separate control transformers is fed to the respective magnetic servo amplifier (either roll or pitch). The magnetic amplifier, in turn, supplies power to the servo motor (B3002 or B3001), commanding a specific direction of rotation determined by the error signal.

The servo motor, through its associated gear train, moves the antenna to its corrected position. A tachometer supplies a rate feedback signal to its respective magnetic amplifier. At the same time, another gear train nulls the error signal at the control transformer through a separate 1:2 gear box train.

The purpose of the rate feedback loops provided by the roll and pitch tachometers is to eliminate oscillations. A tachometer is an instrument for measuring angular speed in revolutions per minute. The tachometer feedback improves the overall stability of the servo system. This action is treated presently.

The roll and pitch magnetic amplifiers are almost identical units. The following discussion applies to both systems unless otherwise noted.

The synchro error voltage from the roll or pitch control transformers (T3001 or T3002) is fed to a phase-sensitive detector (fig. 6-18). With the exception of the fixed frequency reference voltage at terminals 6-7 of T2025, the operation of the phase-sensitive detector is identical to the detector circuit explained in the discussion of the antenna speed control system. The reference voltage for the phase detector is obtained from the 115-volt, 60-cps roll and pitch reference bus.

The d-c output error voltage from the phase detector appears between the center-tap of the two secondaries of T2025. The phase-sensitive detector load is connected between these points, and consists of the 12-23, the 14-25, and the 16-27 series, aiding connected control windings of the magnetic preamplifier, L2021. Current limiting resistor, R2037, in series with the error indicator, is connected across the preamplifier control windings.

The tachometer feedback current (from either G3001 or G3002, fig. 5-15, A) is fed to one set of the control windings of the magnetic preamplifier, L2021. Interference between the phase detector and tachometer circuits is prevented by the isolation network.

The isolation network impedance in the tachometer input circuit also serves the purpose of changing the tachometer voltage from rate feedback to acceleration feedback. Rate and acceleration feedback are discussed in the following paragraphs.

Without the use of a rate-to-acceleration feedback circuit, when the load is turning at a constant speed, a steady d-c voltage (proportional to speed) would be fed back to the preamplifier in opposition to the error signal. Therefore, a greater error signal would be required to overcome the d-c voltage and produce a certain unbalance in the amplifier and a certain speed of the servo motor than would be required if there were no feedback.

In order to produce a greater error signal there must be a greater error. Since the feedback voltage is proportional to speed, the error is also proportional to speed. This is called a velocity error, or rate feedback.

To eliminate this velocity error, a group of capacitors, comprising C2022 thru C2025, are connected in the feedback circuit. These capacitors block the d-c component of the voltage. The a-c component of the voltage at the frequency of hunting (several cycles per second) is fed back without much loss. (The capacitors have a high capacitance and low reactance.)

With this circuit arrangement, only changes in speed will produce a feedback signal through the capacitors. This action is called acceleration feedback, which further improves the stability of the servo system.

The bridge rectifier consisting of CR2053, CR2054, CR2056, and CR2058, serves as a high impedance shunt across the feedback capacitors for low amplitude tachometer feedback voltages. For high amplitude feedback voltages, CR2057 has a low impedance which causes the rectifier to act as a low impedance shunt across the feedback capacitors. This feature results in the tachometer feedback signal having a greater damping effect when the servo system is correcting rapidly than when the system is correcting slowly. CR2057 is selected for its Zener characteristic, which means that the diode will have a very high front-to-back resistance ratio for low amplitude a-c signals, and a very low front-to-back resistance ratio for signals above a certain amplitude.

As stated earlier, the error signal from the phase-sensitive detector is developed across the three control windings of the preamplifier which are connected in series aiding. A rectified bias current is conducted through the 18-28 bias windings of L2021 by the action of the bridge rectifier comprising CR2029 thru CR2032. R2027 and R2028 in parallel are current limiting resistors in the bias circuit. R2025 and R2026 are balancing resistors in the separate halves of the bias windings. C2021 filters the bias voltage to produce the required steady bias current.

An error signal in the control windings of L2021 will cause an impedance increase in one of the load windings (either upper or lower), while simultaneously lowering the impedance in the other load winding. The L2021 output, after being rectified in separate bridge circuits, causes a greater controlling signal to be fed to one of the control windings of the magnetic amplifier (L2022) than to the other.

The two load windings of the magnetic amplifier, L2022, in conjunction with the 5-4 and 4-3 terminals of T2024, form a bridge circuit. The control phase winding of the roll or pitch motor is connected from the transformer center-tap to the center-tap of the two load winding circuits. During quiescence (no-error signal), the a-c input at T2024 is rectified in the bridge circuits comprising CR2045 thru CR2048, and CR2049 thru CR2052. The impedance of the 18-19 and 29-28 load windings will be equal. Since the transformer, T2024, voltage is equal across both legs of the bridge, and since the impedance in each of the legs is the same, no difference in potential will exist across the control phase winding of the motor and the motor does not rotate.

When an error signal is introduced at the 14-25 and 23-12 control windings of the L2022 magnetic amplifier, the signal will tend to saturate the core of one-half of the output bridge circuit (for instance, the top half), while simultaneously desaturating the core of the other load winding (the bottom half). The impedance of the saturated winding is lower than that of the desaturated winding, and the bridge is unbalanced. Thus, a difference in potential exists across the motor control phase winding, and the motor rotates in a given direction. The 5-4 section of T2024 will provide the source voltage for the motor winding.

When the bridge is unbalanced, the motor essentially is placed across one-half of the transformer (T2024) in series with one load winding of the bridge, the current through the motor will be alternating, changing as the transformer secondary voltage changes. If the input error signal from the magnetic preamplifier tends to saturate the core of the lower load winding, the impedance of this winding will be decreased, while the impedance of the upper load winding is simultaneously increased. In this condition, the source voltage for the motor is derived from the 4-3 section of T2024. Note that the instantaneous direction of current through the control phase of the motor will be reversed, since the instantaneous voltages across the separate halves of the transformer (T2024) are 180° out-of-phase. An instantaneous change in the control winding current with respect to the fixed phase winding will reverse the direction of rotation of the motor.

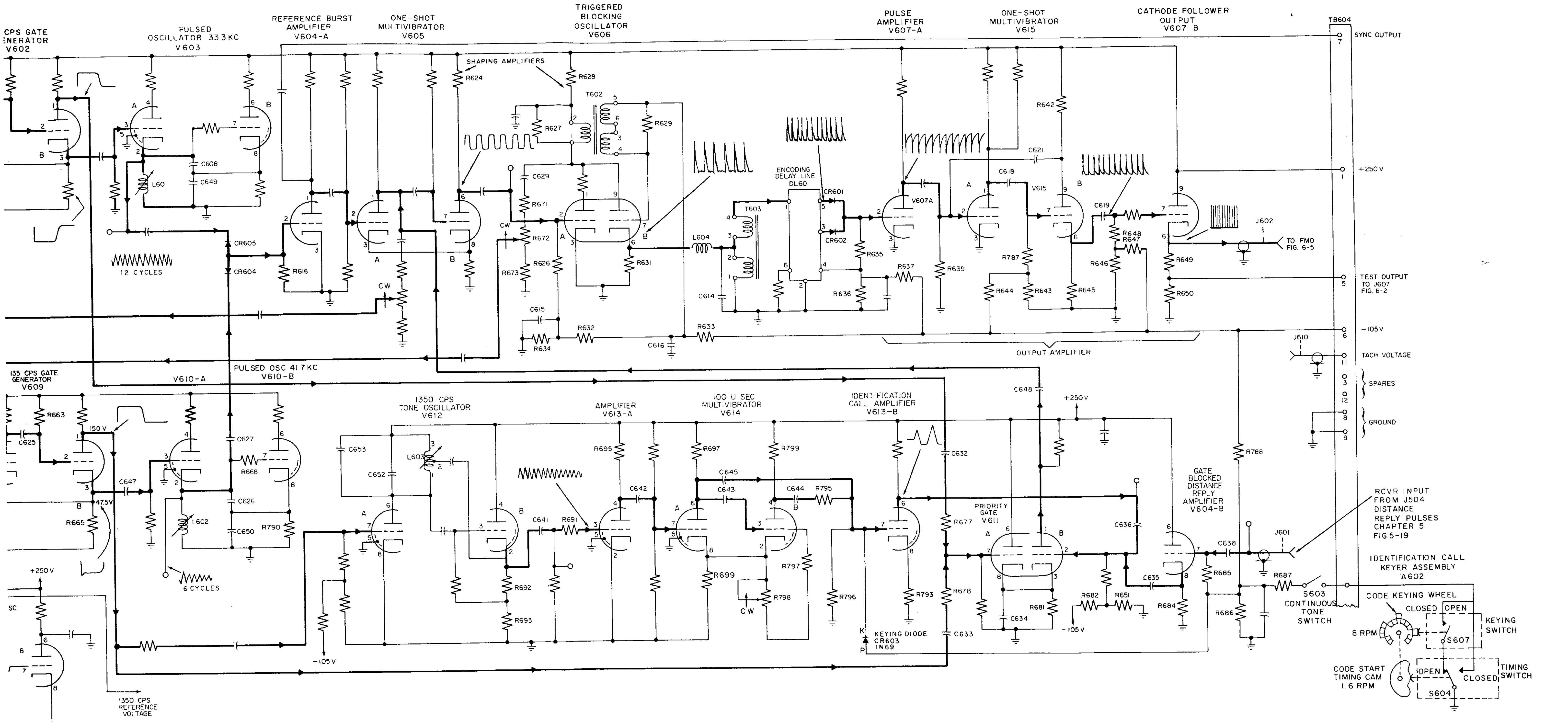
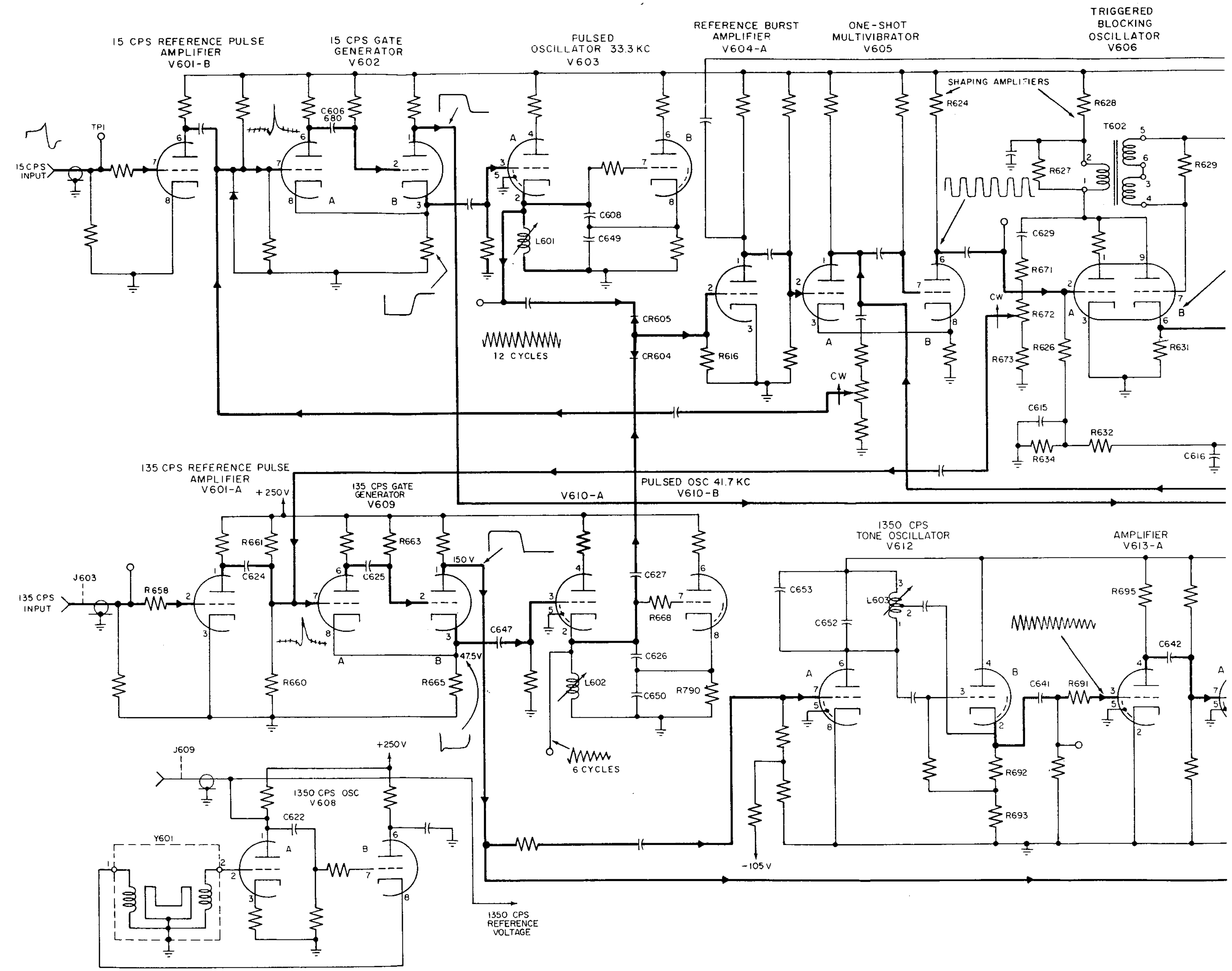
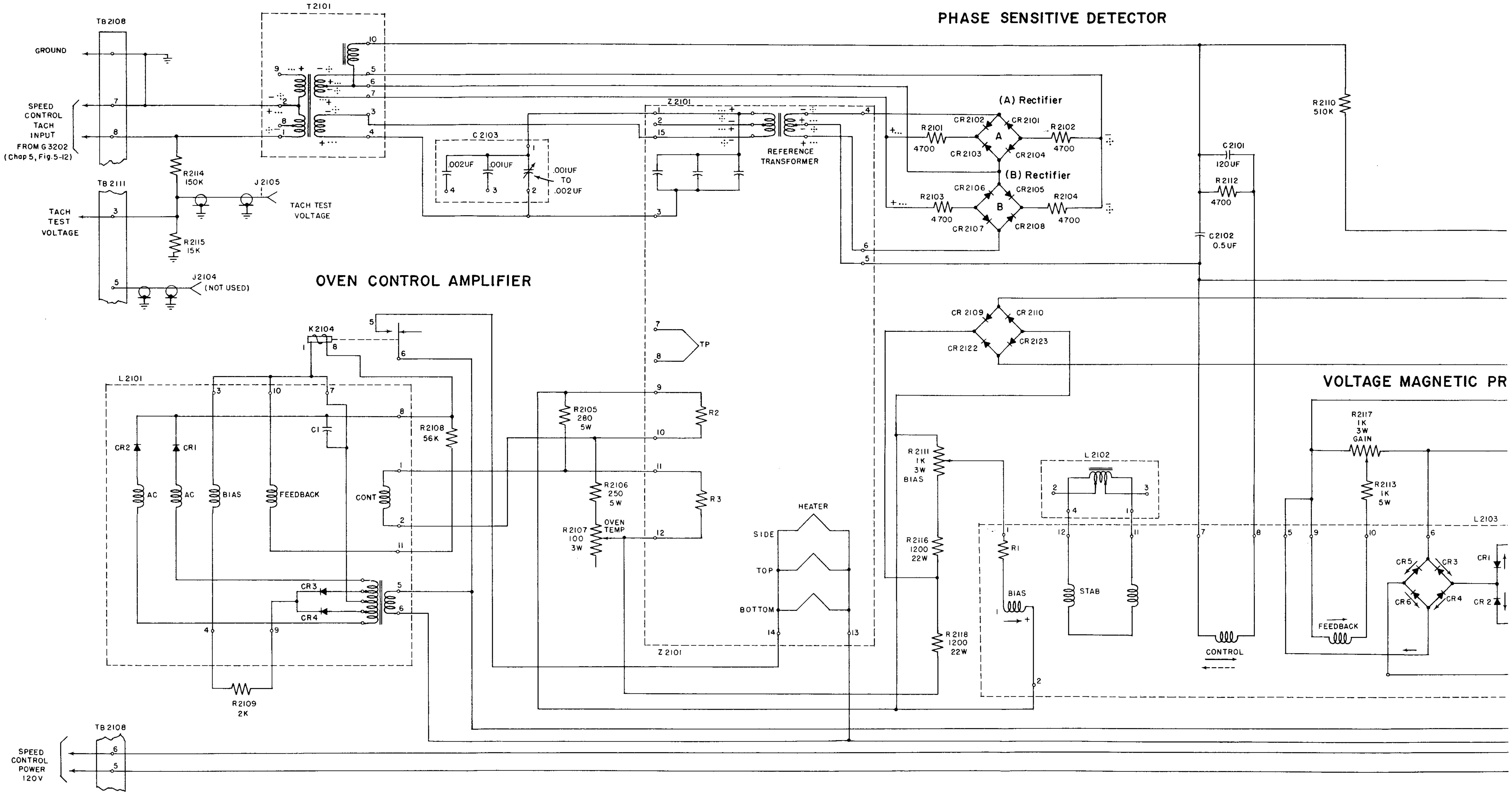


Figure 6-2.—Coder-indicator, schematic diagram.



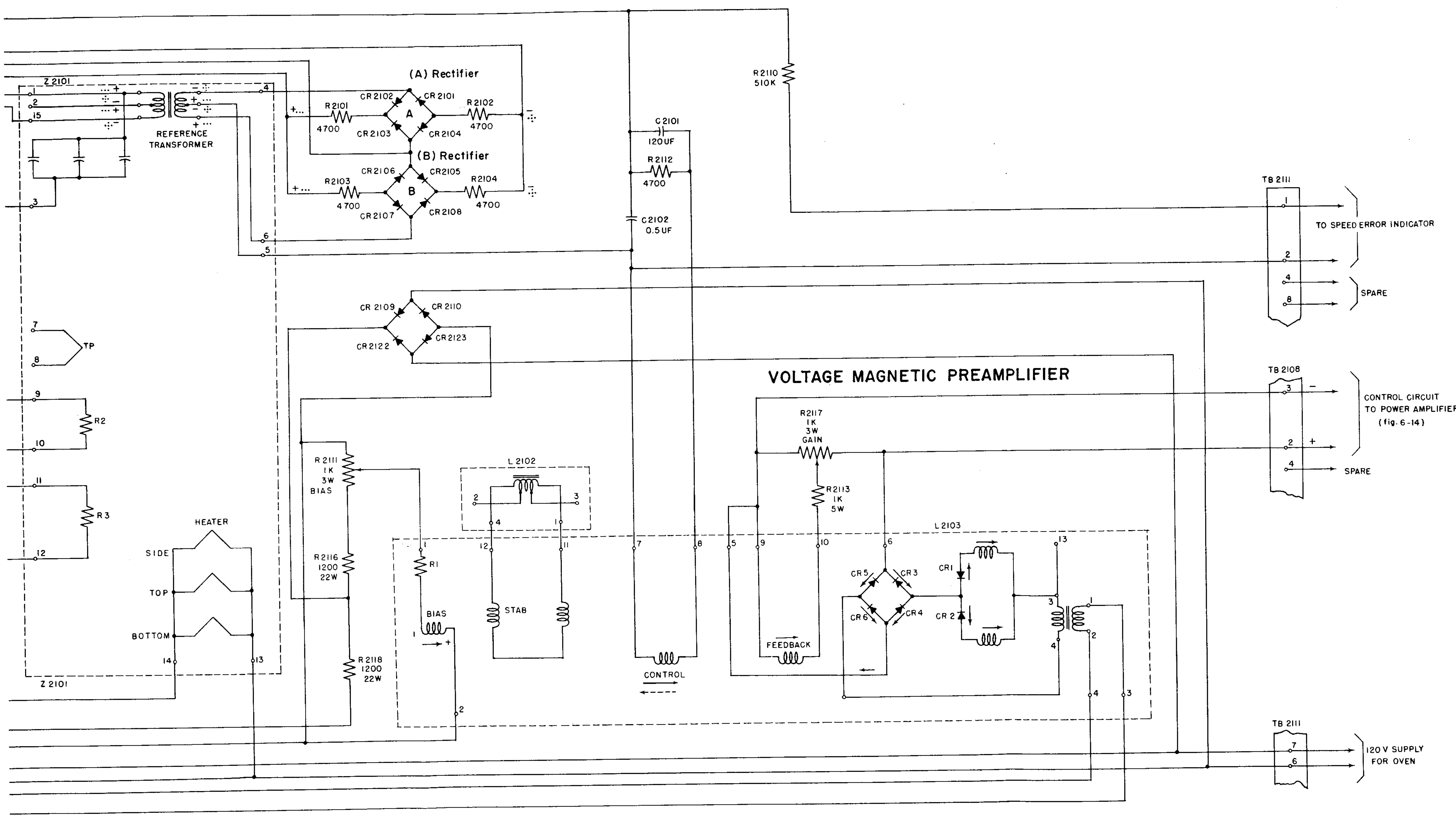
15 CPS REFERENCE PULSE AMPLIFIER V601-B
15 CPS GATE GENERATOR V602
PULSED OSCILLATOR 33.3 KC V603
REFERENCE BURST AMPLIFIER V604-A
ONE-SHOT MULTIVIBRATOR V605
TRIGGERED BLOCKING OSCILLATOR V606
SHAPING AMPLIFIERS
15 CPS INPUT
TP1
12 CYCLES
135 CPS REFERENCE PULSE AMPLIFIER V601-A
135 CPS GATE GENERATOR V609
PULSED OSC 41.7 KC
V610-A
V610-B
135 CPS INPUT
J603
+250 V
150 V
47.5 V
6 CYCLES
1350 CPS TONE OSCILLATOR V612
AMPLIFIER V613-A
J609
1350 CPS OSC V608
Y601
1350 CPS REFERENCE VOLTAGE
-105 V

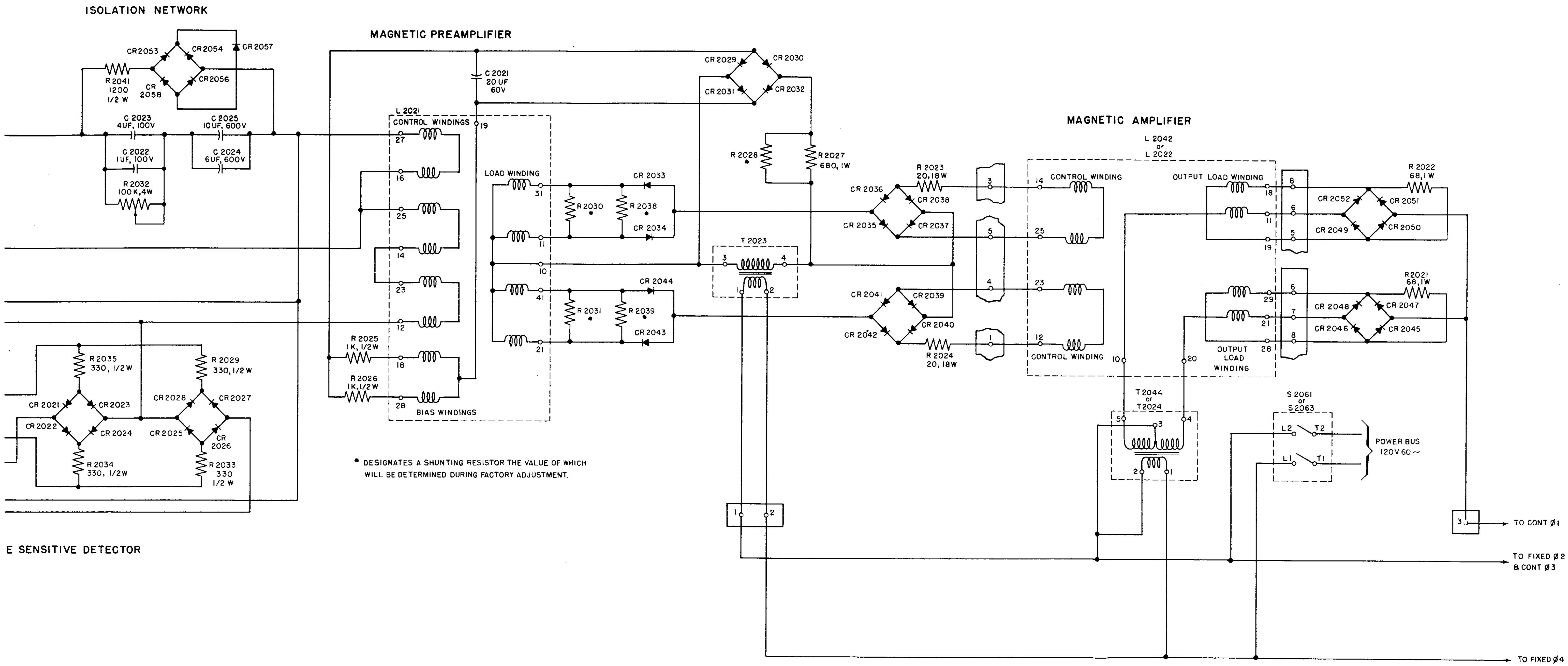


32.102

Figure 6-12.—Antenna speed control, phase-sensitive detector and amplifiers.

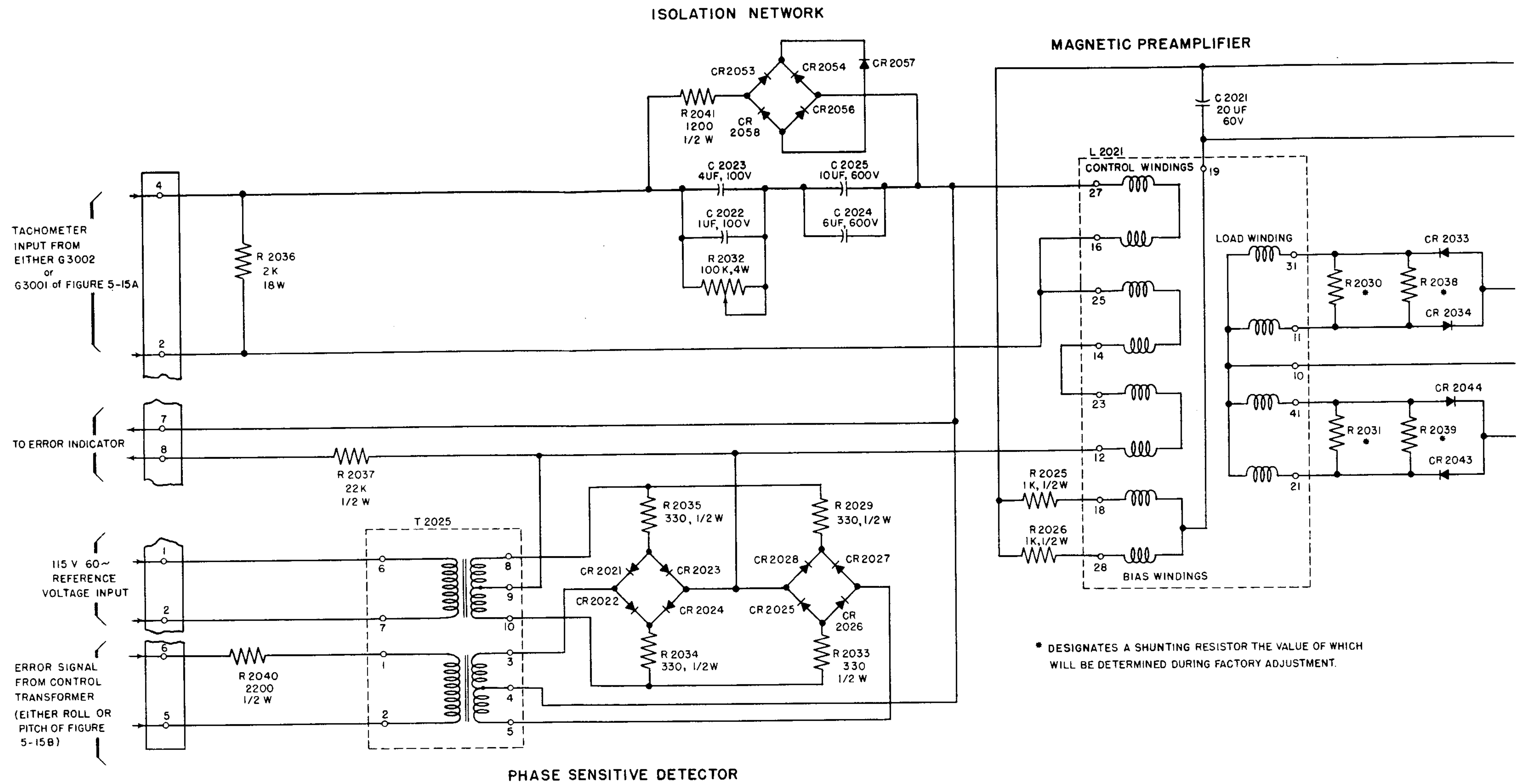
PHASE SENSITIVE DETECTOR





32.108

Figure 6-18.—Roll or pitch magnetic amplifier.



ISOLATION NETWORK
MAGNETIC PREAMPLIFIER
CR2053
CR2054
CR2057
R2041 1200 1/2 W
CR 2058
CR2056
C2023 4UF, 100V
C2025 10UF, 600V
C2022 1UF, 100V
C2024 6UF, 600V
R2032 100K, 4W
C2021 20 UF 60V
L2021
CONTROL WINDINGS
BIAS WINDINGS
LOAD WINDING
TACHOMETER INPUT FROM EITHER G3002 or G3001 of FIGURE 5-15A
R 2036 2K 18W
TO ERROR INDICATOR
R2037 22K 1/2 W
115 V 60~ REFERENCE VOLTAGE INPUT
T2025
R2035 330, 1/2 W
R2029 330, 1/2 W
CR2021
CR2023
CR2028
CR2027
CR2022
CR2024
CR2025
CR 2026
R2034 330, 1/2W
R2033 330 1/2 W
ERROR SIGNAL FROM CONTROL TRANSFORMER (EITHER ROLL OR PITCH OF FIGURE 5-15B)
R2040 2200 1/2 W
R2025 1K, 1/2W
R2026 1K, 1/2W
R2030
R2038
R2031
R2039
CR2033
CR2034
CR2044
CR2043
PHASE SENSITIVE DETECTOR
• DESIGNATES A SHUNTING RESISTOR THE VALUE OF WHICH WILL BE DETERMINED DURING FACTORY ADJUSTMENT.