NATIONAL BUREAU OF STANDARDS REPORT

9886

CIRCUIT DESCRIPTION OF A PROTOTYPE CABLE FAULT LOCATOR

NBS

U.S. DEPARTMENT OF COMMERCE

NATIONAL BUREAU OF STANDARDS

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CIRCUIT DESCRIPTION OF A PROTOTYPE CABLE FAULT LOCATOR

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for

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ABSTRACT

This report describes the circuitry and electrical performance of a prototype Cable Fault Locator test set. The primary function of the test set is to locate the position of faults in single conductor, unshielded, buried cable such as that used in series-string lighting installations. The test set consists of two major parts: an eleven-watt audio-frequency signal generator which is used to supply a test current to the faulted cable, and a hand-carried detector unit which is used to measure the relative strength of the magnetic field produced by the test current. If, for example, the fault is known to be a short to ground, it is located by carrying the detector unit along the cable route and looking for the place where the magnetic field drops sharply.

The test set described is an updated version of a test set produced by Texas Instruments Incorporated in 1955. The new test set is fully transistorized and is well protected from overloads and operator errors. The signal generator supplies a square-wave signal at either 150, 270, or 570 Hz. The operator can select either a continuous test signal; or, as an aid to identifying the test signal under poor signal conditions, an interrupted test signal can be chosen. A multiply-tapped output transformer is used to match the signal generator to its load. The detector unit is a three-band tuned amplifier. It has a "Q" of 13, maximum voltage gain of 250 000, is powered by four 1.5 volt penlite cells, and drives both headphones and an indicating meter.

PREFACE

This report is written primarily for the reader who is interested in the circuitry used in the Cable Fault Locator test set, but attention is also given to the requirements of those who operate and maintain the equipment. Section I contains background information about the uses and features of the equipment, and should be of interest to anyone having contact with the test set. The information contained in Section I and in the Controls and Characteristics portion of Sections II and III constitutes an informal set of operating instructions. Sections II and III contain the detailed circuit descriptions of the Test Signal Oscillator and of the Detector Unit, respectively. These two sections are written primarily for the circuit designer, but are believed to be readable enough to be helpful to repair personnel. The waveform and voltage measurements, performance data, and parts lists should be especially useful to the repairman.

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Section 1

INTRODUCTION

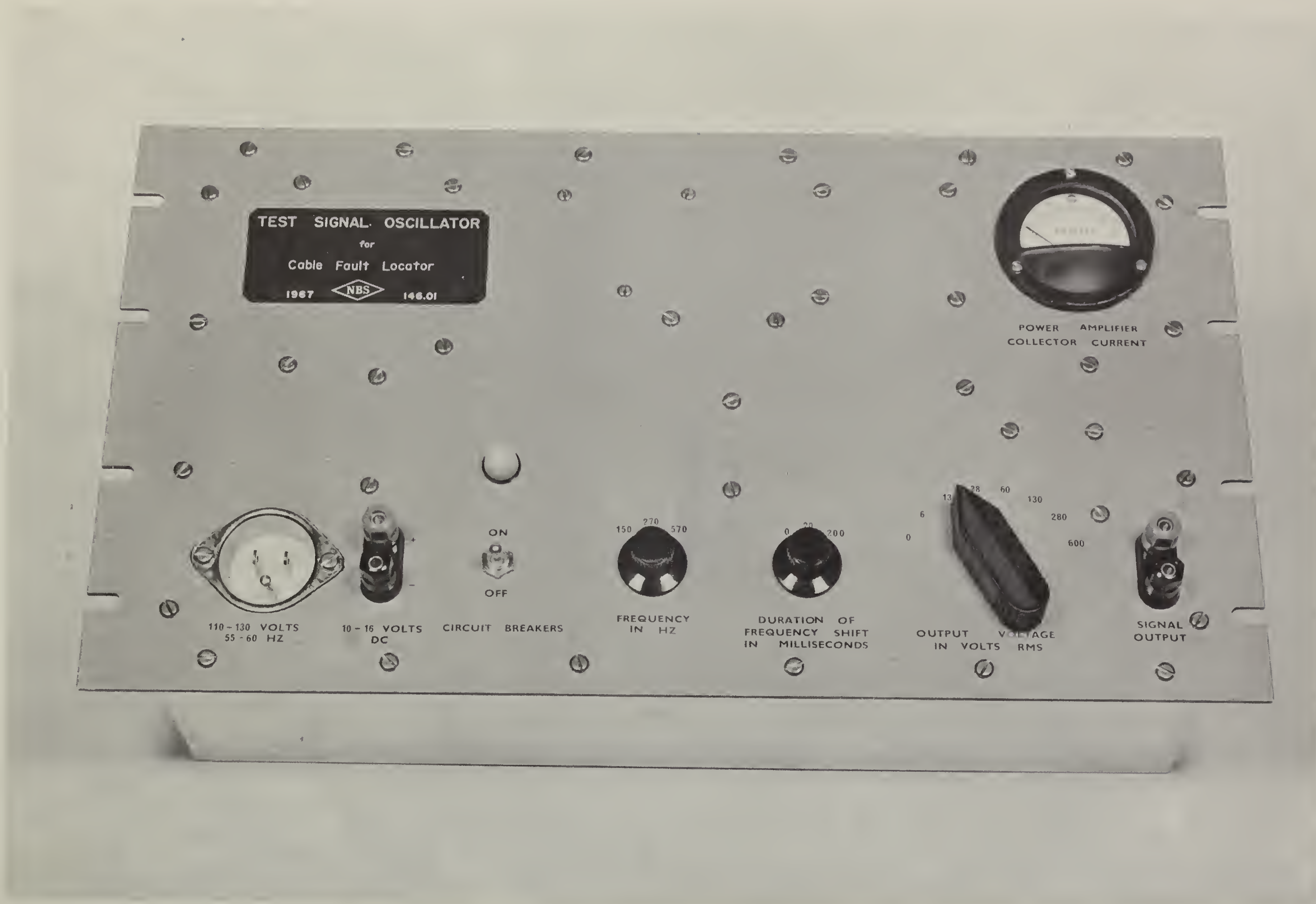


FIGURE 1-1. Photograph of the front panel of the Test Signal Oscillator.

FIGURE 1-2. Photograph of the interior of the Test Signal Oscillator.


SIGNAL STRENGTH
BATTERY TEST
OFF
BATTERY TEST
GAIN
MAX
MIN
DETECTOR UNIT
for
Cable Fault
Locator

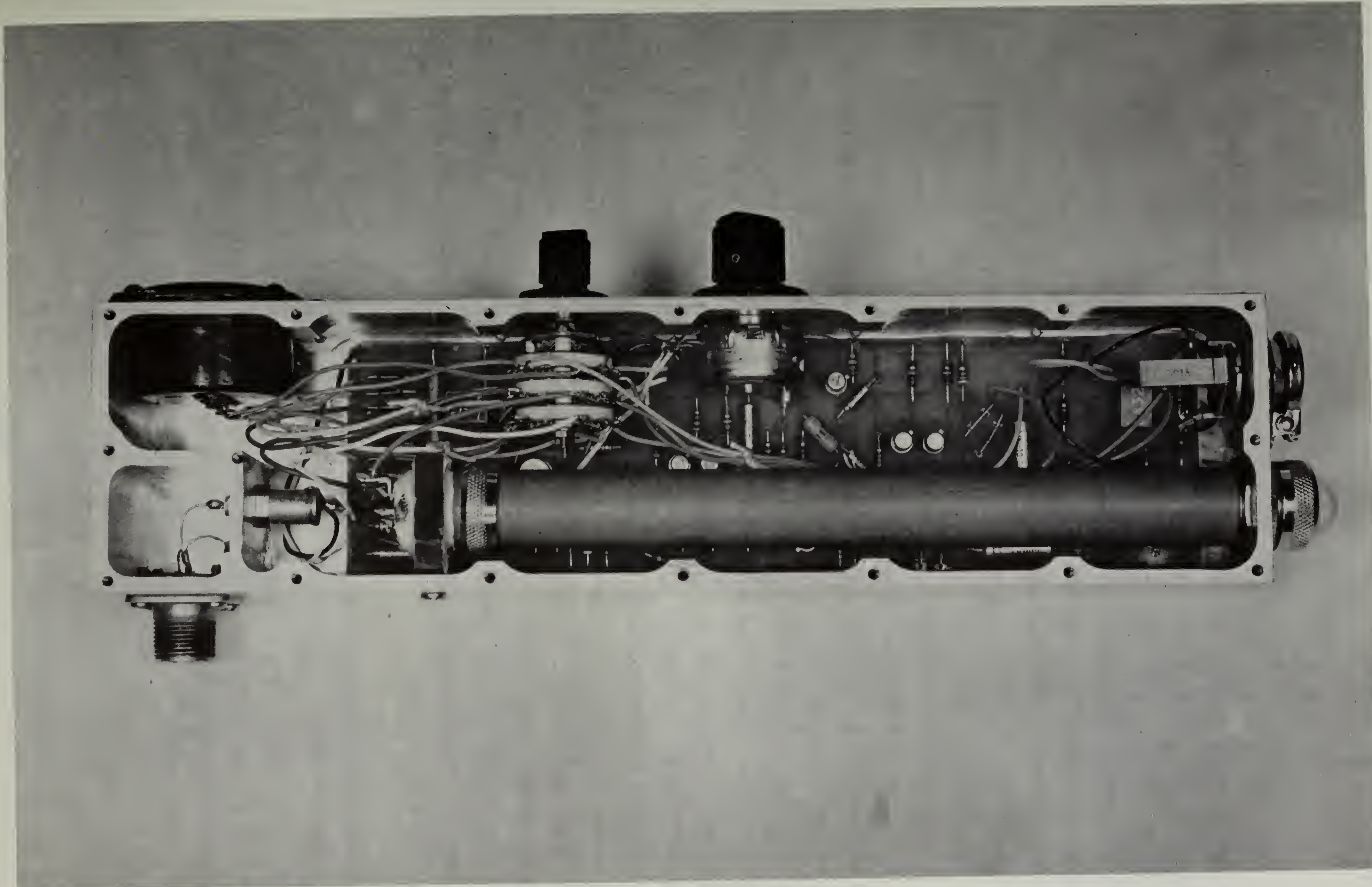


FIGURE 1-4. Photograph of the printed circuit board and interior of the Detector Unit.

INTRODUCTION

The Cable Fault Locator described in this report is a two unit test set whose principal use is to locate the position of faults in single-conductor buried electric cable. The two units which comprise the Cable Fault Locator (CFL) are the Test Signal Oscillator (TSO) and the Detector Unit (DU). Photographs of these units are shown in Figures 1-1, 1-2, 1-3, and 1-4. Figure 1-5 illustrates the technique used to locate the position of a cable fault, in this case a short to ground. The cable under test is disconnected from its normal power source and connected to the TSO instead. The TSO supplies a test current of known frequency to the cable. The DU senses the magnetic field produced by the test current and provides visual (meter) and audible (headphones) outputs simultaneously. The DU is carried along the cable route and indicates a drop in signal strength as it passes beyond the fault. Since the DU senses the magnetic field produced by the test current, any situation which reduces the magnetic field intensity available to the DU reduces the effectiveness of the test procedure. For example, if the cable is in a metal conduit, has metal armor, or has its return circuit nearby, then the magnetic field intensity is reduced. Under favorable conditions the CFL is expected to locate a cable fault to ground to within two feet of the fault, while an open-circuit type of fault would be located to within approximately twenty feet of the fault. A comprehensive guide to the use of the CFL is: "Guide to Use of the AN/TSM-11 Cable Test-Detecting Set".[1] The techniques of fault location which were developed for the AN/TSM-11 Test-Detecting Set are valid also for the CFL.

The CFL is a prototype for an improved version of the AN/TSM-11 Cable Test-Detecting Set which was built in 1955 by Texas Instruments Incorporated. Most of the improvements incorporated into the new design were made either at the suggestion of those who had used the earlier units in the field, or because improved electronic components and circuits have become available. Figure 1-6 lists some of the principal improvements provided by the CFL.

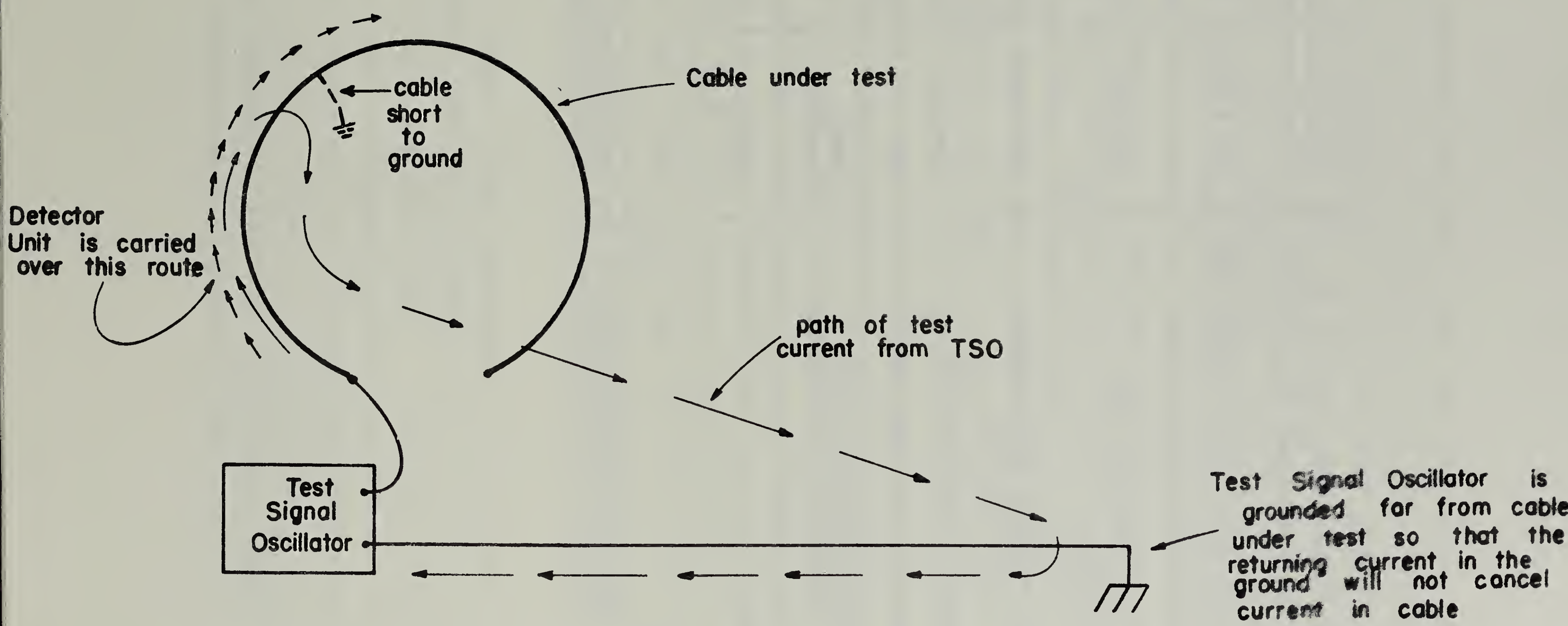


FIGURE 1-5. How Cable Fault Locator test set is used to locate position of a short to ground.

Item	AN/TSM-11	CFL
Test Signal Oscillator power output, max	6 watts	approximately 11 watt
Test Signal Oscillator voltage output, max	300 V rms	600 V rms
Test Signal Oscillator output metered	no	yes
Detector Unit voltage gain, max	200,000	250,000
"Q" of Detector Unit	6	12.7
Detector Unit response, meter reading versus input signal amplitude	non-linear	linear
Battery type used in Detector Unit	special	1.5V penlite cells
Detector Unit battery voltage metered	no	yes
Incorrect battery polarity applied to Test Signal Oscillator or to Detector Unit	damage possible	no damage
Test frequencies	250 Hz	150, 270, 570 Hz

Figure 1-6. Electrical characteristics of the new Cable Fault Locator (CFL) compared to the characteristics of the old AN/TSM-11 test set.

The CFL can be operated at any one of three test frequencies, instead of just one frequency as was the case for the old test set. All three frequencies are selected to be as far away as possible from 60 Hz power line harmonics. The three frequencies used are 150, 270, and 570 Hz. The test frequency desired is selected by a front panel switch on the TSO. The DU is tuned to the same frequency by a similar switch on its front panel. The 150 Hz frequency was chosen as the lowest frequency likely to be useful. A low testing frequency has the advantage that there is relatively little coupling of the test signal into cables which lie adjacent to, or cross, the cable under test. The disadvantages of the 150 Hz test signal are that the adjacent power frequency harmonics are relatively strong, the induced voltage in the magnetic pickup probe of the DU is weak because the rate of change of the magnetic field is low, and the human ear is not very sensitive to low-frequency signals (see Figure 1-7). The 570 Hz frequency is the most sensitive in all respects. The power-frequency harmonics are weaker near this frequency, the human ear is more sensitive, and more voltage is induced in the pickup probe for a given current in the cable under test. The 570 Hz frequency probably gives the best results in locating open-circuit faults because the higher frequency produces larger capacitive currents. The 270 Hz test frequency has characteristics which lie between those of the 150 Hz and 570 Hz frequencies, and is close to the frequency used in the old test set. The use of three frequencies was suggested by those who had used the old test set in the field, and the usefulness of multiple frequencies can be decided only by field testing of the new test set.

The CFL is able to employ either unmodulated or modulated test signals. The DURATION OF FREQUENCY SHIFT IN MILLISECONDS switch on the front panel of the TSO selects one of three modulation modes: 1) continuous uninterrupted test signal; 2) test signal interrupted for 20 ms approximately 130 times per minute; 3) test signal interrupted for 200 ms approximately 95 times per minute. The interrupted signals help the operator to distinguish the test signal from background noise and interference. The 20 ms interruption is expected to be preferable for most purposes because it enables the test signal to be identified

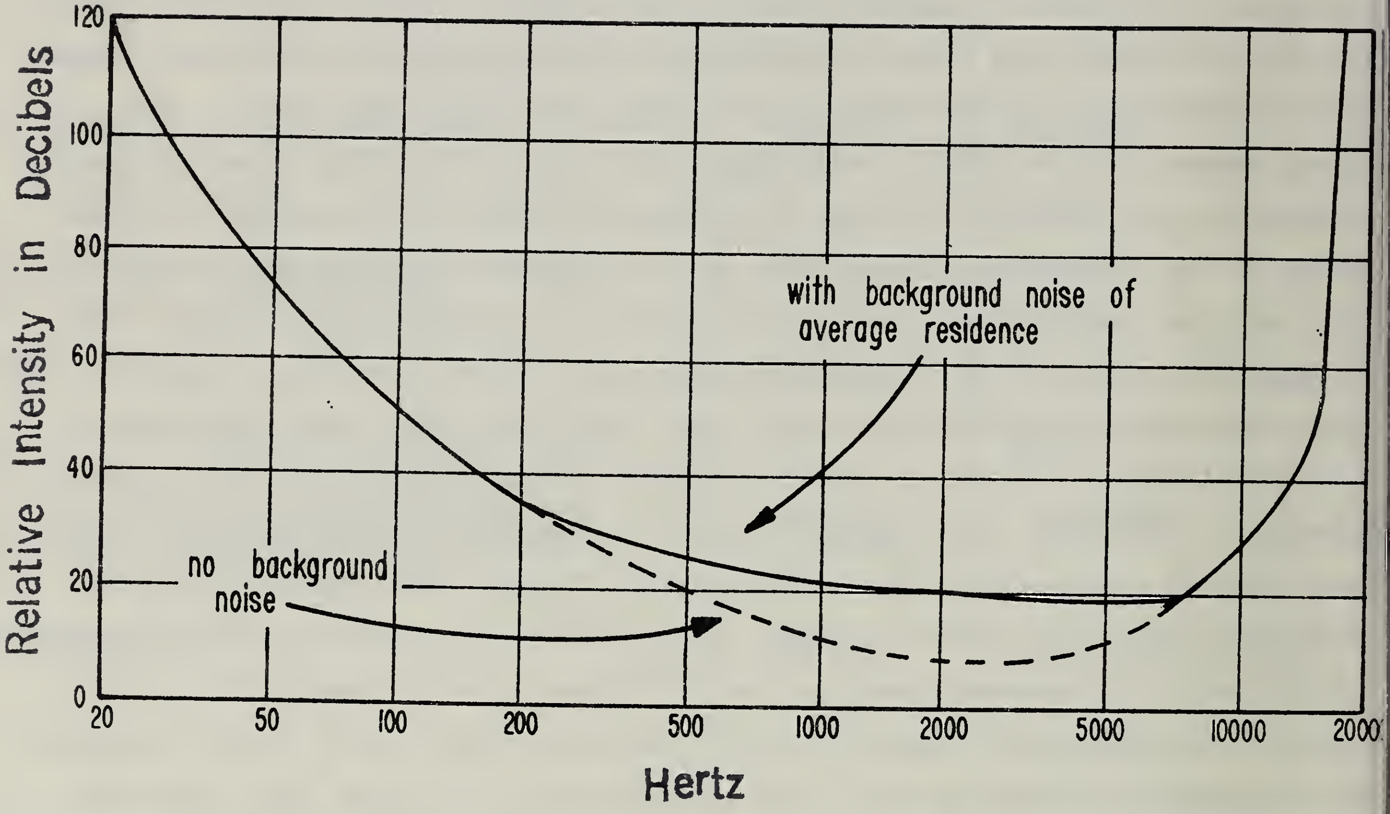


FIGURE 1-7. Threshold of hearing for average listener. The average listener can hear sounds whose intensity lies above the curve. The graph is based on Figure 14.4 of Reference 2.

by both ear and eye. The ear can hear the interruption and the eye can detect the resulting small dip in the meter reading. The dip in the meter reading is small enough so that the meter is still easily read. The ear is able to detect signals which are far too weak to be detected by the meter. When the signal is very weak or when the signal-to-noise ratio is low, using the 200 ms interruption should help the ear to detect and identify the test signal. The uninterrupted test signal is useful when measuring the output frequency of the TSO.

The CFL instruments are designed to be operated and stored at temperatures between -20°C and +60°C.

Section 2

TEST SIGNAL OSCILLATOR

The Test Signal Oscillator (TSO) is a power oscillator which supplies a square-wave output signal. The frequency, voltage, and modulation of the output signal are selected by front panel switches. Approximately 11 watts is delivered to matched loads. A multiple-tap output transformer matches the TSO to its load. Selection of the optimum tap is quickly accomplished by observing the indication of a front panel meter.

CONTROLS and CHARACTERISTICS

The TSO may be powered from either an ac source of 110-130 V rms at 55-65 Hz, or from 10-16 Vdc. The maximum current drawn from an ac power source is 0.4 A rms, or 1.9 A dc if a dc power source is used. Two input power connectors, one for ac and one for dc, are mounted on the front panel. The TSO is protected from incorrect polarity of the dc power source by a series diode. It is permissible to have both the ac and the dc power sources connected simultaneously. The pilot lamp is connected to an unregulated power supply in the TSO, so the lamp brightness can be used to estimate the voltage of the power source. The ground wire of the ac power input is the only connection made to the case of the TSO.

CIRCUIT BREAKERS	The ON-OFF switch is also the reset switch for three mechanically ganged but electrically independent circuit breakers. If any one of the breakers should trip, the other two will also open. The circuit breakers are reset by turning the switch to OFF and then back to ON.
FREQUENCY IN HZ	This switch selects the frequency of the test signal at the SIGNAL OUTPUT connector. The test frequencies are 150, 270, and 570 Hz. Frequency tolerance is ±2%.
DURATION OF FREQUENCY SHIFT IN MILLISECONDS	This switch selects one of three test modes: continuous single-frequency signal; test signal interrupted for 20 ms approximately 130 times per minute; or test signal interrupt for 200 ms approximately 95 times per minute. The interrupti is accomplished by momentarily increasing the test signal frequency to roughly 1.7 times its nominal value.
OUTPUT VOLTAGE IN VOLTS RMS	This switch determines the nominal open-circuit voltage at the SIGNAL OUTPUT terminals by selecting the appropriate tap on output transformer T2. The nominal voltages are 0, 6, 13, 28, 60, 130, 280, and 600 volts rms, when the power source is either 130 volts ac or 15 volts dc. Maximum power into the load is achieved by selecting the switch position which provides the largest on-scale reading of the POWER AMPLIFIER COLLECTOR CURRENT meter.
SIGNAL OUTPUT	The test signal supplied by these terminals is a square wave whose frequency, modulation, and amplitude are determined by the three controls described immediately above. Any passive load may be connected to these terminals. The TSO is protect from short-circuiting of these terminals. However, the TSO may be damaged if more than six volts ac or dc is developed across these terminals by an external source. The black terminal is the circuit ground and is isolated from the instrument case.
POWER AMPLIFIER COLLECTOR CURRENT	This ammeter measures the total dc current drawn by the Powe Amplifier stage (Q17-Q20). The Power Amplifier is operating within its design rating as long as this meter reads on-scal An off-scale reading does not endanger the Power Amplifier bu may cause the circuit breakers to trip.

DESIGN CONSIDERATIONS

To the greatest extent possible the circuits selected for use in the TSO use standard components, require only one circuit adjustment, and yield easily to analysis. The transformers and filter chokes are the only components which were custom built for the TSO. All other electrical components are standard catalog items. The benefits derived by employing circuits amenable to analysis are twofold. First, component values can be selected by calculation rather than trial and error; and second, comparison of calculated performance with measurements made on the finished circuit provides an excellent method for detecting both design and construction errors.

Because the TSO will normally be used at military airfields, an attempt was made to design the circuits with components which are readily available from military supply sources. To this end, all semiconductor components are chosen from the "Military Standard Preferred and Guidance Lists of Semiconductor Devices", MIL-STD-701E, 8 December 1965. Restricting the semiconductor complement to this list did not cause any serious circuit design problems, although it did increase the cost of the transistors and require using four power transistors in the Power Amplifier stage where two non-military transistors would perform as well. It was not possible to obtain delivery in a reasonable time for the military quality semiconductors, so ordinary commercial units with the same device number are employed in their place.

The packaging selected for the TSO is simple but sturdy. The layout of both the chassis and the printed circuit board is roomy and all components are reasonably accessible. The front panel is water tight. The electrical connectors and the panel meter are sealed to the panel with silicone adhesive and probably cannot be removed without destroying this seal. The pilot light, switches, and machine screws are sealed by gaskets and can be easily removed and re-used if required. On the other hand, none of the joints or seams of the protective aluminum case have been made water tight.

CIRCUIT DESCRIPTION

A block diagram of the Test Signal Oscillator (TSO) is shown in Figure 2-1. The Power Supply section provides unregulated dc voltage to the Power Amplifier and regulated 6.8 V dc to the rest of the instrument. The Power Supply can be energized by either a 110-130 V ac line or by a 10-16 volt battery or dc generator. The Oscillator section operates at twice the frequency appearing at the instrument's SIGNAL OUTPUT terminals. The Oscillator feeds the Frequency Divider which divides the oscillator frequency by two. The Frequency Divider feeds the Driver stage which in turn supplies the required base current drive to the Power Amplifier. The Power Amplifier is a push-pull class B amplifier and uses an output transformer with multiple secondary taps to provide impedance matching to the load. The Frequency Shift Multivibrator is a free-running multivibrator which periodically turns on transistor Q7, causing the frequency of the Oscillator to increase.

Power Supply (Q1, Q2, Q3):

The Power Supply circuitry (shown in Figure 2-3) consists of a center-tapped full-wave rectifier, a choke input LC filter, and a voltage regulator. Three magnetic circuit breakers, ganged to each other and to the power (ON-OFF) switch, provide overload protection. Voltage for the pilot lamp and the Power Amplifier is taken directly from the output of the LC filter while all the remaining circuits are powered by the 6.8 volts supplied by the voltage regulator.

The voltage regulator is composed of transistors Q1, Q2, and Q3. In order to enable the regulator circuit to function when the voltage drop across pass transistor Q3 is low, constant-current generator Q1, rather than a resistor, is used to provide the base drive to Q3. C10, shown in Figure 2-4, supplies transient switching current to the Driver stage (Q13, Q16). The voltage regulator has been observed to oscillate under certain load conditions when C10 was between 0.02 and 2.0 μF.

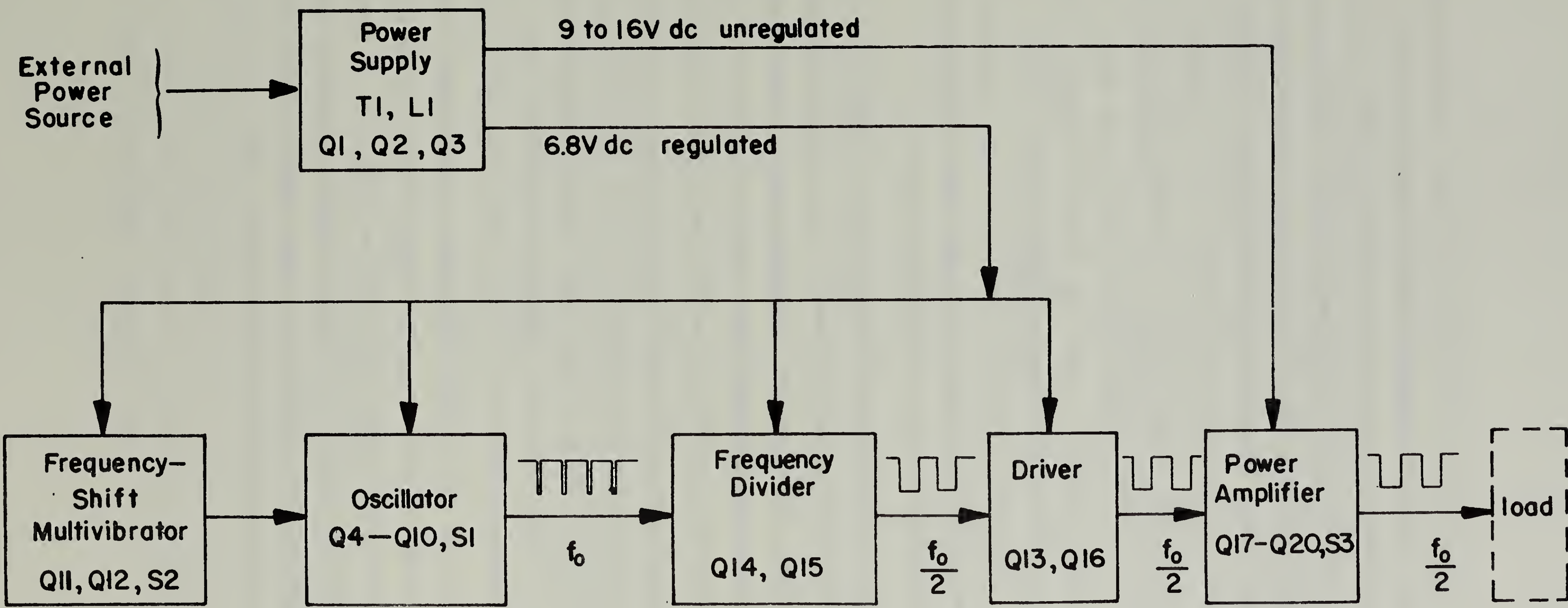


Figure 2-1. Block diagram of Test Signal Oscillator.

The Power Supply may be energized by either 110-130 volts at 55-65 Hz, or by 10-16 volts dc. Although the TSO requires that the ac and the dc power sources be connected to the proper input terminals, it is permissible to have both power sources connected simultaneously. When both ac and dc power are applied, the TSO draws power from the source providing the higher voltage at the collector of Q3. When the ac supply happens to develop the higher voltage, CR4 prevents current from entering the dc source. If polarity is accidently reversed when connecting to a dc power source, the blocking action of CR4 will prevent damage to the TSO.

When a dc power source is used, the initial charging current drawn by C1 might trip the 2.0 ampere breaker if CR3 were not present to prevent the current surge. Diode CR3 is shunted by R4 in order to maintain voltage on C1 when a dc power source is used. Aluminum electrolytic capacitors last longer when stored with a bias voltage than without it.

The thermistor in series with the primary of the power transformer is required to prevent the 0.4 ampere breaker from being tripped by the turn-on current transient. Without the thermistor the transformer would saturate and draw a large current transient when the power switch is turned on, unless the switch happened to be turned on near a peak of the ac line voltage waveform

It is believed that the TSO can safely withstand the application of power voltages up to twice the maximum rated values, although no tests were conducted to verify this belief. Whenever the input supply voltage is high enough to make the voltage at the collector of Q3 rise above 18 volts, the zener diodes (CR14, CR15) in the Power Amplifier circuit will conduct. If the excessive voltage is supplied by a dc source, then the current drawn by the zener diodes will trip the 1.6 ampere breaker. When the contacts of the 1.6 ampere breaker open, CR16 provides a conducting path for the inductive current in L2 and T2, thereby protecting the contacts of the 1.6 ampere breaker. If the excessive voltage is supplied by an ac source, then saturation of the iron core in T1, and/or the load imposed by zener diodes CR14 and CR15 will trip the 0.4 ampere breaker.

<u>Oscillator (Q4 - Q10)</u>:

A block diagram of the oscillator is shown in Figure 2-2. The oscillation cycle begins with C2 charging toward the 6.8 volt supply voltage. When the voltage across C2 reaches one-half the supply voltage, the one-shot is triggered via Q5 and Q8. The one-shot turns on reset transistor Q4 for 40 μs, discharging C2 to ground. When Q4 turns off, C2 again charges toward the supply voltage and the cycle repeats.

The oscillator frequency, f_0, is twice the frequency appearing at the output of the TSO and can be calculated from the equation

$$f_0 = \frac{10^6}{0.693RC + 40} \text{ Hz} \qquad (1)$$

where R = the value in ohms of R11, R12, or R13

C = the value of C2 in μF (=0.22 μF).

The quantity 0.693RC is the time interval required to charge C2 to one-half the supply voltage, and the number 40 is the time interval (in μs) used to discharge C2. The derivation of Equation 1 is based on the conditions that C2 charges to exactly one-half the supply voltage and is then discharged completely to ground. In the circuit actually used, however, tolerance errors in R18, R19, and C2; voltage offset between the bases of Q5a and Q5b; and failure of C2 to discharge fully to ground may cause the actual frequency generated to differ from the calculated value by several percent. The frequency is adjusted to the desired (calculated) value by placing a trimming resistor, R17, in parallel with either R18 or R19, as required (see Figure 2-4).

In order to ensure good frequency stability the RC timing circuit (C2; R11 or R12 or R13) is designed so that at least 99% of the current in the charging resistor is used to charge C2. The emitter junction of Q5a is reversed biased during most of the charging interval, so the transistor specified should have no more than 3 μA leakage at room temperature with 3.5 volts base-to-emitter bias. As much as 10 μA base drive may be required by Q5a in order to initiate the discharge cycle. But currents of this magnitude are drawn only during the last 50 mV of capacitor charging and thus do not significantly affect the overall charging time.

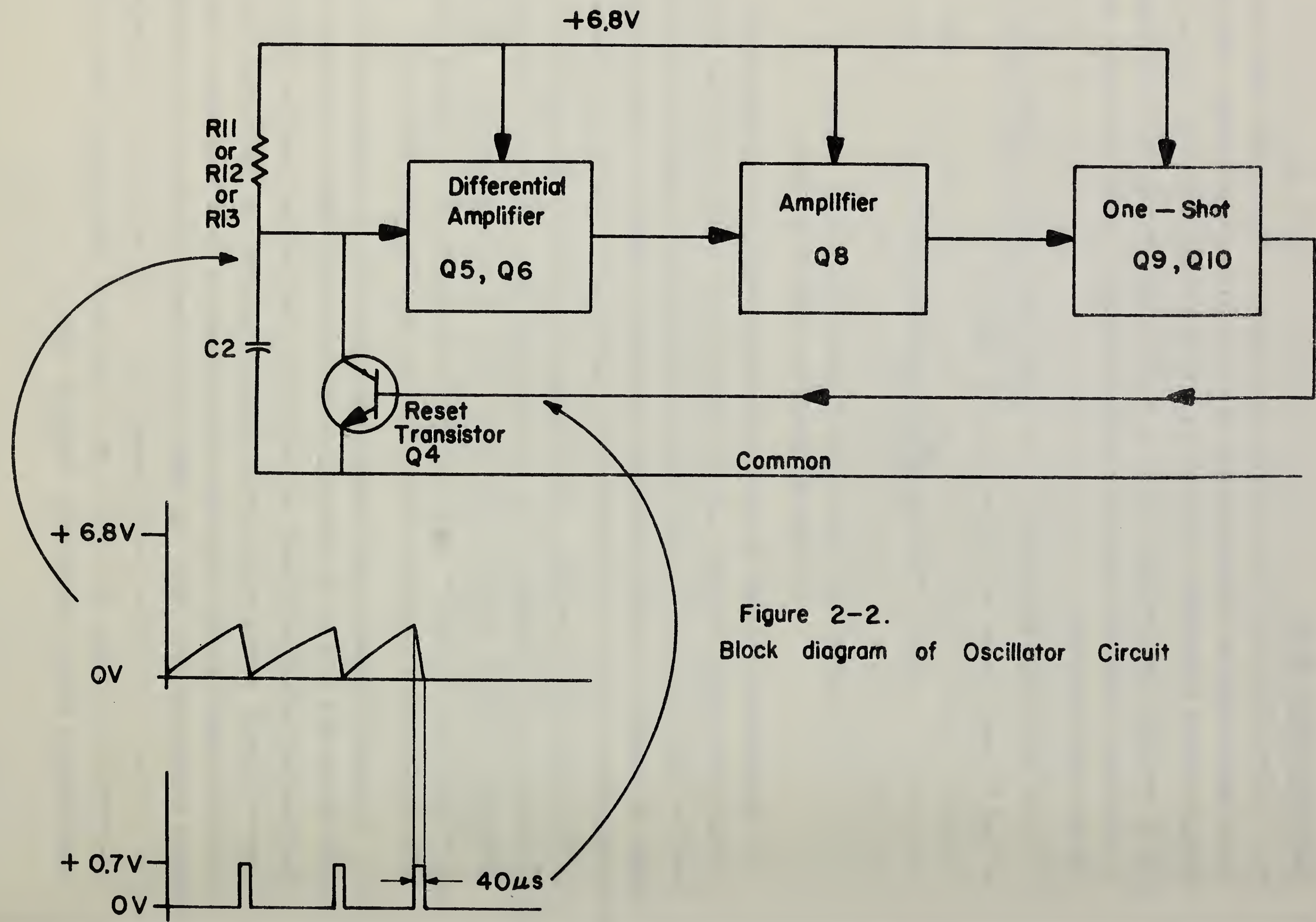


Figure 2-2.
Block diagram of Oscillator Circuit

A constant-current source, Q6, is used to supply emitter current to Q5 in order to make the oscillator circuit start at low Power Supply voltages. The oscillator frequency can be electronically shifted by turning on Q7, as is described in the section dealing with the Frequency Shift Multivibrator.

The one-shot circuit is used to drive the Frequency Divider circuit as well as transistor Q4. When the one-shot is in its quiescent state, Q9 is off and Q10 is conducting, thus keeping Q4 biased off. The one-shot switches state when Q9 is driven into conduction by Q8. The rapid fall of the Q9 collector voltage is used to trigger the Frequency Divider circuit, while the increasing Q10 collector voltage turns on Q4. The one-shot remains in its switched condition for 40 μs. The collector voltage of Q9 is limited to around 4.5 volts in order to prevent the emitter-base junction of Q10 from receiving excessive reverse voltage via C3.

Care was taken to ensure that the one-shot would trigger reliably, since even one failure to respond to a trigger signal from Q8 would prevent Q4 from discharging C2 and the oscillator would hang up. If poorly designed, the one-shot could fail to respond to a slowly rising trigger from Q8 because the presence of C3 in its regenerative feedback loop would block the feedback signal. In the circuit used here regeneration is assured for any trigger signal by making the high-frequency loop gain of the one-shot large enough to ensure spontaneous regeneration when Q9 is brought into conduction. Shunting feedback resistor R25 with a capacitor, and not driving Q10 too far into saturation in its quiescent state assure the necessary high-frequency loop gain.

Frequency-Shift Multivibrator (Q11, Q12):

Either a continuous or an interrupted output signal is available from the TSO. The interrupted signal is obtained by momentarily shifting the signal to a higher frequency. It is the function of the Frequency-Shift Multivibrator to establish the duration and repetition rate of the interruption. The frequency shift is accomplished by permitting the Frequency-Shift Multivibrator to free-run and periodically drive Q7 into conduction (see Figure 2-4). When Q7 conducts, the base voltage of Q5b drops and causes the Oscillator frequency to increase. A single-frequency, uninterrupted output signal is obtained by disabling the multivibrator, causing Q7 to remain off.

The circuitry of the Frequency-Shift Multivibrator is very similar to that of an ordinary free-running multivibrator. When switch S2 is in the 0 ms position the regenerative feedback path of the multivibrator is broken and both transistors conduct indefinitely. When conducting, the collector of Q11 is only 0.5 volts above ground and the series chain of CR10, R28, R29 assures that Q7 is biased off. Tantalum capacitor C5 is reverse biased when the multivibrator is in this inactive state. CR11 is required to limit the magnitude of the reverse bias voltage to a few tenths of a volt so that C5 will not be damaged. If CR11 were omitted, the bases of both transistors would be approximately 0.7 volts above ground and each collector would be approximately 0.1 volts above ground, causing a net reverse voltage of 0.6 volts to appear across C5. The effect of adding CR11 to the circuit is to prevent the collector of Q11 from dropping more than about 0.2 volts below the Q11 base, which means the reverse voltage on C5 is also limited to around 0.2 volts. If the Q11 collector voltage tries to drop below the Q11 base voltage, CR11 will conduct and drain off the excess base drive current. The solid tantalum capacitor used for C5 may be able to stand a 0.7 volt reverse bias without ill effect, but an old tantalum capacitor of unknown construction used in the breadboard circuit developed excessive leakage current with 0.7 volts reverse bias.

When switch S2 is in the 20 ms or the 200 ms position the regenerative feedback path is no longer broken and the multivibrator spontaneously begins to free-run. Reliable starting is assured because the loop gain of the multivibrator is greater than unity when both Q11 and Q12 are on. Diode CR11 keeps Q11 out of saturation, ensuring that this stage functions as an amplifier with reasonable gain. Biasing Q12 so that the ratio I_C/I_B is reasonably high ensures that Q12 is not so badly saturated that the gain of the Q11 stage is lost by attenuation in Q12.

In order to describe the operation of the free-running multivibrator, let us assume that Q11 is initially on, Q12 is off. Q11 is held on by the base current supplied by R32, and Q12 is held off by a reverse base voltage provided by C5. While the transistors are in this stage C5 charges slowly toward the supply voltage through R31. When C5 has charged sufficiently to bring the base voltage of Q12 to around 0.5 volts above ground, Q12 begins to conduct and initiate a regenerative cycle which quickly drives the transistor pair into their other stable state. The regenerative cycle proceeds as follows: As Q12 is brought into conduction its collector voltage drops. The drop in collector voltage is coupled via C6 (or C7) to the base of Q11, causing Q11 to begin to turn off. As Q11 turns off, the current in collector resistor R30 is diverted into C5 and the base of Q12. This current turns Q12 on more strongly, reinforcing the initial turn-on current and causing further regeneration. After the transistors are firmly planted in their new state, current continues to flow in R30 until C5 is fully charged, and C6 (or C7) gradually charges towards the supply voltage to initiate another regenerative cycle.

Ideally, the off transistor will remain off until its timing capacitor has charged sufficiently to bring it back into conduction, at which point regenerative feedback quickly flips the circuit into its other state. The oscillogram of the Q12 base voltage displays the ideal situation (see Figure 2-10). Note that regeneration occurs when the base voltage has risen to about +0.5 volts, which is the voltage required to drive the transistor into conduction. On the other hand the oscillogram of the Q11 base voltage (Figure 2-9) shows that the transition takes place when the base voltage has risen to only +0.25 volts, too small a voltage to turn on Q11.

The premature transition is caused by negative-going transients on the 6.8 volt supply line. These transients are generated by the switching of the Driver circuit (Q13 and Q16) and are conducted to the base of Q12 via R30 and C5. When the negative-going transient reaches the base of Q12 it momentarily turns off Q12 and thereby initiates the transition. Q12 is sensitive to transients because the path (C5, R30, R31) from its base to the power line has a much lower impedance for ac signals than for dc signals. Accurate timing is not required of the Frequency-Shift Multivibrator so this premature transition causes no real problem.

Frequency Divider (Q14, Q15):

The Frequency Divider circuit is a conventional binary divider whose output frequency is one-half its input frequency. The frequency of the divider output is the same as that appearing at the SIGNAL OUTPUT terminals of the TSO. The circuit triggers reliably when the leading (negative-going) edge of the input signal from Q9 has a slope of at least 0.8 V/μs and a frequency of less than 6000 Hz.

The Frequency Divider circuit switches state each time it receives a negative-going input pulse. In order to describe the switching process, let us assume that Q14 is initially conducting and that Q15 is off (see Figure 2-4). The collector voltage of Q15 is high, causing CR13 to be strongly reverse biased. The reverse-biased diode isolates the base of Q15 from input signals of either polarity. On the other hand the collector of Q14 is near ground potential, causing CR12 to be slightly forward biased. CR12 blocks positive-going input signals, as was the case with CR13, but allows a negative-going input voltage to reach the base of Q14. When a negative-going voltage arrives at the junction of C8 and C9, CR12 is driven into conduction and Q14 turns off. As Q14 turns off, its collector voltage rises and turns on Q15 via R41. As Q15 turns on, its collector voltage drops and removes the base drive to Q14, so Q14 remains off after the trigger signal is gone. The transistors remain in their new state until the next trigger signal arrives. The next trigger signal must not arrive, however, until C8 and C9 have had time to at least partially charge toward their new steady-state voltages.

Driver circuit (Q13, Q16):

The Driver circuit (see Figure 2-4) provides a push-pull drive signal to the Power Amplifier. The driver transistors are used as switches and are driven by the Divider circuit. Normally, Q13 will be off when Q16 is on, and vice versa. But during the switching transition the transistors momentarily conduct simultaneously. Stored charge in the base region of the conducting transistor prevents it from turning off as promptly as the "off" transistor turns on, even though both transistors receive their input signal at the same instant. The additional power supply current drawn during the interval of simultaneous conduction is supplied by C10.

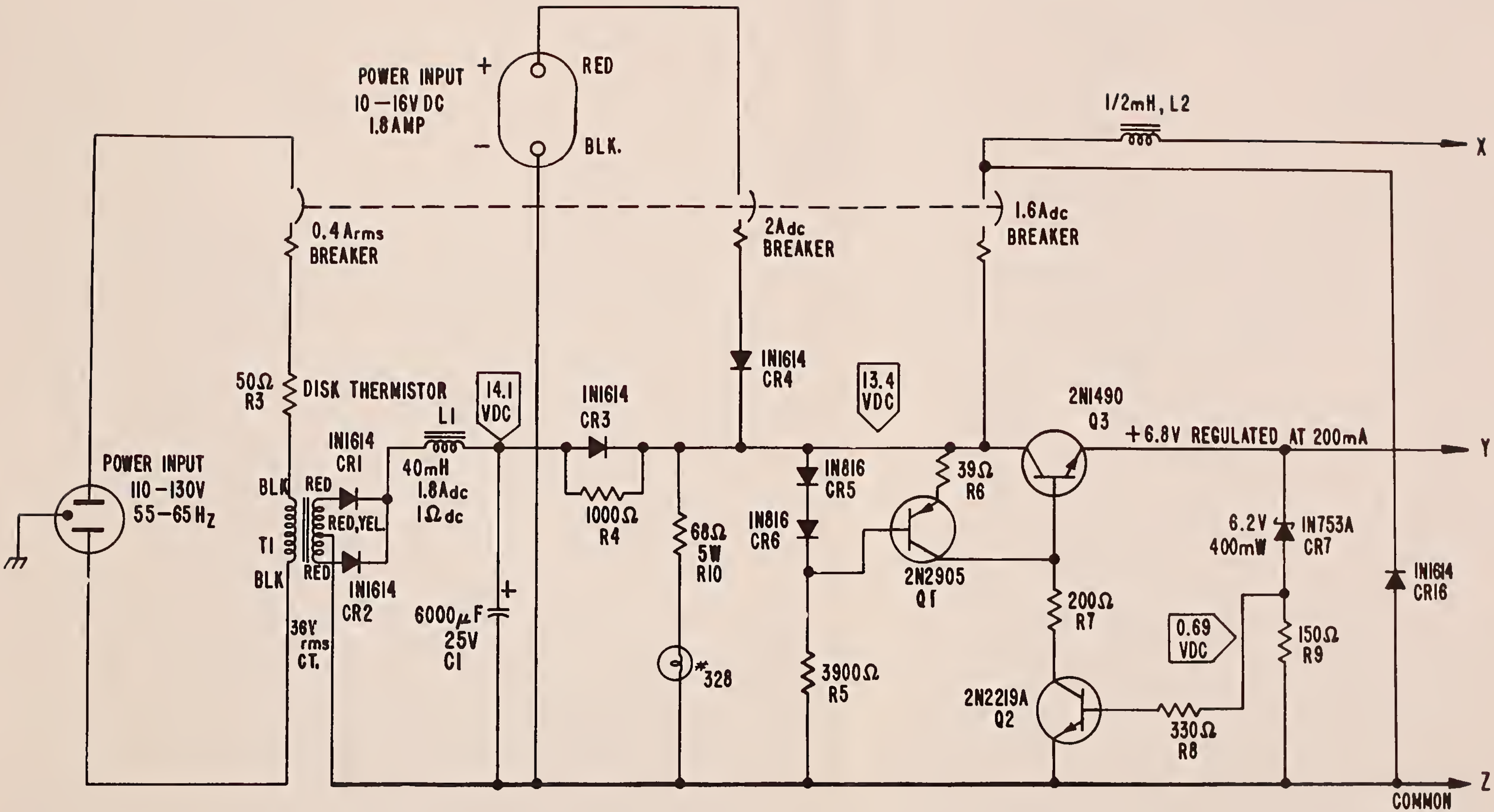
Power Amplifier (Q17, Q18, Q19, Q20):

The Power Amplifier is a push-pull class B square-wave amplifier which drives its load via multiple-tap output transformer T2. The circuit diagram for the Power Amplifier is shown in Figure 2-4. Optimum matching of the Power Amplifier to its load is obtained by selecting the tap (switch S3) which produces the largest on-scale reading of the 0-1.5 ampere meter. The amplifier is fully protected from overloads. If the dc current drawn by the amplifier significantly exceeds the design maximum value of 1.5 amperes for any reason, such as short-circuited SIGNAL OUTPUT terminals, the 1.6 ampere circuit breaker will trip. If the voltage of the source which powers the TSO is excessive, zener diodes CR14 and CR15 will conduct during the positive peak of the square wave and again cause the 1.6 ampere breaker to trip. The primary purpose of the zener diodes, however, is to clip any voltage spikes which may appear across the power transistors during normal operation. The four power transistors and the two 50 watt zener diodes are mounted on a heat sink which is large enough to ensure that the circuit breaker will trip before any of these components overheat.

The 2 ohm resistor, R51, in series with the secondary of the output transformer enables the Power Amplifier to drive a short-circuit load from the 6 volt transformer tap. When a higher voltage tap is used a short circuit will overload the amplifier. As mentioned in the preceding paragraph, short-circuit current drawn by the external load is expected to cause the dc current drawn by the Power Amplifier to become large enough to trip the 1.6 ampere circuit breaker. The 1.6 ampere circuit breaker is guaranteed to trip at 2.0 amperes, so the amplifier must be designed to draw at least 2.0 amperes under short-circuit conditions. Because rather large current and voltage drive to the base is required to ensure obtaining two amperes collector current in a single 2N1490 transistor, the power transistors are operated in pairs. The increased current gain and reduced base voltage drive obtained by dividing the load current markedly simplifies the Driver circuit. The power handled by the Power Amplifier is low enough so that only one transistor per side is required to handle the power requirements, hence it is unnecessary to force the load current to divide equally between the two paralleled transistors.

It is worth noting that the use of dc coupling in the Frequency Divider - Driver - Power Amplifier chain means that at any given time one side of the push-pull Power Amplifier is conducting. If for any reason the conducting state fails to alternate between the push-pull halves, the resulting hang-up will quickly cause the primary of T2 to saturate and short-circuit the Power Supply. Two details in the circuitry of the TSO prevent this characteristic from causing trouble. The first circuit detail is the 1.6 ampere circuit breaker in series with the primary of T2, which protects the TSO from damage if the Power Amplifier should hang up for any reason. Hang-up would result if the Oscillator circuitry should fail to oscillate, or from any other malfunction which disables the push-pull base drive to the Power Amplifier. The second circuit detail is concerned with obtaining reliable starting of the TSO when its power switch is first turned on. The potential trouble here is the possibility of the Power Supply voltage rising so slowly that the Power Amplifier will drive T2 into saturation before the Oscillator and the Frequency Divider circuits can start up and provide the required push-pull drive to the Power Amplifier. For this reason the TSO is designed so that higher Power Supply voltage is required to bring the Driver and Power Amplifier circuits into conduction than is required to bring the Oscillator and Frequency Divider circuits into operation, thus assuring satisfactory turn-on of the TSO. However, under normal circumstances the 6.8 volt supply line rises to full voltage in less than 0.02 seconds, which is fast enough to obtain reliable instrument turn-on regardless of the order in which the various circuits become operational. In other words, when operated from low voltages T2 cannot saturate in only 0.02 seconds. A switch with make-before-break contacts is used for S1 in order to prevent the Oscillator Circ from stopping during switching.

As was the case in the Driver circuit, stored charge in the base of the "on" transistors causes momentary simultaneous conduction of all four transistors in the Power Amplifier during the switching transition. This simultaneous conduction would momentarily short-circuit the Power Supply if inductor L2 were not present. High current pulses are undesirable because they may trip the 1.6 ampere circuit breaker and also make life more strenuous for the transistors.



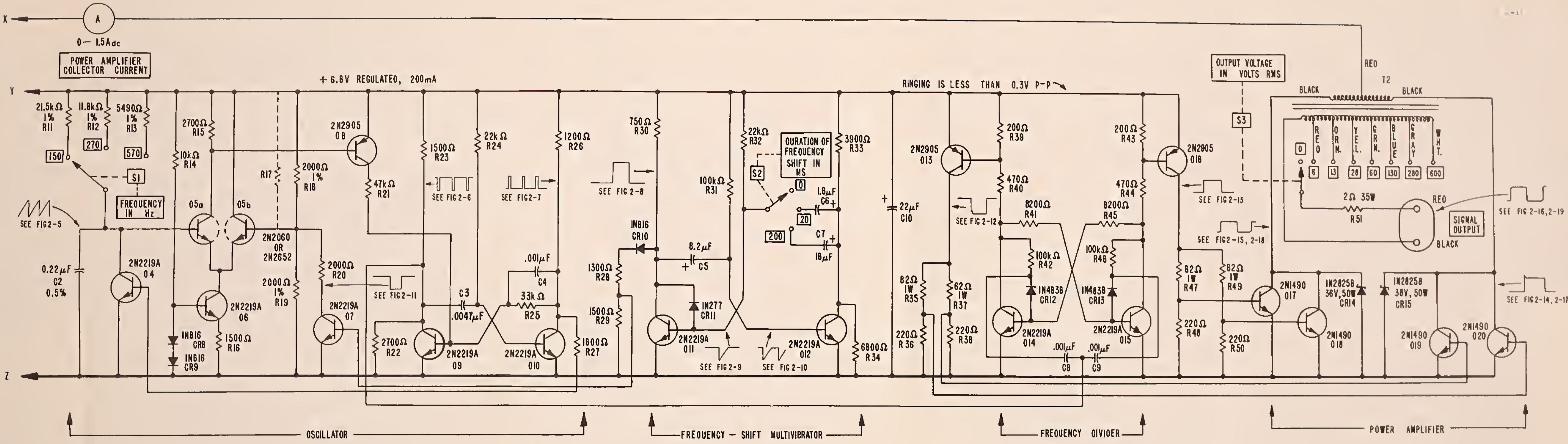
Notes:

1) The components which are mounted on the printed circuit board are CR5, CR6, CR7, Q1, Q2, R3, R5, R6, R7, R8, R9.

2) The dc voltages are measured with respect to the circuit common. The SIGNAL OUTPUT terminals are open-circuited and a 120V, 60 Hz power source is used.

FIGURE 2-3

POWER SUPPLY
OF
TEST SIGNAL OSCILLATOR



Notes:

1) Switches are shown in their extreme ccw position as viewed from the front panel.

2) Components which are not mounted on the printed-circuit board include CR14, CR15, Q17, Q18, Q19, Q20, and R51.

TEST SIGNAL OSCILLATOR
(SEE FIGURE 2-3 FOR POWER SUPPLY CIRCUIT)

FIGURE 2-4

DESCRIPTION OF CIRCUIT COMPONENTS

Test Signal Oscillator

Ammeter

The ammeter is a Honeywell model HS2X, 0-1.5 ampere dc meter. This is a 2-1/2" unit which is ruggedized and hermetically sealed.

Capacitors

C1 is an aluminum electrolytic capacitor, Mallory type CG63U25K1. All other polarized capacitors are tantalum electrolytic with voltage ratings of 15 or 20 V dc from the Sprague 150D series (Military CS13 series).

C2 is a polycarbonate capacitor, Component Research Company, Inc. part number 05PL224D. All other unpolarized capacitors are Sprague series 192P "Pacer" units having ±10% tolerance and voltage rating of either 80 or 200 Vdc.

Circuit Breaker Assembly

The circuit breaker assembly is manufactured by Airpax Electronics, Inc., Cambridge Division, Cambridge, Maryland. The assembly consists of three electrically independent breakers which are mechanically ganged to each other and to the toggle switch. The assembly is a member of the Airpax AP112 family of electromagnetic circuit protectors, with the three electrically independent poles specified as follows:

Pole 1	1.6 amperes dc	Delay 1
Pole 2	2.0 amperes dc	Delay 1
Pole 3	0.4 amperes 60 Hz	Delay 1

Resistors

Unless otherwise noted, resistors are ±5% tolerance carbon composition with 0.5 watt power rating.

Resistors which are specified by the circuit diagram to have ±1% tolerance also have temperature coefficients no greater than 50 ppm/°C. Commercial units which are suitable include Corning Electronic Components types NC4 and NC5, and IRC type MEA-T-2.

R3 is a disk thermistor, 50 ohms at 25°C, 0.4" diameter, Fenwal LB15J1.

R10 is a power resistor from Ohmite 995-5B series.

R51 is a power resistor from the Dale HG-25 series.

Switches

S1 is a shorting (make-before-break) rotary switch.

S2 may be either a shorting or a non-shorting rotary switch.

S3 is a heavy-duty rotary switch with non-shorting contacts, Centralab catalog number JV-9033.

Transformers and Chokes

All transformers and chokes were custom made to the following specifications:

L1 Filter choke, 40 mH at 1.8 A dc, dc resistance less than 1 ohm. Approximate body size is 2.7 x 1.9 x 2.3" high.

L2 Iron core choke, 0.5 mH at 1.5 A dc, dc resistance less than 0.2 ohms. Approximate body size is 1.4 x 1.1 x 1.2" high.

T1 Power transformer, primary to operate at 130 V rms and 55 Hz, secondary to supply 18 V rms on each side of center tap at 1.8 A rms. Approximate body size is 3.1 x 2.3 x 2.6" high.

T2 Push-pull output transformer for square waves between 150 Hz and 570 Hz. Secondary to have taps to provide open circuit voltages of 6, 13, 28, 60, 130, 280, 600 V rms when the power supply voltage at the center tap of the primary is 14 V dc. The voltage at each tap is permitted to drop 20% when loaded so that the primary draws 1.5 A dc. Good square-wave shape is not required. Approximate body size is 2.8 x 2.5 x 3.4" high.

The above items should be able to operate continuously at 75°C ambient.

Miscellaneous

The sockets used for power transistors Q3, Q17, Q18, Q19, Q20 are Augat Inc. part number 8038-1G1.

The sockets used for zener diodes CR14 and CR15 are Robinson Nugent part number NS-502-8.

WAVEFORMS

The oscillograms on the following four pages are the voltage waveforms at selected points in the TSO. The oscillograms supplement the preceding circuit descriptions and also should be helpful in trouble-shooting.

Test conditions which apply to all oscillograms are:

FREQUENCY IN HERTZ switch set at 570 Hz.

Power source is 120 volts at 60 Hz.

Vertical deflection labeling is in volts.

Notations used in describing the test conditions of the individual oscillograms are:

S2 . . . . refers to the setting of the DURATION OF FREQUENCY SHIFT IN MILLISECONDS switch.

S3 . . . . refers to the setting of the OUTPUT VOLTAGE IN VOLTS RMS switch.

LOAD . . . . means the load connected to the SIGNAL OUTPUT terminals. A 99 ohm load causes full-scale deflection of the POWER AMPLIFIER COLLECTOR CURRENT meter.

SYNC -- Q11 . . . means that the horizontal sweep of the oscilloscope is triggered by the negative-going transition at the collector of Q11.

SYNC -- Q13 . . . means that the horizontal sweep is triggered by the positive-going transition at the collector of Q13.

SWEEP. . . . . . . means the sweep speed of the oscilloscope horizontal deflection.

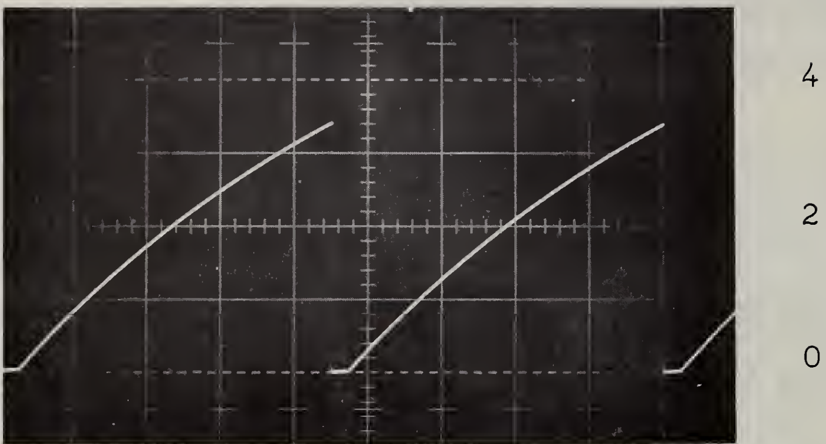


FIGURE 2-5

WAVEFORM at base of Q5a

S2 ---- 0 ms
S3 ---- any
LOAD -- any
SYNC -- Q13
SWEEP - 200μs/cm

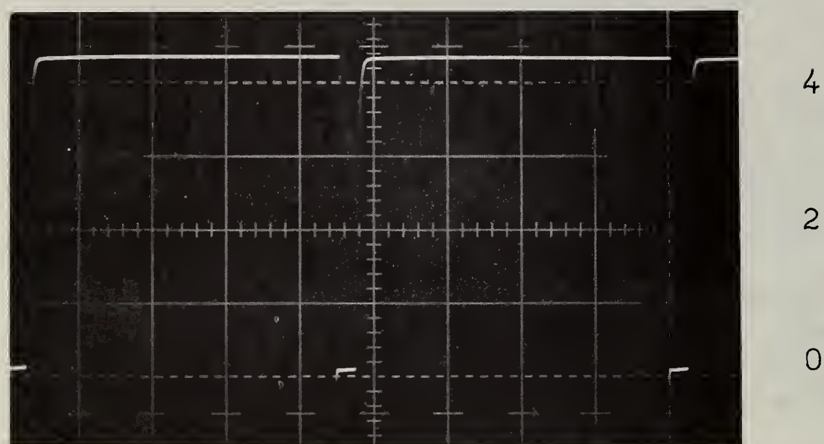


FIGURE 2-6

WAVEFORM at collector of Q9

S2 ---- 0 ms
S3 ---- any
LOAD -- any
SYNC -- Q13
SWEEP - 200μs/cm

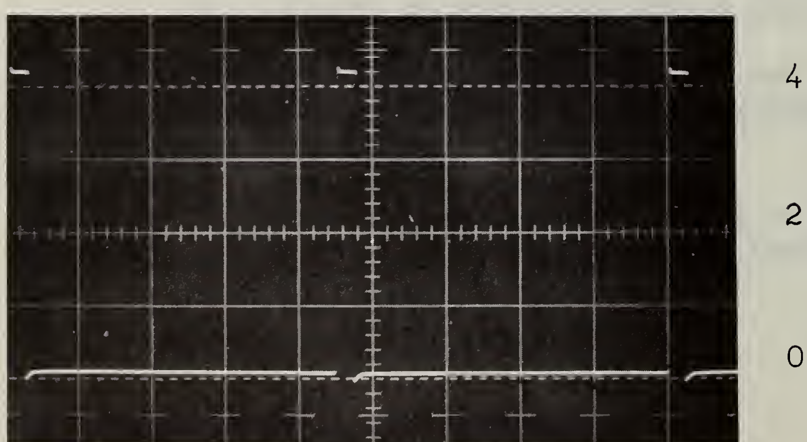


FIGURE 2-7

WAVEFORM at collector of Q10

S2 ---- 0 ms
S3 ---- any
LOAD -- any
SYNC -- Q13
SWEEP - 200μs/cm

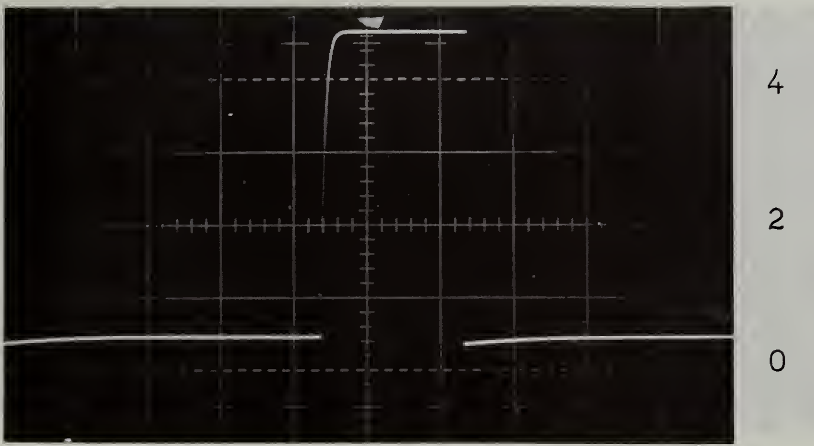


FIGURE 2-8

WAVEFORM at collector of Q11

S2 ---- 200 ms
S3 ---- any
LOAD -- any
SYNC -- Q11
SWEEP - 0.1s/cm

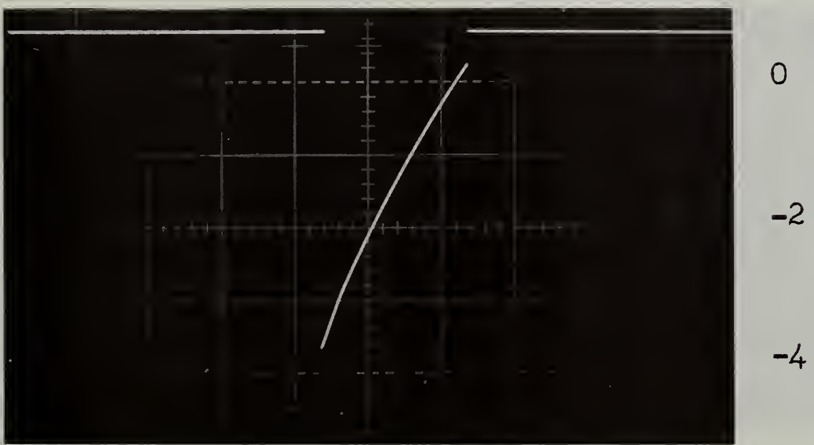


FIGURE 2-9

WAVEFORM at base of Q11

S2 ---- 200 ms
S3 ---- any
LOAD -- any
SYNC -- Q11
SWEEP - 0.1s/cm

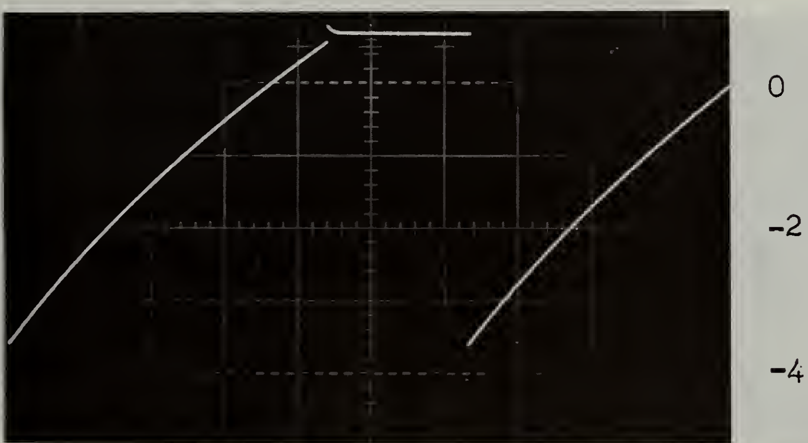


FIGURE 2-10

WAVEFORM at base of Q12

S2 ---- 200 ms
S3 ---- any
LOAD -- any
SYNC -- Q11
SWEEP - 0.1s/cm

FIGURE 2-11

WAVEFORM at collector of Q7

S2 ---- 200 ms
S3 ---- any
LOAD -- any
SYNC -- Q11
SWEEP - 0.1s/cm

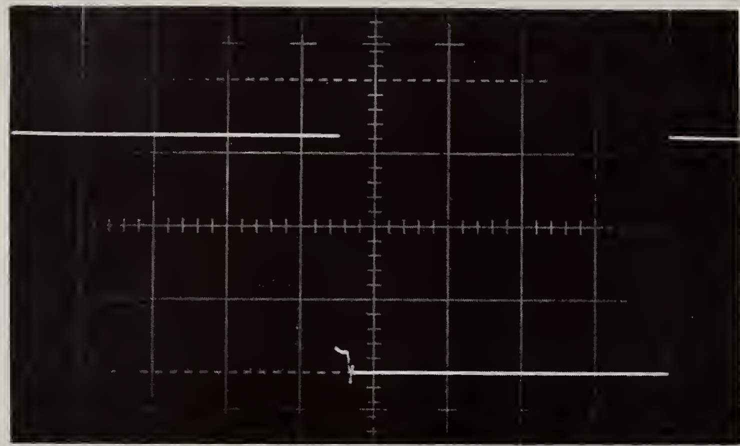


FIGURE 2-12

WAVEFORM at collector of Q13

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- not known
SYNC -- Q13
SWEEP - 200µs/cm

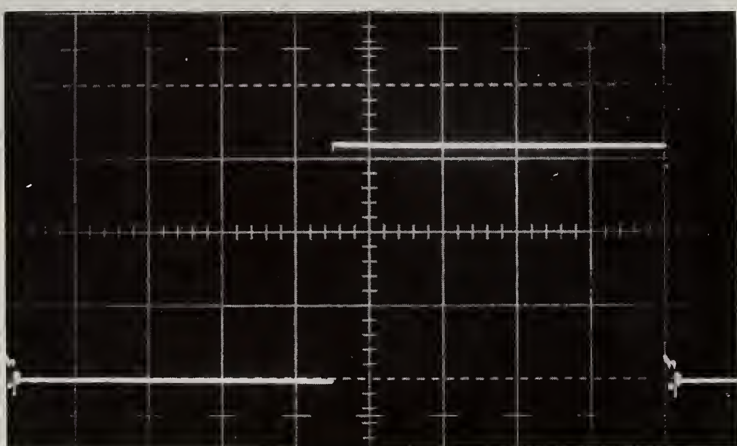


FIGURE 2-13

WAVEFORM at collector of Q16

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- not known
SYNC -- Q13
SWEEP - 200µs/cm

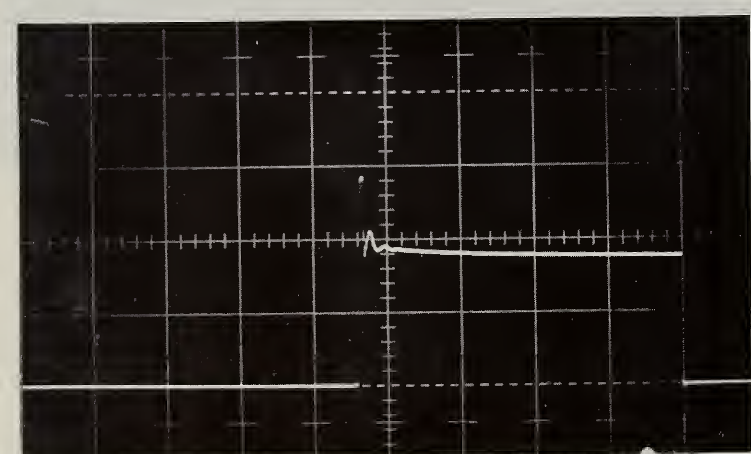


FIGURE 2-14

WAVEFORM at collector of Q20

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- 99 ohms
SYNC -- Q13
SWEEP - 200µs/cm

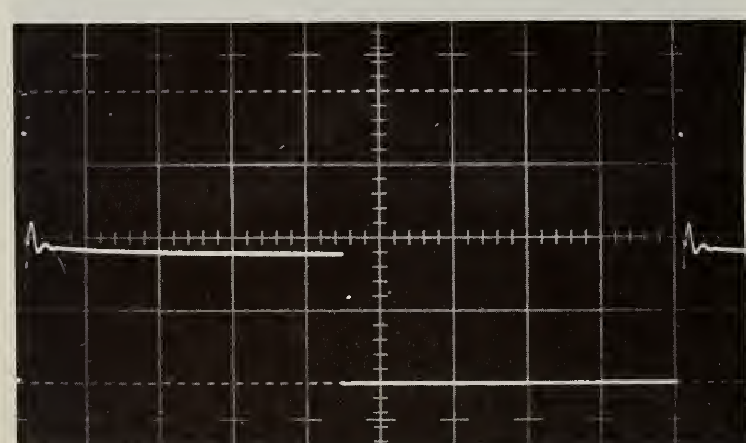
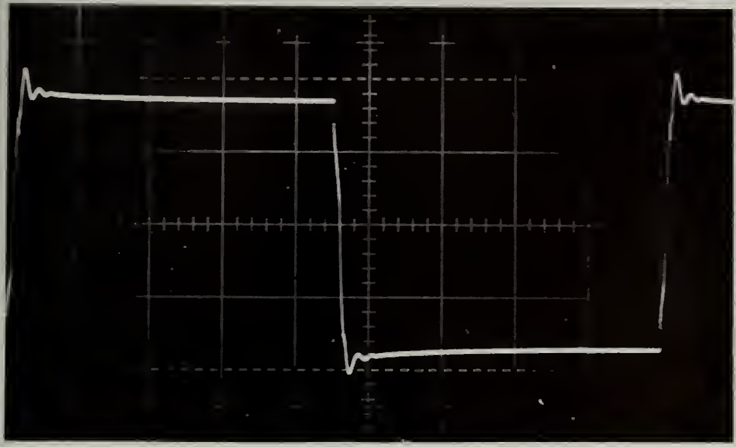


FIGURE 2-15

WAVEFORM at collector of Q17

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- 99 ohms
SYNC -- Q13
SWEEP - 200µs/cm



+40

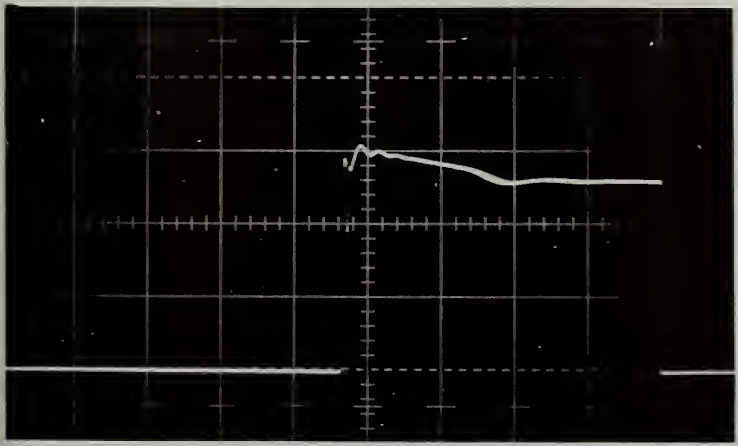
0

-40

FIGURE 2-16

WAVEFORM at SIGNAL OUTPUT terminal

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- 99 ohms
SYNC -- not known
SWEEP - 200µs/cm



40

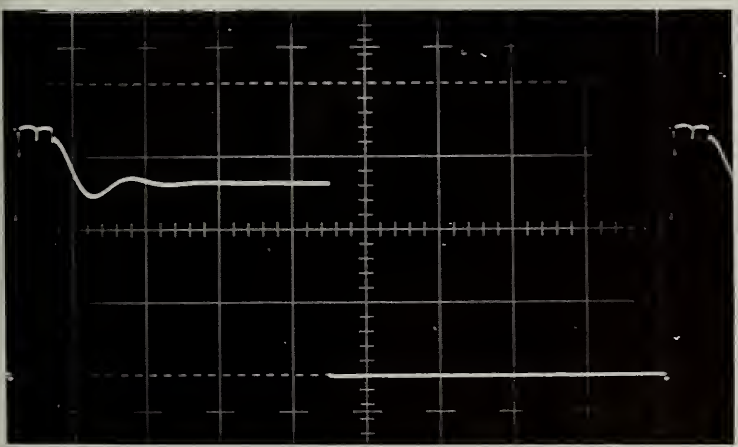
20

0

FIGURE 2-17

WAVEFORM at collector of Q20

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- open circuit
SYNC -- Q13
SWEEP - 200µs/cm



40

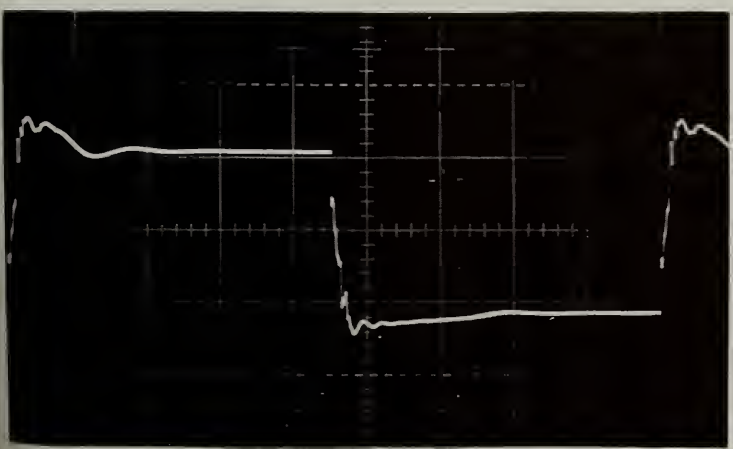
20

0

FIGURE 2-18

WAVEFORM at collector of Q17

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- open circuit
SYNC -- Q13
SWEEP - 200µs/cm



+100

0

-100

FIGURE 2-19

WAVEFORM at SIGNAL OUTPUT terminal

S2 ---- 0 ms
S3 ---- 60 volts
LOAD -- open circuit
SYNC -- not known
SWEEP - 200µs/cm

Section 3

DETECTOR UNIT

The Detector Unit (DU) is a hand-carried magnetic field monitor. Voltage induced in its pickup probe by fluctuating magnetic fields drives a tuned amplifier which in turn drives both an indicating meter and headphones. The amplifier passband and voltage gain are selected by front panel controls.

CONTROLS and CHARACTERISTICS

OFF	When the rotary switch is in this position the battery pack is disconnected from the circuit and the DU is inoperative.
150 Hz 270 Hz 570 Hz	These switch positions select the frequency to which the DU is tuned. The Meter indicates relative signal strength.
BATTERY TEST	This switch position connects the Meter to measure the voltage of the battery pack. The battery voltage is satisfactory when the meter reading is at least 70% of full scale (i.e. battery pack voltage at least 4.5 volts). Except for the change in the Meter function the DU circuits operate as usual as a 570 Hz tuned amplifier.
GAIN	This control varies the voltage gain from the PROBE input connector to the HEADPHONES jack from approximately 250 000 down to less than 600 when a 500 ohm load is used. For a given setting of the GAIN control the meter reading is proportional to the signal voltage induced in the Probe, provided the GAIN control pointer is within the wide, white circular arc marked on the panel. For lower gain settings the DU may saturate and cause the meter indication to be nonlinear.
Meter	The Meter indicates either relative signal strength (see GAIN control, above) or battery voltage, depending on the position of the rotary switch. The meter reading is not affected by loading of the HEADPHONES jack.
HEADPHONES	The signal available at this jack is the amplified Probe voltage. Four milliwatts is delivered to a 500 ohm load when the signal-strength meter reads full scale. 500 ohms is the preferred load impedance but any other load, including a short circuit, can be safetly driven from this jack. Output impedance is roughly 350 ohms.
BATTERIES	Access to the four penlite cells is obtained by unscrewing the threaded insert which is located at one end of the case. Any type of "AA" size cell (9/16" dia x 1-31/32" long) which supplies between 1.1 and 1.7 volts can be used. Carbon-zinc cells designed for use in transistor radios (Eveready #1015 or equivalent) are recommended for use at ambient temperatures between -15 and +40°C (+5 to +104°F).

DESIGN CONSIDERATIONS

The circuit design of the Detector Unit follows the same guidelines used for the Test Signal Oscillator. The DU requires no circuit adjustments, most of its circuits can be easily analyzed, and except for its pickup Probe standard components are used. Most of the semiconductor components are silicon, but germanium semiconductors are used where their smaller base-to-emitter voltage results in better circuit performance. All diode and transistor types used are chosen from the "Military Standard Preferred and Guidance Lists of Semiconductor Devices", MIL-STD-701E, 8 December 1965. Ordinary commercial grade semiconductors were installed in the DU if military grade units could not be obtained.

The mechanical design of the DU is intended to enhance accessibility of the circuit components rather than to obtain a very small instrument package. The frame of the DU is machined from a solid block of 6061-T6 aluminum alloy. Two removable panels provide access to the circuit components. Two edges of the printed circuit board rest in a slot milled in the frame. A machine screw near a third edge holds the board in place. Connecting wires are long enough to permit the instrument to operate with the printed circuit board removed from the case. The wiring layout on the circuit board is very similar to the schematic diagram layout and all components are labeled, hence trouble shooting is made easier. All components and mounting hardware on the front panel are watertight. The other fittings and joints are not necessarily watertight.

CIRCUIT DESCRIPTION

A block diagram of the Detector Unit is shown in Figure 3-1. Audio-frequency current in the cable under test produces a magnetic field which induces a small voltage in the Probe. This voltage is amplified by the Tuned Input Amplifier. The frequency response of the Tuned Input Amplifier approximates that of a single-tuned RLC circuit with a "Q" of 12.7. Since the voltage gain is small for undesired frequencies, saturation of the amplifier by strong but unwanted pickup from nearby power lines is unlikely. GAIN control R17 is used to attenuate the output signal from the Tuned Input Amplifier in order to prevent saturation of the Intermediate Amplifier which follows. The Intermediate Amplifier provides additional voltage gain and drives the Power Amplifier and the Operational Rectifier. The Power Amplifier drives the Headphones. The Operational Rectifier provides full-wave rectification of the ac signal and drives the Meter.

<u>Probe</u>:

The Probe is a coil of wire wound on a high-permeability rod and mounted at the end of a telescoping tube. At 1000 Hz the Probe impedance is equivalent to 5.7 mH in series with a 5 ohm resistor. The Probe assembly is the one originally used with the old AN/TSM-11 Test Set.

<u>Tuned Input Amplifier (Q1 - Q5)</u>:

A block diagram of the Tuned Input Amplifier is shown in Figure 3-2. The low-pass filter prevents radio-frequency signals from entering the amplifier. Input transformer T1 improves the signal-to-noise ratio by providing a degree of matching between the amplifier and the low-impedance Probe. The parallel-T feedback network produces the frequency-selective characteristic of the amplifier (see Figure 3-3).

The impedance seen looking into the PROBE connector is approximately 8 mH in series with 5 ohms resistance. This approximation is useful from dc to 1000 Hz. Radio-frequency signals which may be picked up by the Probe or its leads are prevented from entering the transistor circuitry by the low-pass filter. Because the amplifier gain is very low at radio frequencies,

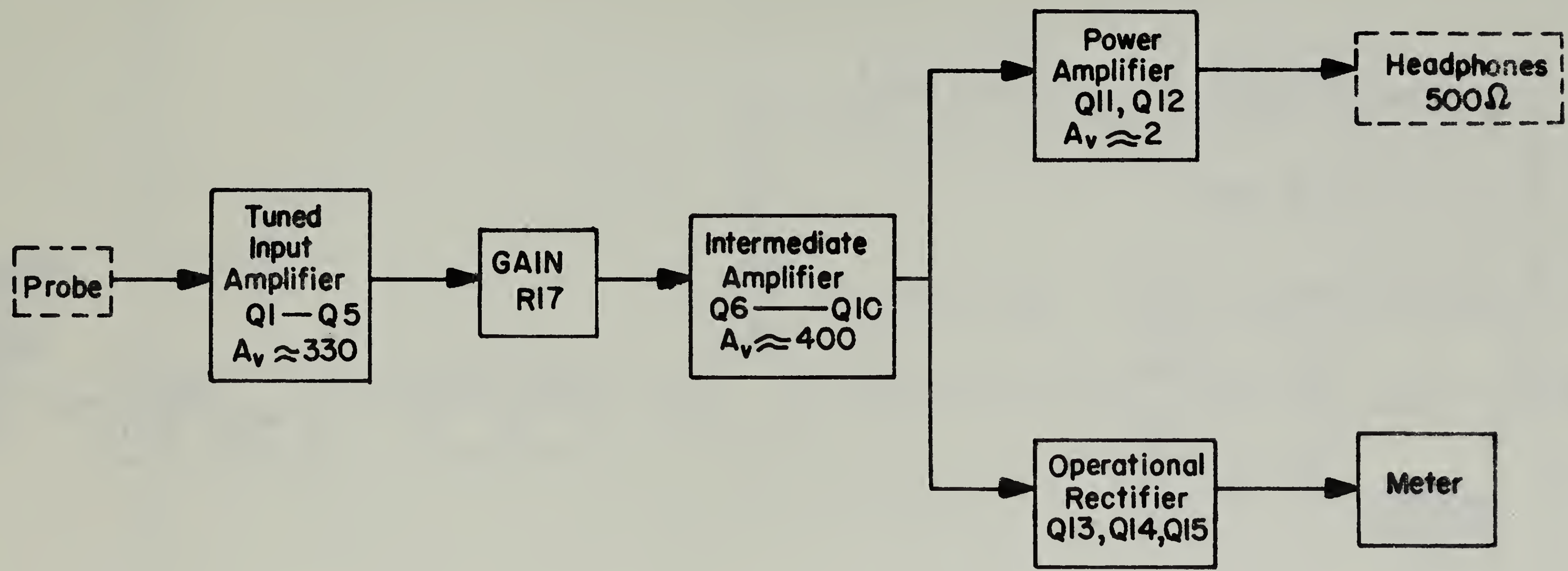


FIGURE 3-1. Block diagram of Detector Unit. Items within dotted boxes plug into the main chassis.

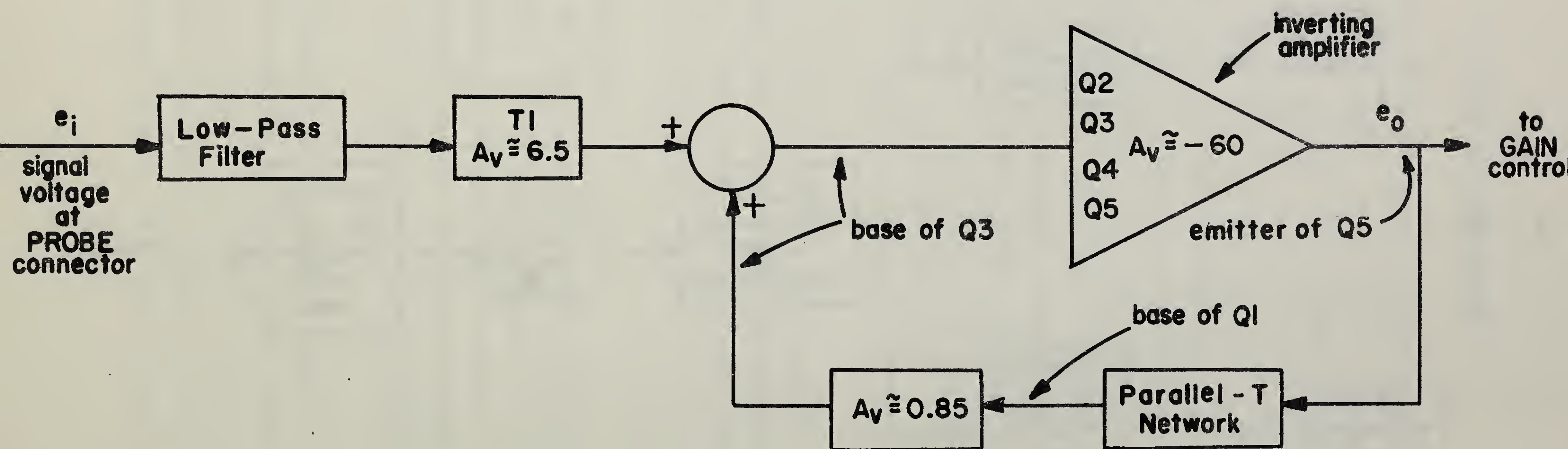


Figure 3-2. Block diagram of Tuned Input Amplifier.

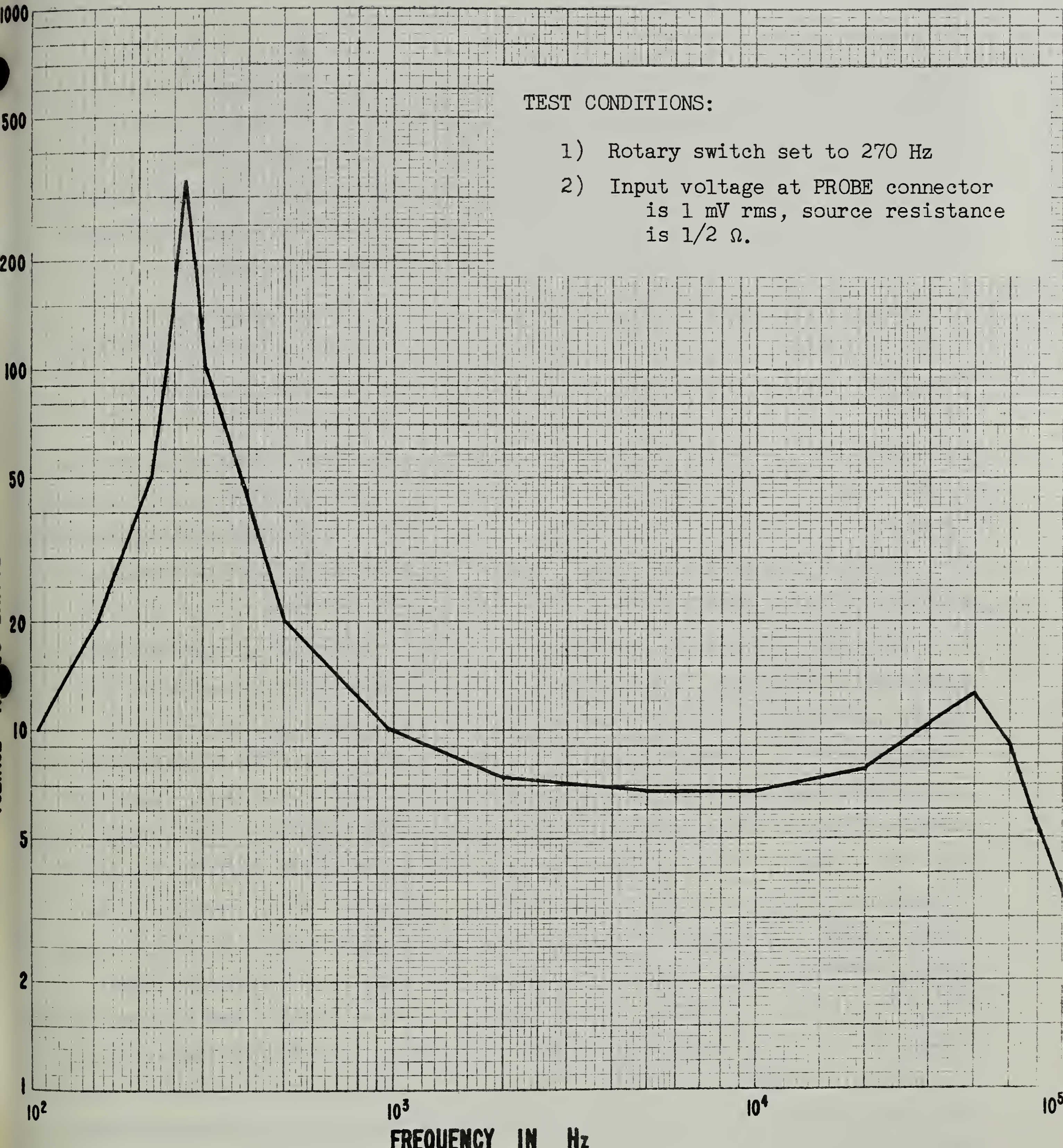


Figure 3-3. Frequency response of Tuned Input Amplifier.

direct amplification of high-frequency signals is not a source of trouble. Instead, the amplifier's sensitivity to rf signals is caused by the fact that the base-emitter junction of a transistor is normally operated as a forward-biased diode and is able to rectify signals having frequencies of tens of MHz. The dc and low-frequency components of the rectified signal constitute undersired signals to which the amplifier can respond.

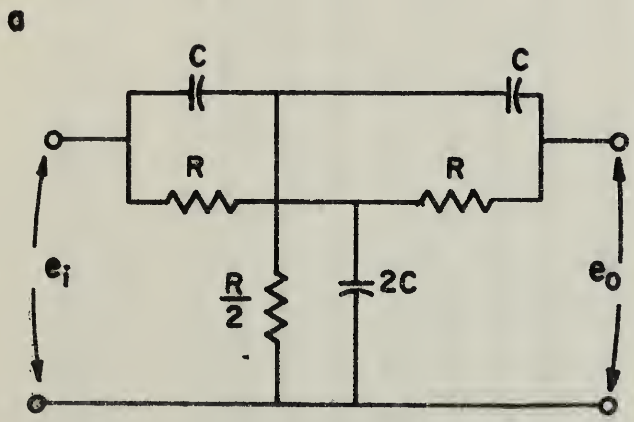
Input transformer T1 gives a voltage step-up of approximately 6.5. A higher voltage step-up would improve the signal-to-noise ratio but reduce the input signal voltage which could be handled without overloading the amplifier. Since the amplifier has no gain control ahead of it, it must be able to handle large input signals without saturating. The largest input signal which will not cause saturation is equal to the maximum output voltage swing (at Q5 emitter) divided by the voltage gain at the frequency of interest. The amplifier can handle sine-wave input signals of at least 3 mV rms in its passband and much larger signals outside of its passband.

The heart of the inverting amplifier is Q3, which is the only transistor providing voltage gain. The remaining three transistors are necessary to obtain an amplifier which is compatible with its surrounding circuitry. Constant-current source Q2 (see circuit diagram, Figure 3-11) decouples the base of Q5 from the dc supply and also increases the signal swing available at the collector of Q3. CR2 functions as a voltage regulator which establishes the base voltage of Q2. Use of this diode is necessary to make the current supplied by the Q2 collector insensitive to ripple on the dc supply. Power supply ripple of 1% causes less than .02% ripple across CR2. Use of current sink Q4 increases the output-signal swing available from emitter follower Q5. The quiescent voltage at the Q5 emitter is equal to the sum of the voltages across CR1 and the emitter junctions of Q1 and Q3. The output signal at the Q5 emitter is capacitively coupled to the GAIN control R17. If the dc voltage were not blocked, the Intermediate Amplifier would be subject to a momentary overload whenever the setting of R17 is changed.

The frequency selective characteristic of the Tuned Input Amplifier is obtained by placing a parallel-T network in the feedback path of the inverting amplifier. The parameters chosen here for the parallel-T produce a network which ideally has zero transmission at one frequency (see Figure 3-4b). At this frequency there is very little negative feedback and the closed-loop gain of the inverting amplifier is equal to its open-loop gain. When the signal is shifted to either side of this frequency, transmission through the parallel-T network rapidly increases and the resulting negative feedback reduces the amplifier gain. The parallel-T network should be driven from a low-impedance source and terminated with a high-impedance load, otherwise the selectivity of the Tuned Input Amplifier will suffer.[3] Emitter followers Q5 and Q1 provide the required impedance levels.

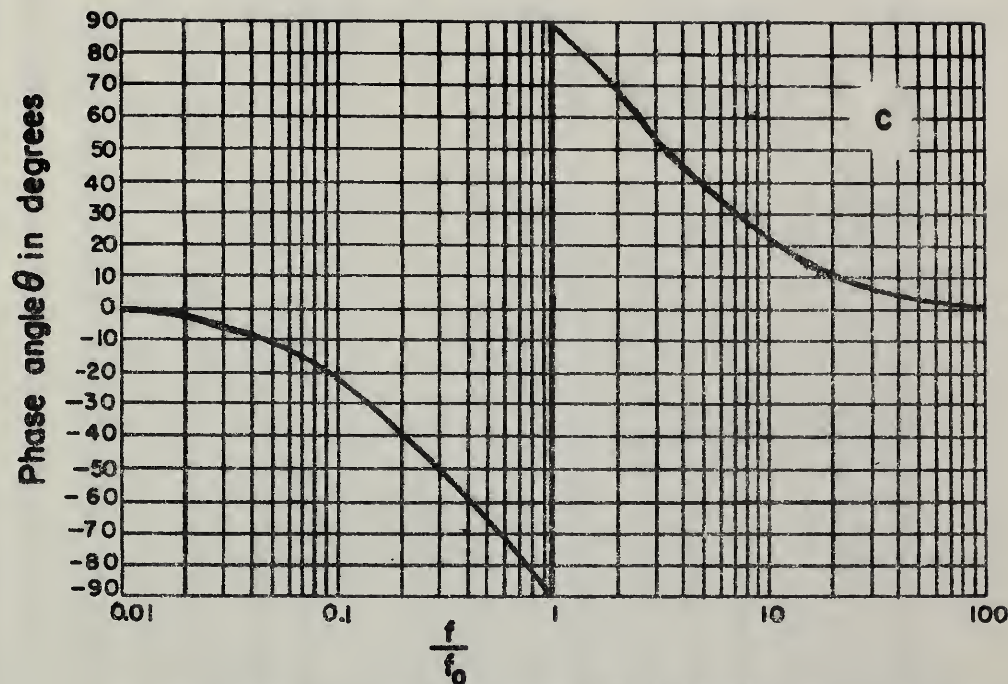
Although the parallel-T network is the predominant frequency determining element, phase shift in any other element of the feedback loop also influences the frequency at which the Tuned Input Amplifier response peaks. Phase-shift effects are negligible for the 150 Hz and the 270 Hz bands but cause the amplifier peak for the 570 Hz band to occur about 1.8% below the null frequency of the parallel-T network. This undesired phase shift is produced by C4, the high-frequency roll-off capacitor. The cause of the 1.8% frequency shift was not discovered until after the DU had been sent to the Visual Landing Aids Laboratory for field tests and therefore was not corrected. The suggested remedy is to change the position of C4 so that it is connected between the collector of Q3 and the circuit ground. This will reduce the amplifier phase shift and greatly reduce the frequency shift now present.

If the "Q" of the selective amplifier is defined as the ratio of the center frequency and the 3 db bandwidth, then the "Q" is approximately equal to 0.25A, where A is the magnitude of the loop voltage gain with the parallel-T network ignored. Hence, $A = 60 \times 0.85 = 51$ and the "Q" is near 12.7. The moderate open-loop voltage gain of 51 is doubly desirable because it makes oscillation unlikely and keeps the frequency response broad enough so that nominal changes in component values will not seriously alter the passband.



$$f_o = \frac{1}{2\pi RC}$$

= null frequency in Hz



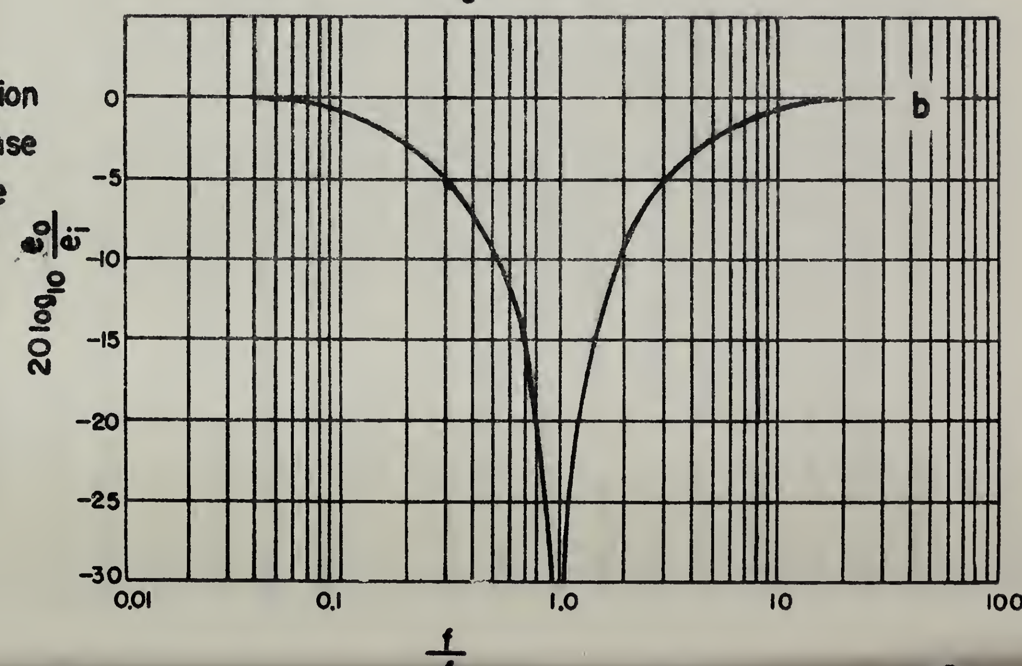


Figure 3-4. Parallel-T network

a) circuit and null frequency equation
b) amplitude versus frequency response
c) phase versus frequency response

The curves in b and c apply when the e_o terminal drives a high impedance load.

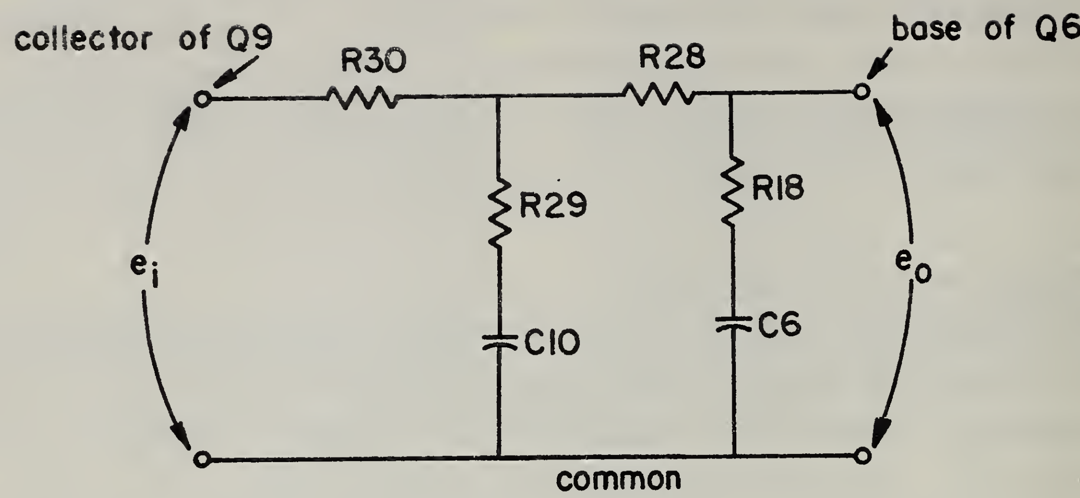
The parallel-T type of frequency selective network was chosen because there is adequate design information available, [3, 4, 5] its components are physically small, band-switching is not too troublesome, and low temperature coefficient resistors and capacitors are readily available to make the network fairly temperature insensitive.

Intermediate Amplifier (Q6 - Q10):

The Intermediate Amplifier is a dc coupled amplifier utilizing negative feedback to provide ac gain stabilization and to establish the quiescent operating point. The circuit diagram of the Intermediate Amplifier is shown in Figure 3-11. The nominal voltage gain from the arm of R17 to the collector of Q9 is 400 and is down 10% at approximately 90 and 2500 Hz.

A large signal swing at the amplifier output is obtained by biasing the collector of Q9 near one-half of the supply voltage and by employing constant-current source Q10. The quiescent output voltage is determined by the voltage established at the base of Q7, since the voltage at the base of Q6 must adjust itself to very nearly the same value. The resistance of R28 and R30 is low enough to ensure that the dc voltage at the base of Q6 is within 0.2 volts of the Q9 collector voltage. Q8 may conduct strongly when power is switched on, so R23 is necessary to limit the Q8 collector current. C9 serves to reduce the high-frequency loop gain of the amplifier and thus prevent high-frequency oscillation. R25 and C8 provide power supply decoupling.

Signal voltage at the base of Q8 produces feedback to the bases of input transistors Q6 and Q7. The desired feedback path is from the base of Q8, through Q9 and the feedback network, to the base of Q6. This path provides negative feedback (see Figure 3-5). Undesired feedback to the base of Q7 is caused by the Q8 emitter current flowing in the nonzero power-line impedance (see Figure 3-6). The voltage developed across the power-line impedance is coupled through R22 to the base of Q7 and is positive feedback at low frequencies. Low-frequency oscillation will result unless the negative feedback signal at the base of Q6 is large enough to override the positive feedback. Early versions of the amplifier would frequently break into low-

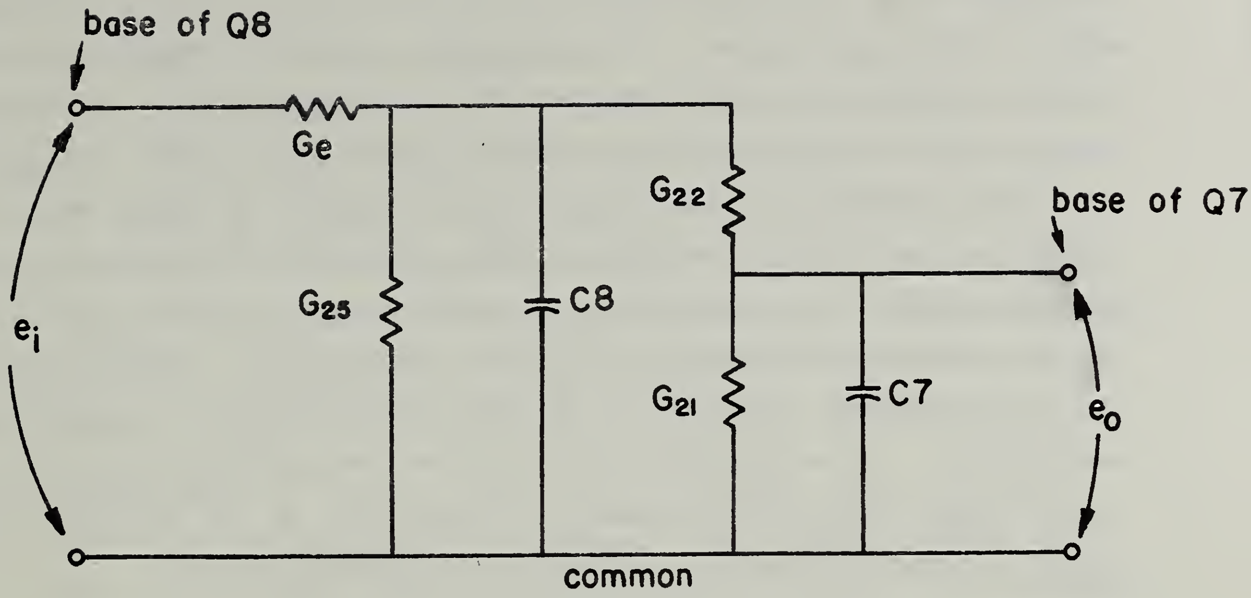


$$\frac{e_o}{e_i} = \frac{(R_{29} + 1/j\omega C_{10})\,(R_{18} + 1/j\omega C_6)}{(R_{30} + R_{29} + 1/j\omega C_{10})\,(R_{29} + R_{28} + R_{18} + 1/j\omega C_{10} + 1/j\omega C_6) - (R_{29} + 1/j\omega C_{10}}$$

$j = \sqrt{-1}$ $\qquad\qquad$ $\omega = 2\pi f$

The electrical values of the above components are given in the schematic diagram shown in Figure 3-11.

FIGURE 3-5. Transfer function of the principal feedback path. The above transfer function is valid if the feedback path terminates at a high-impedance load. The network is loaded, however, by the input impedance of Q6. This loading causes the magnitude of e_o/e_i to drop to around 0.85 of the open-circuit value above 20 Hz, and to around 0.4 of the open-circuit value below 0.5 Hz.



$$\frac{e_o}{e_i} = \frac{G_e G_{22}}{(G_e + G_{25} + G_{22} + j\omega C_8)(G_{22} + G_{21} + j\omega C_7) - G_{22}^2}$$

$$j = \sqrt{-1} \qquad \omega = 2\pi f$$

The electrical values of the components shown above are given in the schematic diagram shown in Figure 3-11. The symbol G means that the resistor values are to be expressed in mhos (units of conductance) instead of in ohms. G_e is the intrinsic conductance of the Q8 emitter junction. Its value in mhos is approximately equal to the emitter current, expressed in milliamperes, divided by 30; which is .011 mho for the case at hand.

Figure 3-6. Simplified circuit of parasitic feedback path. The above simplified circuit shows the principal elements of the parasitic feedback path. Although the effects of battery resistance and of loading by the base of Q7 are omitted, the behavior of the complete circuit is described without serious error.

frequency oscillation at roughly 1 Hz when the dc power was applied. The amplifier could be stabilized by momentarily shorting the output to ground with a large capacitance. In other words, stability criteria which would prevent oscillation under steady-state conditions did not necessarily prevent the circuit from breaking into sustained oscillation when initially turned on. Since analysis of the steady-state stability criteria was of little use, values for the circuit parameters which affected stability were established experimentally. In particular, changing either C7 or C10 by a factor of ten in either direction does not cause oscillation. Increasing the value of C10 or decreasing the value of C7 moves the circuit in the direction of oscillation. A large voltage gain at low frequencies is very desirable in the Q9 stage, for this increases the magnitude of the negative feedback signal without increasing the positive feedback signal.

Power Amplifier (Q11, Q12):

The Power Amplifier is a push-pull circuit designed to drive 500 ohm headphones. The amplifier can, however, safely drive any other load impedance, including a short circuit. It delivers 4 mW into a 500 ohm resistive load when the signal strength meter reads full-scale. The circuit diagram of the Power Amplifier is shown in Figure 3-11. The voltage gain of the amplifier from the collector of Q9 to the HEADPHONES jack is approximately 1.7 when loaded with 500 ohms.

The voltage developed across R32 and CR5 biases Q11 and Q12 slightly ON when no signal is present. Diode CR5 is used to ensure that satisfactory bias voltage is maintained in spite of changes in battery voltage and temperature. The voltage developed across R32 increases the bias voltage slightly so that the emitter current in Q11 and Q12 is large enough to establish a reasonably low value of small-signal internal emitter resistance. If the internal emitter resistance is too high, the voltage gain from the base of Q11 (and Q12) to its emitter will suffer. Emitter resistors R34 and R35 ensure that the quiescent emitter current will be satisfactory without requiring that the base-emitter voltage of Q11 and Q12 be individually matched to the voltage drop across D5. R34 and R35 differ in resistance in order to compensate for a difference in the dc resistance of the two halves of the center-tapped winding of T3. R33 is required to damp out a resonance of C11 with the inductance of T2 which occurs at approximately 150 Hz.

<u>Operational Rectifier and Meter (Q13 - Q15)</u>:

The Operational Rectifier and Meter form an ac voltmeter. The meter reading is proportional to the ac voltage applied to R36. A block diagram of the circuit is shown in Figure 3-7. The operation of the circuit is straight-forward. The ac input voltage e_i produces a current in the input resistor which is equal to e_i/R36. Input transistor Q13 draws very little of this signal current, so the instantaneous output voltage e_o must assume the value required to draw this current through the full-wave bridge rectifier in the feedback path of the amplifier. Thus the current waveform in the meter is the current in R36 which has undergone full-wave rectification.

The above description of circuit operation is accurate only if the amplifier has high gain and high input impedance. The current in R36 can be equal to e_i/R36 only if the voltage swing at point A is small with respect to e_i. The amplifier gain is adequate if this situation exists. Amplifier input impedance is large enough when the current drawn by its input terminal is negligible with respect to the current in R36. The gain and input impedance of the three-transistor amplifier used is adequate for driving the small 100 μA meter. The meter movement itself is the largest cause of non-linearity.

A few properties of the Operational Rectifier and Meter circuit should be noted. The dc bias current for Q13 flows through the 0-100 μA meter. This current is typically 0.3 μA, and is always less than 1 μA. Bias network R39, R40 (see Figure 3-11) allows ripple on the dc supply line to appear at the base of Q14. Filtering out this ripple by shunting R40 with a capacitor would improve the amplifier somewhat, but the improvement is not needed here. C14 is used to slow down the response of the meter. The meter needle would otherwise dip annoyingly when the TSO is supplying the 20 ms interrupted signal. The meter still responds too quickly, however, to be useful when the TSO supplies its 200 ms interrupted signal. Note that the voltage of C14 reverse biases the bridge diodes, requiring the amplifier output voltage to overcome this reverse bias before signal current can flow through the bridge network.

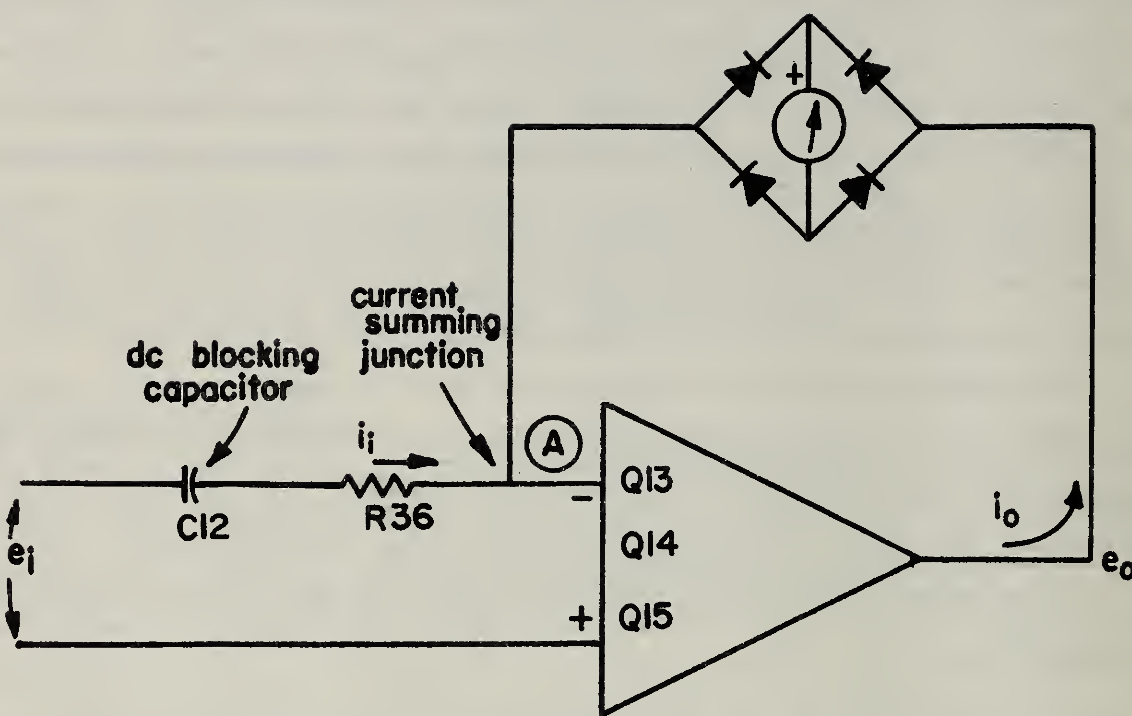


FIGURE 3-7. Block diagram of Operational Rectifier.

Power Supply (Q16):

The Power Supply circuit (Figure 3-11) consists of Q16, C15, R42, and the battery pack. Q16 is used as a diode to protect the DU from damage if the supply voltage is reversed. The diode-connected germanium transistor has substantially less forward voltage drop than even a high-conductance germanium diode. C15 maintains low ac impedance on the supply line. R42 provides protection for Q16 by limiting the turn-on charging current to C15. The small-signal resistance of the battery pack can exceed 300 ohms without affecting the operation of the DU. In other words, as long as the battery pack voltage tests OK on the built-in meter, the internal resistance of the cells is not excessive.

"AA" size cells (9/16" dia x 1-31/32" long) were selected as the power source because they are commonly available and provide long service life. These cells are available in several different chemical systems. The table shown below provides a comparison of several battery systems available.[6] Carbon-zinc cells designed for use in transistor radios (Eveready #1015 or equivalent) are recommended for general purpose use. The table can be used to select cells to meet special requirements, such as need for extra-long life or high-temperature operation.

Comparison of Battery Systems

Chemical System	Eveready #	Approximate Operating Life in Hours*	Suggested Temperature Range in °C
carbon-zinc (transistor radio)	1015	130	-15 to +40
carbon-zinc (flashlight service)	915	90	-15 to +40
mercury	E502 or E9	270	+5 to above +50
alkaline	E91	100	-25 to above +50
nickel-cadmium (rechargeable, sealed)	C450	45	-20 to +45

*Operating life is based on a load of 150 ohms across a single cell, cell used two hours per day at 21°C, and 1.1 volts at end of life.

TEST DATA

Detector Unit

This section is a collection of data obtained from tests made on an operating DU. The data were obtained from measurements made on only one DU.

DC Voltage Measurements:

The dc voltage measurements are included as a part of the circuit diagram, Figure 3-11.

AC Voltage Measurements:

A series of ac voltage measurements made with an input signal of fixed frequency and amplitude is given in Figure 3-8.

Frequency Response:

Frequency response curves for the DU are given in Figures 3-3, 3-9, and 3-10.

Noise:

With the PROBE connector open-circuited and the GAIN control set to its extreme clockwise (maximum gain) position, the noise voltage appearing across a 500 ohm resistive load at the HEADPHONES jack is approximately 18, 11, and 8 mV rms for the 150, 270, and 570 Hz bands respectively. The noise voltage is around 1.4 mV rms for all band settings when the GAIN control is set to its extreme counterclockwise (minimum gain) position.

0	Location	mV rms	Waveform
1	PROBE connector, pin A	1.00	268 Hz sine wave
2	T1, pin 5	0.94	268 Hz sine wave
3	Q1 collector	.09	268 Hz sine wave, distorted
4	Q3 emitter	3.6	268 Hz sine wave
5	Q3 base	5.6	268 Hz sine wave
6	Q3 collector	344	268 Hz sine wave
7	Q5 emitter	340	268 Hz sine wave
8	R17 sliding contact	2.0	268 Hz sine wave
9	Q6 base	.09	268 Hz sine wave
10	Q6 collector	1.16	268 Hz sine wave
11	Q7 base	< .02	
12	Q9 base	11.3	268 Hz sine wave, positive peak flattened
13	Q9 collector	830	268 Hz sine wave
14	T2, pin 4	940	268 Hz sine wave
15	Q11 base	2.14	268 Hz sine wave
16	Q11 emitter	1.83	268 Hz sine wave, negative peak flattened
17	Q11 collector	4.3	536 Hz sine wave plus some 268 Hz signal
18	T3, pin 1	1620	268 Hz sine wave
19	T3, pin 4	1480	268 Hz sine wave
20	Q13 base	3.2	268 Hz sine wave plus large 536 Hz signal
21	Q15 collector	400	268 Hz rounded square wave, 860 mV p-p
22	Q16 emitter	3.7	536 Hz sine wave plus some 268 Hz signal

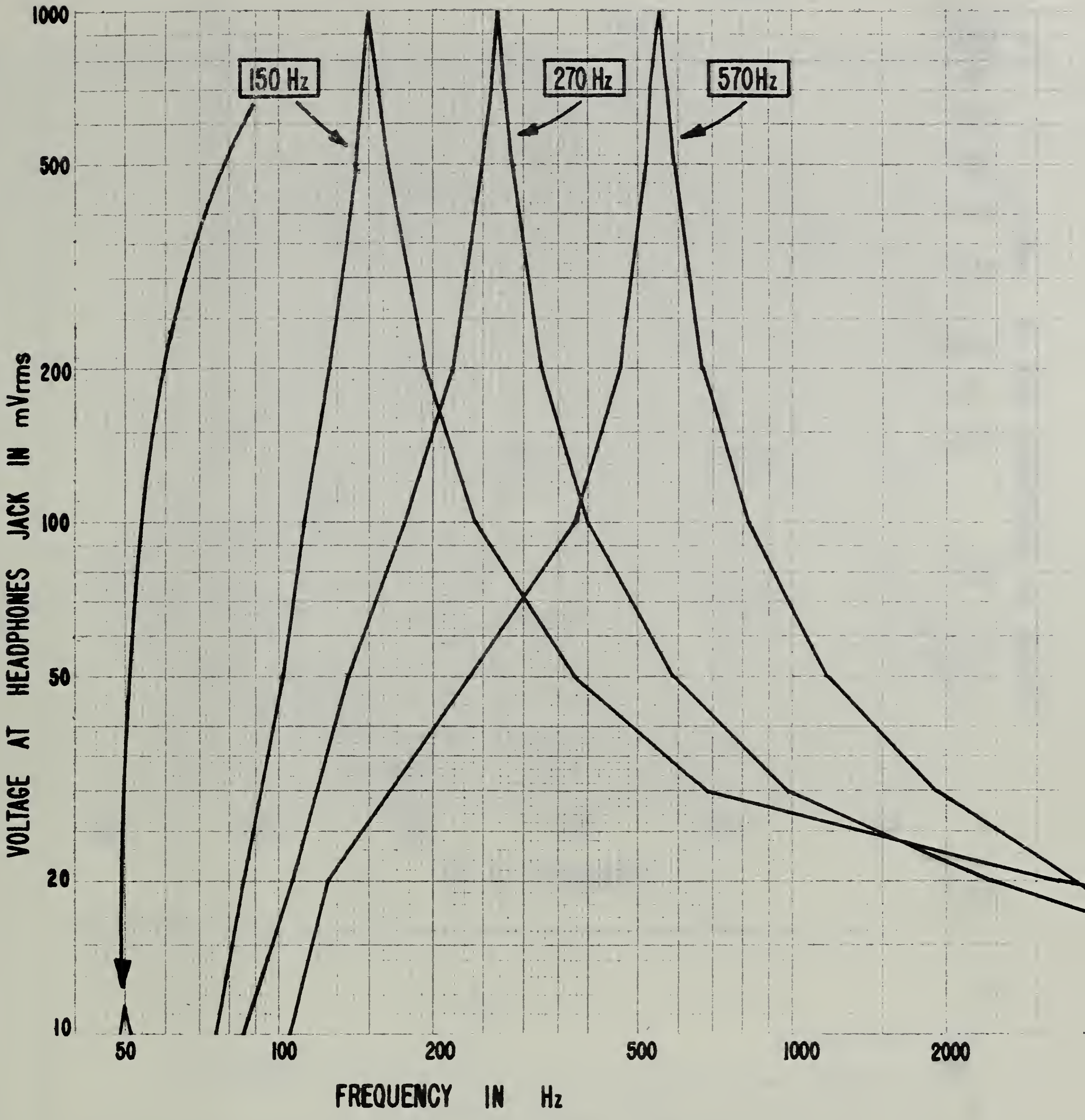
Test Conditions:

1) Battery pack voltage is 6 V dc.
2) Rotary selector switch set at 270 Hz.
3) 500 ohm load connected to the HEADPHONES jack.
4) 1.00 mV rms sine wave applied to the PROBE connector from a 0.5 ohm source; frequency is adjusted to maximize the panel meter reading.
5) GAIN control adjusted to produce full-scale deflection of the meter.

Notes:

1) Ballantine Laboratories Model 320 true rms voltmeter used to measure ac voltages.
2) Input impedance of the voltmeter is 10 megohms shunted by approximately 90 pF (includes capacitance of shielded input cable used).
3) All voltages are measured with respect to ground.
4) If an ac operated power supply is used instead of the battery pack, measurements 3 and 11 will probably be masked by voltage fluctuations from the supply.

FIGURE 3-8. Typical ac voltage measurements for the Detector Unit.



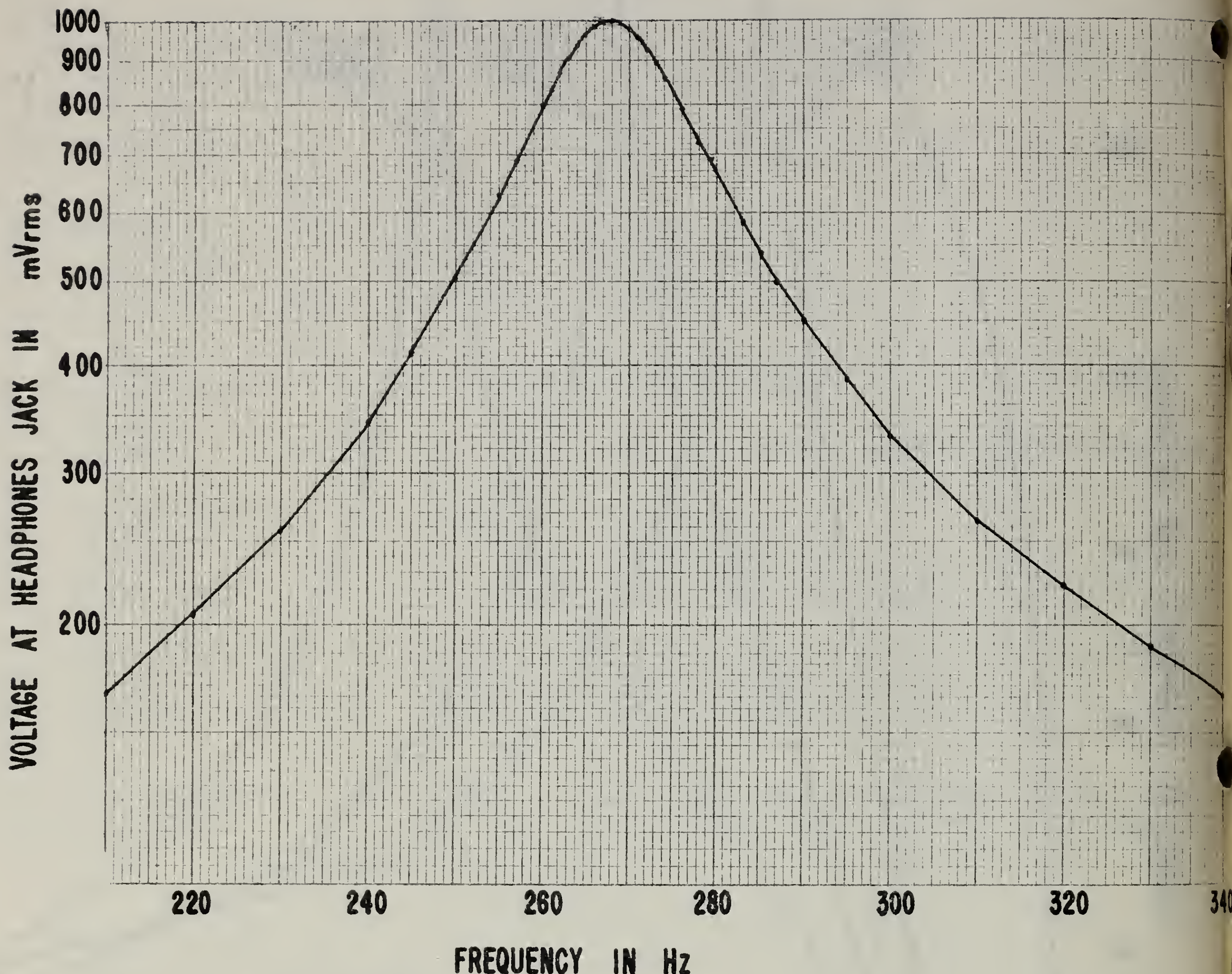
Test Conditions:

1) A 1 mV rms sine wave test signal is applied to the PROBE input connector from a 0.5 ohm source.

2) A 500 ohm resistive load is connected at the HEADPHONES jack.

3) The GAIN control is adjusted to produce 1.00 V rms at the HEADPHONES jack at the response peak of each frequency band.

FIGURE 3-9. Frequency response of each band of the Detector Unit.



Test Conditions:

1) Battery pack voltage is 6.0 V dc.
2) Rotary selector switch set at 270 Hz.
3) 500 ohm resistive load connected to the HEADPHONES jack.
4) 1.0 mV rms sine wave applied to the PROBE connector from a 0.5 ohm source
5) GAIN control adjusted to produce 1.00 V rms at the HEADPHONES jack at the frequency which maximizes the output voltage.

FIGURE 3-10. Frequency response of the Detector Unit near the response peak.

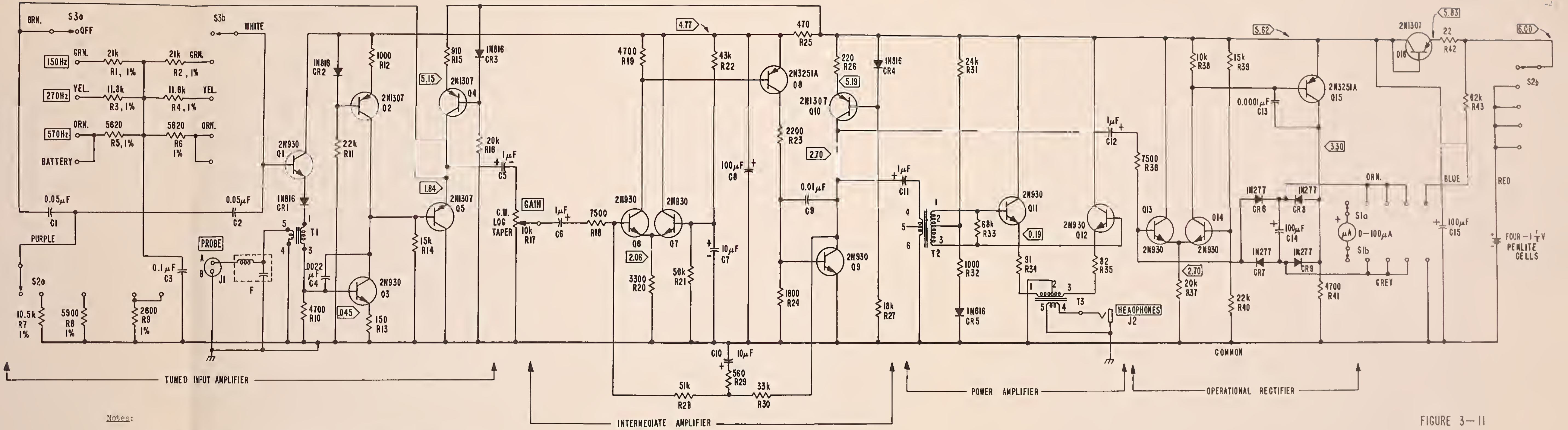


FIGURE 3—11

DETECTOR UNIT

Notes:

1) All switches are mounted on a common shaft and are mechanically ganged. S1 is mounted the closest to the front panel, S3 is farthest from the front panel.

2) Switches are shown in their extreme ccw position as viewed from the front panel.

3) The pointer boxes ⊏⊃ contain dc voltage measurements, in volts, made at the indicated points under the following test conditions:

   1. Battery pack voltage is 6.00 V dc.
   2. PROBE connector is open-circuited (i.e. no ac input signal).
   3. Rotary switch is set to 270 Hz.
   4. GAIN control is at its extreme ccw (minimum gain) position.
   5. Voltages are measured with respect to the circuit common with a high-impedance voltmeter which does not measurably load the circuit.

DESCRIPTION OF CIRCUIT COMPONENTS

Detector Unit

Capacitors:

Polarized capacitors are solid tantalum electrolytics from the Sprague 150D series (military series CS13). Voltage ratings are 35, 20, and 10 volts dc for the 1, 10, and 100 μF capacitors, respectively.

C1, C2, and C3 are 50 volt dc polycarbonate capacitors whose capacitance at -25 and +85°C is within 0.5% of their +25°C value. C1 and C2 are Component Research Company, Inc. part number 05PL224D, and C3 is part number 05PL104D.

C4 and C9 are miniature tubular capacitors from the Sprague 192P "Pacer" series and have ±10% tolerance and an 80 volt dc working voltage.

C13 is a disk ceramic capacitor.

Resistors:

Unless otherwise noted, 0.25 watt carbon composition resistors having ±5% tolerance are used.

Resistors which are specified by the circuit diagram to have 1% tolerance are Corning Electronic Components type NC4. These are 0.1 watt tin oxide resistors having temperature coefficients no greater than ±50 ppm/°C.

R17 is a 2 watt molded carbon potentiometer having a clockwise logarithmic taper, "O" ring seals on shaft and bushing; Allen-Bradley JA1P056P103AA.

Transformers:

T1 is a Stancor PCT-54 miniature transistor transformer. This transformer has a 12 ohm winding (2.7 ohms dc and 8 mH) and a 600 ohm center tapped winding (92 ohms dc).

T2 is a Triad SP-13 miniature transistor transformer. This transformer has an 800 ohm winding and a 20,000 ohm center-tapped winding.

T3 is a Stancor PCT-62 or a Triad SP-52 miniature transistor transformer. This transformer has a 600 ohm winding (97 ohms dc) and a 1500 ohm center-tapped winding (167 ohms dc).

Miscellaneous:

F, the low-pass rf suppression filter is Erie Technological Products, Inc. part number 1200-019. This is an L-section attenuator made up of a capacitor of around 2.2µF and an inductor of around 80µH and 1.8 ohms dc.

The meter is a Honeywell model HS1, 0-100 µA meter. This is a 1.5" unit which is ruggedized and hermetically sealed.

The PROBE connector, J1, is military type MS-3102A-10SL-4P (Amphenol, Cannon, etc.)

The three-deck rotary switch, S1a through S3b, is assembled from the following Centralab components: Two PS-22 switch sections, one PS-23 switch section, and one P-504 shaft and index assembly. The PS-23 switch section is mounted closest to the knob end of the shaft. All of the switch sections are 2-pole, 5-position units. The PS-22 switch has shorting (make-before-break) contacts, while the PS-23 switch has non-shorting (break-before-make) contacts.

Section 4

MISCELLANEOUS

IMPROVEMENTS SUGGESTED for the CABLE FAULT LOCATOR

Test Signal Oscillator:

1) Increasing the secondary voltage of power transformer T1 to 20 V rms each side of the center tap when the primary voltage is 130 V rms will enable the instrument to operate on input voltages between 100 and 130 V rms, and also increase the output power available from the TSO by around 20%.

Detector Unit:

1) Connecting C4 between the Q3 collector and ground will bring the 570 Hz band much closer to its stated center frequency.

2) The noise generated by the Tuned Input Amplifier can be reduced. Transistor Q1 is the predominant noise source and the use of a lower noise transistor is desirable. Employing a lower noise transistor for Q3 causes a small additional reduction in the amplifier noise. For example, using 2N4384 transistors for Q1 and Q3 results in typical noise voltages of 6, 5, and 3.5 mV rms for the 150, 270, and 570 Hz bands respectively. This compares with typical noise voltages of 18, 11, and 8 mV rms when 2N930 transistors are used. The preceding noise voltages are measured across a 500 ohm resistive load at the HEADPHONES jack when the GAIN control is at its extreme clockwise (maximum gain) position and the PROBE connector is open-circuited.

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