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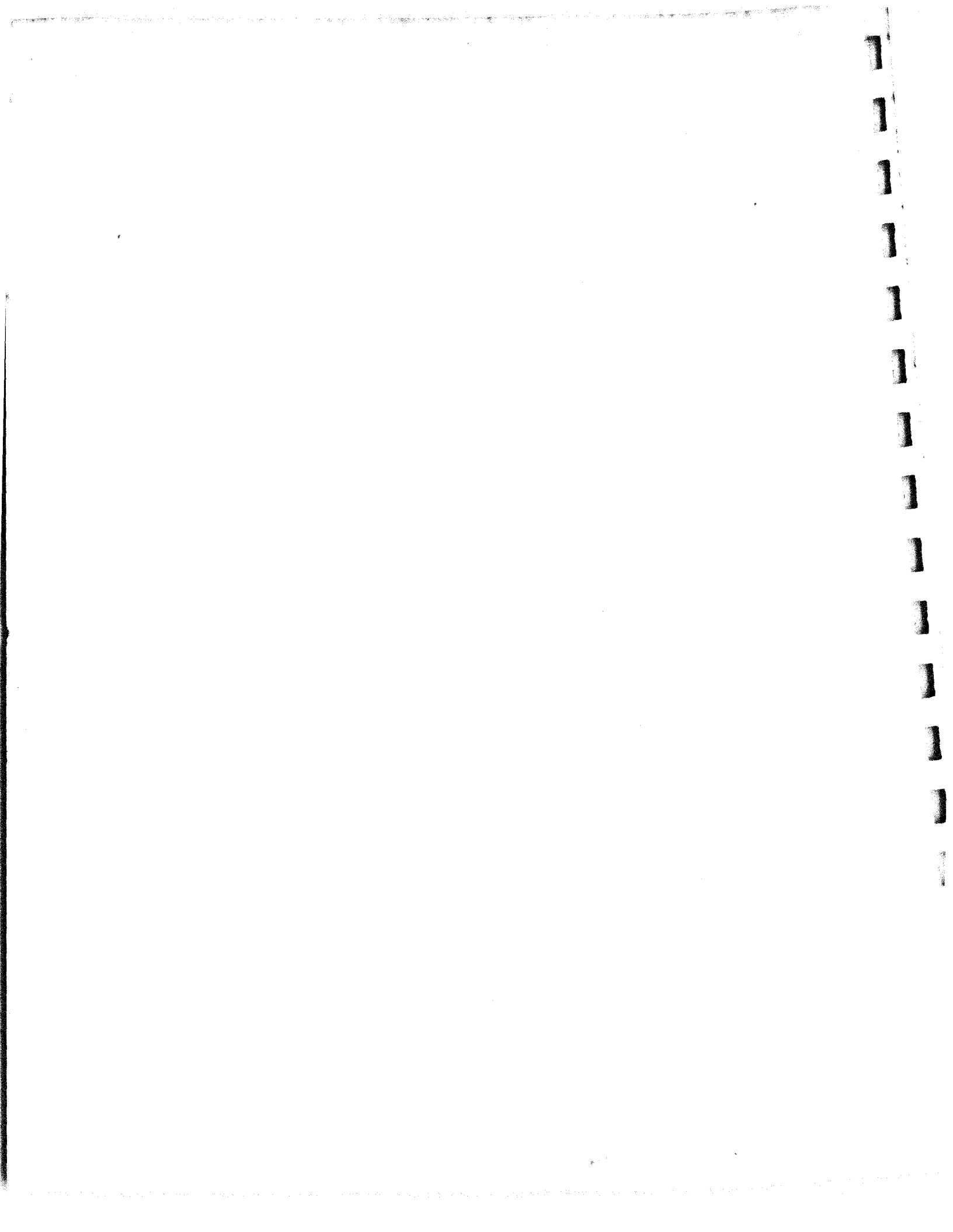


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FEDERAL AVIATION AGENCY
OFFICE OF PERSONNEL AND TRAINING
FAA ACADEMY



RECEIVERS

THIRD EDITION

FIRST PRINTING

COMMUNICATIONS EQUIPMENT BRANCH
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FEDERAL AVIATION AGENCY ACADEMY

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and Systems Maintenance Service.

COMMUNICATIONS RECEIVERS

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COMMUNICATION RECEIVERS

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RECEIVER—COURSE OUTLINE

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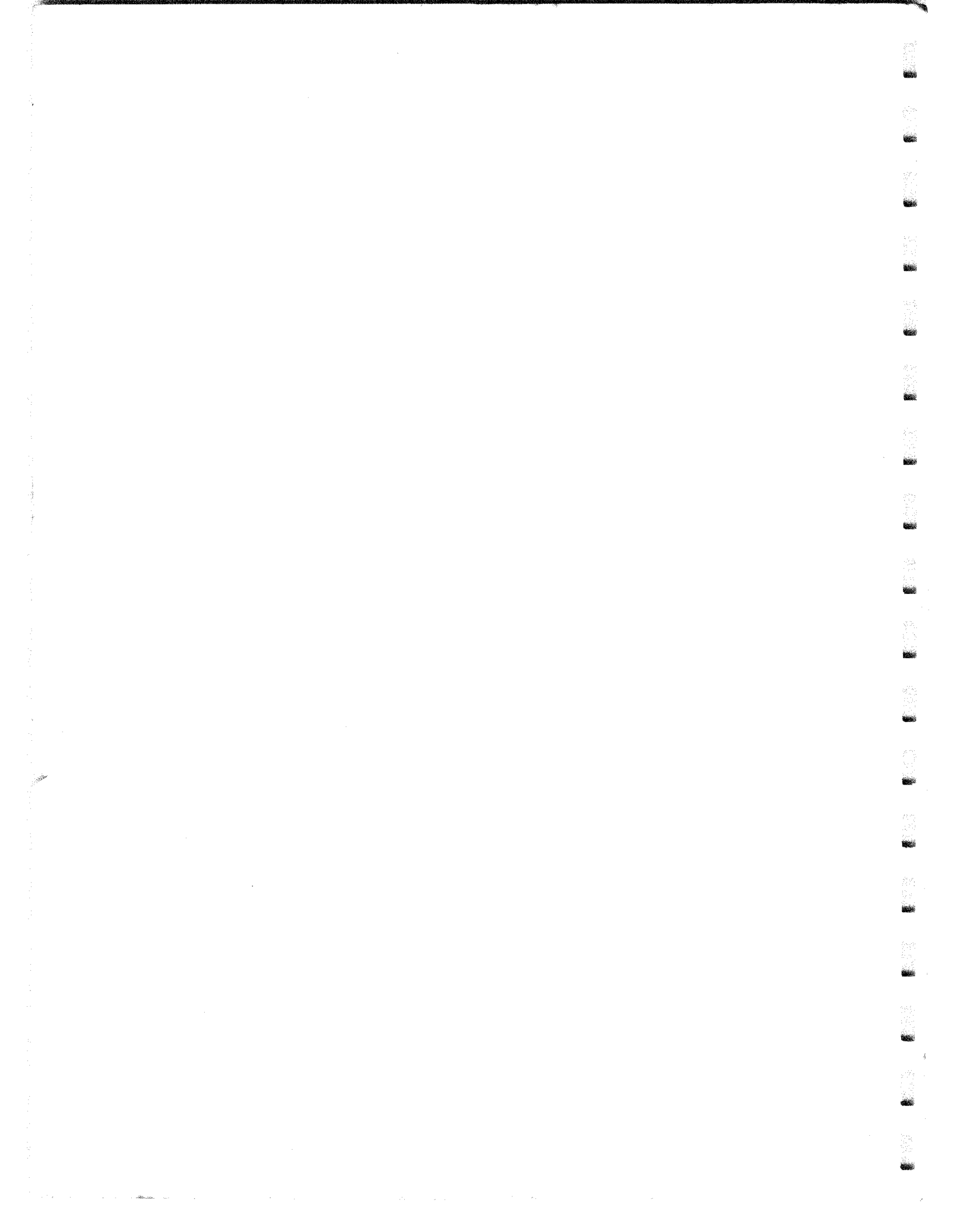
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10.0 TENTH DAY

A. *Final Examination*



CHAPTER 1

CA-1510 STANDARD SIGNAL GENERATOR

1-1. INTRODUCTION

The material presented in the following chapters is for the purposes of familiarizing the student with the many Communications Receivers used by the FAA and to familiarize the student with alignment procedures and methods of making performance measurements to insure continuous operation and optimum performance of these various receivers.

Since the average student has an understanding of the general characteristics of receiving equipment used for the reception of amplitude modulated signals, only a brief review of these concepts will be given. Detailed description of the special circuits found in each receiver will be adequately covered.

Before the characteristics of the individual receivers, the methods of making performance measurements, and alignment of various receiver stages can be presented, it is necessary to become familiar with the test equipment to be used in the laboratory. Of the test equipment required, the CA-1510 Standard Signal Generator is by far the most complex and used most frequently. A detailed discussion of this signal generator and methods of interpolation applicable to the VFO tuning mechanism follows. It is imperative the student obtain a thorough understanding of the operation of the signal generator because every laboratory experiment on amplitude modulation receivers depends upon proper operation of the signal generator, if correct results are to be obtained.

1-2. GENERAL DESCRIPTION

The type CA-1510 Standard Signal Generator is of a portable type to be used for testing and calibrating radio receiving equipment within the frequency range of 80 kc to 200 mc. It is free of any spurious FM and the crystal

controlled provision allows the generator to be used as an accurate frequency standard for the presetting of receiver channel frequencies.

The signal generator is equipped with 10 installed crystals, RG-9A/U RF output cable with type N male connectors, clip lead adaptor and a 6-db pad.

A. Physical Description

The signal generator is constructed of two separate chassis-panel assemblies, both mounted side by side in the single carrying case. The left unit contains the power supply, audio circuits, variable frequency oscillator and meters. The right unit contains all of the RF circuits enclosed in double shielding, the crystal compartment holding ten installed crystals and one spare holder, crystal access door and the RF attenuator. One power cable and one coaxial line interconnect the two units within the cabinet. Refer to Figure 1-1 for front view.

The signal generator is equipped with two carrying handles and a front cover which is held in place with 4 latches. The inside of the front cover contains clips for holding the power and output cables, two RF adaptors, instruction books and up to 25 additional crystals.

B. Electrical Outline

Refer to Figure 1-2, Functional Diagram, and Figure 1-13, Schematic Diagram. The signal generator contains a crystal oscillator circuit covering the range of 3 to 6.25 mc. Five frequency doublers are gang-tuned to the output frequency of the oscillator. This provides a maximum frequency multiplication of 32 times under normal operating conditions. The output frequency range of the first doubler is 6 to 12.5 mc; the second doubler frequency

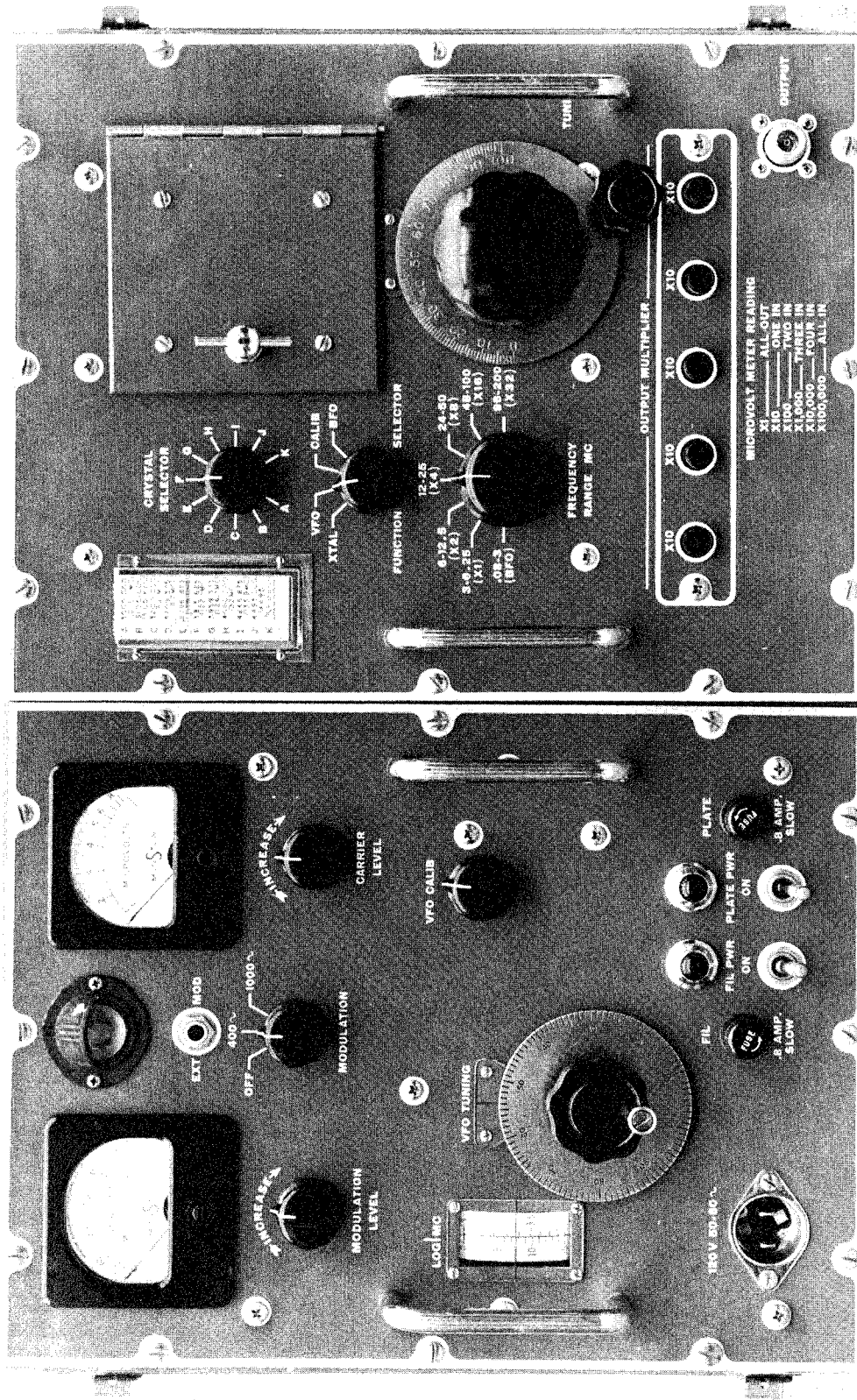


Figure 1-1. CA-1510 Signal Generator, Front View

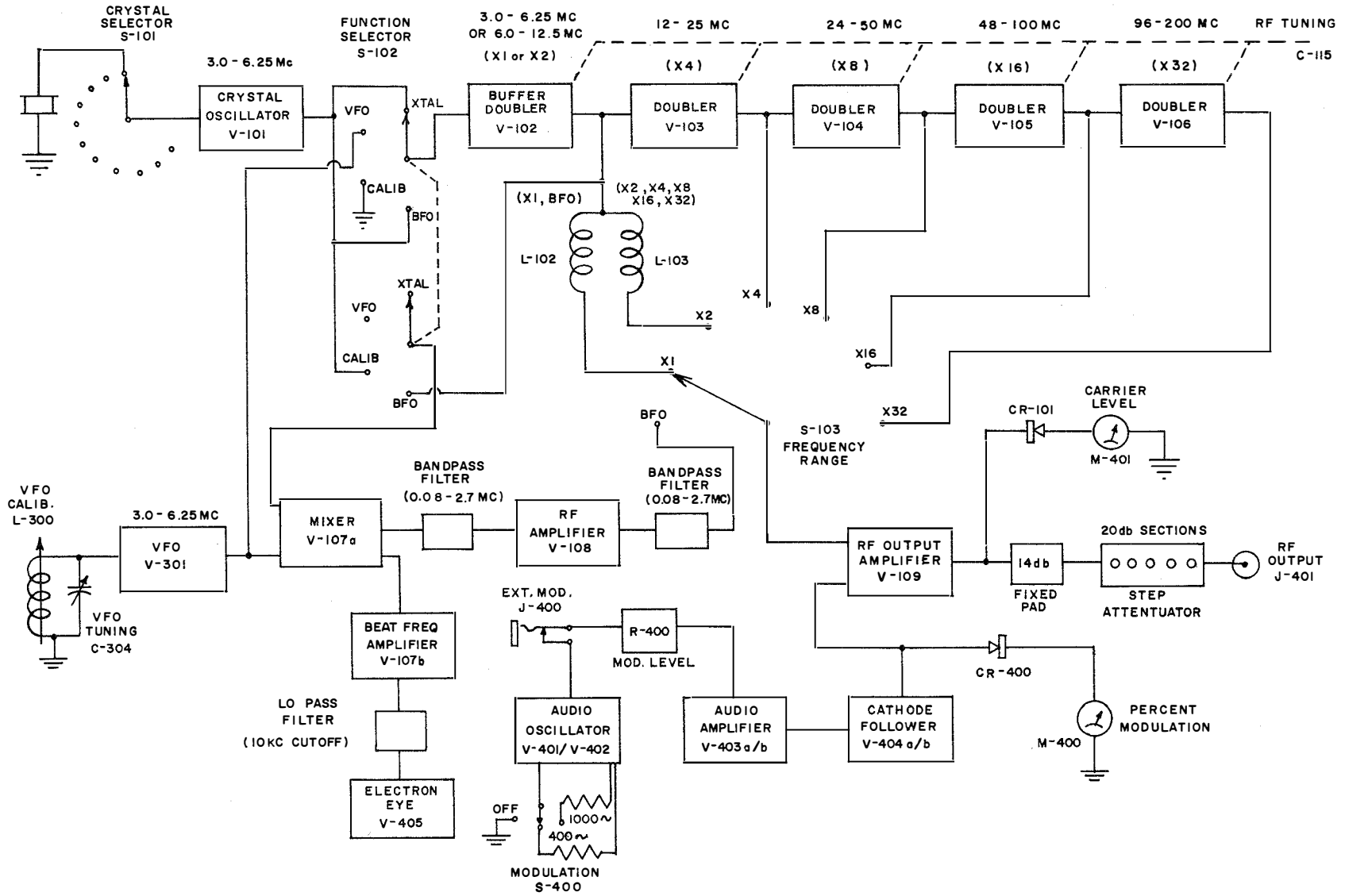


Figure 1-2. CA-1510 Signal Generator Functional Diagram

range is 12 to 25 mc; the third doubler frequency range is 24 to 50 mc; the fourth doubler frequency range is 48 to 100 mc; and the fifth doubler frequency range is 96 to 200 mc. The first of these frequency multipliers can also be operated as a buffer stage.

A separate untuned RF amplifier-modulator tube can be driven by the output of any of the frequency multipliers or buffer, depending upon frequency range selected. This stage can deliver modulated or unmodulated RF output between 80 kc and 200 mc.

A variable frequency oscillator covering the range of 3 mc to 6.25 mc can be selected to replace the crystal oscillator when variable frequency output between 3 mc and 200 mc is desired.

The variable frequency oscillator and the crystal oscillator outputs can be mixed and the difference frequency selected to provide RF output between 80 kc and 2.7 mc.

The RF mixer is also used to calibrate the variable frequency oscillator with any crystal inserted in the generator. An electron eye tube is provided for an indicator of proper calibration.

A resistive step attenuator system and RF voltmeter are used to provide calibrated output voltage.

An internal audio oscillator will provide either 400 cps or 1000 cps for modulating the RF carrier generated. Provision is made for external modulation also.

1-3. DETAILED CIRCUIT THEORY OF OPERATION

It is realized the CA-1510 Standard Signal Generator is but one of many RF signal generators used throughout the FAA for alignment, calibration and testing of communications and monitor receiving equipment. The circuit analysis and operation techniques of the CA-1510 should be applicable in part to other signal generators. It could be considered as a combined Servo Model 2400 crystal controlled signal generator and a Measurements Corporation Model 80 variable frequency signal generator, both of which may be found in the laboratory. Each mode of operation will be discussed separately.

A. Crystal Controlled Operation

The crystal oscillator consists of, V-101, 6CB6 pentode, operating as a modified Colpitts oscillator in which no tuned circuit is required. The crystal oscillator employs quartz crystals mounted in FT-243 holders. The frequency range is 3.0 to 6.25 mc. C-102 is preset at the factory to provide a total shunt capacity to the crystal in use of 32 μf . Mountings for 11 crystals are provided, accessible from the front panel. Figure 1-3 is a table showing Crystal Selector Switch position and the frequency of the crystal installed. Crystal Selector Switch, S-101, selects the desired crystal. The output of the oscillator is fed through Function Selector Switch, S-102, to either the first buffer-multiplier, V-102, or the mixer, V-107, as required.

Position	Frequency - kc
A	3000.000
B	3105.000*
C	3200.000
D	3709.375
E	3796.875
F	3815.625
G	3959.375
H	4000.000
I	4246.875
J	4495.000
K	spare

Figure 1-3. Crystal Positions

While the screen grid of the oscillator tube is provided 30v dc by means of voltage divider R-102 and R-138, the plate voltage is dependent upon the position of Function Selector Switch, S-102. In the BFO and Crystal position, S-102-B-1 connects the plate load resistor, R-103, to the Carrier Level Control, which provides regulated dc adjustable from 0 to 150 volts. S-102-A-1 connects the oscilla-

* This crystal may be replaced with 3023.500 kc in the field facility.

tor output to the buffer-multiplier. In the Calibrate position, S-102-B-1 connects the plate load to a regulated 150v dc non-adjustable source. The oscillator output is connected to the control grid of the mixer, V-107, through S-102-A-2. In the VFO position no plate voltage is applied and there is no output. Only Crystal operation will be considered at this time.

The buffer-frequency multiplier stage consists of V-102, 6CB6 pentode, which is driven by either crystal or variable frequency oscillator as selected by Function Selector Switch, S-102. The plate circuit is tuned by section C-115-A of the five-gang capacitor C-115 and inductance L-102 or L-103 as selected by the Frequency Range Switch, S-103-C.

When Frequency Range Switch, S-103, is positioned to 3 to 6.25 mc, L-102 is connected to the plate of V-102. V-102 provides buffer action with the output coupled through capacitor C-109 to the output amplifier-modulator, V-109. The buffer plate circuit is tuned to crystal frequency between 3 and 6.25 mc by C-115-A, a section of the ganged Tuning Control. The screen grid voltage is obtained from the Carrier Level Control, which provides regulated dc from 0 to 150v dc. The 3 to 6.25 mc output is varied by variation of the buffer screen voltage.

When the Frequency Range Switch is positioned to 6 to 12.5 mc, L-103 is connected to the plate of V-102, this stage then acts as a frequency multiplier and provides this frequency range of output signal through capacitor C-112 to V-109. The screen voltage is still applied from the Carrier Level Control. The plate circuit is normally tuned by C-115-A to the second harmonic of the crystal frequency providing doubler operation.

Although the frequency multipliers' tuning range is limited to 2.1:1 to minimize spurious responses, it is possible to operate the stage other than a doubler when in the 6 to 12.5 mc position of the Frequency Range Switch. However, should this stage be operated as a frequency multiplier other than a doubler, it would be difficult to determine the output frequency of the signal generator. To assist the operator in tuning this stage for doubler action, the Tuning Control is provided with

a calibrated dial from 0 to 100 divisions. The Tuning Control dial reading of 5 corresponds closely to the lower end of the output frequency range, in this case 6 mc, and a dial reading of 95 corresponds closely to the higher limit of the output frequency range, in this case 12.5 mc. This dial indication should be observed when adjusting the Tuning Control for any output frequency.

By studying carefully the following examples of false resonance points, the student should be able to avoid such spurious responses.

It is desired to produce crystal controlled operation on 6 mc by using a 3 mc crystal, since a 6 mc crystal is not normally provided. The 3 mc crystal is selected and the signal generator adjusted for Crystal operation. The 6 to 12.5 mc frequency range is selected. The selection of 6 to 12.5 mc will simply mean V-102 is operating as a frequency multiplier with its plate circuit tuneable from 6 to 12.5 mc. It does not mean the plate circuit is automatically tuned to twice the crystal frequency. The plate circuit will resonate to 6 mc, the second harmonic of the 3 mc crystal. The 6 mc position of the Tuning Control is near a dial indication of 5. This provides proper operation.

When V-102 is operating as a frequency multiplier, it is producing harmonic distortion due to its Class C operation. This harmonic distortion is not limited to only the second harmonic; the third and fourth harmonics are produced and are of sufficient strength to give plate tank resonance indications. If the Tuning Control is adjusted near a dial reading of 50, a resonance indication will be obtained on the Carrier Level Meter. This will be the middle of the frequency range of 6 to 12.5 mc and will correspond to 9 mc, the third harmonic of the 3 mc crystal.

Should the Tuning Control be adjusted near a dial reading of 90, a resonance indication for 12 mc will be obtained. 12 mc is the fourth harmonic of the 3 mc crystal. Only the resonance indication at a Tuning Control dial position of 5 provided the desired output frequency of 6 mc.

For Crystal operation on all frequency ranges except the 3 to 6.25 mc range, V-102 operates as a frequency multiplier with L-103

connected in the plate circuit. The CA-1510 cannot be operated crystal controlled for the 80 kc to 2.7 mc range. For all frequency ranges above 6 to 12.5 mc, the output of the first frequency multiplier is connected through capacitor C-113 to the control grid of the second multiplier tube, V-103.

The 6 to 12.5 mc output drives a series of four frequency doublers arranged in cascade, V-103, V-104, V-105 and V-106. All are 6CB6 pentodes with their tuned plate circuits ganged and tracked with the plate circuit of V-102 such that each succeeding stage is tuned to twice the frequency of the preceding stage. For Frequency Range Switch positions of 12 to 25 mc, 24 to 50 mc, 48 to 100 mc and 96 to 200 mc, the appropriate output frequency is selected through capacitors C-121, C-126, C-134 or C-142 to the RF amplifier-modulator tube, V-109. Depending upon frequency range selected, the screen voltage of these frequency multipliers may be adjusted by the Carrier Level Control, thereby controlling the level of RF output.

The carrier output frequency is selected by the Frequency Range Switch, sections S-103-D and S-103-E, and applied to the output grid-modulated RF amplifier stage, V-109, 6CL6 pentode. This stage provides modulation as desired and transforms the RF voltage of the desired frequency from its high impedance source to the low impedance of the first attenuator section consisting of resistors R-144, R-145 and R-146. Negative fixed bias for the RF amplifier-modulator is developed across the ground return resistor, R-438, in the power supply and applied to the control grid through the isolating resistors, R-416 and R-141. Audio modulation is applied from the audio circuits through capacitor C-407 to the control grid of RF amplifier-modulator, V-109.

The modulated RF output from the plate of V-109 is applied to the first attenuator section through capacitor C-159. This section comprising of resistors R-144, R-145 and R-146 transforms the output RF amplifier plate im-

pedance to 52 ohms and produces a fixed voltage reduction of 5:1 or -14 db. The following five attenuator sections are removable 52-ohm networks, each with a voltage attenuation ratio of 10:1 or -20 db. The level of the RF voltage applied to the attenuator is adjustable by varying the output of the RF stages preceding the output RF amplifier by means of the Carrier Level Control.

The Carrier Level Meter, M-401, in conjunction with a crystal diode CR-101, Type 1N21B, meters the attenuator input level in terms of microvolts output across a 52-ohm terminating load with all attenuator sections inserted, buttons out. Attenuator sections are removed from the circuit by depressing push buttons, each button pushed increases the output voltage by a factor of ten.

To adjust the output of the signal generator to 1 μ v across a 52-ohm load for example, the output signal undergoes various attenuation circuits. The input to the attenuator consisting of R-144, R-145 and R-146 is adjusted to 0.5v rms by means of the Carrier Level Control. This signal level is variable from 0.05 to 0.5 volts. It was previously mentioned this attenuator reduces the signal 5:1, or for 0.5v input across R-144, there will be an output of 0.1v across R-146 when the signal generator is terminated into 52 ohms. Refer to Figure 1-4

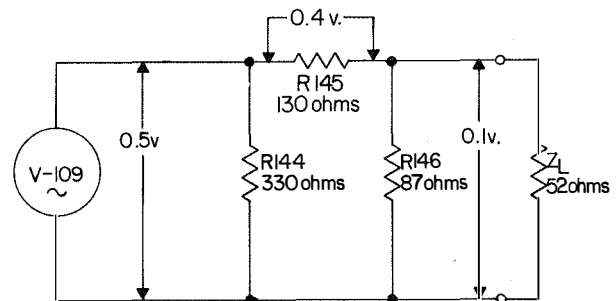


Figure 1-4. First Attenuator Section

Considering the parallel combination of Z_L and R-146:

$$\frac{R_1 R_2}{R_1 + R_2} = R_T \quad (1-1)$$

Substituting values and solving:

$$\frac{52 \times 87}{52 + 87} = 32.5 \text{ ohms}$$

This effective resistance is in series with R-145, 130 ohms. Series resistances being additive, $130 + 32.5 = 162.5$ ohms. Since this value of resistances is in parallel with R-144, the voltage across R-144, 0.5v, is across this 162.5 ohms of effective resistance. Of this value, the voltage across R-145 may be calculated:

$$\frac{130}{162.5} \times 0.5 \text{ v} = 0.4 \text{ v}$$

The remaining 0.1v will appear across R-146 and the parallel Z_L . Therefore it can be seen this network produced a 5:1 voltage reduction.

By a similar procedure it can be proven each Step Attenuator will produce a 10:1 signal voltage reduction. If all five Step Attenuator buttons are out, the attenuators are in the output circuit producing an over-all signal reduction of 100,000:1. If 0.1v rms is the input to the first Step Attenuator, 1 μ v will be the output to a 52-ohm load. This is the value the Carrier Level Meter will indicate and it is full scale deflection. For every Step Attenuator button pushed in, the output would increase 10 fold.

It was previously stated the first attenuator section, R-144, R-145 and R-146, also transforms the output RF amplifier plate impedance to 52 ohms. This is the purpose for it being used; the 5:1 signal reduction is a result. The application of Thevenin's theorem will verify this reduction to 52 ohms. It implies the effective circuit resistance as seen by the load, Z_L , looking toward the generator can be determined by short-circuiting all voltage sources and calculating the resultant resistance. If the voltage source V-109, of Figure 1-4 was short-circuited, R-144 would be short-circuited, R-145 and R-146 would then be connected in parallel instead of series. Application of Equation (1-1) to this combination would result in 52.1 ohms. By proper operation of the Carrier Level Control and the Step

Attenuators, the output of the signal generator is adjustable from 0.1 to 100,000 microvolts to an external 52-ohm termination.

Not always is the input impedance of the receiver circuit under test equal to 52 ohms. This input impedance is the load impedance for the section of RG-9A/U transmission line connecting the receiver and the signal generator output. If the transmission line is terminated into its characteristic impedance, in this case 52 ohms, the impedance at any point along the transmission line will be equal to the characteristic impedance of the transmission line. This will result in a 52-ohm load for the generator output, the value of load required if the Carrier Level Meter is to accurately indicate the signal level input to the receiver circuit.

Suppose the input impedance of the receiver circuit under test were 500 ohms. The impedance reflected to the output of the signal generator would be a value other than 52 ohms in all probability. This can be calculated if the length of transmission line, θ , the value of load impedance, Z_L and the characteristic impedance of the transmission line are known by application of Equation (1-2). The value of θ would be in degrees rather than inches.

$$Z_s = Z_o \frac{Z_L + jZ_o \tan \theta}{Z_o + jZ_L \tan \theta} \quad (1-2)$$

The load impedance for the generator would be the sending end impedance of the transmission line, Z_s . If the generator is terminated in a load other than 52 ohms, the Carrier Level Meter indication of the generator output is in error.

It is a logical assumption the input impedance of all receiver circuits will not be 52 ohms. To reduce the erroneous indications this could produce, a 6-db pad is normally used on the receiver end of the transmission line connecting the signal generator output to the receiver circuit. It is to be noted, the CA-1510 instruction material indicates the 6-db pad should be connected at the receiver end of the transmission line. Instructions for other signal generators may indicate the use of the pad at the generator end of the transmission line.

Figure 1-5 shows the circuitry of the 6-db pad used. If the output of the 6-db pad is terminated into Z_L of 52 ohms, the input impedance, Z_S , is 52 ohms and will properly terminate the transmission line, thereby producing a 52-ohm load for the generator. However, the voltage attenuation is 2:1 or 6-db loss. This would require the output on the Carrier Level Meter to be read as one-half its indicated value to determine the receiver input when using the 6-db pad. An example would be when the Carrier Level Meter shows full scale deflection or $1 \mu\text{v}$, it should be read as one-half that value or $0.5 \mu\text{v}$, as this would be the level of signal applied to the receiver. Of course the Step Attenuator buttons are all released or out.

An open circuit load for the 6-db pad would provide an input impedance for the pad of 89 ohms. For a short circuit load, the input impedance for the 6-db pad is 31 ohms. The 6-db pad has reduced load variations from an open circuit to a short circuit into variations from 89 ohms to 31 ohms. This greatly improves the accuracy of the indicated receiver input signal level.

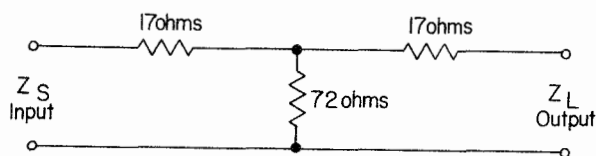


Figure 1-5. 6-db Pad

B. VFO Operation

Normally when performing alignment or calibration of various receiver circuitry, crystal controlled operation is desired due to the frequency stability of 0.001%. Periodically the crystal frequencies should be measured by a frequency standard of some type to insure operation at indicated frequency.

There are occasions when it is desired to have a variable frequency signal source. This would be desirable when measuring receiver selectivity characteristics or measuring attenuation of undesired frequencies such as an image frequency.

In VFO operation, the crystal oscillator is replaced with a grounded-plate Hartley oscil-

lator, V-301, Type 6CB6 pentode. The frequency range is 3.0 to 6.25 mc. The screen grid of V-301 is the plate of the oscillator. The oscillator output is electron coupled to the plate circuit of V-301. The plate load is R-127 and the developed signal is coupled through S-102-A-1 to the buffer-frequency multiplier, V-102, or through capacitor C-145 to the cathode of the mixer, V-107. The oscillator tank circuit is tuned by directly calibrated variable capacitor C-304 and is temperature compensated by capacitors C-300 to C-303. A powdered-iron-core slug within the coil L-300 is adjustable from the front panel for calibration purposes. The variable frequency oscillator can be accurately calibrated using any of the crystal positions as check points. The procedure can be more clearly understood by an example.

It is desired to use the signal generator on 3.21 mc. Since a crystal is not normally available at that frequency, VFO operation is essential. A crystal is available at 3.2 mc. The VFO will be set to a dial indication of 3.2 mc and calibrated against the 3.2 mc crystal by adjusting the VFO Calibrate Control, L-300. The VFO should be recalibrated at a new check point should it be desired to produce a VFO output removed by 500 kc or more from this calibration point of 3.2 mc to insure frequency accuracy. The VFO frequency stability is 0.1%.

To calibrate the VFO, the Function Selector Switch is positioned to calibrate and the Crystal Selector Switch is positioned to the 3.2 mc crystal. Crystal frequency output is applied through capacitor C-146 to the grid of the mixer V-107A. VFO output is applied to the mixer cathode through C-145. With the VFO Tuning adjusted to 3.2 mc, the VFO Calibrate Control is then adjusted. It should be noted this control will adjust through several complete revolutions from one limit of adjustment to the other.

When the VFO frequency approaches the crystal frequency, an audio beat note is obtained at the cathode of the mixer, V-107A. This is applied through C-148 to audio amplifier V-107B, the second half of the dual triode, Type 12AT7. The amplified beat note is applied through C-149 to the grid of the electron

eye indicator tube, V-405, Type 6E5. When frequencies are less than 10 kc, the eye closes. As the difference frequency is reduced to approximately 20 cycles, the eye flutters. At zero beat, the indicator eye opens. This position of tuning is very sensitive and should not be confused with an indication of frequencies in excess of 10 kc, in which case the eye also opens. The VFO is properly calibrated.

The Function Selector Switch is returned to VFO. By an interpolation procedure explained later, the VFO is set to the frequency originally desired of 3.21 mc. The VFO output is fed to V-102, which would be operated as a buffer, and through V-109 to the output circuit. It should be remembered the VFO Tuning Control only tunes the VFO. The buffer is tuned by the Tuning Control, C-115. Every time the VFO is retuned, the buffer and any frequency multipliers used must be retuned by C-115. The frequency multipliers operate the same for VFO operation as Crystal controlled operation.

The electron eye indicator circuit will be further explained. Normally the only bias on the left hand side is developed across R-429 in the control grid circuit. For this bias to be appreciable, there must be grid current. This requires application of signal. This signal comes from the mixer and audio amplifier circuits of V-107. C-202 and R-201 produce a low pass filter allowing only signals of less than 10 kc to reach the control grid of the electron eye tube. Therefore, no grid current flows through R-429 and little or no bias is developed for the left half of the tube. This section of the tube conducts heavily, producing a voltage drop across R-428. This reduces the voltage on the control grid of the right half of V-405, producing the deflection of electrons and the shadow or open eye.

As the frequency reduces below 10 kc, signal arrives at pin 3 of V-405, producing grid current on the positive peaks of the signal. This grid current produces grid-leak bias, which reduces the conduction of the left half of V-405 and reduces the voltage drop across R-428. This results in less electron stream deflection and the eye closes.

When the frequency reduces to approximately 20 cps, the grid-leak capacitor dis-

charges between positive peaks of signal and produces varying bias conditions on V-405 causing the eye to open and close, or flutter. The approximate RC time constant of the grid-leak combination of R-429 and C-149 is .01 seconds. In five time constants, or .05 seconds, the capacitor is nearly discharged. 0.05 seconds is also the time interval of one cycle of a 20 cps signal, therefore, the capacitor has ample time to nearly discharge between application of positive signal peaks.

At zero beat, there isn't any signal, no grid current for bias, and the eye is open.

C. Beat Frequency Operation

For RF output between 80 kc and 2.7 mc, a mixer circuit V-107A, Type 12AT7, is included to convert the output of the variable frequency oscillator, V-301, to a lower frequency. This is accomplished by beating with a 3 mc signal from the crystal oscillator. The 3.0 mc crystal position is selected. Turn the Function Selector Switch to crystal. The RF Tuning control is adjusted for resonance at the crystal frequency. Turn both the Function Selector and Frequency Range Switches to BFO. 3.0 mc output is obtained from the buffer stage V-102, and applied to the mixer grid through capacitor C-146. The 3 to 6.25 mc variable frequency output is applied through capacitor C-145 to the cathode of the mixer. The input signal frequencies, the sum and the difference frequencies from the mixer output are applied through capacitor C-147 to the low pass filter consisting of L-112, L-113, L-114, C-150, and C-151. This has a cutoff frequency of 2.7 mc, passing the difference frequency and rejecting the fundamental and sum frequencies. The 80 kc to 2.7 mc beat frequency is applied to the grid of low frequency RF amplifier, V-108, Type 6CB6. The amplified output is applied through capacitor C-153 and additional filter L-115, C-154 and C-165 to the output modulator V-109. The output level is adjusted by the action of Carrier Level Control, R-435, controlling the screen voltage of amplifier V-108.

D. Audio System

Modulating frequencies of 400 cps and 1000 cps may be generated within the CA-1510

signal generator. The audio oscillator producing these frequencies is of the Wein-bridge type.

Provision for application of external modulating frequencies is provided by means of the External Modulation Jack, J-400. This is a normally closed jack that will open the path for the internally generated signal when a headphone plug is inserted into the jack.

Irrespective of the modulating signal origination, an input signal of 0.5v rms is required for the grid of V-403A to produce 50% modulation. V-403 is a 12AU7 dual triode in a conventional two-stage amplifier. The percent of modulation is controlled by adjusting the input signal level by means of the Modulation Level Control, R-400. The audio amplifier output is applied to the grids of parallel-connected V-404, a type 12AT7 dual triode operated as a cathode follower. C-402 and R-405 provide inverse feedback for the amplifier. Low impedance output is obtained from the cathode circuit consisting of R-410, R-411 and R-412 in series. Capacitor C-407 couples the audio to the RF amplifier-modulator.

A Percent Modulation indication is produced on meter M-400. The level of the audio signal applied to the RF amplifier-modulator is rectified by a 1N34 crystal, CR-400, filtered and applied to the meter. The modulation is variable from 0 to 50% for desired indication.

Since the Wein-bridge oscillator is commonly used in other FAA equipment as well as the CA-1510, the theory of operation will be discussed. Figure 1-6 is a simplified Wein-bridge oscillator. The screen grid and suppressor grid circuitry has been omitted from the circuitry actually used in the CA-1510 to simplify the figure. The component designations are those actually used in the CA-1510.

The Wein-bridge oscillator employs a frequency selective bridge circuit as the resistance-capacitance feedback network. No use is made of a tank circuit consisting of inductance and capacitance to control the frequency.

Tube V-401 is the oscillator tube while V-402 provides amplification and phase inversion for feedback. Thus even without the bridge circuit, this system oscillates, since any signal that appears at the control grid of V-401 is amplified and inverted by both V-401 and V-402. The voltage fed back to the control grid of V-401 must reinforce the initial signal, which causes oscillations to be set up and maintained. However, this system amplifies voltages of a very wide frequency range. Voltages of most any frequency or combination of frequencies can cause oscillation. The bridge circuit is used to eliminate feedback voltages of all frequencies except the single frequency desired in the output.

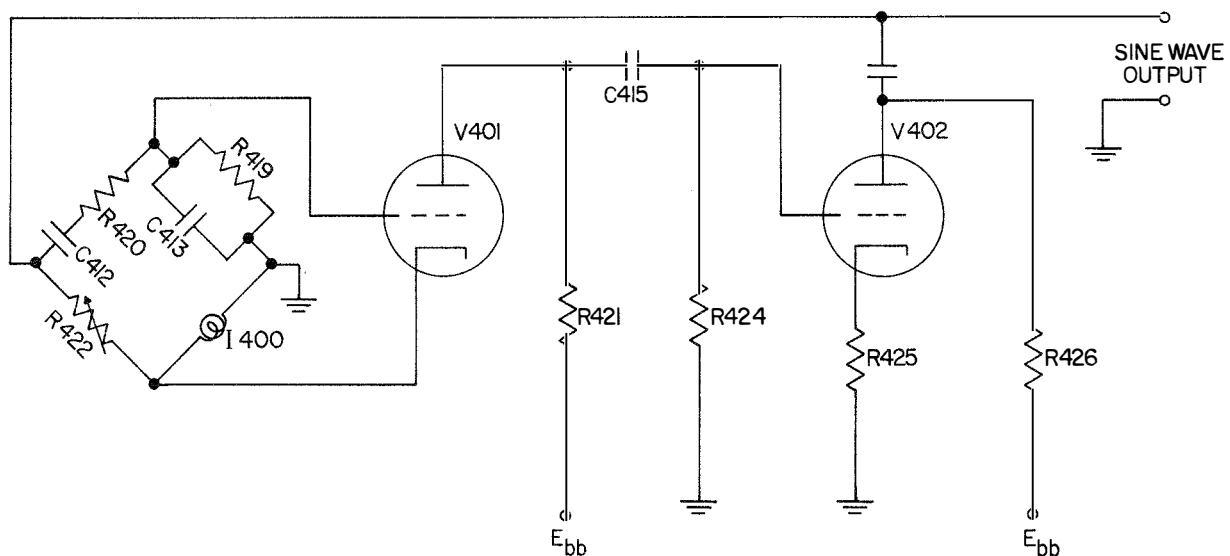


Figure 1-6. Wein-Bridge Oscillator, Simplified Circuit

For a circuit to oscillate, the regenerative feedback must compensate for the oscillatory circuit losses. To accomplish this, the amplitude and phase of the feedback voltage must be properly adjusted. Figure 1-7 shows the change in amplitude and phase of the feedback for variations in frequency. It should be observed the negative or degenerative feedback is constant even though the frequency varies. This is due to the resistive nature of the degenerative feedback circuit consisting of R-422 and I-400. There is only one frequency at which the positive feedback voltage equals the negative feedback and the phase is proper to reinforce the grid signal. This frequency is designated f_r .

The positive or regenerative feedback voltage is provided by the voltage divider consisting of C-412, R-420, C-413 and R-419. If the frequency is very high, the reactance of the capacitors approaches zero. In this case R-419 is shunted by a very low reactance, making the voltage between the control grid of V-401 and ground almost zero. Conversely, if the frequency is reduced toward zero, the reactance of the capacitors approaches infinity. Nearly all feedback voltage would be developed across C-412 and the voltage between the control grid and ground decreases to near zero value. Although the reactance of capacitor C-413 also approaches infinity, it is shunted by R-419 and the total impedance of the parallel combination must be less than the smaller branch impedance, which would be R-419 in this case.

At some intermediate frequency the positive feedback voltage is a maximum, as shown in Figure 1-7. The curve is rather flat in the vicinity of f_r but the phase shift that occurs in the positive feedback circuit permits only a single frequency to be generated.

The voltage across R-419 is in phase with the output voltage of V-402 when the reactance of C-412 is equal in value to R-420 and the reactance of C-413 is equal in value to R-419. If the frequency of the output of V-402 increases, the voltage across R-419 lags the voltage at the plate of V-402. If the frequency decreases, the voltage across R-419 leads the output voltage at the plate of V-402.

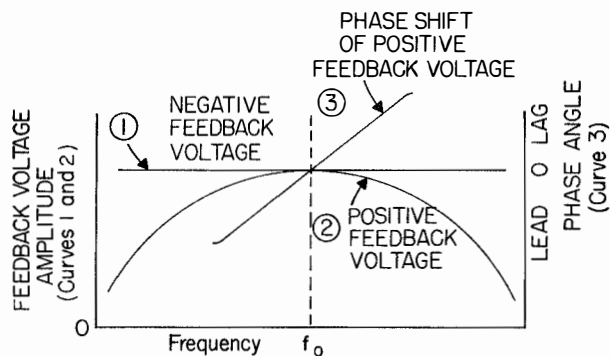


Figure 1-7. Feedback E vs F in Wein-Bridge Oscillator

The frequency at which the circuit oscillates is:

$$f_r = \frac{1}{2\pi\sqrt{(R-420)(C-412)(R-419)(C-413)}} \quad (1-3)$$

At this frequency the positive voltage on the grid V-401 just equals or barely exceeds the negative feedback voltage on the cathode and the positive feedback voltage is of the proper phase to sustain oscillation.

To change the frequency of oscillation from 400 cycles, to 1000 cycles, R-420 is paralleled by R-418, and R-419 is paralleled by R-417. This decreases the effective circuit resistance and it will require a higher frequency to reduce the reactance of C-412 and C-413 proportionally. The lamp I-400 is used as the cathode resistance of V-401 in order to stabilize the amplitude of oscillation. This is necessary if the Percent Modulation indication is to remain constant. If for some reason the amplitude of oscillation tends to increase, the current through the lamp will increase, the filament of the lamp becomes hotter making its resistance larger. A greater degenerative voltage is developed across the lamp, which reduces the gain of V-401 providing a nearly constant amplitude of output voltage. Since the waveform is sinusoidal only at a small amplitude of output from V-401, the lamp serves also to prevent distortion of the sinusoidal waveform of the output.

Should it become necessary to replace lamp I-400 use exact replacement. In some applica-

tions it has been found lamps by different manufacturers are sufficiently different in characteristics to upset the circuit. R-422 is provided to compensate for these changes and should be adjusted for maximum distortion-free oscillations. This controls the amount of degeneration. The Wein-bridge oscillator has many advantages over other types of audio oscillators. For instance, it may conveniently be made to produce a wide range of frequencies. The waveshape is very nearly a true sine wave. Frequency stability is excellent. Finally, the output amplitude is nearly constant over a very wide frequency range.

E. Power Supply

Plate voltage for the signal generator is obtained from full-wave rectifier V-406, Type 5U4G. Regulation of the high voltage output is obtained by amperite tube V-407, Type 7A20, in the primary circuit of power transformer T-461. This corrects for line voltage variations of 105 to 130v ac. Further regulation of the 150 volt lines is obtained by voltage regulators V-408 and V-409, Type 0D3/VR150. Resistor R-438 in the negative return provides a source of negative voltage for RF amplifier-modulator bias.

1-4. PROVISIONS FOR OBTAINING DESIRED FREQUENCIES

The CA-1510 signal generator has a non-linear VFO frequency scale on its tuning dial, actually a drum, which is coarsely divided into 0.1 mc graduations from 3.0 to 6.3 mc. Immediately adjacent to the frequency scale is a linear logging scale engraved on the same drum. The logging scale divisions on the tuning drum are effectively divided into 100 equal parts by reference to a tuning dial, the periphery of which is linearly divided into equal divisions and which is geared to the drum in such a ratio that one complete rotation of the dial moves the drum surface a distance equal to the length of one linear (logging) graduation. See Figure 1-8.

By interpolation, the desired dial reading for the VFO tuning drum for a given frequency can be determined by referring the frequency scale to the logging scale. VFO frequencies

may be set to within 1 kc at the 6 mc end of the scale and to within 600 cycles at the 3.0 mc end of the scale by properly interpolating the scale readings.

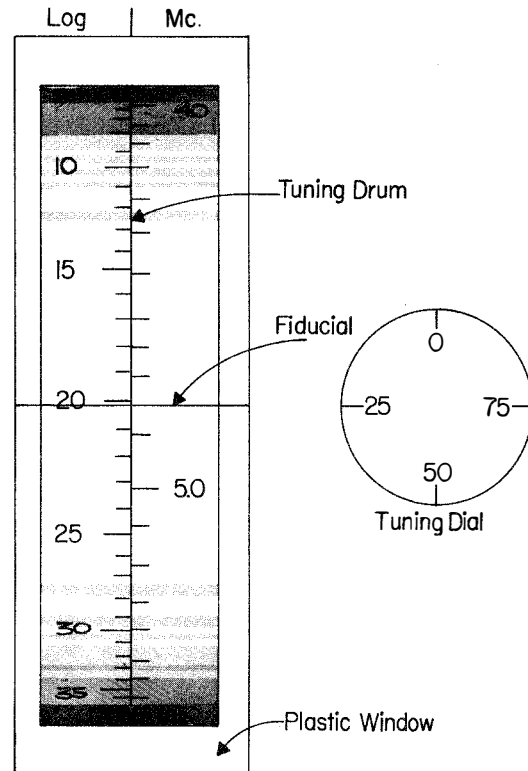


Figure 1-8. Tuning Dial, CA-1510 Signal Generator

A. Frequency Interpolation for 3.217 mc

It is desired to accurately determine the logging dial setting for a frequency of 3.217 mc. The nearest frequencies readable below and above the desired frequency are 3.2 mc and 3.3 mc, the frequency scale being divided into 0.1 mc increments. Suppose now that by setting 3.2 mc under the fiducial, the pointer line or hairline, it is found the logging dial reading is 467 and that when 3.3 is under the fiducial the logging dial reading is 656. By reference to Figure 1-9, expand the drum dial scale until the total length of the line as shown is representative of the distance from 3.2 to 3.3 mc. Let distance "a" equal the difference in frequency from 3.2 mc to 3.217 mc, distance "b" equal the difference in frequency from 3.2 mc to 3.3 mc, distance "c" equal the number of logging dial divisions

representing the frequency difference "a" and distance "d" equal the number of logging dial divisions representing the frequency difference "b". Then it should be quite apparent that $c/d = a/b$, or $c = ad/b$. By substituting the values indicated, where $a = 0.017$, $b = 0.100$ and $d = 656 - 467 = 189$, and solving, $c = (.017) (189) / (.100) = 32.2$ dial divisions. The dial reading for 3.217 mc then would be $467 + 32 = 499$.

B. Frequency Interpolation for 122.2 mc

Inasmuch as the oscillator of the generator operates in a frequency range of 3.0 to 6.25 mc, then 122.2 mc must be a harmonic of a frequency in the oscillator range. By proper setting of the frequency range switch on the generator, the 1st, 2nd, 4th, 8th, 16th and 32nd harmonics of the oscillator may be obtained. In this instance the desired frequency of 122.2 mc, if divided by 32, indicates a subharmonic frequency of 3.81875 mc. If the logging dial readings at the frequencies of 3.8 mc and 3.9 mc are 1405 and 1527 dial divisions respectively, by the interpolative process as given previously, the logging dial setting for the desired frequency is 1428 dial divisions. See Figure 1-10.

1-5. CAUTIONS TO BE OBSERVED IN USING THE CA-1510 GENERATOR

Calibrations of the VFO at 3.000 mc, 3.105 mc and 3.200 mc are valid only at frequencies close to the calibration frequencies. Do not calibrate the VFO on any of the above frequencies and expect calibration to be valid at a frequency of 3.400 mc, for instance.

Frequency indications on the tuning drum are not accurate at the indicated frequencies of 3.400 and 5.400 mc. Determination of the logging dial readings for these frequencies require special procedures.

Closely observe the setting of the TUNING control at the right hand side of the instrument panel with respect to the output frequency. False tuning may be obtained in many instances if the TUNING control is at scalar indications not commensurate with the placement of the output frequency within the frequency range being used.

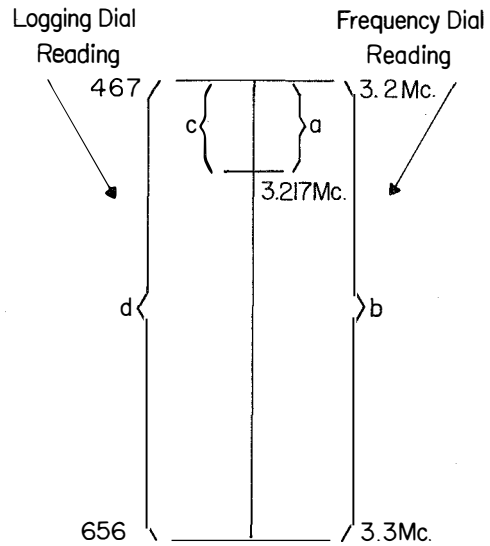


Figure 1-9. Tuning Dial, Sample Reading

Be certain that all switches are positioned to settings compatible with the mode of operation.

Note that the VHF frequency produced by the crystal E as indicated on the generator front panel is wrong. This frequency should be listed as 121.5 mc.

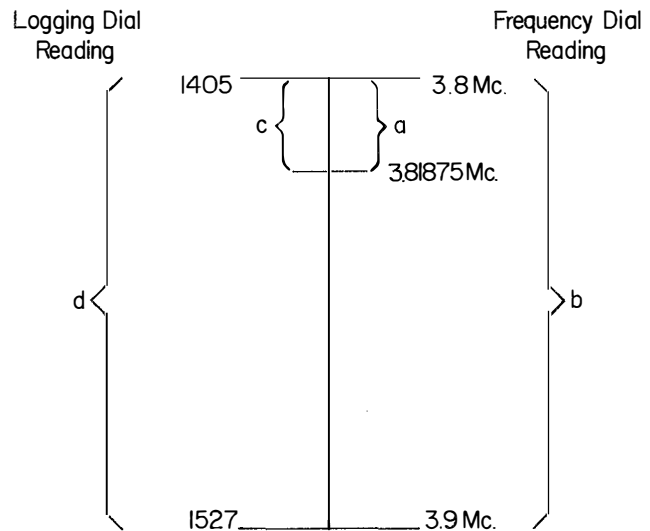


Figure 1-10. Tuning Dial, Sample Reading

1-6. SAMPLE PROBLEMS

Problem I: The instruction book for the CA-1510 signal generator indicates that frequencies between 0.080 mc and 2.7 mc may be obtained by beating the variable frequency oscillator against the crystal controlled oscillator with the crystal controlled oscillator operating on 3.000 mc. If a frequency of 455 kc (0.455 mc) were desired from the signal generator this desired frequency could be obtained by operating the VFO at 3.455 mc, beating this frequency against the crystal oscillator operating at 3.000 mc and using the 0.455 mc beat frequency thus obtained. ($3.455 - 3.000 \text{ mc} = 0.455 \text{ mc}$). If the logging dial readings for 3.400 mc and 3.500 mc are 826 and 976 dial divisions respectively, interpolate for the dial

settings needed to obtain the desired frequency. Correct answer is 908 dial divisions. Use Figure 1-11 and show computations.

Problem II: It is desired to determine the unknown frequency on which a VHF receiver is operating in the 100 to 150 mc range. The frequency range switch of the generator is set so that the generator is operating on the 32nd harmonic of the oscillator and the VFO is tuned through its tuning range until the receiver is receiving the generator output signal. The logging dial reading at this frequency is 1416. Referring back to the data given in section 1-4, **B** determine the VFO frequency and output signal frequency of the generator. Use Figure 1-12 and show computations. Answer is 121.9 mc.

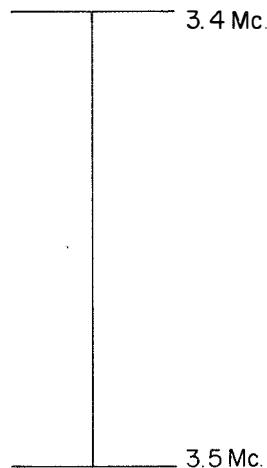


Figure 1-11. Tuning Dial, Sample Reading

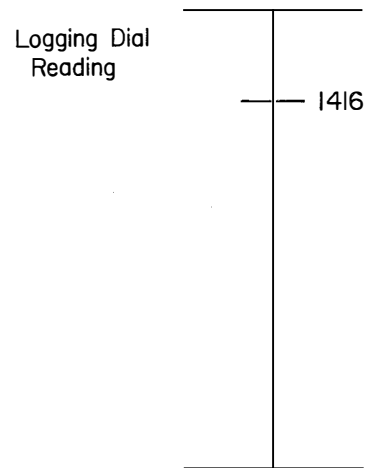


Figure 1-12. Tuning Dial, Sample Reading

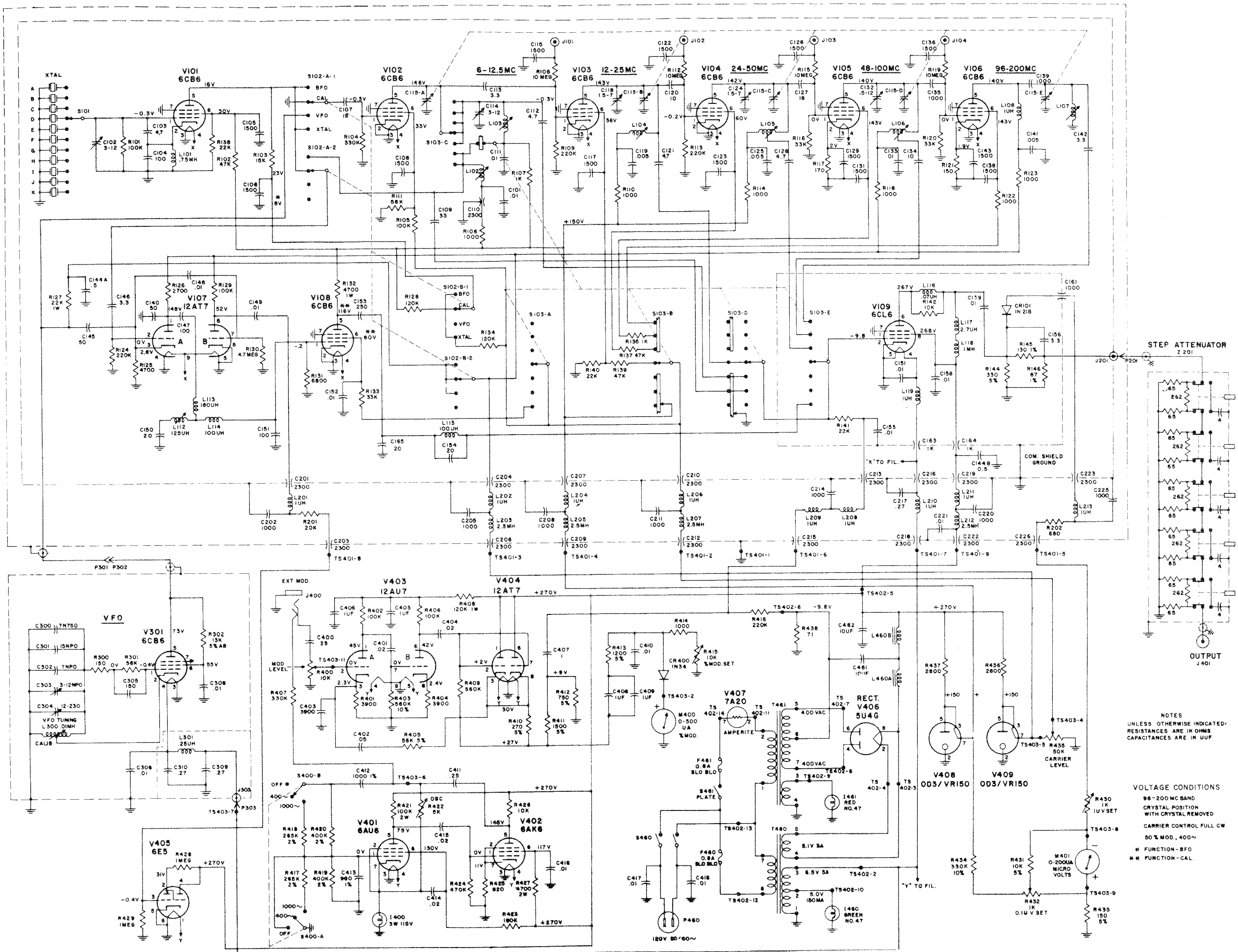


Figure 1-13. CA-1510 Signal Generator

CHAPTER 2

PERFORMANCE MEASUREMENTS

To enable the FAA to provide reliable facilities for Air/Ground Communications, a system of periodic routine maintenance procedures has been developed. It is desirable that the electronic technician receive adequate training in the two weeks of the Receiver Course to enable him to keep receiver failures to a minimum and to enable him to rapidly restore a receiver to normal operation in the event of failure. It is desired that receiving facilities provide reliable reception 100% of the total operating time. The electronic technician should not develop a defeatist attitude but should strive for the goal of 100% reliable operation. Many facilities more complex than a receiver are high in the 90% bracket for reliable operation.

Many students with a background in radio and television servicing find it difficult to accept the need for routine maintenance and performance measurements, but they are to be reminded these methods have been tested and have proven reliable time and again. New methods are developed and tested to keep abreast of improved equipment design.

The source of much of the information in this chapter will be *Maintenance of VHF and UHF Receiving Equipment Handbook AF P 6500.28 (MTC 3.3.1.3)*. Upon returning to his station, the student should become completely familiar with the contents of this Handbook, other Washington instructions, regional maintenance manuals and local Airways Technical Field Office instructions to insure that the various maintenance operational requirements are met.

2-1. HOUSEKEEPING

It is expected the electronic technician and student will employ such practices as are nec-

essary to insure that the receiving equipment, test equipment, and maintenance quarters of classroom will present at all times a clean, well-ordered appearance, reflecting the careful application of skillful workmanship.

In the laboratory the student is expected to employ these practices in his daily work. The working area should appear orderly. Equipment must be returned to normal storage areas upon completion of the laboratory period. Test equipment and tools should be used properly and not damaged unnecessarily. Equipment damaged due to carelessness will reflect on daily workmanship grades. Good habits of workmanship developed while in the laboratory will be an asset to every student when he returns to his station. By all means write on paper and *do not write on the tops of the laboratory tables*. Return equipment instruction books to the normal storage place as soon as you finish with a piece of equipment. The instruction books will be required by students who will be performing the experiment the next day.

2-2. ROUTINE MAINTENANCE

The need for routine maintenance has been discussed in the introduction. The desired frequency of routine maintenance may vary somewhat with location. The frequency of maintenance outline in this chapter was primarily extracted from Handbook AF P 6500.28 (MTC 3.3.1.3).

When any maintenance is being performed on a piece of electronic equipment, appropriate operations personnel should be advised.

A. Daily

When the technician responsible for the proper operation of the Air/Ground equipment

first comes on duty, it is advisable to check with operations personnel to determine whether there has been any outage or difficulty noted relative to the equipment operation. It is good practice to check the receiving equipment to determine if the receiver is operative by disabling the muting circuit momentarily and listening to the audio output of the receiver to determine if normal noise level is present. *Corrective action should be taken immediately if any abnormal condition is detected.*

Another method to check receiver operation is to use a spot frequency generator or a regular signal generator to momentarily produce a signal on each receive frequency and listen to the applicable receiver output for the modulating tone. *Keep in mind the source of signal should not be on longer than is necessary as it will produce possible interference to any aircraft signal on that particular frequency.*

B. Weekly

In addition to the daily items, the technician should perform operational checks on Air/Ground receivers in accordance with procedures outlined in Paragraph 6.10 of Handbook AF P 6500.28 (MTC 3.3.1.3), particularly noting the check on aural quality (by listening to voice transmissions). Indicator and dial lamps should be replaced as required.

C. Monthly

Service the receiver without removing from rack, if practicable. Clean dust from racks and receivers. Check operation of controls. Make over-all sensitivity test; if receiver sensitivity is out-of-tolerance, check tubes or otherwise correct as required. Measure and record *sensitivity, signal-to-noise ratio, squelch threshold and differential, and AVC threshold and level control*. Examine for noisy and frozen controls, loose knobs, etc.

Check dial calibration on tunable receivers; correct as required and post new settings if changed. Check aural quality by listening to voice transmission.

Examine associated antenna structures and supports. Remove and disassemble antennas for investigation if doubt exists concerning their

condition; clean insulators; replace cracked insulators; replace fatigue gaskets, lead washers and bushings as required. Make sure antennas are legibly marked for easy identification.

Check transmission lines for slackness, proper support and deterioration. See that transmission lines are marked as required. Test transmission lines for leakage with a megger. (It is also desirable to check transmission lines for continuity with an ohmmeter to detect any deterioration in shielding and to detect poor contact at fittings; however, this test is not specified in the Handbook.

Check horizontal wire antennas. Adjust sag as required. Check rustproofing of hardware and renew as required. See that turnbuckles are safety wired to prevent the turnbuckle from unscrewing during vibration.

Inspect antenna tuning houses. Clean as required. Check for security from entry of dust and moisture; particularly inspect gaskets for deterioration. Readjust tuning house circuits as required.

D. Quarterly

Although not specifically required, some stations follow the practice of a quarterly tube check of all receiving type tubes that are not checked on a more frequent basis.

NOTE: Indiscriminate tube checking is generally regarded as a highly questionable practice by the majority of technicians and engineers. Checking tubes will probably cause more trouble than it prevents, for two reasons: (1) An old tube that is working satisfactorily, even if low on emission, will be more reliable than a new tube. (2) Miniature and sub-miniature tube sockets do not stand up well when tubes are removed frequently.

The defects that would be most readily detected by a tube checker would be *low emission, shorted elements, open filament, tube noise, some intermittent conditions and greatly abnormal transconductance*. Not all defective tubes will be detected simply by testing in a tube checker. Tube substitution is commonly employed. If a questionable tube is removed from service, it should be destroyed. The new

tube should be tested in a tube checker before installing into the equipment.

Service the receiver on the bench. Clean thoroughly. Check gain controls, switches and fuses. Make sensitivity and other tests as required to assure optimum performance.

E. Annually

Clean greased surfaces and relubricate sparingly with white vaseline. Check electrolytic capacitors for leakage, reduced capacity, power factor and open circuit. Make such tests as are required to assure optimum performance, such as maximum power output, audio response, measure first oscillator frequency, and selectivity (bandwidth, symmetry, and channel frequency).

Remove and disassemble VHF antennas for thorough inspection and repair as required.

It should be remembered that this system of scheduled maintenance is a *minimum* requirement prescribed by the Handbook, and an increased schedule of maintenance may be desired to assure optimum performance of equipment. Due to climatic conditions, an adequate schedule in one area could easily prove inadequate in another area.

2-3. TEST EQUIPMENT REQUIREMENTS

It is realized that various types of test equipment will be found in the numerous field stations. Following is a list of test equipment that would be desired to assist the technician in maintaining the receiving equipment and the associated antennas.

- Receiver Test Set, CA-3452 or equal
- Signal Generator, CA-1510, HP-608D and HP-606A or equal
- Insulation Tester, DC, Biddle Model 7675-R or equal
- Bridge, Capacitance-Resistance, Sprague T0-4 or equal
- Tube Checker, Hickok Model 750 or equal
- Volt-ohm-milliammeter, Triplet 630NA or equal
- VTVM, AC-DC-RF, RCA WV-98A or equal
- Power Level Indicator, GR-583A or equal
- Headphones

Regardless of the type of test equipment available, any piece of test equipment needs proper care to provide the desired results. The technician is reminded that the pieces of test equipment are the tools of his trade.

A. Signal Generator

Signal generators used for the alignment and testing of Air/Ground communications receivers must provide an accurate source of signal frequency between 200 Kc and 400 Mc. Generally two separate signal generators are required to cover this frequency range. The *CA-1510 Signal Generator*, available at some stations, covers 80 Kc to 200 Mc. This is adequate for LF, MF, HF and VHF receivers. Generally a *Measurements Model 80 Signal Generator* is available at the same location to cover the range of 200-400 Mc. A more common combination of signal generators to cover the range of 200 Kc to 400 Mc is the *Hewlett Packard 606A* with a frequency range of 50 Kc to 65 Mc, and the *Hewlett Packard 608D* with a frequency range of 10-420 Mc.

It is further desired that the generator output frequency be stable. For that reason use a crystal-controlled signal generator whenever possible. If it isn't possible to use a crystal-controlled signal generator, be sure the tunable signal generator has been allowed sufficient time to reach normal operating temperature before calibrating. The output frequency should be accurately established by using a crystal check point, which is available on some signal generators like the HP 608D, or by using a BC-221 or Gertsch Frequency Meter.

Some very conscientious technicians may be observed using a VFO type signal generator for conducting monthly performance measurements when a crystal-controlled type signal generator sits nearby. Unless an accurate source of signal frequency is used while making receiver performance measurements, *what is to insure that the receiver is operating on the desired frequency?* On a periodic basis transmitter crystals and crystals in signal generators are checked to insure the generated frequency meets the tolerance specified. Each time a crystal-controlled signal generator is connected to a receiver for purposes of measuring the re-

ceiver sensitivity, the receiver frequency is also checked. If one of the crystals in the receiver changes frequency, it would be reflected in lowered sensitivity for the receiver. This would not be the case if a VFO signal generator was connected to the receiver input and the VFO was adjusted to provide maximum receiver output.

Another signal generator requirement is low stray radiation. For radio servicing applications a particular signal generator may have been acceptable, but its stray radiation may become a problem when checking the sensitivity of the high-gain communications receivers.

The indicated output of the signal generator must be accurately adjustable to the desired range of output levels. Generally $0.1 \mu\text{v}$ to $100,000 \mu\text{v}$ levels are desired. Most signal generators are calibrated to deliver a given output into a specific load. To insure that the specified load is provided irrespective of the receiver input impedance, a dummy antenna is used. This dummy antenna is a 68-ohm non-inductive resistor for high frequency receivers having nominal input impedances below 150 ohms, or a 200-ohm noninductive resistor for receivers having nominal input impedances above 150 ohms. When using the CA-1510 Signal Generator or similar signal generators having a 50-ohm output impedance with a receiver having a 50-ohm input impedance, no dummy antenna is required. This would be a matched condition. To insure that as near perfect an impedance match as possible is maintained at all times irrespective of the receiver input impedance, a 6-db pad is commonly used. In the Receiver Laboratory the 6-db pad will be connected to the load end of the transmission line. Some signal generator instruction books may indicate the 6-db pad should be connected at the signal generator end of the interconnecting transmission line; in that case, it is suggested you follow the prevailing instructions.

Another signal generator requirement is provision for modulating the output RF energy with a desired frequency and at the desired level. The percentage of distortion should be at a minimum. It is suggested the accuracy of the modulation level indicator on the signal

generator be checked periodically to insure the desired level of modulation is being obtained.

When in use, the signal generator case should be bonded to the receiver chassis, and both generator and receiver should be connected to a common ground. If for no other reason, this common ground is a good safety precaution to prevent shock when touching the two pieces of equipment simultaneously.

B. Power Level Indicator

There are various power level indicators, but the two types that will be used in the Receiver Laboratory are the *General Radio 583-A* and the *Daven OP-182* Power Output Meters. Figure 2-1 shows a front view of the Daven Power Output Meter and Fig. 2-2 is a schematic of the same power level indicator.

The output meters used in this course are direct reading audio frequency power meters. They are designed to measure the power output of the various receivers, audio amplifiers, filters and similar equipment over the entire audio frequency range. Various receiver tests such as noise levels, bandwidth, selectivity, fidelity, sensitivity and signal-to-noise ratio may be readily made using the decibel scale of this instrument. The circuit of the output meter consists of a load impedance adjusting network, a meter and a meter multiplier control which are designated 100, 113 and 106 on the schematic. The load impedance adjusting network provides for loads varying in steps from 2.5 to 20,000 ohms. The basic meter covers the range from 1 to 50 milliwatts and from 0 to 17 db using a one milliwatt reference. The meter multiplier extends the power range by ratios of 0.1, 1, 10 and 100 to 1. The meter multiplier also extends the db reading by -10, 0, +10 and +20 db.

Refer to Fig. 2-2 for the discussion of the circuitry involved. The input is shunted by a 40-section noninductive resistance, 101, controlled by a switch, 100. Up to the 300-ohm position, the resistance is bridged by a high-impedance coil, 102. The switch also controls the tap setting of the high-impedance coil; therefore, the power dissipated in the resistance is directly proportional to the voltage applied to the meter by the high-impedance coil.

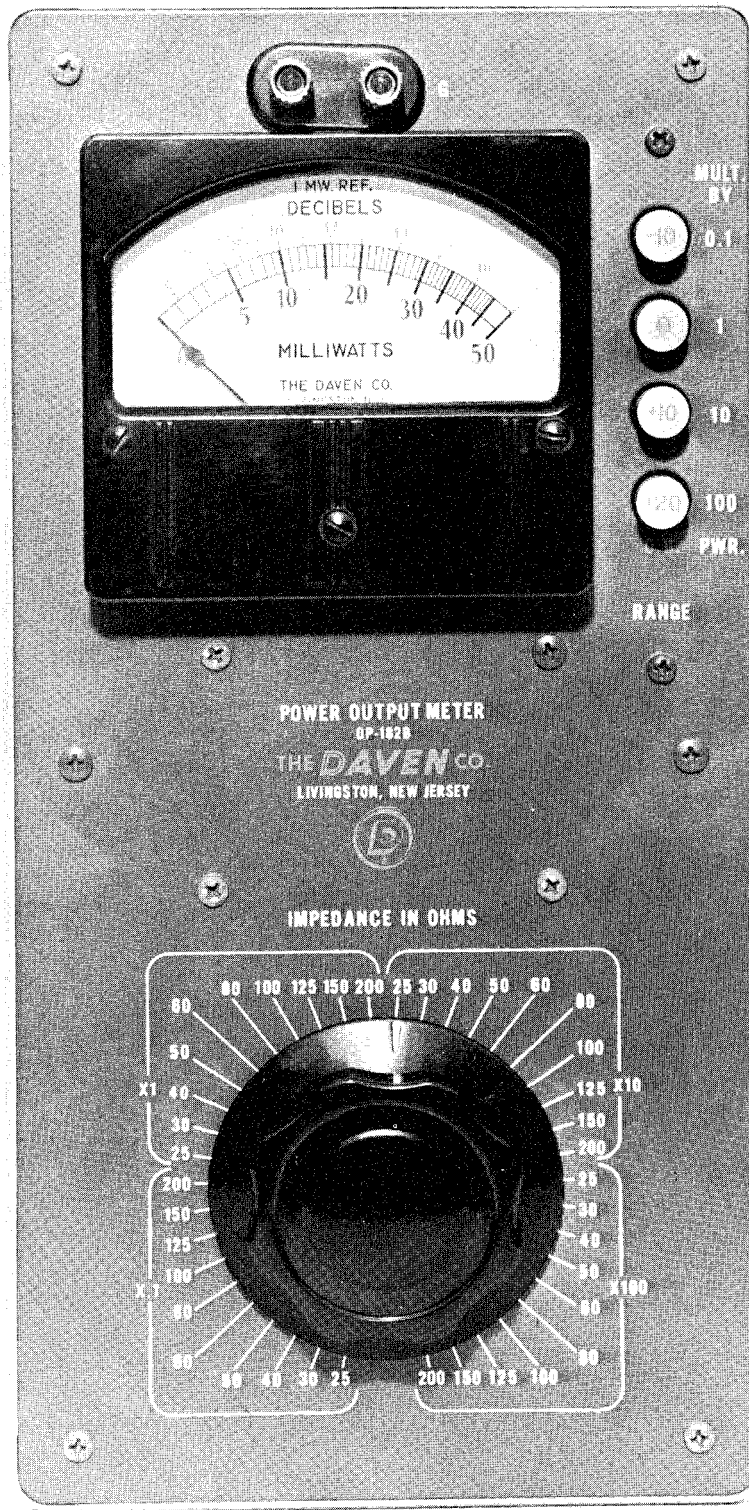


FIGURE 2-1. OP-182B Power Output Meter, Front View.

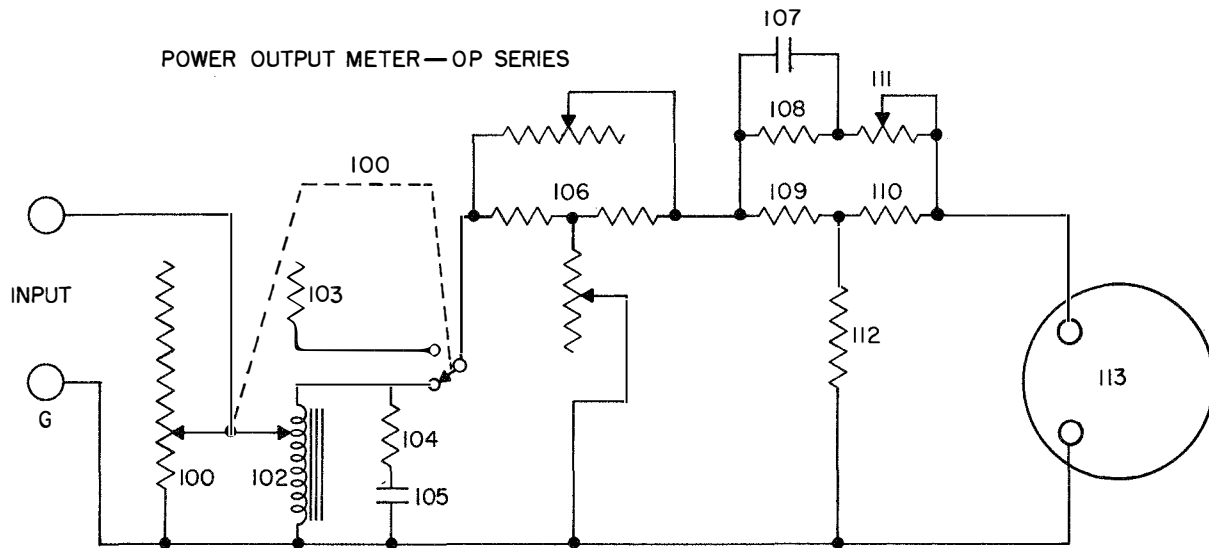


FIGURE 2-2. OP-182B Power Output Meter, Schematic.

To insure that the input impedance remains essentially noninductive, the impedance value of the coil is quite high compared to the resistance of 101.

Above the 300-ohm position of the switch, the coil is disconnected from the circuit. This resistance controls the voltage appearing across the meter. The tap settings on the resistance and the action of switching from the coil to the resistance are controlled by switch 100.

The RC network consisting of resistor 104 and capacitor 105 is shunted across the coil to compensate for losses in the coil at low frequencies.

The indicating meter is a rectifier type ac voltmeter calibrated in milliwatts and decibels. The meter multiplier control, 106, is a "T" type network calibrated in power ratio and decibels. This control is used to extend the range of the meter. Resistors 108, 109, 110 and 112; a rheostat, 111; and a capacitor, 107, form a combination calibration and high-frequency-compensation network.

When using the power output meter, set the meter multiplier control at maximum position, set the impedance control at the position corresponding to the output impedance of the device under test and connect the output of the receiver or other electronic device to the input of the power output meter. With the receiver in operation, adjust the meter multi-

plier control to provide an indication on the meter as near center scale as possible. Multiply the meter scale indication by the meter multiplier power ratio setting to obtain the output in milliwatts of the equipment under test.

When using the power output meter to indicate audio levels in db, the db in reference to a level of 1 milliwatt is obtained by taking the meter scale indication in db plus or minus the db setting of the meter multiplier control.

There are various precautions to be observed when using the power output meter. The meter multiplier control should be set to maximum for all initial measurements to prevent damage to the meter. It is easier to reduce the meter-multiplier-control setting to a level to provide the desired indication than it is to repair a meter that has been damaged from overload. Also, keep the hands clear of the input terminals of the meter; voltages in excess of 100 volts may be present between the two input terminals.

In the Receiver Laboratory the student will be required to calculate all db levels even though the db levels could be obtained directly from the power output meter. This practice will insure the student can calculate db if necessary, and by comparing the calculated value with the value indicated on the power output meter the student should learn how to read the power output meter accurately.

Although the previous discussion pertained to the Daven OP-182 Power Output Meter, the operational characteristics of the General Radio 583-A Output Meter is similar with the exception that two impedance controls are employed instead of one.

C. Volt-Ohm-Milliammeter

Perhaps two of the most common volt-ohm-milliammeters used in the field and in the various laboratories of the FAA Academy are the Triplet 630 and the Simpson 260. Their circuitry is standard for this type of meter; therefore, the details of the circuitry will not be discussed in any detail.

The scales of the two meters are not the same. If the markings on the scales are not self-explanatory, ask an instructor for clarification; if in the field, refer to the applicable instruction book for the instrument.

Care should be taken of the instrument at all times. When not in use, the instrument should be stored in a protected location to prevent damage. Periodically, on a routine basis, the batteries in the meter should be tested to insure that the various scales of the ohmmeter portion will zero properly. One of the most common over-sights in the proper maintenance of test equipment is the neglect of the batteries in various volt-ohm-milliammeters.

The Triplet 630 meter has an "OFF" position on the selector switch. When the meter is momentarily unemployed for making measurements, turn the meter selector switch to the OFF position. This will reduce the chances of damaging the meter when it is connected to various circuits. Also, if the value of the voltage under test is not known, start out with the highest scale possible and reduce the selection until the desired range of indication is obtained. Although the Simpson 260 does not have an OFF position on the selector switch, the meter will be provided reasonable protection if the meter selector switch is returned to the 1000 v dc position when not in use.

There is at least one thing worse than burning out a meter and that is burning out a meter or otherwise damaging it and not initiating corrective action to return the meter to normal operating condition. The next time the meter

is needed may be an emergency situation. The equipment restoration may be unnecessarily delayed due to negligence in getting the meter repaired. If the limitations of the meter have been greatly exceeded and yet the meter is not burned out, it may still be desirable to have the meter recalibrated to insure correct indication.

Most meters of this type employ rectifiers when indicating levels of ac voltage. The meter is calibrated on the assumption the ac voltages being measured are sinusoidal in nature. When the voltage is not sinusoidal in nature, as would be the case where voltage regulators are installed between the commercial power source and the equipment, the meter indication would be erroneous.

It is commonly known that the rms value of a given sinusoidal waveform may be calculated from E_{max} divided by the square root of 2; however, a particular sawtooth waveform would require E_{max} divided by the square root of 3 to calculate the rms value. This fact would require the recalibration of the meter if it were to accurately reflect the rms value of a complex waveform.

If it is desired to measure the ac voltage applied to the filaments of a tube that is provided *regulated* ac power, use an iron-vane voltmeter similar to a Weston 528. An electro-dynamometer type instrument may also be used.

It should be kept in mind that these meter circuits are sometimes very frequency conscious and that a multimeter should be calibrated for the frequency of the voltage to be measured if reliable voltage values are to be obtained.

Another precaution to be observed when using the volt-ohm-milliammeter for measuring operating circuit voltages is the shunting effect produced by the meter on the circuit being tested. The input impedance of the volt-ohm-milliammeter is 20,000-ohms per volt when making dc measurements. This comes from the fact a 50 microampere meter movement is employed and 20,000 ohms of resistance must be added in series with the meter movement for every volt in the circuit being measured to prevent the meter indication from exceeding full scale. Should the meter be on the 2.5 v dc

scale to measure the voltage across a 100,000 ohm resistance, the shunting effect of the meter would be equal to $2.5 \times 20,000$ ohms/v or 50,000 ohms. The resultant effective resistance would be 33,000 ohms which would be quite a departure from the normal value of 100,000 ohms. Should the meter be used on the 1000 v dc scale, the shunting resistance from the meter would be 20 megohms. It can be seen that this value of shunt resistance would not affect a 100,000-ohm load appreciably, but difficulty would be encountered when reading 2.5 volts on the 1000 volt scale.

If the Triplet 630 meter is used for ac voltage measurement, it is a *5000-ohm-per-volt meter*. If the Simpson 260 meter is used for ac voltage measurement, it is a *1000-ohm-per-volt meter*. It can be seen the shunting effect of either meter is greatly increased when switched to ac operation from dc operation. This is due to the fact a different bank of resistors are used on ac than on dc. Also, the dry-disk rectifier is used when on ac.

When it is desired to measure AVC voltage or oscillator bias, the shunting effect of this type meter would be most undesirable. The use of a vacuum tube voltmeter would overcome this deficiency.

D. Vacuum Tube Voltmeter

Various types of vacuum tube voltmeters, (VTVM), are used in the Receiver Laboratory. These VTVM's are typical of the meters found at the various field stations. The VTVM will be used when measuring oscillator bias, frequency multiplier bias, AVC and other voltages where the shunting effect of the volt-ohm-milliammeter would result in erroneous indications.

1. WESTON MODEL 976

The Weston Model 976 Electronic Multimeter is one of the instruments available at the various field stations; however, they may be assigned to one of the navigational aid sites.

Although this instrument has an input impedance of 15 megohms when used to make dc voltage measurements, there is no isolation resistor installed in the test leads used with the

meter. When measuring oscillator bias, the added capacitance and inductance of this test lead frequently detunes the oscillator stages; for that reason, an isolation resistor has been added in the probe end of the red lead normally used with this type meter. Usually, a one-megohm resistor is used for isolation, but the desired results were not obtained in the Receiver Laboratory with the Weston Model 976 until a 10-megohm resistor was used. Only 60% of the actual voltage is developed across the input impedance of the meter as 40% is developed across the 10-megohm resistor. This results in a great error for the meter indication; however, if only relative values are desired for indicating a resonant circuit condition, the use of the 10-megohm isolation resistor is permissible.

The input impedance of this meter for ac measurements is 5 megohms and 150 pf of capacity.

2. SYLVANIA POLYMER TYPE 221Z

The Sylvania Polymer has an input impedance of 16 megohms when used for measuring dc voltages. When the dc probe is used, it contains a one-megohm isolation resistor. Little difficulty should be encountered when using this meter provided the dc probe is used for dc voltage measurements.

When the meter with a shielded lead is used for ac voltage measurements, the input impedance is 2.7 megohms and 125 pf of capacity.

3. RCA VOLTOHMYST TYPE WV-98A

The RCA Voltohmyst is a more compact unit than either of the types previously mentioned. The DC/AC-Ohms probe should be used with this instrument. When in the DC position, a built-in one-megohm isolation resistor is placed in series with the probe tip. The switch should be in the AC-Ohms position for making ac measurements.

The input impedance of the instrument is 11 megohms for dc voltage measurements. For ac measurements the input impedance varies from 0.83 megohms and 70 pf of capacity to 1.5 megohms and 60 pf of capacity. This

change in impedance occurs with a change in voltage-range settings.

The technician should bear in mind that most VTVM's measure *peak* ac voltages which are translated by the meter scales to rms values.

Some VTVM's have a peak-to-peak voltage scale for the convenience of the user.

The preceding discussion of test instruments should call to the technician's mind the shortcomings and advantages of various types of instruments and indicate to him when, and under what conditions, specific instruments should be used.

E. Miscellaneous Test Equipment

Although various additional test equipment may be used to good advantage when servicing receivers, no effort will be made to discuss each type individually. The oscilloscope is adequately covered in the Test Equipment course. It is reasonable to assume the average technician can use a tube tester properly.

A useful item that is frequently overlooked in receiver servicing is a pair of headphones. The technician is encouraged to use headphones freely as they readily indicate interfering signals, false resonance points and audio distortion.

2-4. CONDUCTING PERFORMANCE MEASUREMENTS

The various performance measurements will now be described in detail. For the tests that follow a *Standard Test Voltage* (also called *Standard Signal*) will be used unless otherwise specified. A Standard Test Voltage (abbreviated STV) is defined in Receiving Equipment Handbook AF P 6500.28 as "an RF signal, amplitude modulated 30% by 400 cps". Since military communications equipment commonly uses 400 cps for the power supply, 400 cps would be undesirable as a standard modulating frequency; therefore, for military type equipment 1000 cps is generally used.

At no time should equipment be allowed to operate just at the limits of the specified tolerance. The technician should be familiar with the receiver capabilities outlined in the applicable Handbook. If the receiver is maintained near these optimum capabilities, the specified tolerance will be easily exceeded.

By comparing receiver performance with past performance recorded in the Facility Record Book, any pattern of general receiver deterioration should be readily detected.

A. Over-all Sensitivity

A measurement of receiver sensitivity will determine whether or not the receiver gain is near optimum. Many of the FAA communications receivers have an over-all gain in excess of 100 db. Any deterioration due to low tube emission, low operating voltages or similar defects will be reflected in lower gain or poorer sensitivity.

1. GENERAL CONDITIONS OF TEST

For fixed-tuned receivers and for tunable receivers normally used to guard a single frequency, the sensitivity will be measured at the normal operating frequency. For "spare" tunable receivers the measurement will be made at the middle dial setting of each band. Receivers having both crystal and VFO control will be measured under both conditions. All tunable receivers used for general coverage should be checked at all band ends to assure normal response throughout the tuning range. The AVC feature will be disabled where this can be done by a switch or by grounding the AVC line without affecting the other circuits; otherwise AVC will be left in operation. If the receiver has a switch for the purpose, the squelch will be disabled. Crystal filter, if any, will be turned off. The tone control setting, where this is provided, will be that which least attenuates the audio signal. Receiver output power will be measured across the 20,000 ohm output terminals unless the receiver has no such output in which case the measurement will be made across the receiver output terminals in use. In either case the receiver output will be loaded by an equivalent resistance in the power level indicator or the receiver test set.

Prior to the sensitivity measurements, the receiver input circuit trimmer, if one is included, will be adjusted for maximum sensitivity to the signal generator at the test frequency.

With the signal generator connected to the receiver input through the dummy load, where required, the generator output will be adjusted

to produce 0.5 watt at the receiver output. Precise receiver and signal generator resonance must be maintained.

Adjust the signal generator output for a standard test voltage (STV) input to the receiver as specified for the particular receiver under test; this is 5 μv STV for all FAA VHF communications receivers. Reduce the AF gain control for a power output as specified under sensitivity requirements for the particular receiver under test; 0.5 watt for all FAA VHF communications receivers. Next, remove modulation from the signal generator and note receiver output level. Without further change to the AF gain control, vary the output level of the signal generator until a power ratio of 10:1 (10 db) is obtained when the modulation is turned alternately on and off. The sensitivity of the receiver is the standard test voltage level required to obtain a signal plus-noise (modulation on) to noise alone (modulation off) ratio of 10:1 (10 db). Note that this is a power ratio; it corresponds to a voltage ratio of 3.17:1 (10 db). The *operating tolerance* for a sensitivity measurement on all FAA VHF communications receivers is stated in Handbook AF P 6500.28 as "Not more than 5.0 μv STV to produce 0.5 watt output into a 20,000 ohm load with 10 db SNR".

Most receivers' sensitivity is dependent upon the value of line voltage provided the receiver. It is suggested the line voltage be measured each time the receiver sensitivity is measured. If the line voltage departs appreciably from 115 v ac, the receiver sensitivity can be expected to vary accordingly. See Fig. 2-3. If the line voltage is consistently too high or too low, arrangements should be made with the local power company to have the voltage error corrected.

Poor receiver performance is generally first noticed by poor receiver sensitivity; more than 5 microvolts required to produce 0.5 watt output. Poor receiver sensitivity is perhaps the most common defect noted on inspection reports covering receiving equipment.

B. Signal-to-Noise Ratio

This measurement would be more accurately stated as signal-plus-noise-to-noise ratio. The

measurements that will be conducted will use the rms value of signal and noise.

Several years ago sensitivity measurements were considered quite indicative of the receiver's ability to respond to a weak signal. This was true because most communications were in the low frequency and high frequency spectrums. At these frequencies most noise is external to the receiver due to atmospheric or man-made noises. Today most communications are in the VHF or UHF spectrums. At these frequencies thermal noise and other noises generated internally are the greatest source of receiver noise. Although the receiver may have sufficient gain (sensitivity) to detect a weak signal, the internal noise of the receiver may be too great to hear the signal above the noise. For that reason the receiver's signal-to-noise ratio generally expressed in db is becoming more important as an indicator of receiver performance. It is not necessary to measure the signal-to-noise ratio of the receiver as a separate check if it meets the tolerance for sensitivity, because the method of measuring sensitivity is to determine the RF voltage required to obtain a predetermined AF output at a specified signal-to-noise ratio; therefore, an increase in noise level will be detected in the sensitivity measurement. A signal-to-noise ratio measurement has been included for use when it is necessary for any purpose to determine this ratio.

1. CONDITIONS OF TEST

The equipment arrangements and receiver control settings are as specified for the over-all sensitivity measurement. The signal generator output is adjusted to the maximum STV for the receiver as specified by the sensitivity tolerance for the particular receiver under test. The receiver AF gain control is adjusted to give 0.5 watts output from the receiver. Modulation is then removed from the signal generator and the output level of the receiver is measured in watts. The signal-to-noise ratio is then computed from the formula:

$$\text{db} = 10 \log_{10} \frac{P_2}{P_1} \quad (2-1)$$

A signal-to-noise ratio of 16 db or greater is considered acceptable for the RCP, RCQ and RHZ receivers. A signal-to-noise ratio of 10 db or greater is considered acceptable for the RUQ, RVA and all other FAA VHF receivers. For all other receivers refer to the receiver instruction book. Actually the RUQ, RVA and other FAA VHF receivers generally have a signal-to-noise ratio between 13 db and 16 db when properly maintained.

In the Receiver Laboratory the student will be required to calculate the signal-to-noise ratio in db. All work must be shown. The indications on the power level indicator may be used to check the calculated values for accuracy. This calculation is required because there may be occasions, when the power level indicator is not available, that a resistance equal to the receiver output impedance and an ac voltmeter must be used for conducting performance measurements. Also, some students may not know how to read a power level indicator properly and the ability to calculate the db change will prove a valuable check for accuracy whenever there is doubt.

C. CONS Threshold and Differential

The Carrier Operated Noise Suppression (abbreviated CONS) circuit is commonly referred to as Squelch or Muting. The purpose of any similar circuit designations is to hold the receiver silent, or nearly so, during no-signal conditions. Since it is desired that the receiver respond to even very weak signals, the purpose of this test is to determine the signal level required to render the receiver operative for reception.

1. CONDITIONS OF TEST

The power level indicator should be set to properly terminate the receiver, generally 20,000 ohms; the AVC should be operative; the CW oscillator and crystal filter should be OFF; the AF gain control should be fully advanced. The signal generator controls should be set to provide a Standard Test Voltage at the receiver's operating frequency; however, the RF output of the signal generator should be zero.

With the signal generator properly connected to the receiver input, turn squelch switch ON while observing the noise output of the receiver. If the squelch does not operate to quiet the receiver, reduce the RF gain control to the point at which it just operates. Now gradually increase the signal generator output from zero and note the STV required to deactivate the squelch. This RF output recorded in microvolts is the CONS threshold. *Caution should be used to prevent overloading the power level meter because the output changes from a value near zero to possibly several hundred milliwatts at the instant the receiver becomes operative for reception.* The CONS shall open (cease to mute the audio output) at an RF input of not more than 2 microvolts STV for FAA VHF communications receivers.

Reduce the signal generator output and note the STV at which the squelch again operates. The ratio of this voltage to the voltage required to open the CONS is the squelch differential. This should be approximately a one-third decrease of the voltage required to open the squelch.

If the external and/or internal noise level is high enough to require the RF gain control to be retarded to close the squelch (with no applied signal), it should be remembered that the receiver sensitivity is also reduced. It is suggested the receiver sensitivity be re-checked at this reduced setting of the RF gain control to insure the receiver is still operating within the specified tolerance. Before making any such reduction in RF gain control setting, it is imperative the technician determine that the receiver is operating normally in all other respects and that there are no defective receiver components.

D. AVC Action

The purpose of the automatic volume control circuit is to maintain the audio output of the receiver at a nearly constant level while the RF signal input varies between wide limits. This feature assures that the output will remain near the level established when the operator initially adjusted the manual audio gain controls. The

audio gain controls available to the operator are generally controls other than the AF gain control on the receiver. The AF gain control on the receiver is generally operated at the fully advanced position, full gain.

1. CONDITIONS OF AVC THRESHOLD TEST

The receiver AF and RF gain controls should be fully advanced; the CONS and CW oscillator should be OFF; AVC should be operative; the tone control should be set in the position giving minimum attenuation; and the crystal filter should be properly terminated into a 20,000-ohm setting of the power level indicator. The signal generator should be properly connected to the receiver input, tuned to the proper frequency and adjusted to provide a Standard Test Voltage at zero microvolts.

Increase the signal generator output while observing the audio output level for the point at which AVC throttling action begins. This is the point at which a linear increase in input signal no longer produces a linear increase in output signal level. The AVC threshold is the STV input at which AVC threshold occurs. For FAA VHF communications receivers the AVC threshold should occur between 0.5 and 8.0 microvolts STV.

2. CONDITIONS OF AVC LEVEL CONTROL TEST

Adjust the signal generator output to provide 50 microvolts STV at the receiver input. Reduce the receiver AF gain control to produce 50 milliwatts audio output. Retarding the AF gain control is necessary to insure that any leveling off of the receiver audio output is due to AVC and not due to over-driving the audio amplifier section of the receiver.

Leave all receiver controls unchanged and adjust the signal generator to provide receiver inputs of 7.5, 500, 5000, 50,000 and 500,000 microvolts STV, noting the corresponding receiver output level for each step. This procedure is applicable to all FAA VHF communication type receivers.

Determine the ratio of maximum to minimum receiver power output expressed in db. Output power of FAA VHF communication re-

ceivers shall not vary more than 4.5 db for an input signal variation of 7.5 microvolts to 0.5 volts STV.

E. Maximum Power Output

The purpose of this test is to determine if the receiver is capable of delivering the maximum audio power output for which it was designed.

1. CONDITIONS OF TEST (MAXIMUM POWER OUTPUT)

The signal generator should be connected to the receiver input and should be tuned to provide a standard test voltage at the frequency of operation for fixed-frequency receivers. In the case of standby tunable receivers, the measurement is to be performed at the mid-dial setting of the most used tuning band. The receiver output should be terminated into the proper value of load impedance. The receiver CONS and CW oscillator should be OFF, the AVC circuit should be operative; the tone control should be set to give minimum attenuation; the crystal filter should be OFF, the RF and AF gain controls should be fully advanced.

While maintaining precise resonance between the receiver and signal generator, adjust the signal generator attenuator to the position giving maximum audio output as indicated on the power output meter. Record the maximum power output obtained and the receiver input required to obtain this audio output. The maximum power output may occur at receiver input levels as high as 100,000 microvolts.

If an AVC-MVC switch or similar means of disabling the AVC is provided and the receiver performance indicates a need for such a test, the AVC may be disabled by turning the AVC-MVC switch to the MVC position and the maximum power output of the receiver measured under these conditions. The test is conducted in the same manner as though the AVC were in operation.

For all FAA VHF receivers the maximum audio output shall not be less than 0.8 watts.

NOTE: Some receivers may produce an output in excess of 5.0 watts; care should be taken to assure that maximum capabilities of the power output meter are not exceeded.

2. CONDITIONS OF TEST (MAXIMUM UNDISTORTED POWER OUTPUT)

Additionally, it may be desirable to measure the Maximum Undistorted Power Output of the receiver. This may be accomplished in conjunction with the preceding Maximum Power Output performance measurement, by connecting an oscilloscope across the receiver output terminals and observing the output waveform while adjusting the attenuator of the signal generator until the audio pattern begins to distort, or until maximum receiver output is reached, whichever occurs first. At this point the receiver output level is measured and this level in watts is the receiver's maximum undistorted power output. Some receiver types may show distortion on the oscilloscope at all output levels. Listening to a modulated signal will indicate whether the distortion is significant, and whether it will be objectionable in normal receiver use.

There are no tolerances specified for this test; however, the results of this test should be very close to those specified for maximum power output.

F. Selectivity

Good receiver selectivity is the ability of the receiver to reject signals close to the frequency of the desired signal. It is impractical to cascade several stages of variable frequency (tunable) amplifiers. The bandwidth is narrower for a given Q in a low-frequency than in a high-frequency amplifier (bandwidth varies directly with changes in the resonant frequency and inversely with changes in the circuit element Q , $BW = f_r/Q$); therefore the bandwidth and selectivity characteristics of the receiver are usually established by the lowest frequency intermediate frequency amplifier. A typical exception would be the RIU receiver which is a TRF receiver and has no intermediate frequency stages, as the term intermediate frequency is normally applied. In this case the fixed-tuned RF amplifiers determine the selectivity. Since the RIU operating frequencies are in the 200 to 400 Kc range, good selectivity is obtained by the several fixed tuned RF stages.

Closely related to selectivity are considerations of bandwidth, symmetry, and channel frequency. Too broad a bandwidth and off-frequency operations are caused by component failure, improper alignment, or detuning by tube replacement and is usually reflected in reduced receiver sensitivity at the receiver channel frequency. A too-narrow bandwidth is usually caused by improper alignment and is generally reflected in missed calls from aircraft.

1. CONDITIONS OF TEST

The following procedure closely parallels the recommended procedure in Handbook AF P 6500.28, which specifies injecting the signal generator output at the receiver's operating (channel) frequency. Since the selectivity characteristics of the IF amplifier determine the over-all selectivity of most superheterodyne receivers, the technician could, as an alternative procedure, perform the selectivity measurement by injecting the intermediate frequency at the mixer and following essentially the same procedure as outlined for the operating frequency. The results of these two methods should be nearly identical.

2. CONDITIONS OF TEST (SELECTIVITY MEASUREMENTS)

With the signal generator tuned exactly to the channel frequency of the receiver by using the Gertsch, or equal, frequency meter, turn off the modulation and adjust the output level of the signal generator until 1.0 volt of AVC is obtained at the AVC jack. The signal input level to the receiver is the reference signal for the selectivity measurements that follow.

Increase the output of the signal generator 6 db above the reference level (twice the voltage). Tune the signal generator above and below the channel frequency until the reference level of 1.0 volt AVC is again obtained, carefully measuring the frequency in each case with the frequency meter. Compute the bandwidth at the 6 db points by taking the difference of the two frequencies. Compute the percentage of non-symmetry by using the following formula (or an approved simplification):

$$\% \text{ Non symmetry} = \frac{2[(f_0 - f_1) - (f_2 - f_0)]}{(f_0 - f_1) + (f_2 - f_0)} \times 100$$

Where: f_0 = Channel frequency

f_1 = Frequency below f_0
for 6 db down

f_2 = Frequency above f_0
for 6 db down

The percentage of non-symmetry for an operating FAA VHF communications receiver must not exceed 20%.

Increase the output of the signal generator 60 db above the reference level (1000 times the voltage). Tune the signal generator above and below the channel frequency until the reference level of 1.0 volt AVC is again obtained, carefully measuring the frequency again with the frequency meter. Compute the bandwidth at the 60 db points by taking the difference of the two frequencies. Compute the percentage of non-symmetry by the same method as for the 6 db procedure.

Operating tolerances of the selectivity characteristics for FAA VHF Fixed Tuned Receivers:

	VHF BROAD BAND	VHF NARROW BAND
6 db Bandwidth	45 Kc Minimum	34 Kc Minimum
60 db Bandwidth	220 Kc Maximum	118 Kc Maximum

Although receiver selectivity measurements generally are not required as frequently as some other performance measurements, the selectivity should be measured whenever the intermediate frequency stages are realigned or whenever problems are encountered in receiving calls from aircraft. With the increasing congestion of the VHF spectrum, and the advent of channel-splitting, receiver selectivity is becoming increasingly important. The mere fact that selectivity measurements are not presently being conducted at a particular station is no assurance that they may not be required in the future.

G. Audio Frequency Response

It is essential that the receiver be capable of amplifying frequencies in the range of 300 to 3,000 cps without appreciable distortion in order to provide satisfactory voice quality for communications purposes.

1. CONDITIONS OF TEST

Connect a tunable audio oscillator to the RF signal generator external modulation jack. Set the signal generator for 50 microvolts input to the receiver antenna terminals, and adjust the audio generator for 30% amplitude modulation at 1,000 cps. Set receiver AF gain control for an output of one watt, as indicated on the audio power level indicator.

Adjust the audio oscillator frequency control from 100 cps to 10,000 cps, maintaining a constant 30% amplitude modulation of the RF signal. Note and record the output meter reading for several frequencies in this range.

The operating tolerance for all FAA VHF communications receivers is: Within 4 db 300 to 3,000 cps. Gain continually decreases as input frequency increases above 3,000 cps or decreases below 300 cps.

Although not mentioned in Handbook AF P 6500.28, an oscilloscope may be connected to the audio output of the receiver to indicate distortion. Sometimes headphones are helpful in detecting the presence of audio distortion, and in identifying the desired audio signal.

H. First Oscillator Frequency Measurement

The increased utilization of the VHF spectrum for Air/Ground communications has necessitated a decrease in the bandwidth of the communications receivers used in this band. It is essential, therefore, that the receiver channel frequency be maintained within narrowly prescribed limits. Since the first mixer crystal is the chief potential source of error in the receiver channel frequency, the first oscillator frequency should be measured periodically to determine if the channel crystal frequency is within the required tolerance. Although the procedure that follows is specifically for VHF narrow band receivers, a similar procedure

should be employed to check other communications receivers that show indications of being off frequency.

1. CONDITIONS OF TEST

Connect a one or two pf capacitor to the control grid of the first mixer (or remove tube shield from frequency multiplier tube and wrap insulated test lead around tube's glass envelope). Connect the other side of the capacitor (or test lead) to the Gertsch, or equal, frequency meter RF input cable. Measure the frequency of the oscillator-multiplier-injection amplifier. The frequency thus measured must be within $\pm 0.002\%$ of the appropriate crystal frequency.

$$\% \text{ frequency error} = \frac{\text{difference from specified frequency}}{\text{specified frequency}} \times 100$$

I. Final Check

It is necessary that adjustments be made to a receiver following completion of performance checks in order to assure normal operation. The following series of steps may be followed as a recommended procedure for returning a receiver to service:

1. TRIMMER GAIN

Connect the receiver antenna to the input jack and a signal generator to a spare or standby antenna. Adjust the signal generator to the exact operating frequency of the receiver by tuning the generator frequency control for maximum receiver output. Adjust the RF output of the signal generator until receiver output caused by signal voltage exceeds noise level by at least 50 mw. Reduce AF gain control until 50 mw of output is indicated and adjust the antenna trimmer for maximum receiver output.

2. RECEIVER CONTROLS

Adjust the RF, AF and Audio Quieting controls for the levels required for normal facility operation.

3. FINAL MEASUREMENTS

Remove the output meter (or VTVM), connect audio lines to receiver terminations, and bridge with a VOM (or VTVM). Increase signal generator output into the spare antenna and record:

- a. Signal generator output required to open squelch.
- b. Signal generator output required to reach AVC threshold.
- c. Audio output on VOM (or VTVM) at the AVC threshold.

4. OPERATIONAL CHECK

Remove test equipment and restore receiver to service. Check voice quality by listening to voice transmissions. Using the signal generator into the same spare antenna, the over-all operation may be checked periodically without disturbing receiving controls or wiring by repeating the "Final Measurements" outlined in the preceding paragraph. Results will indicate antenna faults, reduced receiver gain, inadequate squelch operation or audio output level that may require adjustment or servicing.

2-5. ADDITIONAL PERFORMANCE MEASUREMENTS

There are several performance measurements that provide additional useful information concerning the operational characteristics of a receiver. Two of these, *Image Rejection Ratio* and *Receiver Internal Noise*, are described in the following paragraphs.

A. Image Rejection Ratio

The mixer stage in a superheterodyne receiver will produce an output signal at the intermediate frequency when the RF input signal differs from the local-oscillator injection frequency by the intermediate frequency. For a specific local-oscillator injection frequency there are two frequencies at which an incoming RF signal will produce the intermediate frequency; one RF signal is above and the other is below the local-oscillator injection frequency. One of these frequencies is the desired signal frequency and the other is the image frequency. When the receiver is tuned to the desired signal,

a signal appearing at the image frequency will interfere with the desired signal unless the image frequency is sufficiently attenuated.

Periodically, it may be desirable to compare the signal levels required to produce a given receiver output at the desired signal frequency and the image frequency. The ratio of these two signal input levels is recorded in db.

1. CONDITIONS OF TEST

The equipment arrangement for this measurement is as for measuring over-all sensitivity with the receiver RF gain fully advanced. The signal generator output is increased until the receiver output rises above the noise level. This should occur in the vicinity of 50 or 100 milliwatts. The receiver input signal level is recorded at this time and labeled E_1 . The signal generator is then tuned to the image frequency and its output increased until the receiver output again equals that noted for E_1 input at the signal frequency. The receiver input signal level is again recorded and labeled E_2 . The image-rejection ratio in decibels is computed from E_2/E_1 . Refer to the applicable receiver instruction book for the tolerances. A value in excess of 80 db is generally acceptable.

B. Receiver Internal Noise

This test is a measure of the noise inherent in the receiver under no-signal conditions.

1. CONDITIONS OF TEST

Since most literature on sources of receiver internal noise indicates the input resistance of the first RF amplifier is a prime contributing factor, the input terminal of the receiver should be terminated into a resistance equal to the receiver input impedance. For example a coax fitting containing a 50-ohm resistor will be used in lieu of either an antenna or short circuit for measuring internal noise of most FAA VHF communications receivers. With the receiver controls set as for measuring over-all receiver sensitivity and the proper resistance connected to the receiver input, the receiver output in milliwatts is then measured and recorded.

The tolerance for most receivers indicates the internal noise should be approximately 1.0 to 10 milliwatts.

2-6. SUMMARY

An effort has been made to explain the need for performance measurements, the routine maintenance required, the equipment requirements and the methods employed to conduct the various receiver performance measurements. If a conscientious routine maintenance program is followed, good receiver performance is assured.

In the Receiver Laboratory every performance measurement described will be accomplished at least once and the majority of them will be accomplished on each of several receivers. An effort is made to give you experience on a typical variety of FAA receivers. A typical example of the receivers assigned in the Receiver Laboratory would be the RUQ/RVA, RV-5/RV-8, RV-6/9, RIU, and RF-2 receivers.

2-7. PRACTICE PROBLEMS

$$\text{db} = 10 \log P_2/P_1$$

$$\text{db} = 20 \log E_2/E_1 + 10 \log R_1/R_2$$

$$\text{db} = 20 \log I_2/I_1 + 10 \log R_2/R_1$$

Solve the following problems. Answers are provided at the end of the problems.

- Express the following power ratios in db:
 - 25/1
 - 95/2
 - 1/195
 - 1/38
- Express the following voltage ratios in db:
 - 57/1
 - 13.9/1
 - 1/115
 - 1/76
- If 0 db is referred to 6 mw, how much voltage across a 600-ohm load does this represent? How much current flows through the load?

4. If 0 db is referred to 1 mw, how much voltage across a 600-ohm load does this represent? How much current flows through the load?
5. A receiver has an over-all gain of 90 db. What power ratio does this represent?
6. A radio receiver has an input impedance of 70 ohms and an output impedance of 20,000 ohms. $2.5 \mu\text{v}$ of RF signal applied to the receiver input produces 1.5 watts of audio output.
- What is the voltage developed across 20,000-ohm load?
 - What is the db gain of the receiver?
7. The only available power output meter was damaged; therefore, a 20,000-ohm resistor and a multimeter were used to measure the signal-to-noise ratio of a receiver. With $5 \mu\text{v}$ of *Standard Test Voltage* input to the receiver, the RF gain control was adjusted to provide 100 volts ac across the 20,000-ohm load. When the modulation of the RF signal was turned off, the meter indicated 7.1 volts ac.
- What is the power output of the signal plus noise?
 - What is the power output of the noise alone?
 - What is the signal-to-noise ratio in db?
8. An attenuator in a VHF communication receiver consists of a voltage divider network of three resistors in series. The values of the resistors are 68,000 - ohms, 33,000-ohms and 33,000-ohms. When the voltage is taken across all three resistors, the attenuation is 0 db.
- What is the attenuation in db when the voltage is taken across only the two 33,000-ohm resistors?
 - What is the attenuation in db when the voltage is taken across only one 33,000-ohm resistor?
9. While measuring the AVC response of a receiver, the following data was recorded.
- | RECEIVER INPUT | RECEIVER OUTPUT |
|-----------------------|-----------------|
| $7.5 \mu\text{v}$ | 38 mw |
| $50 \mu\text{v}$ | 50 mw |
| $500 \mu\text{v}$ | 52 mw |
| $5000 \mu\text{v}$ | 55 mw |
| $50,000 \mu\text{v}$ | 60 mw |
| $500,000 \mu\text{v}$ | 58 mw |
- What is the db change in receiver input?
 - What is the over-all db change in receiver output?
10. The sensitivity of the 3.2 Mc intermediate frequency section of an RUQ receiver is $110 \mu\text{v}$ to produce 1 volt AVC. It is desired to check the selectivity at the 6-db, 20-db, 40-db and 60-db points of the selectivity curve.
- What is the signal input required for 6 db?
 - What is the signal input required for 20 db?
 - What is the signal input required for 40 db?
 - What is the signal input required for 60 db?
11. When using the Daven OP-182B power output meter, shown in Fig. 2-1 of this chapter, for measuring receiver signal-to-noise ratio, the meter indicated 5 milliwatts of signal pulse noise. The meter multiplier, 106, was on the $\times 100$ position. With modulation removed from the receiver input signal, the meter indicated 25 milliwatts when the meter multiplier was changed to the $\times 1$ position. What is the signal-to-noise ratio in db?

12. From the following data taken with the Daven OP-182B power output meter, calculate the db level with respect to 1 milliwatt.

METER INDICATION	METER MULTIPLIER	DB SETTING	DB LEVEL
(a) 14 db	+20	
(b) 8 db	+10	
(c) 12 db	-10	
(d) 16 db	0	

B. Answers to Practice Problems

1. (a) 14.0 db
(b) 16.8 db
(c) -22.9 db
(d) -15.8 db
2. (a) 35.1 db
(b) 22.9 db
(c) -41.2 db
(d) -37.6 db
3. 1.9 volts
3.16 ma
4. 0.775 volts
1.29 ma

5. 10^9
6. (a) 173 volts
(b) 132 db
7. (a) 500 mw
(b) 2.5 mw
(c) 23 db
8. (a) 6 db
(b) 12 db
9. (a) 96.5 db
(b) 2 db
10. (a) 220 μv
(b) 1100 μv
(c) 11,000 μv
(d) 110,000 μv
11. 13 db
12. (a) 34 db
(b) 18 db
(c) 2 db
(d) 16 db

CHAPTER 3

GENERAL RECEIVER THEORY

3-1. INTRODUCTION

A receiver is any device which accepts and demodulates RF signals to obtain the intelligence contained in the signal. Radio receivers vary widely in their complexity and basic design depending upon the intended application and upon economic factors. A simple radio receiver can consist of an earphone and a silicon or germanium crystal as a carrier rectifier or demodulator, and the associated antenna may consist of a length of wire. Such a receiver is insensitive and offers little discrimination between signals in the same portion of the frequency spectrum.

On the other hand, a dual-diversity receiver, which may employ double or triple detection, may occupy several relay racks and cost many thousands of dollars.

3-2. CLASSIFICATION OF RECEIVERS

Radio receivers may be classified in many ways. One classification may include the type of signal to which the receiver will respond; this signal classification would include a continuous wave, an amplitude modulated wave, a frequency modulated wave and a pulse modulated wave. Another receiver classification would pertain to the frequency spectrum in which the receiver operates; this frequency classification would include low frequency (30 kc to 300 kc), medium frequency (300 kc to 3 mc), high frequency (3 mc to 30 mc), very high frequency (30 mc to 300 mc) or ultra high frequency (300 mc to 3000 mc). Receivers may be classified according to their use or purpose; the receiver may be used to receive voice or music broadcasts, for point-to-point communications, for monitoring, for remote control, for radar or for television.

Receivers are frequently classified by their circuitry; some common circuit classifications are autodyne, regenerative, superregenerative, tuned radio frequency, superheterodyne, multi-band, tunable, fixed-tuned and crystal-controlled.

This chapter will relate mainly to communications and monitoring receivers using amplitude modulation, but these receivers will include most of the circuit classifications and operate in the frequency range from 200 kc to 400 mc. The reception of frequency modulation is covered in Chapter 6 of this material.

3-3. GENERAL REQUIREMENTS

The most important characteristics or requirements in a communications receiver are the sensitivity, the selectivity and the noise figure.

A. *Sensitivity*

The sensitivity of a receiver is the measure of the ability of the receiver to respond to weak signals.

A receiver intercepts many radiated RF signals. These radiations may originate from transmitters operating anywhere, or the signals may originate from transmitters of widely varying power output. Therefore, the radiations intercepted by the receiver antenna and presented to the receiver are of widely varying signal strength. A sensitive receiver will provide good reception of the desired signals whether the signals are weak or strong. The receiver's sensitivity is primarily dependent upon the over-all gain of the various amplifier stages in the receiver.

B. Selectivity

The selectivity of a receiver is the measure of the ability of the receiver to accept the desired signal and reject adjacent channel signals.

Selectivity may be illustrated as a curve which shows the strength of carrier input required at the resonant frequency to produce a specified receiver output as compared to the input signal strength required to produce the same output at specified frequencies above and below resonance. A typical selectivity curve is shown in Figure 3-1. The selectivity is primarily dependent upon the characteristics and number of tuned circuits employed in the receiver.

C. Noise Figure

The noise figure of a receiver may be described as the ratio of the noise appearing in the output of the receiver to the noise appearing at the input to the receiver. The noise figure determines the weakest signal that can be received and not be concealed by the noise generated in the receiver. Factors which affect the noise figure are discussed further in Chapter 5. In addition to these three general requirements, there are several other requirements which may vary with importance from one type of receiver to another. Some of these requirements would be fidelity, stability and undistorted power output.

3-4. TUNED RADIO FREQUENCY RECEIVER

A tuned radio frequency (TRF) receiver consists of several cascaded RF amplifier stages which are all tuned to the signal frequency and which are followed by a detector and the audio or video amplifiers. The number of RF stages depends upon the gain required to amplify the weakest signal to be received to a level suitable for detection. Figure 3-2 shows a block diagram of a TRF receiver.

The RF amplifiers employed are usually of the untuned-primary tuned-secondary type; however, double-tuned coupling circuits may be used when fixed-tuned operation is being

employed. For tunable operation, double-tuned RF amplifiers would complicate the tuning and tracking circuits.

The detector is generally of the linear type, diode detector or linear plate detector; however, a square-law detector may be used if little RF amplification is desired.

The characteristics of a TRF receiver are simplicity of design, no image frequency, poorer selectivity and difficult to track when tuned over an appreciable bandwidth. These characteristics are true for the comparison of the TRF receiver with the superheterodyne receiver.

The RIS, RIV and RIU receivers are examples of TRF receivers used in the FAA. The disadvantages of the average TRF receiver are not too apparent in the FAA receivers because the receivers are operated fixed-tuned on some specific frequency in the 200 kc to 400 kc band. If it is desired to change the frequency of operation for one of these receivers, it is necessary to realign all of the RF amplifier stages to the new frequency. Double-tuned RF amplifiers are employed in these receivers; this feature improves the selectivity. The receivers are used to monitor a particular facility, therefore, fixed-tuned operation is satisfactory, and the problem of tracking is overcome. Additionally, at these low frequencies nothing is to be gained by employing the superheterodyne principle.

A. Detectors

There are many different types of detectors; however, they all operate fundamentally in the same way. All detectors accomplish the three steps necessary for demodulation; these steps are rectification, filtering and separation of the audio frequency component from the RF component.

The output of an ideal detector reproduces exactly the modulation existing on the RF carrier. Failure to accomplish this purpose results in distortion. When the detector output contains frequencies which were not in the modulation envelope, the result is amplitude distortion. When the detector discriminates between modulation frequencies, the result is frequency distortion. When the

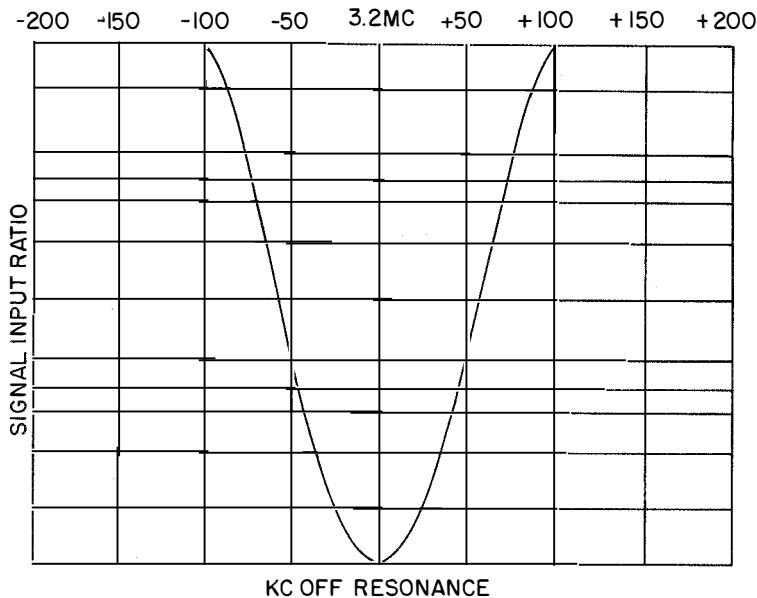


Figure 3-1. Typical Selectivity Curve

detector changes the phase relationship between different frequency components of the modulation envelope, the result is phase distortion.

The discussion of detectors that follows will consider detectors employed for amplitude-modulated signals.

1. CHARACTERISTICS OF AMPLITUDE-MODULATED SIGNALS

If a radio wave is to convey a message, there must be some means of transmitting the desired information in such a way that the message can be recovered with acceptable fidelity at the receiving point. One way to

accomplish this transmission is to vary the amplitude of the radiated RF signals in accordance with the intelligence that it is desired to transmit. This method is called amplitude modulation. Figure 3-3 shows an amplitude modulated carrier.

There are two common methods of visualizing amplitude modulation. In one method the modulation process is visualized as a wave of constant frequency with varying amplitude. The other method expands the radiated wave into components of the carrier and the side-band frequencies. While studying amplitude modulation in the Transmitter Course, these two methods will be combined into one presentation.

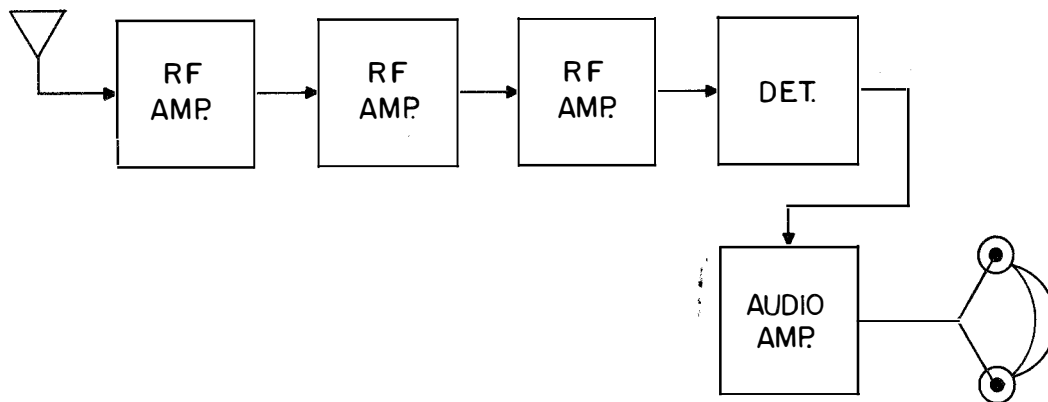


Figure 3-2. TRF Receiver Block Diagram

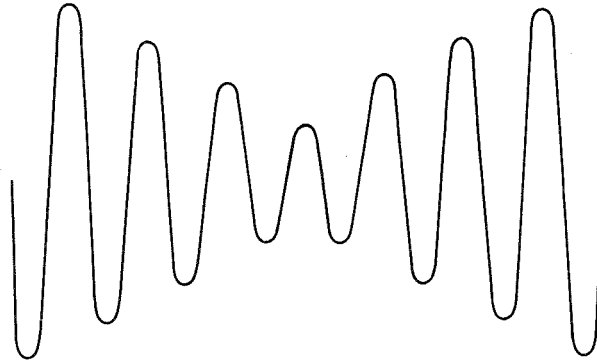


Figure 3-3. Amplitude Modulated Carrier

Amplitude modulation as shown in Figure 3-3 serves to illustrate how one can visualize a wave form as having its amplitude varied by the application of an audio frequency upon the RF carrier.

The sideband viewpoint illustrates that the modulated signal appears as three components or three different frequencies for a tone-modulated RF carrier. Assuming a carrier frequency of 1500 kc modulated by a single audio tone of 5 kc, the illustrations of Figure 3-4 show that the audio component will result in the generation of two new frequencies. These new frequencies are known as the upper and lower sidebands. The upper sideband is the sum of the carrier frequency and the audio frequency or 1505 kc. The lower sideband is

the difference between the carrier frequency and the audio frequency or 1495 kc. Only the carrier, the upper sideband and the lower sideband will be radiated due to the resonant properties of the power amplifier plate tank circuit and the associated antenna. The summation of these three radiated signals will result in the modulated envelope shown in Figure 3-4 (e).

When several audio frequencies simultaneously modulate an RF carrier, several pairs of sidebands are generated, and the combination of all the generated frequencies with the carrier produces a complex resultant modulation envelope. Should any of these various sideband frequencies be rejected by tuned circuits prior to detection, there will not be

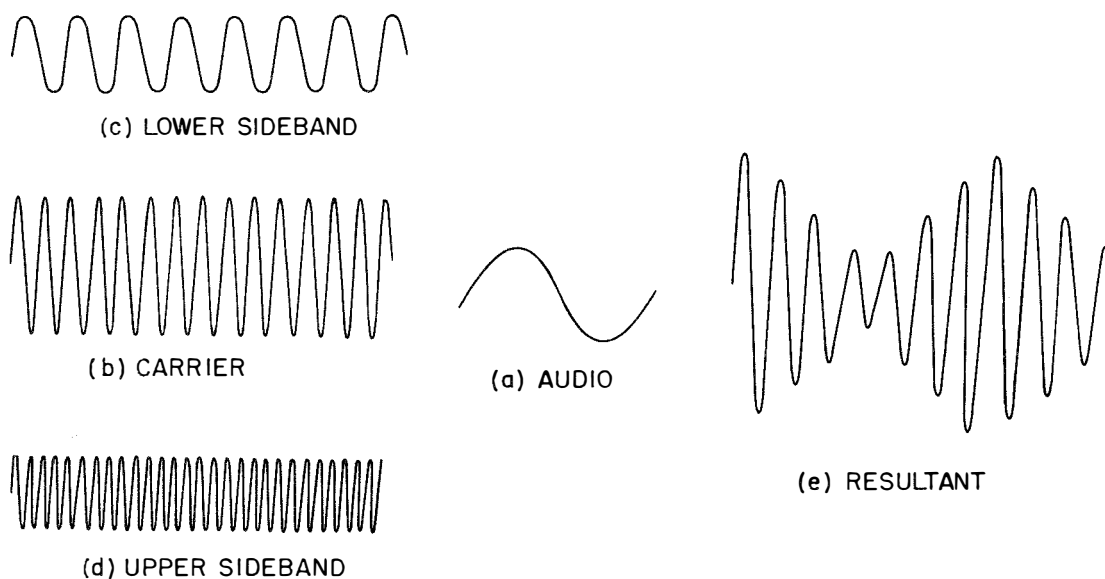


Figure 3-4. Frequency Considerations of Amplituded Modulated Wave

true reproduction of the modulating signals. The bandpass of the various receiver tuned circuits should be adequate to accommodate the desired carrier and its associated sidebands. As the frequency of the modulating signal increases, the bandpass of the receiver must be increased.

Another factor to consider is the amplitude of the modulating signal. As the amplitude of the modulating signal is increased, the depth or percentage of modulation is also increased.

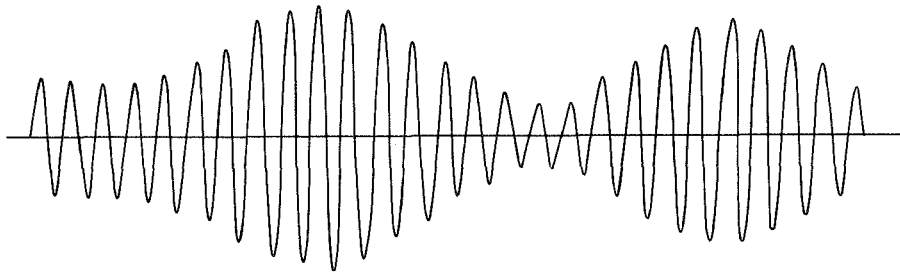
2. DIODE DETECTOR

A diode rectifier is the most common de-

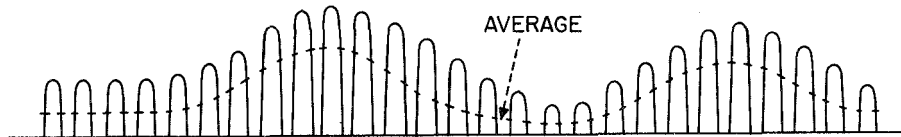
tector of amplitude-modulated signals. Ordinarily, this diode rectifier is a diode vacuum tube; however, crystal rectifiers are being used in increasing numbers. Crystal and vacuum tube diodes behave in similar fashion in most circuits. Generally, a crystal diode can replace a vacuum tube diode without any appreciable change in circuit behavior, and a vacuum tube diode can replace a crystal diode except at extremely high frequencies where vacuum tube diodes are much more subject to transit-time effects than are crystal diodes. A simple diode detector circuit with diagram showing the operation is shown in Figure 3-5.



(a) Circuit of Simple Diode Detector



(b) Input Voltage Applied to Diode



(c) Current through Diode

Figure 3-5. Diode Detector and Waveforms

In Figure 3-5a, C is a small capacitor, and R is a relatively high resistance. The combination RC circuit is the load impedance across which the output of the diode is developed. At each positive peak of the applied RF wave the diode conducts, and capacitor C is charged to a value almost equal to the peak of the RF cycle. Between peaks of the RF wave, some of the charge on the capacitor leaks off through the resistance. This loss in charge will be replaced at the peak of the next RF cycle provided the peaks of the two RF cycles are equal. If the peaks of the RF wave vary, the charge on the capacitor will also vary; this variation is shown in Figure 3-5b. The voltage developed across the RC combination reproduces the modulation envelope, provided the time constant of the RC combination is short when considering the frequency of the modulating signal, but the time constant is long when considering the frequency of the RF carrier. These RC circuit characteristics allow the voltage across the capacitor to vary at an audio rate but will not allow the voltage across the capacitor to vary appreciably at an RF rate.

The current that flows through the diode is in the form of pulses as shown in Figure 3-5c. The magnitude of these pulses must be sufficient to charge capacitor C to a voltage which is almost equal to the peak of the applied voltage. The average value of these pulses of current produces the output voltage across the detector load. This voltage, which should be filtered free of the RF component, is a reproduction of the intelligence which was used to modulate the RF carrier produced in the transmitter. For applied signals larger than one volt, the diode detector functions as a linear detector; for applied signals of less than one volt, the diode detector generally functions as a square-law detector. Under the latter conditions of operation, the average diode current increases as the square of the amplitude of the applied voltage. This change in average diode current is a function of the rate of change of the slope of the diode characteristic at the operating point. This operating point should be located where the diode characteristic curvature is greatest as shown in Figure 3-6.

3. GRID-LEAK DETECTOR

The detection of an amplitude-modulated signal can be accomplished in a triode as well as in a diode. If the detection is accomplished in the grid circuit of the triode, the detector is commonly referred to as a grid-leak detector. In Figure 3-7 the grid and cathode function as the plate and cathode of a diode detector. The detected modulation of the input signal is developed across the RC grid-leak network. As in diode detection, grid-leak detection can be either linear or square-law, depending upon the amplitude of the applied signal.

For linear operation, the signal applied to the grid-leak detector should have an amplitude of several volts. The grid draws current on the positive portions of the incoming signal that drives the grid positive with respect to the cathode. This grid current charges the grid-leak capacitor to nearly the peak value of the applied signal. During the remainder of the cycle of applied signal, a portion of the charge on the capacitor leaks off through the grid-leak resistance. The voltage across the grid-leak RC network consists of a dc component due to rectification and filtering of the carrier, an ac component due to the modulation of the carrier, and a ripple component at the carrier frequency. Since this ripple component would be amplified by the tube and appear in the plate circuit, an additional RC or LC filter is needed to minimize the value of this ripple voltage. The same considerations are used to select the proper value of R and C for use in the grid-leak detector as are used for selecting the proper value of R and C for use in the diode detector. R should be much larger than the forward resistance of the diode or equivalent diode to improve rectification efficiency. C should be small enough to prevent amplitude or phase distortion of the highest modulating frequency but large enough to reduce the carrier frequency ripple voltage appearing across the load resistance to an acceptable value.

The linear grid-leak detector has the advantage over the linear diode detector of greater voltage output due to the amplification characteristic of the triode tube employed as a grid-leak detector.

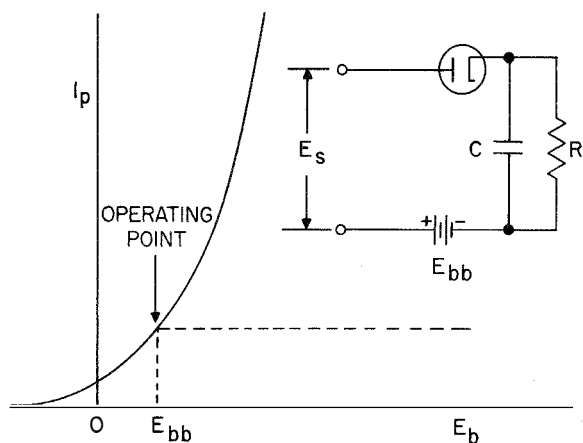


Figure 3-6. Square-Law Detection

When the applied signal amplitude is small, the grid-leak detector operates very similar to the square-law diode detector.

4. PLATE DETECTOR

The detection of an amplitude-modulated carrier can be accomplished in the plate circuit of a vacuum tube, generally a triode. The plate detector can be either a linear detector or a square-law detector. Figure 3-8 shows a circuit diagram of a plate detector.

When the plate detector is used as a linear detector, the tube is biased to projected cutoff, a point just above the maximum curvature on the applicable tube characteristic curve. Refer to Figure 3-9. Without the capacitor in the

plate circuit, a large input signal would cause the plate current to increase almost linearly during the positive half cycle of the applied signal but would cause the plate current to be cut off during the majority of the negative cycle of the applied signal. If a capacitor is placed in parallel with the load resistor as shown in Figure 3-8, the capacitor will take a charge that is a function of the tube resistance and the plate load resistance.

If the tube is a triode, a large plate load resistance, as compared to the tube resistance, would be used. This resistance relationship will result in an average voltage across the capacitor which is slightly greater than the minimum instantaneous plate voltage that would exist across the tube if the capacitor were removed. The average voltage across the capacitor will vary inversely with the average value of plate current. It is in this manner that the modulation on the carrier is detected in the plate circuit.

The square-law plate detector is very similar to the linear plate detector except that the tube bias is adjusted to the point of maximum curvature on the tube characteristic curve.

This point of operation for weak signals will result in a maximum change in average plate current for a variation in amplitude of the modulated input signal.

Although a plate detector has an advantage over a diode detector due to the amplification

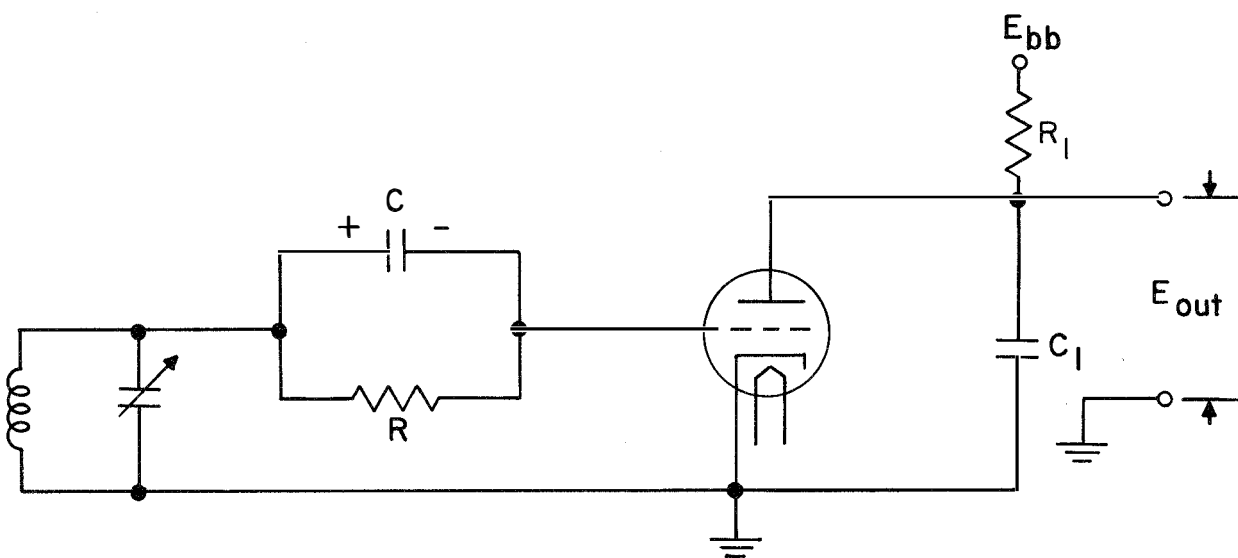


Figure 3-7. Simplified Grid-Leak Detector

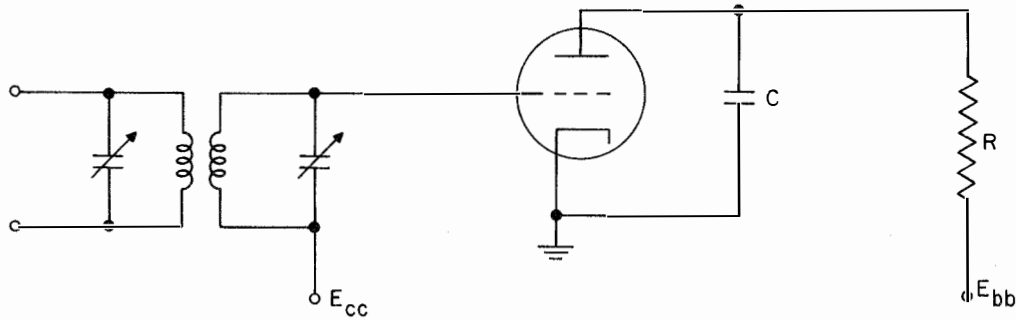


Figure 3-8. Circuit of Plate Detector

of the triode producing an over-all gain through the detector, this gain is somewhat less than when the same tube is used as an amplifier.

The diode detector is most frequently employed in FAA communications receivers because of the ability of the detector to respond favorably to wide variations in input signal levels. This variation is apparent when it is realized the receiver must respond to

communications from an aircraft sitting on the ramp as well as communications from an aircraft located several miles from the receiver. In addition to distance, the power output of the respective transmitters may differ greatly.

B. Amplifiers

The amplifiers associated with the TRF receiver are of two types, audio frequency and

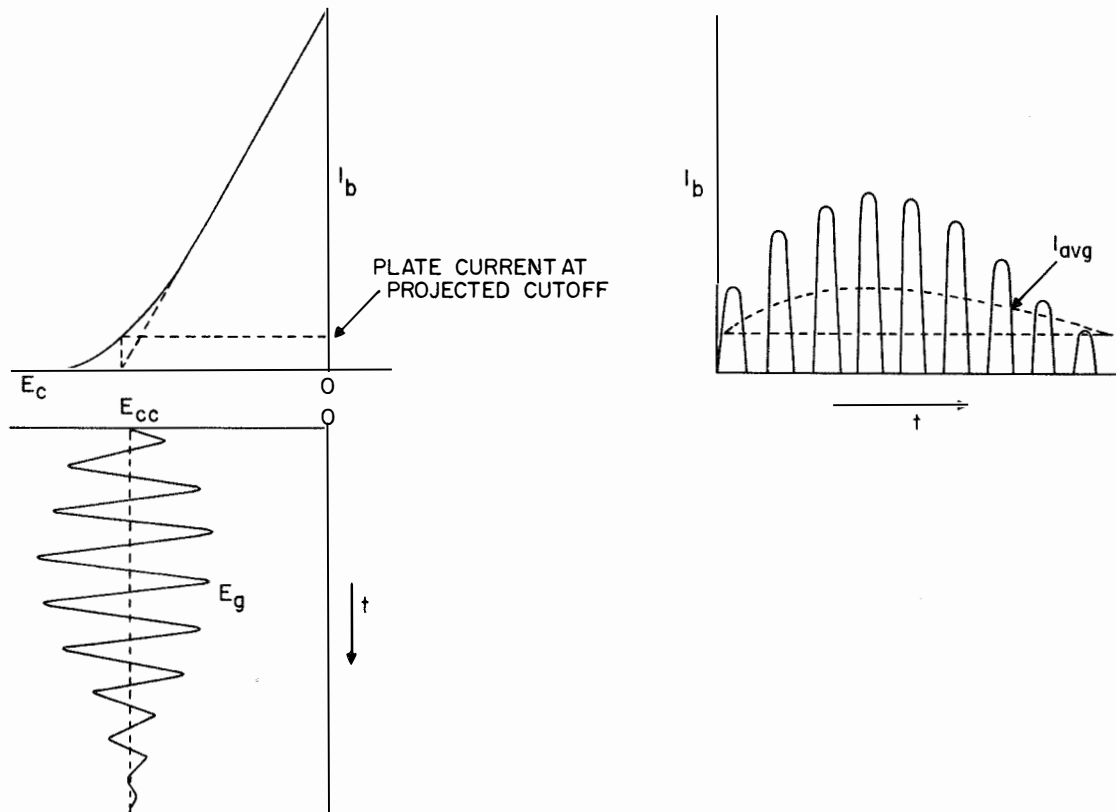


Figure 3-9. Waveforms in a Linear Plate Detector

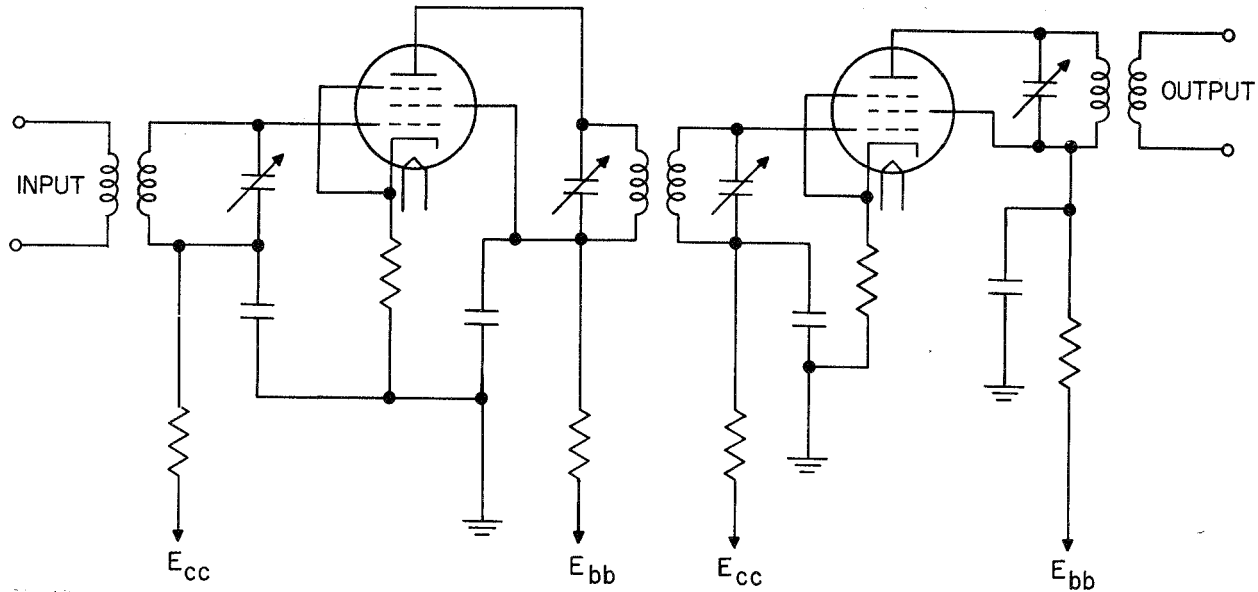


Figure 3-10. Simplified Schematic of Double-Tuned RF Amplifier

radio frequency. Generally, more gain per stage would be accomplished if audio amplifiers were used; however, audio amplifiers would not improve the selectivity of the receiver. To obtain the desired selectivity, a number of tuned radio frequency stages normally precede the detector. The gain and selectivity of the RF amplifiers are dependent upon the Q of the tuned circuits employed. At this time only double-tuned RF amplifiers will be considered because this is the type of amplifiers used in FAA TRF monitor receivers.

A typical double-tuned RF amplifier is shown in Figure 3-10. The primary and secondary circuits are both resonant to the same frequency and are inductively coupled to the adjacent circuits. The desired selectivity curve is achieved through proper adjustment of the

primary and secondary resonant circuits and the degree of coupling between the primary and secondary. The double-tuned circuits are sometimes divided into high-Q and low-Q circuits. A high-Q circuit is considered to be a circuit having a bandwidth, at the half power point, that is one-tenth or less than one-tenth of the resonant frequency.

A low-Q circuit is one that has a bandwidth, at the half power point, that is greater than one-tenth of the resonant frequency.

Although the coefficient of coupling is adequately covered in Chapter 4 of this material, it is necessary to understand the effects on the secondary voltage of varying the degree of coupling between primary and secondary of circuits tuned to the same frequency. This is shown pictorially in Figure 3-11.

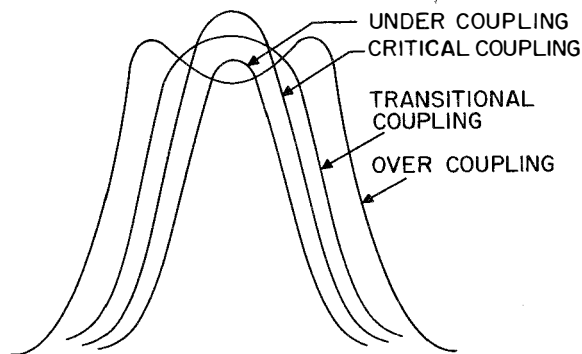


Figure 3-11. Effects of Coupling on Secondary Voltage

To briefly explain each illustrated degree of coupling, the following explanations are given. For each of these explanations it is assumed that the two circuits involved are inductive coupled and are tuned to the same frequency.

Undercoupling — increased coupling will cause an increase in the secondary voltage.

Critical coupling—the circuits have been oriented to provide the maximum possible secondary voltage.

Transitional coupling—the circuits are coupled to provide the flattest secondary response curve possible.

Overcoupled—the degree of coupling is sufficient to produce a secondary response curve which has two peaks.

The coefficient of coupling, k , between the primary and secondary is expressed by:

$$k = \frac{M}{\sqrt{L_p L_s}} \quad (3-1)$$

Where: M is mutual inductance between windings

L_p is primary inductance

L_s is secondary inductance

The amplifier gain, A , obtained at the resonant frequency is given by:

$$A = \frac{g_m k \sqrt{R_p R_s Q_p Q_s}}{1 + k^2 Q_p Q_s} \quad (3-2)$$

It should be apparent from Equation (3-2) that the gain of the amplifier is dependent upon the Q of both the primary and secondary circuits. This equation is greatly simplified for critical coupling because at critical coupling:

$$k_c = \frac{1}{\sqrt{Q_p Q_s}} \quad (3-3)$$

The expression for the midband gain simplifies from Equation (3-2) to:

$$A = \frac{g_m \sqrt{R_p R_s}}{2} \quad (3-4)$$

The over-all bandwidth of several, n , stages is determined by:

$$\Delta f_t = \Delta f \sqrt{2^{1/n} - 1} \quad (3-5)$$

It can be seen from Equation (3-5) that the greater the number of tuned stages, the narrower the bandwidth, and the better is the selectivity.

Although considerable amplification was obtained from the required number of radio-frequency amplifiers employed, additional gain will be required to provide the necessary receiver power output. Most FAA receivers will deliver one watt into a 20,000-ohm load. This power output is the equivalent of 141.4 volts across the 20,000-ohm load. Normally, 141.4 volts would be too great of signal amplitude to be obtained directly from the detector due to the distortion that would be produced. For this reason, one or more audio amplifier stages are required after the detector to amplify the detected signal to the desired output level.

Two types of audio amplifier stages are used, voltage amplifiers and power amplifiers. The voltage amplifiers are used to provide an amplified signal of sufficient magnitude to properly excite the grid of the power amplifier. Generally, medium- μ tubes are used, especially, when the preceding radio frequency amplifiers provide considerable gain. Resistance-capacitance coupling of the voltage amplifier to the power amplifier is conventional. The load resistance of the voltage amplifier should be several times the plate resistance. These voltage amplifiers should produce a minimum of distortion.

A power amplifier must efficiently convert dc power to ac power. Maximum power transfer occurs when the load impedance equals the internal impedance of the source. In practice, however, the load impedance is kept smaller than the plate resistance of the tube to prevent distortion. Power amplifiers normally use tetrodes or pentodes. For tetrodes and pentodes, a load impedance of about one-tenth the value of the plate resistance is used.

For further discussion of audio amplifiers, refer to the course material on Audio Equipment which was provided during the preceding weeks of the Communications Equipment Course. Particular attention should be given to class of operation, methods of obtaining bias and degeneration.

The TRF receivers that are used by the FAA for monitoring purposes do not employ any automatic volume control circuits because this feature would tend to maintain the signal output constant when the input signal decreases below the desired alarm level.

3-5. SUPERHETERODYNE RECEIVER

When receiving communications in the VHF region, it becomes considerably more difficult to obtain the desired selectivity and gain through the radio-frequency amplifier stages. If several radio-frequency amplifiers are employed in an effort to obtain the desired selectivity, the tunable receiver becomes quite complex due to the tuning mechanism required. These design problems can be overcome to a great degree by employing the design of the superheterodyne receiver. Refer to Figure 3-12 for a block diagram of a superheterodyne receiver.

The superheterodyne receiver is the most commonly used type of radio communications receiver. It also has wide application in pulse equipment. At higher frequencies the superheterodyne receiver is generally capable of greater sensitivity and selectivity than the TRF receiver for a given number of stages. These advantages are due to the fact that most of the amplification of the signal can be accomplished in the intermediate-frequency amplifier stages, which do not have to be tunable and which usually operate at a lower frequency than the radio-frequency amplifier stages. This results in the sensitivity and selectivity being more uniform throughout the tuning range of the receiver. A greater gain per stage can be obtained at intermediate frequencies than at radio frequencies.

The basic principle of the superheterodyne receiver is that the received radio-frequency signal is converted to a lower frequency, or

intermediate frequency, by heterodyning the radio-frequency signal with a local-oscillator signal. The difference between the two signals is usually employed for the intermediate frequency. When tuning the receiver, the local oscillator, mixer, and any radio-frequency amplifiers are tuned simultaneously in such a manner that the intermediate frequency remains constant.

To simplify the discussion of the superheterodyne receiver as much as possible, the previously discussed tuned radio frequency receiver will be expanded into the superheterodyne receiver. Since the radio-frequency amplifier stages of the TRF receiver were fixed-tuned, assume that these stages are aligned to 300 kc. It is desired to adapt such circuitry to the reception of some VHF signal, for example 121.5 mc. By heterodyning or mixing 121.5 mc with an oscillator frequency that is removed from 121.5 mc by 300 kc, the tuned circuits of the intermediate-frequency amplifiers will reject all frequencies except 300 kc and a narrow band of frequencies on either side of 300 kc. Adequate gain and selectivity can be obtained through the intermediate-frequency amplifiers.

A. Mixers

In order to convert the received signal at 121.5 mc to a signal at 300 kc, it will be necessary to add a mixer and a local-oscillator circuit to the basic circuitry of the TRF receiver. In the mixer circuit the local-oscillator frequency is modulated by the incoming radio-

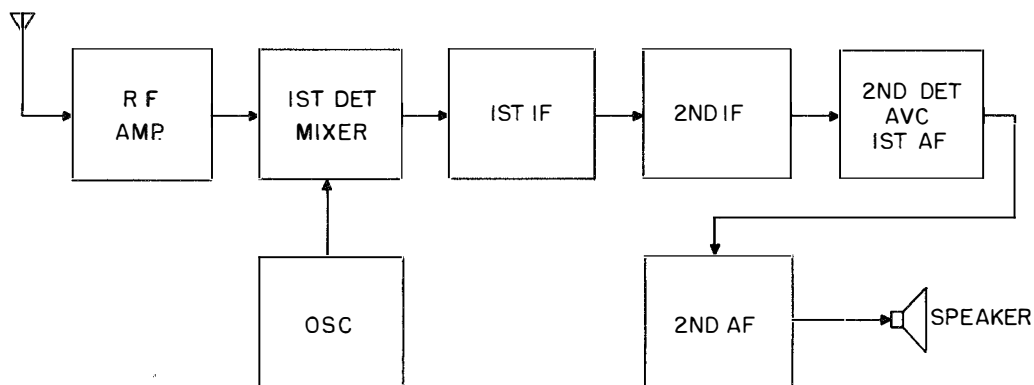


Figure 3-12. Superheterodyne Receiver Block Diagram

frequency signal to produce modulating frequencies which are the sum and difference of the frequencies of the incoming signal and the local oscillator. In the example being used, an oscillator frequency of 121.8 mc combined with the incoming signal of 121.5 mc would result in a difference frequency of 0.3 mc or 300 kc and a sum frequency of 243.3 mc. If the incoming signal is modulated with a 10,000 cycle tone at the originating transmitter, then an upper sideband of 121.51 mc and a lower sideband of 121.49 mc is produced. These sideband frequencies would mix with the local-oscillator frequency also. The upper sideband of 121.51 mc would combine with the local-oscillator frequency to produce a difference frequency of 290 kc, and the lower sideband of 121.49 mc would combine with the local-oscillator frequency to produce a difference frequency of 310 mc. Sum frequencies would also be produced, but since the intermediate-frequency amplifiers will respond to only a narrow band of frequencies in the 300 kc region, the sum frequencies serve no useful purpose. The intelligence or 10,000-cycle tone now appears as modulation of a 300 kc intermediate frequency. The intermediate-frequency amplifiers must have an over-all bandpass of 20 kc at half-power points to satisfactorily pass the band of frequencies from 290 kc to 310 kc. This band of frequencies would then be amplified and detected in the same manner that was previously explained for the TRF receiver.

In receivers designed for broadcast reception, economy and simplicity of design are major factors to be considered. Usually, the signal strength of the station being received is sufficient to override any noise that is generated within the receiver at these operating frequencies. In the VHF spectrum internal receiver noise becomes a much greater problem and will be discussed in considerable detail in Chapter 5. At this time it is sufficient to recall the common use of pentagrid converters for the mixer stage in broadcast receivers, but that the pentagrid converter produces more internal receiver noise in VHF communications receivers than do triode or pentode mixers and, therefore, is less frequently used.

Mixers and converters are necessarily non-linear devices. If they were linear devices, then only the signals applied at the input could be obtained in the output because no distortion would result. It is this distortion that will produce the familiar sum and difference frequencies; the difference frequency is commonly used for the desired intermediate frequency. Normally, a combination of grid-leak and cathode bias is used for a triode or pentode mixer. The peak local-oscillator voltage is approximately equal to the cut-off voltage of the tube and is large compared to the incoming signal voltage. The tube draws plate current for slightly more than one-half the local-oscillator cycle. The conversion gain becomes dependent upon the local-oscillator voltage under these operating conditions. This conversion gain is generally only a small percentage of the gain realized from the same tube when it is used as a conventional amplifier.

The local-oscillator injection voltage may be applied to either the control grid or cathode of a triode mixer. This injected signal can be coupled to the mixer either inductively, capacitively, or conductively. In general, local-oscillator signal injection into the grid circuit is objectionable because of the interaction between the local-oscillator output and the incoming signal. Also, if no RF amplifiers are used between the mixer grid circuit and the antenna, the local-oscillator signal can easily be radiated from the antenna and will cause interference in other nearby receivers. Since the tuned circuit in the mixer grid is resonant to the incoming signal, and the local oscillator is of a different frequency than the incoming signal, the tank circuit will present a reactance to the local-oscillator frequency. As the tank circuit is tuned to a different incoming signal, the tank circuit will provide a different value of reactance to the local-oscillator frequency. This variation in reactance will produce a variation in the level of injected local-oscillator signal, which in turn will vary the conversion gain of the mixer. Also, the reactance that the local-oscillator and incoming signal circuits present to each other will cause detuning effects in both circuits and may result in considerable difficulty in tracking the two circuits.

Cathode injection is generally more suitable because there is less chance of interaction between the circuits of the local-oscillator and the incoming signal, and there is less chance for a signal produced by the local oscillator to be radiated from the antenna.

In pentode mixers the local-oscillator signal can be injected by means of the screen grid or the suppressor grid, but the incoming signal is normally applied between control grid and cathode.

B. Local Oscillator

The local oscillator in the superheterodyne receiver may be a part of the mixer, but in most FAA communications receivers the local-oscillator employs either a separate tube or a portion of a dual triode. There are several factors which must be considered when designing a local oscillator for a particular application.

It was previously stated that the proper amplitude of local-oscillator signal must be injected to obtain the maximum conversion gain from the mixer. In triode, pentode and multi-grid mixer tubes, the local-oscillator peak injection voltage at the mixer should be from three to ten volts. Since the local oscillator is generally loosely coupled to the mixer to minimize interaction, the local-oscillator signal voltage should be five to ten times the required voltage at the mixer. This injection voltage amplitude must be held constant to reduce variations in conversion gain of the mixer.

The frequency of the local oscillator will determine to some extent the type of tube and circuit that is most suitable. At low frequencies the Hartley and Armstrong tuned-grid oscillators are frequently employed. However, these oscillators are not suitable at frequencies above 100 mc due to difficulties encountered with the mutual-inductive coupling employed for feedback. The Colpitts and tuned-plate tuned-grid oscillators are commonly utilized in receivers operating above 100 mc. Where frequency multiplication is required in the local oscillator, the electron-coupled oscillator is advantageous because of the high harmonic content in its output. Where fixed-tuned receivers are used, crystal-controlled oscillators are used because of the

improved frequency stability. It is desirable that frequency variations due to oscillator instability shall not exceed twenty per cent of the intermediate-frequency bandwidth. If the allowable warm-up time of the receiver is insufficient to allow the components in the local-oscillator to reach thermal equilibrium, the frequency drift due to change in component values with temperature may necessitate the use of temperature compensated parts. Since variations in the voltages applied to the tube may also cause frequency variations in the local oscillator, voltage-regulated power supplies are frequently used.

Since the local oscillator is tuned to a frequency differing from the RF signal frequency by the intermediate frequency, the local oscillator must tune over either a greater or smaller percentage of its center frequency than the RF circuits, depending upon whether the local oscillator is tuned below or above the signal frequency. Generally, the lower the oscillator frequency, the more stable the frequency becomes. This advantage is offset in part by a greater percentage change in the local-oscillator frequency, and this condition makes tracking between the local oscillator and the RF signal circuits more difficult. In some FAA communications receivers the local oscillator will operate above the incoming signal, and in other FAA receivers the local oscillator will operate below the incoming signal. This choice is determined by the priority of frequency stability and improved tracking.

When a crystal-controlled oscillator is employed for VHF receivers, it is desirable to utilize a crystal in the 6 mc to 8 mc region and employ the necessary frequency multipliers to increase the frequency to the value required for mixer injection. Crystals in the 6 mc to 8 mc region are less fragile and less expensive than crystals ground to operate at higher frequencies.

C. Image Frequency

In section 3-5, A, titled Mixers, it was explained that an oscillator frequency of 121.8 mc would mix with an incoming signal of 121.5 mc to produce the desired intermediate frequency of 300 kc. If a signal on 122.1 mc was received at the same time, this signal would also mix with the oscillator frequency of 121.8

mc to produce the intermediate frequency of 300 kc. The signal on 122.1 mc would interfere with the reception of the desired signal on 121.5 mc. This interfering signal would be called an image frequency. It is desirable to attenuate this image frequency to reduce the interference produced when receiving the desired signal. One or two tuned RF amplifiers preceding the mixer are commonly employed to attenuate this image frequency. In addition these RF amplifiers will reduce the antenna radiation produced at the local oscillator frequency.

Some of the more common equations that can be used to determine the image frequency follow:

$$f_i = f_o \pm 2f_{ir} \quad (3-6)$$

$$f_i = f_o \pm f_{ir} \quad (3-7)$$

$$f_i = 2f_o - f_s \quad (3-8)$$

Where: f_i = image frequency

f_o = local-oscillator frequency

f_s = incoming signal frequency

f_{ir} = intermediate frequency

From equation 3-6 it can be determined that the image frequency is only removed 600 kc from the incoming signal of 121.5 mc in the preceding example. It would be quite difficult to design two RF amplifiers that would have the needed selectivity to provide the needed attenuation at 122.1 mc while providing adequate reception of 121.5 mc.

One solution to this problem would be to use a higher intermediate frequency. If an intermediate frequency of 18.3 mc was used, the image frequency would be removed 2×18.3 mc or 36.6 mc from the desired frequency of 121.5 mc, and two RF amplifier stages could be designed that would provide considerable attenuation at the image frequency. However, as the image frequency rejection was improved, the selectivity as determined by the intermediate frequency stages became poorer.

The relationship between bandpass and resonant frequency is commonly expressed:

$$\Delta f = \frac{f_r}{Q} \quad (3-9)$$

Where: Δf = bandpass at the half-power points
 f_r = resonant frequency of the tuned circuit

Q = quality factor of the circuit as determined by

$$\frac{X_L}{R}$$

From this equation the general conclusion would be that the lower the resonant frequency, the narrower the bandpass and the sharper the selectivity curve. Therefore, as the intermediate frequency is increased from 300 kc to 18.3 mc in the example, the selectivity became considerably poorer.

D. Double-Conversion Superheterodyne Receiver

In order to obtain the desired selectivity and image rejection, it may be necessary to employ double-conversion. The incoming signal is converted to a high intermediate frequency to obtain the desired image rejection as explained using an 18.3 mc intermediate frequency, and then the high intermediate frequency is converted to a lower intermediate frequency to obtain the desired selectivity. A low intermediate frequency that is more practical than 300 kc is 3.2 mc, the frequency that is used in the RUQ VHF Receiver which is used for FAA communications. Since the signal has been converted twice, there must be two mixers and two local oscillators. Refer to Figure 3-13.

The double-conversion superheterodyne receiver principle is used in the RUQ receiver. The preceding explanation should not be construed to imply that a double-conversion superheterodyne receiver always has better selectivity than the single-conversion type. Both the RCP and RHZ receivers are the single-conversion type, and they have better selectivity than the RUQ. To obtain this selectivity, however, some image rejection was sacrificed.

Another advantage of a double-conversion superheterodyne receiver is that greater overall intermediate-frequency gain can be obtained without regeneration than in a single-conversion superheterodyne receiver due to the frequency isolation between the two intermediate-frequency amplifiers.

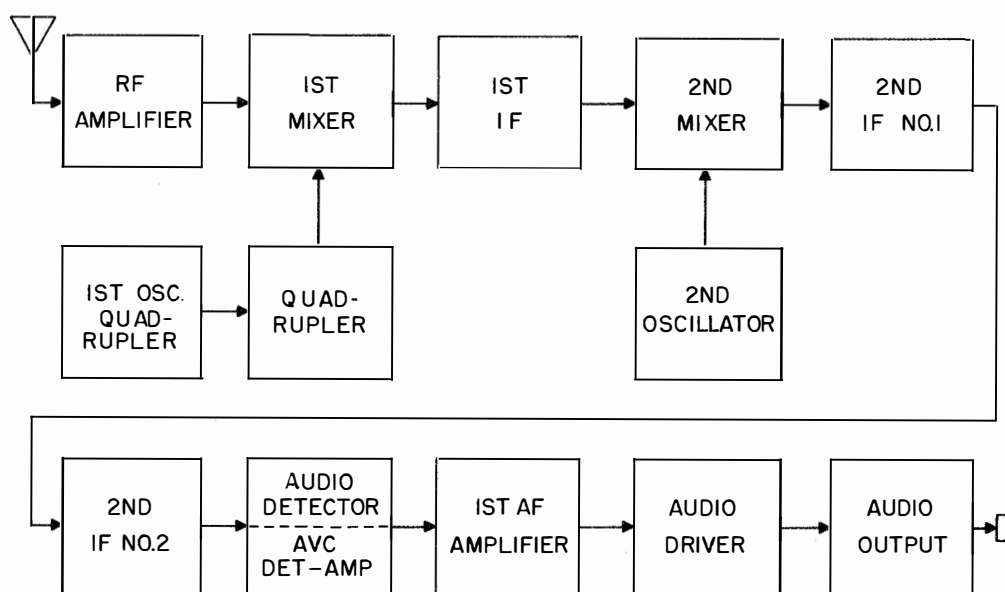


Figure 3-13. Double-Conversion Superheterodyne Receiver Block Diagram

E. Intermediate-Frequency Amplifiers

The desired characteristics of the intermediate-frequency amplifier are determined by the receiver application and the characteristics of the other stages of the receiver.

The required gain must be determined in order to establish the number of stages in the intermediate-frequency amplifier section. The maximum gain of the intermediate-frequency amplifier is normally determined by the amplitude of the smallest mixer output signal that must be amplified and by the signal level required at the audio detector for proper detection. Surplus gain should also be provided to allow for tube aging, tube variations and nonoptimum circuit alignment.

There are numerous factors that affect the selection of the intermediate frequency. The factors that affect the selection of a high or a low frequency were previously explained in image rejection.

The bandwidth and selectivity characteristics of a receiver are usually established by the intermediate-frequency amplifier because the circuit element Q 's are lower for a given bandwidth in a low-frequency amplifier than in a high-frequency amplifier and because it

is impractical to cascade several tunable RF amplifiers. The bandwidth of the intermediate-frequency amplifier is ordinarily selected to make the signal-to-noise ratio as large as possible. Any drift in the local-oscillator frequency must also be accommodated by the intermediate-frequency bandwidth. Where wide bandwidths are required, the intermediate-frequency amplifier tuned circuits may be over-coupled, stagger-tuned, or damped by adding resistance to the circuit. Each of these practices is employed in various FAA receivers.

In addition to providing the desired bandwidth, the intermediate-frequency amplifier should provide off-resonant skirts that have steep slopes to improve the selectivity.

The intermediate-frequency amplifier must not be regenerative. In crystal-controlled receivers the intermediate-frequency amplifier is a primary source of receiver instability. If the amplifier is close to the point of oscillation, serious distortion of the amplifier bandpass can occur even though the amplifier does not actually oscillate. If automatic volume control is applied to the amplifier, the changes in gain of the individual stages can cause the input capacitance of these stages to change sufficiently to detune the circuits a significant

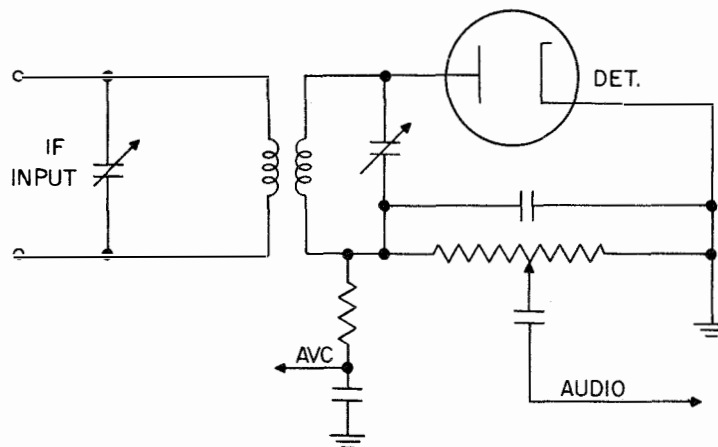


Figure 3-14. Simplified Schematic Diagram of Detector and AVC Circuits

amount. This effect must be compensated for by proper circuit design.

The intermediate-frequency amplifier must be able to handle the largest expected mixer output signal without limiting or overloading. It must have sufficient gain to amplify the smallest desired signal to a usable level for the audio detector. Some form of automatic gain control must be included.

F. Automatic Volume Control

Most amplitude-modulation communications superheterodyne receivers utilize an automatic volume control circuit to maintain the signal voltage essentially constant at the audio detector. Figure 3-14 is a simplified schematic diagram of the avc circuit of an ordinary receiver.

The diode detector rectifies the signal at the output of the last intermediate-frequency transformer. This output contains the audio signal and a dc component, whose amplitude is proportional to the strength of the RF carrier. In order to separate the audio and dc components, an RC filter is employed. The time constant of this RC filter should be between .04 and .09 seconds. The capacitor employed in this filter will bypass the audio component to ground from the avc buss. The resistor will decouple this filter capacitor from the audio gain control. If the time constant of the RC network is too short, the avc voltage would not be free of the audio component.

If the time constant is too long, the avc would no longer be able to compensate for signal fading. This filtered dc component is used to control the gain of the RF amplifiers, mixer, intermediate-frequency amplifiers, and in some instances, the first audio amplifier. If the carrier strength increases, greater negative voltage is developed across the detector load resistance. This increased voltage is fed back to the avc controlled stages where it constitutes a greater negative bias. The controlled tubes are generally remote cut-off pentodes, whose gain decreases with increased bias. If the carrier strength decreases, the avc voltage decreases, and the gain of the avc controlled stages increases. This operation compensates for the drop in signal strength. Figure 3-15 shows a comparison between sharp cut-off and remote cut-off tubes.

There is a disadvantage to this system of automatic gain control; any received signal will cause additional bias to be applied to the avc controlled tubes. This additional bias will reduce the amplification, and the signal will be held to a low value at the audio detector, even though it is already lower than the value necessary for desired output.

To overcome this short-coming of the simple avc circuit, delayed avc is commonly used in communications receivers. In delayed avc there is no avc voltage developed until the incoming signal exceeds a predetermined level, which is established by bias on the avc stage or stages. Delayed avc becomes operative

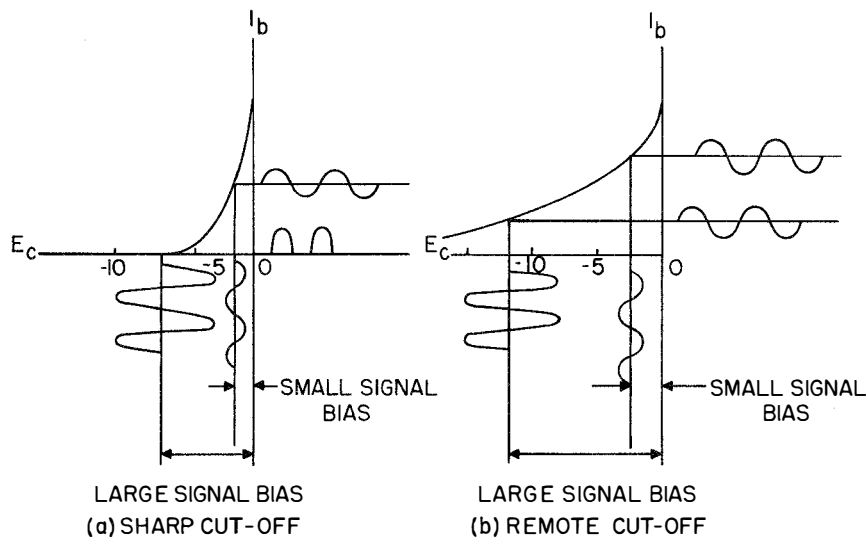


Figure 3-15. Comparison between Sharp Cut-off and Remote Cut-off Tubes

with signal levels in the vicinity of $7 \mu\text{v}$. With delayed avc, the sensitivity of the receiver is greatly improved over the sensitivity obtained with simple avc, particularly for weak signals. In addition, the signal level is maintained at a much higher ratio with respect to the noise level inherent in high-gain receivers.

RC decoupling networks are employed between avc controlled stages to prevent feedback reactions between the various stages due to the common coupling connections. It is important that the values of the capacitors and resistors used in these decoupling filters do not impair the desired avc action, particularly with fading signals.

When delayed avc is being used, a rectifier separate from the audio detector is generally used because the bias required to produce the delay in the avc circuit would also produce a delay in the detection of the audio signal. Refer to Figure 3-16.

Figure 3-17 shows a comparison between receiver output vs. receiver input for receivers having no avc, simple avc and delayed avc. The point where the output no longer increases linearly with input is called the avc threshold.

G. Noise Reduction

Since the average communications receiver has an over-all gain of 100 to 120 db, slight

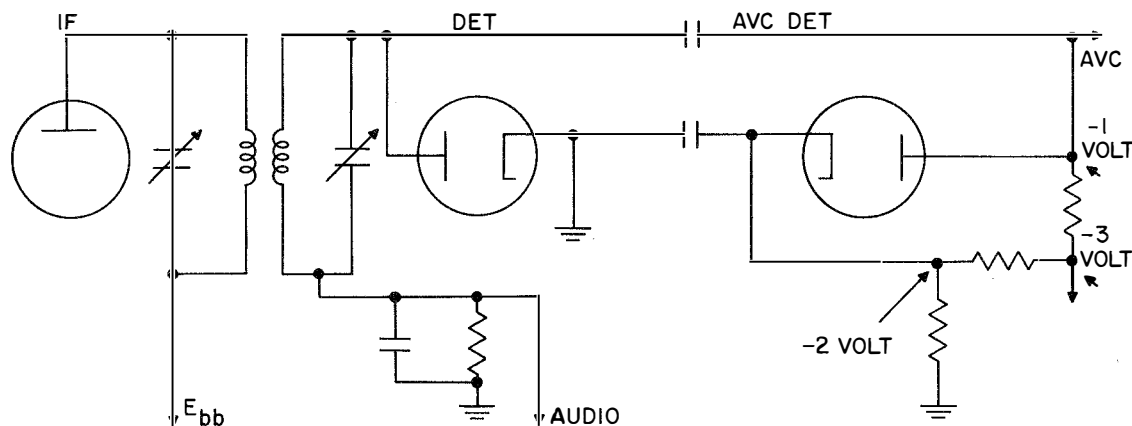


Figure 3-16. Delayed AVC and Audio Detector Circuits



3-6. SUMMARY

There are many operating conditions to consider when deciding the best circuit design of a communications receiver. The TRF receiver is satisfactory for low frequency fixed-tuned operation as monitor receivers.

Although the double-conversion superheterodyne receiver is generally superior to the TRF and single-conversion superheterodyne

receivers, its circuitry is more complex, and expensive. Although the superheterodyne receiver will produce better sensitivity and selectivity than the TRF receiver, there is the problem of image interference with the superheterodyne receiver. Each type of receiver has its particular application, and this application will become more apparent as the circuitry of each type is discussed individually.

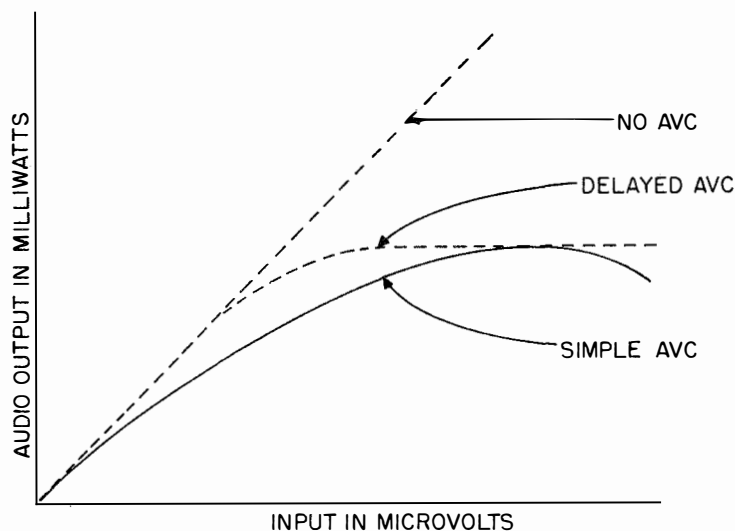


Figure 3-17. Comparison of Receiver Outputs

variations due to noise at the receiver input will cause appreciable disturbances at the receiver output. The best method to reduce noise is to attack it at its source. The sources of noise and methods of reducing noise will be adequately covered in Chapter 5, but some of the common methods of reducing noise after it has been generated will be mentioned at this time since it does affect the receiver circuitry.

The inherent noise of a mixer is greater than the inherent noise of an RF amplifier at VHF. The noise generated in the receiver input circuit is the major noise to be considered since it will be amplified by 100 to 120 db. For this reason, in addition to improved image rejection and less local-oscillator radiation, at least one RF amplifier is commonly added preceding the first mixer circuit.

Various noise reduction circuits may be incorporated into the receiver design to reduce the noise level in the receiver output. Some of these circuits are the series noise limiter, shunt noise limiter, noise-cancellation detector and low-pass filter, all of which are effective while a normal signal is being received. When no signal is being received, a carrier-operated-noise-suppression circuit will greatly reduce the noise present in the receiver output. The

other noise reduction circuits are also effective for no-signal condition.

These circuits will be discussed individually in Chapter 5 or while discussing the circuitry of various FAA communications receivers.

H. Power Supply

In present day communications receivers the power supply is generally on the same chassis with the receiver and is designed to supply the receiver with the necessary alternating and direct current voltages for proper operation from the 120 volt ac line.

The power supply consists of the power transformer, rectifier and filter circuits. The transformer consists of the primary winding, which is connected to the ac line, and the various step-up and step-down secondary windings, which supply the high plate voltages and low filament voltages. In most instances the rectifier will be the full-wave type, although the half-wave type is sometimes used, particularly for the bias supply rectifier where separate tubes are employed for high voltage and bias voltage sources. The filter circuit is generally two sections and may be inductive or capacitive input. Voltage-regulator tubes may be employed to provide improved voltage regulation.

CHAPTER 4

COUPLED CIRCUITS

4.1. GENERAL

Two or more electrical circuits may be coupled together in several ways in order that energy may be transferred from one circuit to another. Inductive and capacitive coupling can be used if the energy to be transferred is comprised of alternating current and voltage. FAA applications make use of ac energy for telephone, power, audio and radio frequencies ranging from 20 cycles for telephone ringing to 24,000 megacycles for Airport Surface Detection Equipment (ASDE).

The properties of inductively and capacitively coupled circuits are used throughout a wide variety of electronic circuits to provide specific characteristics in electronics equipment. Radio receivers and transmitters make use of the properties of such circuits to control characteristics such as selectivity, image rejection, audio fidelity, voltage step-up or step-down, impedance matching, signal-to-noise ratio, etc.

An understanding of the principles of inductance, and capacitance, and the application thereof to various types of circuitry, is required in order for the technician to be able to properly analyze and maintain circuitry which includes inductance and capacitance.

This chapter will consider a general review of circuit theory, circuit characteristics, general applications, specific FAA applications, and design consideration for coupled circuits.

4.2. SCOPE OF CHAPTER

The term "coupled circuits" as used in this chapter will be restricted to those circuits that are coupled by means of inductance and/or capacitance, and primarily confined to inductively coupled circuits. Such coupling is

capable of transferring electrical energy in the alternating current form from one circuit to another, but will not so transfer electrical energy in the direct current form. Only sinusoidal ac will be considered in the analysis of the circuitry discussed in the chapter.

4.3. CIRCUIT PHASE RELATIONSHIP AND CHARACTERISTICS

Alternating current and voltage relationships in an ac circuit vary depending upon the predominant circuit characteristics, that is, the instantaneous values of current may be directly proportional to the applied emf values at any specific instant, or may precede or succeed (lead or lag) proportional emf values with respect to time. Depending upon the effect produced, whether proportional instantaneous current values are caused to lead or to lag the instantaneous applied emf, the impedance of a circuit is also said to give respectively, leading or lagging circuit characteristics, for lagging or leading impedance phase angles.

A. Resistive Circuits

In a purely resistive circuit, the current is limited by the circuit resistance and is instantaneously proportional to the applied emf. See Figure 4-1.

$$I = \frac{E}{R} \quad (4-1)$$

Where: E is the rms value of emf in volts, and R is the resistance in ohms, then I represents the rms value of current in amperes.

$$i = \frac{e}{R} \quad (4-2)$$

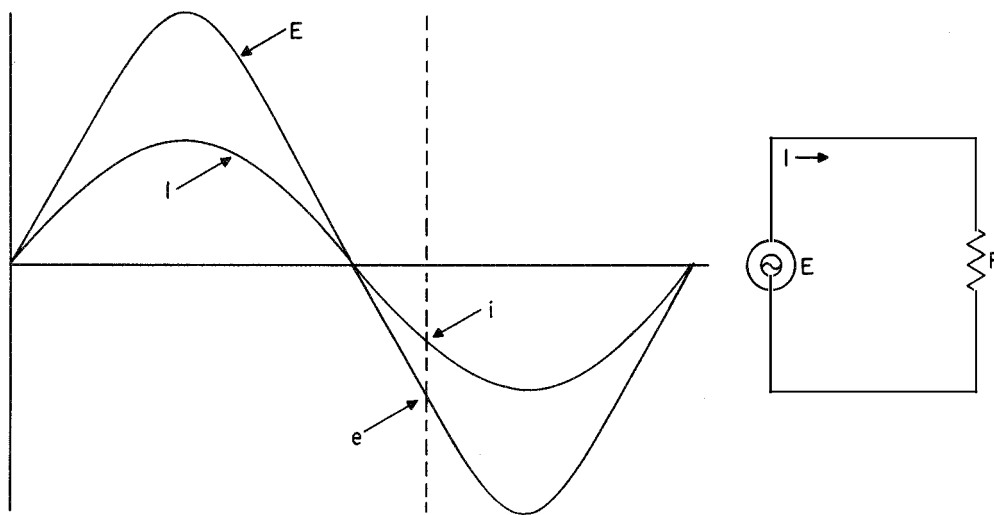


Figure 4-1

Where: i is the instantaneous current value, R is the circuit resistance, and e is the instantaneous applied emf.

Where: L is the inductance in henries, f is the frequency in cycles per second, and X_L is the inductive reactance in ohms.

$$I = \frac{E}{X_L} \quad (44)$$

B. Inductive Circuits

In a purely inductive circuit, the current is limited by the inductive reactance and is instantaneously lagging the applied emf by 90° . See Figure 4-2.

Where: E and I again represent rms values of emf in volts and current in amperes respectively, and X_L is in ohms.

$$i = \frac{-E_{max} \cos \phi}{X_L} = \frac{e}{X_L} \quad (45)$$

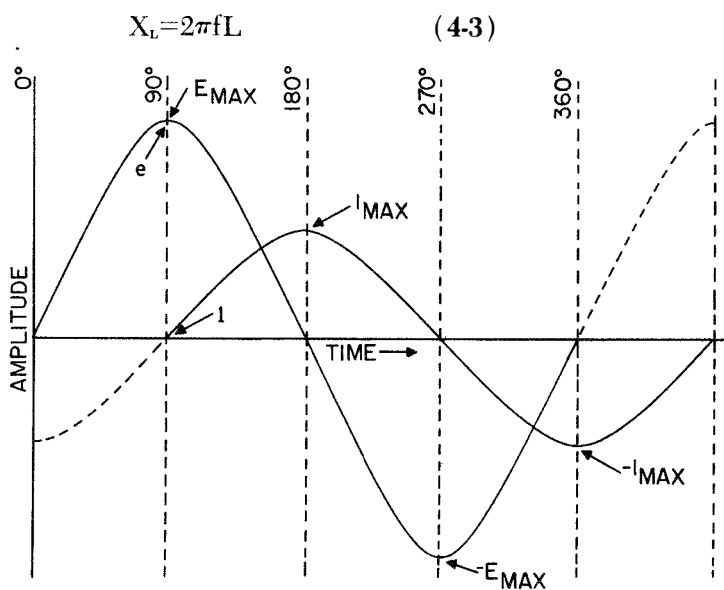
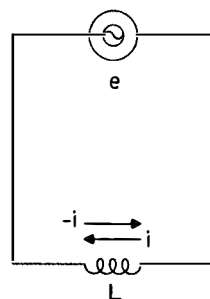


Figure 4-2



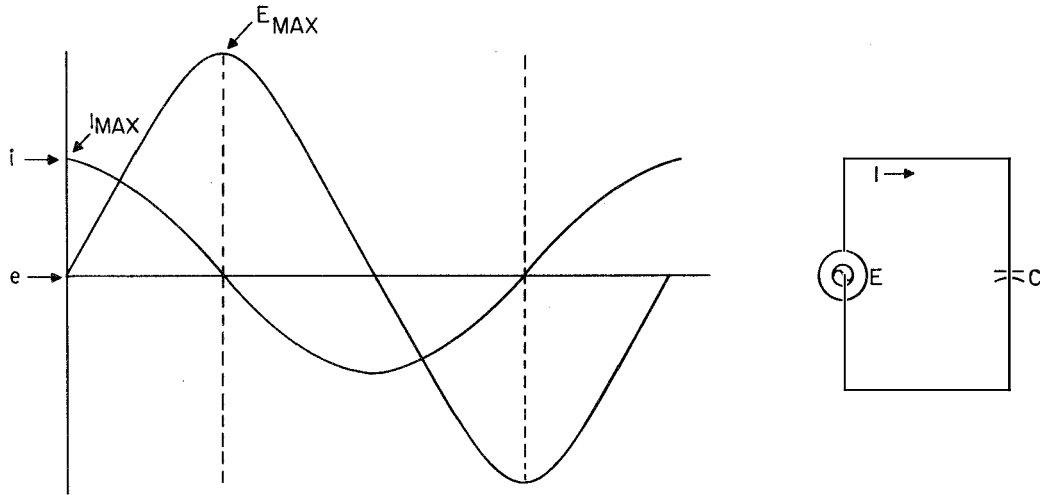


Figure 4-3

Where: i and e represent the current and emf values at a given instant, ϕ is the angle of generation through which e has passed, and E_{max} is the maximum instantaneous value of emf.

C. Capacitive Circuits

In a purely capacitive circuit, the current is limited by the capacitive reactance and is instantaneously leading the applied emf by 90° . See Figure 4-3.

$$X_c = \frac{1}{2\pi f C} \quad (4-6)$$

Where: X_c is the capacitive reactance in ohms, f is the frequency cycles per second, and C is the capacitance in farads.

$$I = \frac{E}{X_c} \quad (4-7)$$

Where: E and I represent the rms values of emf in volts and current in amperes, respectively, and X_c is in ohms.

$$i = \frac{E_{max} \cos \phi}{X_c} = \frac{e}{X_c} \quad (4-8)$$

Where: i and e represent current and emf values at a given instant, ϕ is the angle of generation through

which e has passed, and E_{max} is the maximum instantaneous value of emf.

D. Circuits With Combined LCR

By controlling the relative values of L , C and R in series, parallel, and series-parallel circuits the current may be caused to lead or lag the applied emf by phase angles varying from 0° to 90° . Also, the instantaneous currents in separate portions of the circuit may be caused to vary from 90° to 180° out of phase with each other.

In circuit of Figure 4-4, the current I_L , lags the applied emf by 90° and lags the current I_R , by 90° inasmuch as I_R is in phase with the applied emf, E .

In Figure 4-5a these current values I_R and I_L are depicted as rotating vectors that are rotating in a counterclockwise direction with a generator armature and producing a sinusoidal wave form as in Figure 4-5b. In-as-

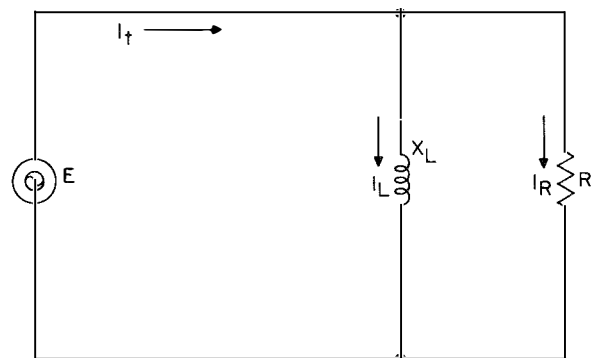


Figure 4-4

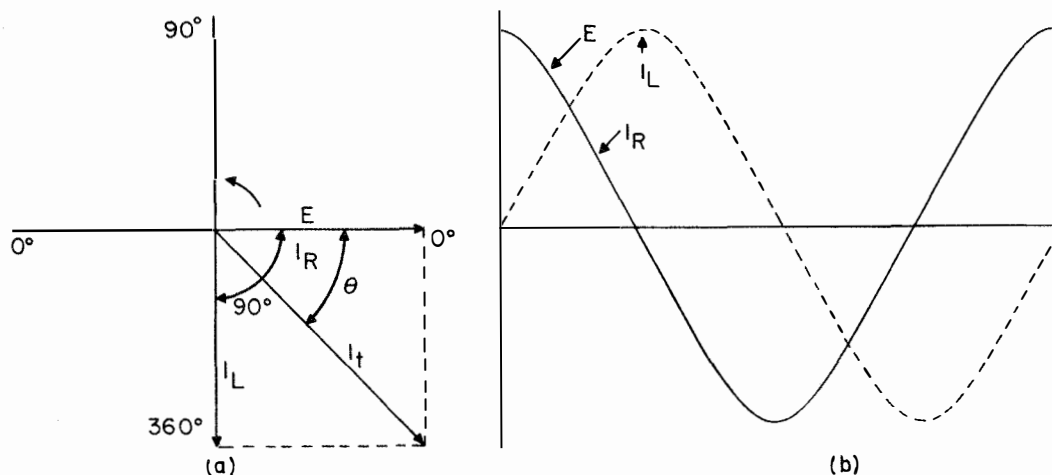


Figure 4-5

much as I_R is in phase with E , then I_R must be maximum when E is maximum, etc., as shown in Figure 4-5a.

I_L , which lags E by 90° , is shown as I_L . The resultant current, then, is

$$I_t = \sqrt{I_R^2 + I_L^2} \quad (4-9)$$

and I_t lags E by a phase angle θ . By substitution,

$$I_t = \sqrt{\frac{E}{R} + \frac{E}{X_L}}$$

The total impedance of the circuit is, $Z_t = \frac{E}{I_t}$, or

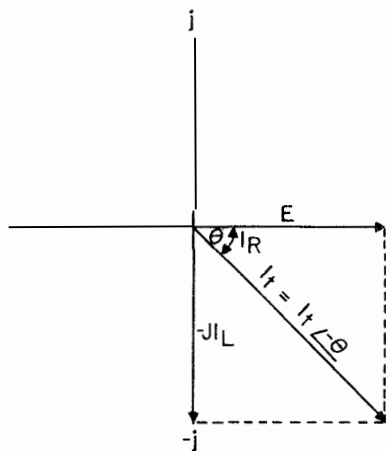


Figure 4-6

$$Z_t = \frac{RjX_L}{R + jX_L} \quad (4-10)$$

Equation 4-10 is the same equation that would be developed by starting from the equation

$$Z = \frac{1}{\frac{1}{Z_R} + \frac{1}{Z_L}}$$

where Z_R and Z_L are the vector impedances of the resistive and inductive branches, respectively.

E. Vector Notation

Emf, current and impedance phase relationships for other combinations of LCR circuits may be similarly derived by the method just used. However, another system of notation which results in overall simplification of computations is more generally used and will be indicated to facilitate subsequent explanations in this chapter.

Let the Cartesian coordinate diagram of Figure 4-5a be re-drawn as Figure 4-6 with revised nomenclature as indicated. Note: $j = \sqrt{-1}$.

I_t then becomes the vector sum of I_R and $-jI_L$, or

$$I_t = I_R - jI_L, \text{ or} \quad (4-11)$$

$$I_t = I_t \cos \theta - jI_t \sin \theta, \text{ or}$$

$$I_t = |I_t| \angle -\theta$$

From the equation, $Z = \frac{E}{I_t}$, then we find that,

$$Z_t = \frac{E}{(I_t \angle -\theta)}, \text{ or, } Z_t = |Z_t| \angle \theta \quad (4-12)$$

The equivalent LR series impedance for Figure 4-4 is,

$$Z_t = Z_t \cos \theta + jZ_t \sin \theta, \text{ or } Z_t = R + jX, \quad (4-13)$$

Where $Z_t \cos \theta = R$ and $Z_t \sin \theta = X$

F. Admittance

To facilitate solution of parallel complex impedance circuits, the ADMITTANCE Y , is the reciprocal of the impedance, Z and its unit is the mho.

$$Y = \frac{1}{Z} \quad (4-14)$$

The circuit admittance is made up of CONDUCTANCE, G , which is the reciprocal of resistance, R , and SUSCEPTANCE, B , which is the reciprocal of the reactance, X .

$$\text{CONDUCTANCE } G = \frac{1}{R}$$

$$\text{SUSCEPTANCE } B = \frac{1}{X}$$

$$\frac{1}{Z} = \frac{1}{R} \pm \frac{1}{jX}$$

$$Y = G \pm jB \quad (4-15)$$

G. Circuit Characteristics

In series complex impedance circuits such as Figure 4-7, the impedance is,

$$Z = R + jX_L - jX_C \quad (4-16)$$

If X_L is greater than X_C , the same current which flows in all parts of the circuit is limited more by X_L than by X_C , and the circuit is inductive, that is, the current lags the applied emf. If X_C is greater than X_L , the circuit is capacitive.

If $X_L = X_C$, the impedance of the circuit effectively consists of R only, and the circuit current, I , is greatest.

The emf existing across each element of the circuit is equal to the product of the circuit

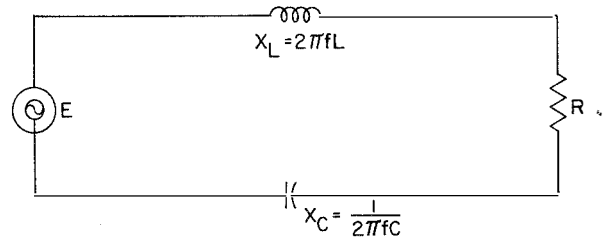


Figure 4-7

current and the impedance of that particular element.

In a parallel complex impedance circuit such as is shown in Figure 4-8 the impedance and admittances are as shown in Table I.

Branch	Impedance Symbol	Impedance Value	Admittance Symbol	Admittance Value
L	Z_L	$0 + jX_L$	Y_L	$\frac{1}{+jX_L} = -jB_L$
C	Z_C	$0 - jX_C$	Y_C	$\frac{1}{-jX_C} = +jB_C$
R	Z_R	$R + j0$	Y_R	$\frac{1}{R} = G$

TABLE I

The total admittance of the circuit of Figure 4-8 is,

$$Y_t = Y_L + Y_C + Y_R = G - jB_L + jB_C \quad (4-17)$$

The total current of the circuit is,

$$I_t = EY_t = EG - jEB_L + jEB_C \quad (4-18)$$

The emf existing across one circuit load element exists across each load element. When the reactance of the inductive branch X_L is smaller than the reactance of the capacitive branch, X_C , the inductive branch carries more current than the capacitive branch, and

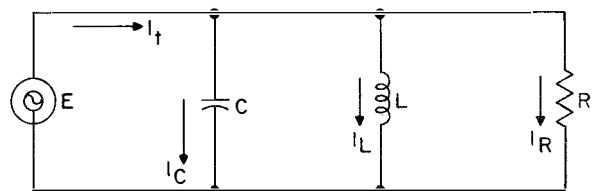


Figure 4-8

the circuit becomes inductive. When $X_L = X_C$, the circuit impedance is equal to R , and the total current, I_t is minimum.

H. Power Consumed in a Circuit

No power is consumed in a purely reactive circuit component. All the power stored therein during one half cycle of alternation is returned to the source each succeeding half cycle of alternation, thus the only power consumed in a circuit is in its resistive component, and is,

$$P = I^2 R = I^2 Z_t \cos \theta, \text{ or } P = EI \cos \theta \quad (4-19)$$

4-4. INDUCTIVE COUPLING

When an electrical conductor is linked by the magnetic field from another conductor which is carrying an electric current, the first conductor is said to be coupled (inductively) to the second conductor. If the proximity and positioning of the first conductor is such that it will be so linked when there is current in the second conductor, even though no current exists at a given time, the conductors are considered to be inductively coupled by reason of such proximity and positioning.

A. Inductance

Inductance is a circuit property or characteristic, and is usually defined in one of two ways: (1) the property of a circuit that op-

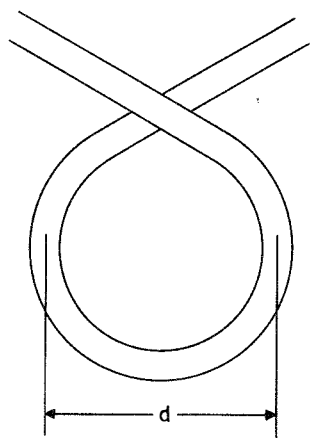


Figure 4-9

poses a change of current therein; and (2) the property of a circuit which causes an emf to be developed by virtue of a change in the circuit current. If a change of current at the rate of 1 ampere per second in a circuit causes an emf of 1 volt to be developed, the circuit is said to have an inductance of 1 henry. The induced emf will oppose the applied emf when the current is increasing and the induced emf will aid the applied emf when the current is decreasing, therefore the induced emf is

$$e = -L \frac{di}{dt} \quad (4-20)$$

Where: e is in volts, L is in henrys, i is in amperes, t is in seconds, $\frac{di}{dt}$ is the rate of change of current with respect to time.

1. SELF-INDUCTANCE

The self-inductance of a single-layer solenoid (a helically-wound air core coil) is very closely equal to,

$$L = N^2 d F \quad (4-21)$$

Where: L is the inductance in microhenrys, N is the number of turns, d is the diameter of the coil measured to center of wire, and F is the form factor as read from a specially prepared chart. This chart is shown on page 16 of the second edition of RADIO ENGINEERING by TERMAN. Two curves are shown in this chart, one is when d is in inches and the other is when d is in centimeters. Figure 4-9 shows the measurement of d .

2. MUTUAL INDUCTANCE

The self inductance of a coil was shown to be directly proportional to N^2 . If a second inductive circuit is brought into close proximity to the first inductive circuit, three general conditions may exist as shown in Figure 4-10. In Figure 4-10a, L_1 and L_2 are shown not

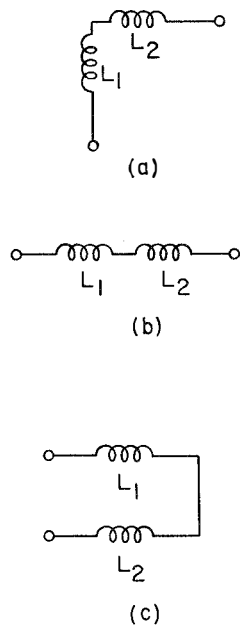


Figure 4-10

inductively coupled. In this instance the total inductance, L_t , is,

$$L_t = L_1 + L_2 \quad (4-23)$$

In Figure 4-10b, L_1 and L_2 are connected in series and closely coupled inductively and,

$$L_t = L_1 + L_2 + 2L_M \quad (4-24)$$

where the mutual inductance, L_M , is equal to

$$\sqrt{L_1 L_2}$$

In Figure 4-10c, L_1 and L_2 are connected in series and closely coupled inductively. However, because the magnetic fields oppose each other,

$$L_t = L_1 + L_2 - 2L_M \quad (4-25)$$

3. COEFFICIENT OF COUPLING

Mutual inductance between two windings is equal to $\sqrt{L_1 L_2}$ only when we have perfect (100%) coupling of two inductive circuits, which is impossible in practice. We can closely approach 100% coupling in many instances with high-permeability transformer cores and carefully wound inductances. In RF coils, the coupling is often much less than

unity (100%), varying from about 0.5% to 50% of the magnetic field from one coil linking the turns of the other coil. This condition of lessened coupling greatly reduces the inductive effect of one coil upon the other, or in other words reduces the mutual inductance, L_M , which then becomes,

$$L_M = k\sqrt{L_1 L_2} \quad (4-26)$$

Where: k is the "coefficient of coupling", and L_1 and L_2 are the two coils.

To determine L_M and k for a loosely-coupled circuit use the following procedure:

1. Measure inductances of L_1 and L_2 separately.
2. Mount L_1 and L_2 in their operating positions, connect them in series-aiding and measure L_t .
3. Compute L_M from equation 4-24.
4. Compute k from equation 4-26.
5. Measure L_t with coils connected in series-opposing.
6. Compute L_M from equation 4-25.
7. Using L_M as found in step 6, compute k from equation 4-26.
8. Compare results of L_M in steps 3 and 6, and of k in steps 4 and 7. The value of L_M should be the same in both cases, and the value of k should be the same in both cases.

B. Energy Transfer

Inductively coupled windings (transformers) are used to transfer ac energy from one circuit, commonly called the primary circuit, to another circuit, commonly called the secondary circuit. Several advantages are thereby obtained, among them being: elimination of dc energy in one of the circuits which might otherwise be present in both circuits if direct metallic coupling were used; desired voltage or current step-up or step-down, ability to match generator and load impedances, and ability to select specified bandwidth of desired frequencies to be transferred.

C. Coupled Impedance

If two coils L_1 and L_2 , of equal inductance, but of separate closed circuits be tightly coupled together ($k=1$), then $L_M = \sqrt{L_1 L_2}$. See

Figure 4-11. If $R_L=R_g$, then $Z_1=R_g+jX_{L1}$, $Z_2=R_L+jX_{L2}$, and $Z_1=Z_2$. Thus $I_1=I_2$, the magnetic fields produced will be equal and opposite in polarity and the total inductance is effectively zero. R_L then appears to be connected directly to the terminals of the generator.

In general, if a primary and secondary circuit are coupled together as in Figure 4-12a, as far as the primary circuit is concerned, the effect of a coupled secondary is as though an apparent impedance of $(\omega L_M)^2/Z_s$ had been coupled in series with the primary. The coupled impedance,

$$Z_c = \frac{(\omega L_M)^2}{Z_s} \quad (4-27)$$

Where: $\omega=2\pi f$, and Z_s is the series impedance of the secondary circuit alone.

Figure 4-12b, is the equivalent circuit for Figure 4-12a and,

$$Z_{in} = Z_p + \frac{(\omega L_M)^2}{Z_s} \quad (4-28)$$

or:

$$Z_{in} = R_p + \frac{(\omega L_M)^2 R_s}{R_s^2 + X_s^2} + j \left[X_p - \frac{(\omega L_M)^2 X_s}{R_s^2 + X_s^2} \right] \quad (4-28a)$$

Where: X_s is positive if the secondary is inductive and negative if the secondary is capacitive.

If Z_s is a complex impedance, the coupled impedance will be the conjugate of Z_s , that is, if the secondary impedance is inductive, the coupled impedance will be capacitive. If Z_s consists of resistance only, the coupled impedance is resistive.

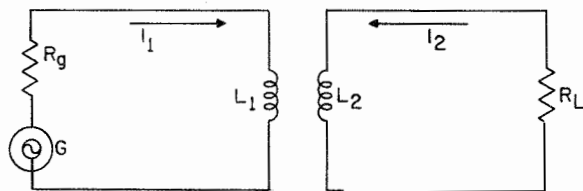


Figure 4-11

D. Impedance Matching

The most effective transfer of energy from generator to load takes place when the generator and load impedances are equal to each other. However, many times the impedance

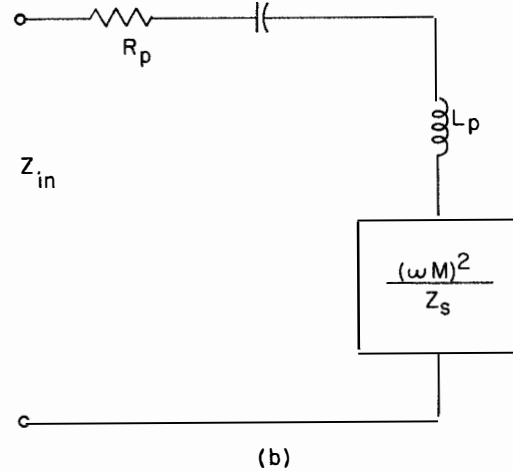
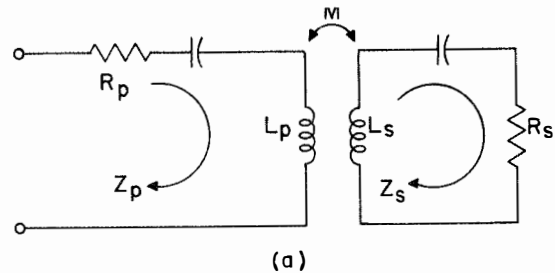


Figure 4-12

of the load is different from that of the generator. In such circumstances it is necessary to cause the load impedance to appear to the generator to be equal to the generator impedance, and one of the devices used to accomplish this is a matching transformer. Refer to Figure 4-13.

If the impedances to be matched are purely resistive, the primary and secondary turns ratios, current ratios, voltage ratios and impedance ratios are in the relationship,

$$\sqrt{\frac{Z_p}{Z_s}} = \frac{E_p}{E_s} = \frac{N_p}{N_s} = \frac{I_s}{I_p} \quad (4-29)$$

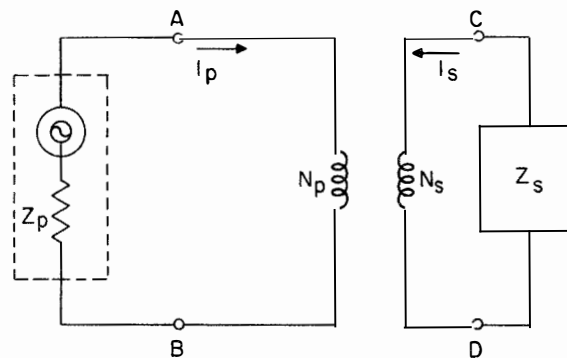


Figure 4-13

Where: p and s refer to primary and secondary circuits, respectively, Z is impedance, E is emf, I is current and N is number of turns.

If the impedances to be matched contain reactance, introduction of the opposite reactance in series with the reactive portion of the circuit, with the proper magnitude, will make the circuit resistive.

4.5. TRANSFORMER APPLICATIONS

Some of the applications of transformers are: to transfer ac energy while isolating two circuits from transfer of dc energy; to "step-up" power transmission voltages to reduce transmission losses; as "step-down" transformers to reduce power line voltages to safe levels for use in homes or to operate low voltage devices such as doorbells; to match impedance between generator and load; to act as part of a filtering circuit to pass or reject certain frequencies; and many other applications.

The power levels handled by transformers varies from a small fraction of a watt to levels approaching 1,125,000 kw. The voltage levels handled by transformers varies from a small fraction of a volt to 750,000 volts for experimental purposes. Frequencies handled by transformers vary from about 20 cycles to several hundred megacycles.

A. Power Transformers

The term "power transformer" covers a wide range of applications. It usually refers to transformers used in connection with commercial ac power applications. It may refer to the large transformers at the power house, which "step-up" the voltage of the generated ac for more economical power transmission; to the "step-down" power transformers adjacent to an individual home; to the many transformers around a factory that are used to supply the desired voltage levels for various equipments; to the rectifier power supply transformers in your television receiver; to filament power supply transformers in transmitters; etc.

Impedance matching as such, is often given little or no consideration in "power" applications inasmuch as power line loads may vary greatly from hour to hour. In order that power

transformers may be designed to produce the most efficient transfer of energy, commercial ac power is provided at specific fixed frequencies, most commonly at 60 cps. Some conversion to 400 cps, 600 cps or 800 cps power is being accomplished for factory consumption to reduce the bulk and weight of transformers and hand-held power tools. Power frequency of 400 cps is used in military aircraft to increase the ratio of power/(size and weight) of electrical equipment as much as possible.

B. Audio Transformers

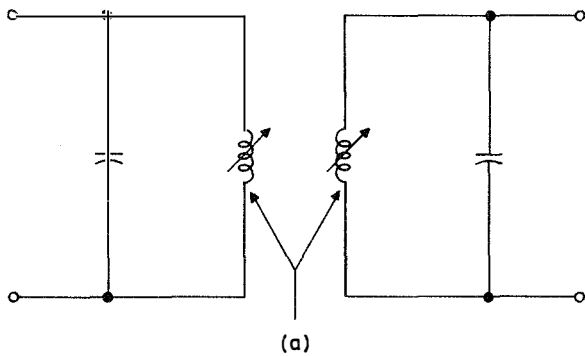
Audio transformers are required to handle moderate ranges of power as compared with the range of power handling capabilities covering the commercial power field. However, audio transformers may be required individually to handle a wide range of frequencies with little or no frequency discrimination. Special designs and manufacturing procedures are thus necessary to insure the required coverage of power and frequency range with optimum fidelity of signal reproduction.

4.6. RF TRANSFORMERS IN TUNED CIRCUITS

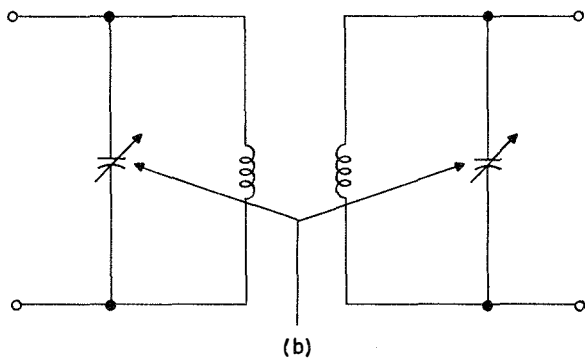
The over-all frequency range that may be encountered in RF transformers in communications equipment will vary from approximately 200 kc to 400 mc. The VLF radio frequency spectrum is considered to go as low in frequency as 30 kc and some VLF RF transformers will be found in land-lines carrier equipment. Individual low power RF transformers are often required to cover approximately a 3.5/1 frequency range as in tunable RF receiver circuits.

Receiver front ends may require that RF transformers operate at power levels less than 1 micro-watt for example (1 micro-volt across 50 ohms), while some transmitters may require output transformers that will handle 100,000 watts of RF power.

In order to obtain selectivity of desired radio frequency signals from an almost unlimited number thereof, it is necessary to design RF circuits to provide for this characteristic, hence most RF transformers are used as components of tuned circuits to provide



(a)



(b)

Figure 4-14

signal selectivity as well as interstage inductive coupling of the desired signal.

A. Fixed-Tuned or Aligned Stages

Fixed-tuned, or aligned stages are used almost exclusively for receiver intermediate fre-

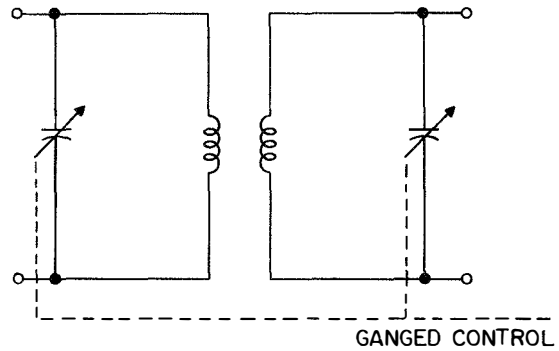


Figure 4-15

quency stages, and for the RF stages in fixed-tuned receivers. Their alignment usually requires the services of a well qualified technician.

Two generally accepted methods of aligning such stages are depicted in Figure 4-14a showing circuitry that is aligned by adjusting the inductance of the transformer, and in Figure 4-14b, showing circuitry that is aligned by adjusting the associated capacitors.

B. Tunable Stages

Tunable RF stages are provided when it is desired to select any one of a band of signal frequencies at will. Although several stages may be so tuned, the individual stages are tracked and aligned so that they may all be tuned simultaneously by means of a common ganged control as indicated in Figure 4-15.

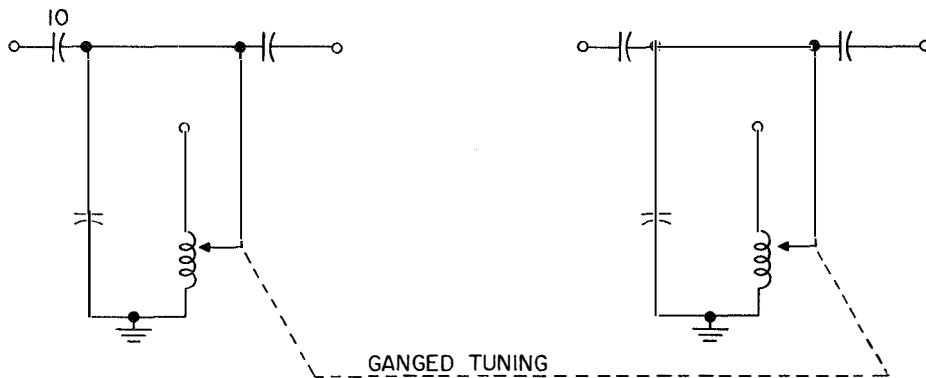


Figure 4-16

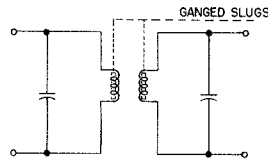


Figure 4-17

In some receivers, such as the RV-5, the tuning is accomplished by means of ganged-tuned variable inductances called inductuners as indicated in Figure 4-16. The inductance is varied by moving a slider around the turns of the inductor to obtain the amount of inductance desired.

A third method of ganged tuning is used, usually in automobile receivers only. As indicated in Figure 4-17, ganged slugs for inductor tuning are used.

C. Departures from Inductive Coupling

In FAA receivers there are several departures from strictly inductive coupling although tuned circuits comprised of inductance and capacitance are used.

In the RV-5 receiver, the RF stage coupling is capacitive. See Figure 4-16. Although tuned circuits are used, the individual inductuners inductances are contained in separate shielded compartments so that no magnetic coupling is obtained between stages. Inasmuch as variable inductances are used, the primary and secondary inductance values vary

with tuning and the mutual inductance would vary considerably as the receiver was tuned over its range from 108 to 152 mc. By contrast, X_c of the $10 \mu\mu\text{f}$ coupling capacitors varies from 147 ohms at 108 mc to 105 ohms at 152 mc, either value being very small when compared to the high impedance circuits with which they are associated, thus the sensitivity of the receiver is affected very little from one end of the band to the other.

The intermediate frequency stages of the RV-5 are also capacitively coupled, the primary and secondary circuits being contained in separate shield cans. See Figure 4-18. With capacitances of 5.6 and $6.0 \mu\mu\text{f}$ used at the intermediate frequency of 4.5 mc, the reactances are approximately 6300 and 5900 ohms, respectively. Again this type of coupling minimizes interaction between the slug-tuned primary and secondary inductances, thus facilitating alignment. Additionally, it provides for a dc-isolated test point for measuring or injecting signal, and reduces the change of circuit impedances when low-impedance test equipment is connected thereto.

The AN/URR-13 UHF receiver uses push-pull RF stages which are capacitively coupled by means of $1.5 \mu\mu\text{f}$ capacitors. See Figure 4-19. Again, the inductances of the tuned stage are magnetically shielded to eliminate inductive coupling.

The RCP receiver uses a combination of inductive and capacitive coupling from the

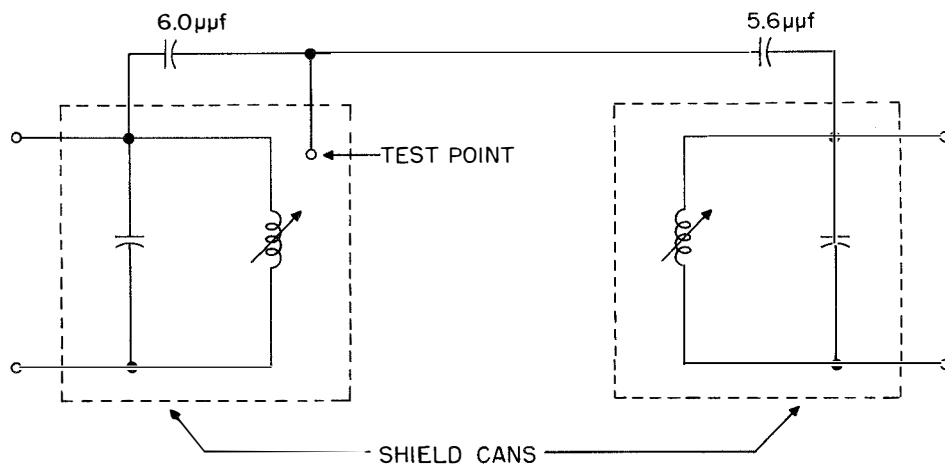


Figure 4-18

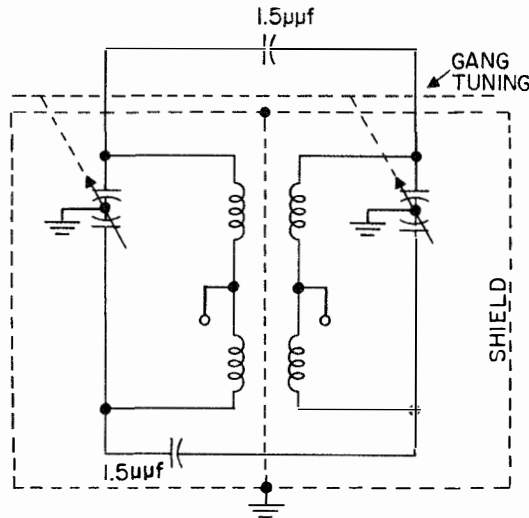


Figure 4-19

plate of its RF amplifier as shown in Figure 4-20. While the untuned primary inductance offers low impedance to RF at the low end of the band, and high impedance at the high end, the capacitance offers high impedance at the low end of the band and low impedance at the high end. Thus relatively constant coupling across the whole band is obtained. The capacitance is comprised of a few turns of open-ended coil wound into the secondary winding.

4.7. RESONANCE

The condition of resonance is obtained in an ac circuit containing L, C and R when that circuit is particularly responsive to a specific frequency in the circuit. In most instances,

that responsive condition is sharply defined and distinguishable from non-resonant conditions, whether in a circuit in which the L, C and R elements are connected in series or in parallel. This peculiarity of response is particularly useful in selecting specific desired frequencies to the exclusion of other existing frequencies.

A. Series Resonance

A series LCR circuit is said to be resonant at the frequency at which $X_L = X_C$. At this frequency the circuit impedance is,

$$Z = R + jX_L - jX_C \text{ or } Z = R. \quad (4-30)$$

See Figure 4-21.

Under this condition, the circuit current, I, is limited by the resistance of the circuit only, and will be very high if R is small. (From equation 4-1, $I = E/R$. However, the emf across L and C individually may be several times the value of E_g .

$$\frac{E_L}{E_g} = \frac{E_C}{E_g} = \frac{X_L}{R} = \frac{X_C}{R} \quad (4-31)$$

Where: E_L and E_C are the emf across the inductance and capacitance, respectively, and E_g is the generator emf.

If, $X_L = X_C$, then $2\pi fL = \frac{1}{2\pi fc}$, and from this it follows that,

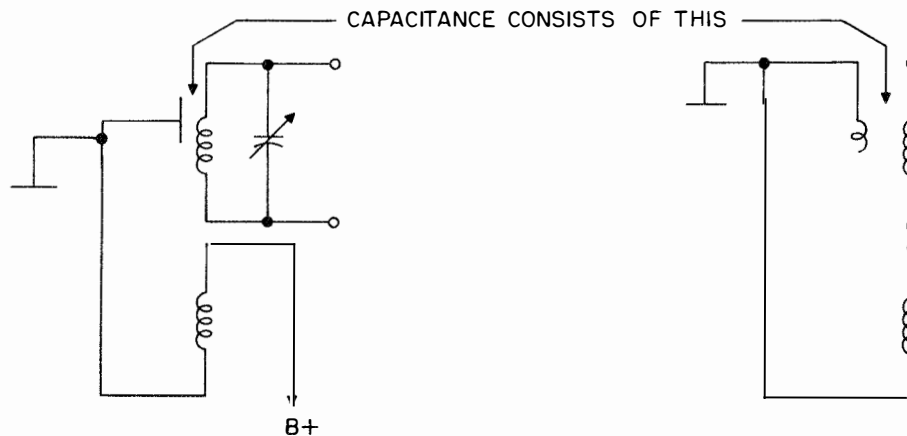


Figure 4-20

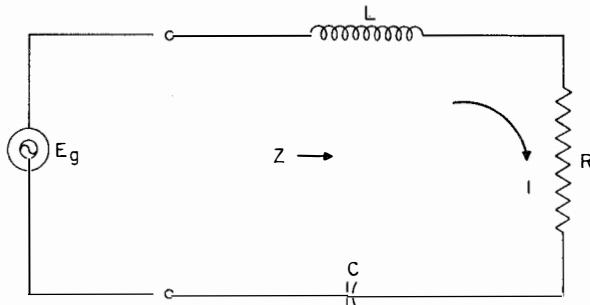


Figure 4-21

$$f_r = \frac{1}{2\pi\sqrt{LC}} \quad (4-32)$$

Where: f_r designates the resonant frequency.

At frequencies lower than f_r , $X_c > X_L$, the circuit is capacitive, and the current leads the applied emf. At frequencies higher than f_r , $X_L > X_c$, the circuit is inductive, and the cur-

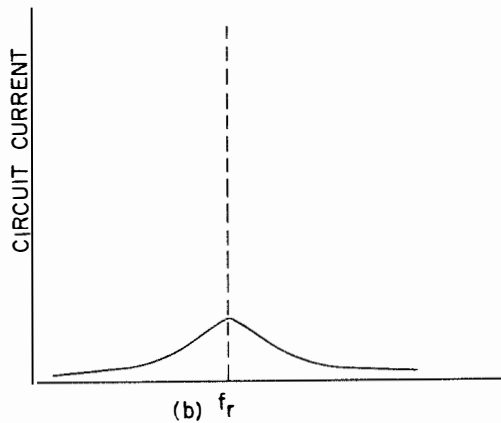
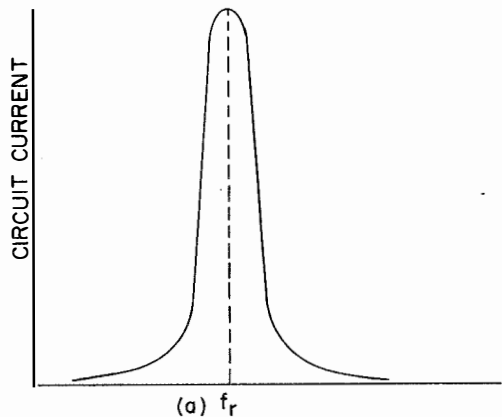


Figure 4-22

rent lags the applied emf.

In a series LCR circuit, the current at resonance varies inversely as the resistance. Thus, as R approaches zero the current approaches infinity. Or, symbolically represented, as $R \rightarrow 0$, $I \rightarrow \infty$.

Figure 4-22 depicts the relative amounts of series circuit current at resonance and off resonance. Part (a) indicates the response curve when the circuit resistance is very small and part (b) indicates the circuit response when the resistance is large.

B. Parallel Resonance

In a parallel resonant circuit, there are three general definitions for resonance as follows:

1. The frequency at which $X_L = X_c$ (used

where $\frac{X_L}{R} > 10$).

2. The frequency at which line current becomes minimum.
3. The frequency at which the parallel circuit acts as a pure resistance.

If the first definition of parallel resonance is used, then the resonant frequency is found from the same formula used for the series resonant circuit, namely equation 4-32, which

states, $f_r = \frac{1}{2\pi\sqrt{LC}}$

For definition 2, the resonant frequency may be found from the equation;

$$f_r = \left(\frac{1}{2\pi}\right) \left[\sqrt{\frac{1}{LC} - \frac{R^2}{L}} \right] \quad (4-33)$$

For definition 3, the resonant frequency may be found from the equation;

$$f_r = \left(\frac{1}{2\pi}\right) \left[\sqrt{\frac{1}{LC} - \frac{R^2 C}{2L^3}} \right] \quad (4-34)$$

In equations 4-33 and 4-34, it can be seen that as R becomes increasingly small, that these equations revert to equation 4-32, and when $X_L/R \cong 10$ the first definition is considered to accurately express resonance for all practical purposes. Referring to Figure 4-23, the total impedance is,

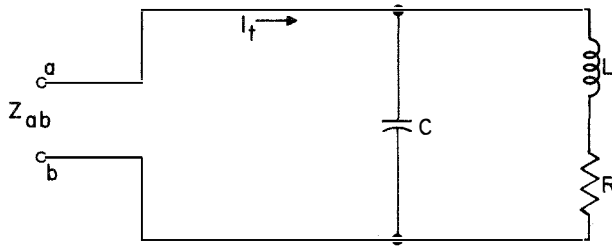


Figure 4-23

$$Z_{ab} = \frac{-jX_c(R+jX_L)}{-jX_c + (R+jX_L)}$$

At resonance, $Z_{ab} = \frac{X^2}{R} - jX_c$ (4-35)

If $X/R=10$, then $Z_{ab}=(100-j10)R=100.98R -5.55^\circ$ which substantiates the statement that as X_L/R becomes greater than 10, that the first definition for parallel resonance holds, and that when X_L/R becomes less than 10, that consideration should be given to definitions 2 and 3.

At frequencies lower than f_r , $X_L < X_c$ and the circuit is inductive inasmuch as more current exists in the inductive leg. At frequencies higher than f_r , $X_L > X_c$ and the circuit is capacitive.

In a parallel circuit such as Figure 4-23, the line current is minimum at resonance, and decreases as R decreases.

In a parallel circuit such as Figure 4-24, the circuit impedance is;

$$Z_{in} = \frac{RX_LX_c}{R(jX_L - jX_c) + X_LX_c}$$
 (4-36)

and at resonance, $Z_{in}=R$. Thus, the line current is minimum when the circuit is

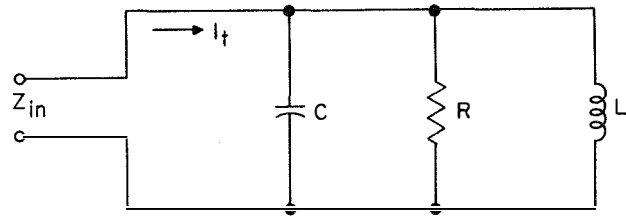


Figure 4-24

resonant, and increases as R decreases. In parallel resonant circuits maximum impedance is obtained at resonance.

C. Transformer-Coupled Resonant Circuits

Although the parallel circuits were described as resonant, the parallel resonant circuits are sometimes called anti-resonant circuits.

Figure 4-25 portrays simple transformer-coupled resonant circuits. The primary tuned circuit as seen from the generator G is an anti-resonant circuit, presenting a high impedance to the generator. In the secondary circuit, L_s assumes the role of the generator G_s and this resonant circuit appears to G_s as a series-resonant circuit having low impedance. Thus, a large circulating current exists through L_s and C_s and the emf existing individually across L_s and C_s is very high. Across L_s and C_s in parallel a high impedance exists as seen from the load Z_L . Thus the secondary, depending upon the point of reference, may appear as both resonant and anti-resonant.

4-8. SELECTIVITY (Bandpass)

Inductive coupling is a factor affecting selectivity, and selectivity is a function of inductive coupling.

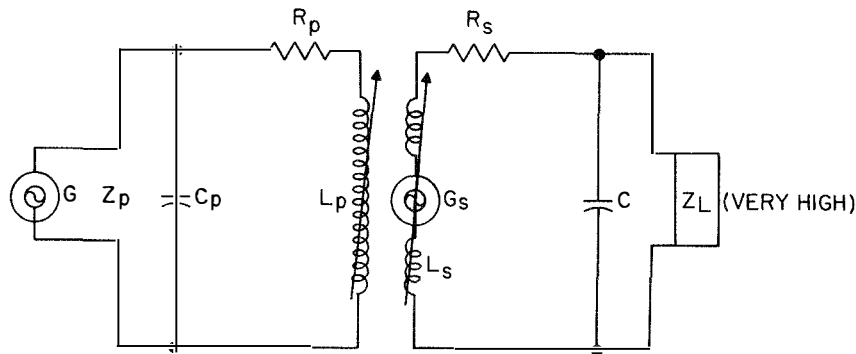


Figure 4-25

As seen in previous paragraphs, circuits consisting of inductance and capacitance (L and C) exhibit special characteristics at certain frequencies, thus rendering them "selective". The term "selectivity" or "bandpass" applies not alone to the reaction of a circuit at one specific frequency but to the concept of the behavior thereof over a range of frequencies as determined by the design of the circuit.

Many factors affect the selectivity of a circuit, among them being the degree of coupling, the Q of the components, Q of the circuit, method of tuning, relative value of L and C, AVC, and cascading of the successive stages.

A. Merit or Q

The "merit" or Q of a circuit, or circuit component, is the ratio of its reactance to its resistance. Inasmuch as high quality capacitors are considered to have negligible resistance, the Q is seldom considered except when testing for quality, but the Q of an inductance is usually of concern because it indicates the ability to store energy instead of consuming it. The merit of a coil is,

$$Q = \frac{X_L}{R} = \frac{I^2 X_L}{I^2 R} = \frac{P_{\text{stored}}}{P_{\text{consumed}}} \quad (4-37)$$

The Q of air core inductances may vary from approximately 900 at 400 kc to approximately 183 at 15 mc, depending on the size and shape of the coil, and the size and type of conductor used to wind the coil. The Q of compact coils may be increased in many instances by use of core material made of finely divided grains of magnetic material bound together with suitable binder.

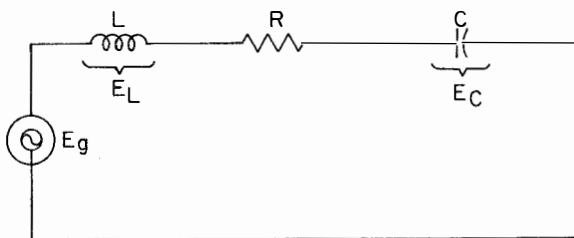


Figure 4-26

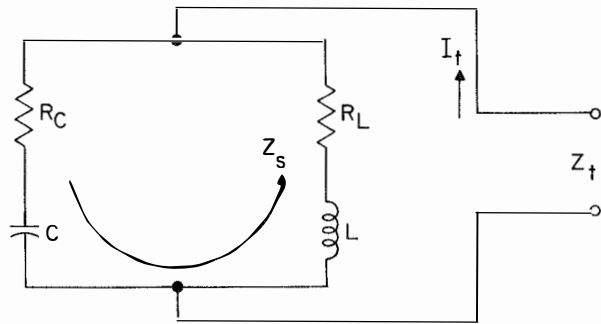


Figure 4-27

In a series resonant circuit (Figure 4-26) R is considered to consist of the resistance of the inductance, the resistance of the conductors, and such resistive components as may be introduced into the circuit. From equations 4-31 and 4-37, we may determine that,

$$E_r = E_c = Q E_i \quad (4-38)$$

In a parallel resonant circuit (Figure 4-27) the Q of the circuit is,

$$Q = \frac{X_L}{R_s} \quad (4-39)$$

Where: R_s is the series resistance of the circulating current path.

Also:

$$Q = \frac{Z}{X_L} \quad (4-40)$$

Where: Z is the parallel impedance of the circuit.

The Q may also be expressed as,

$$Q = \frac{I_L}{I_r} \quad (4-41)$$

1. EFFECT OF Q ON SELECTIVITY

Referring to Figure 4-28, the bandwidth measurements of a series circuit taken at half-power points, i.e., at frequencies above and below resonance where the power dissipated in the circuit is one-half that at resonance, is very closely,

$$\frac{\Delta f}{f_r} = \frac{1}{Q_s} + \frac{\omega^2 L_M^2}{R_p X_s} \quad (4-42)$$

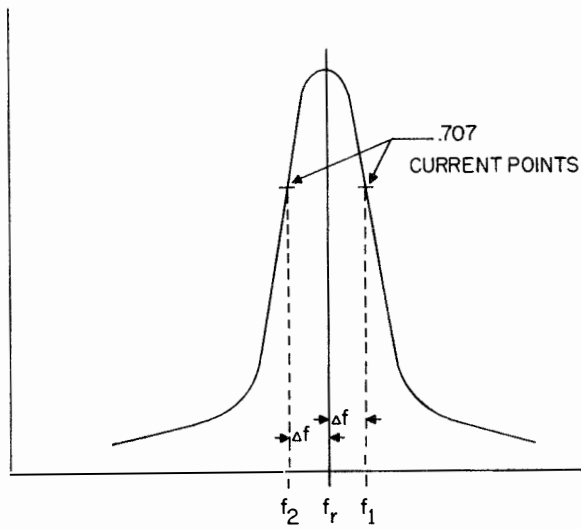


Figure 4-28

or, when critically coupled,

$$\frac{\Delta f}{f_r} = \frac{2}{Q_s} \quad (4-42a)$$

Where: $f = f_r - f_2 = f_1 - f_r$, $Q \geq 10$,

f_r is the resonant frequency,
 R_p is the primary resistance,
 and X_s is the secondary reactance.

Figure 4-28 shows the 0.707 current points which are the half-power points inasmuch as $P = I^2 R = (0.707)^2 R = 0.5^2 R$. The bandwidth

measurements taken at $\Delta f = \frac{f_r}{Q}$ represent the

place where the current is reduced to 0.447 of the value at resonance, or the one-fifth power points. Far off-resonant points, where

$\Delta f = \frac{3f_r}{Q}$ may be determined by the formula,

$$\frac{\text{Off-resonance current}}{\text{Current at resonance}} = \frac{1}{\left[Q\gamma \left(1 - \frac{1}{\gamma^2} \right) \right]} \quad (4-43)$$

Where: $\gamma = \frac{\text{actual frequency}}{\text{resonant frequency}}$

In anti-resonant circuits where $Q > 10$, (See Figure 4-29), the impedance of the circuit

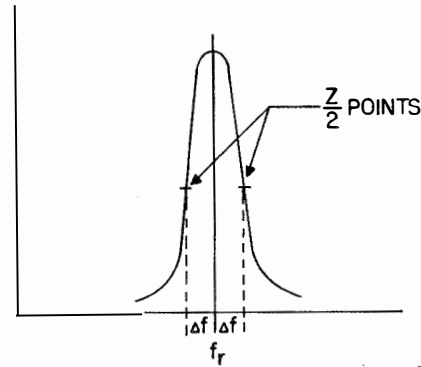
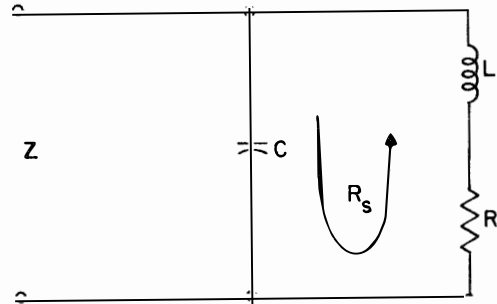


Figure 4-29

instead of the circuit current is considered. At the one-half impedance points,

$$\Delta f = \frac{f_r}{2Q} \quad (4-44)$$

Where: $Q = \frac{X_r}{R_s}$

To obtain impedance values far from resonance,

$$\frac{Z \text{ from resonance}}{Z \text{ at resonance}} = \frac{1}{\left[Q\gamma \left(1 - \frac{1}{\gamma^2} \right) \right]} \quad (4-45)$$

B. Degree of Coupling

In inductively coupled tuned circuits which are tuned to the same frequency, the emf induced in the secondary is a function of frequency and the degree of coupling. Four degrees of coupling are generally recognized, viz., under coupling, critical coupling, transitional coupling, and over coupling. The general response curves, emf vs. frequency, for each type of coupling are shown in Figure 4-31.

When the amount of coupling between the two tuned circuits (Figure 4-30) is small

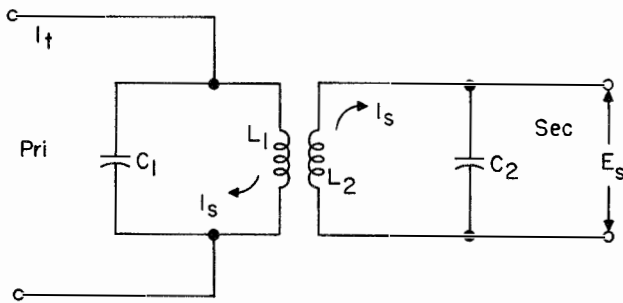


Figure 4-30

enough that an increase of coupling would increase the induced emf across L_2 at resonance, the circuit is said to be *under* coupled.

When the degree of coupling of the two circuits is such that the maximum possible secondary emf is being developed at resonance, the circuits are *critically* coupled.

When two inductively coupled circuits are coupled to such a degree that the flattest possible response curve is obtained, the circuits are said to be *transitionally* coupled.

Two circuits that are tuned to the same frequency and coupled inductively are considered to be *over* coupled if the coupling coefficient, k , is large enough to cause the secondary response curve to be double-peaked. In high Q circuits, the amplitude of the two peaks are very nearly equal. In low Q circuits, the low frequency peak tends to be higher in amplitude than the high frequency peak. In either case, the average amplitude of the two peaks will be very close to the amplitude of the peak when the circuits are transitionally coupled.

The scope of gain-bandwidth considerations is too lengthy to be treated in this chapter, but

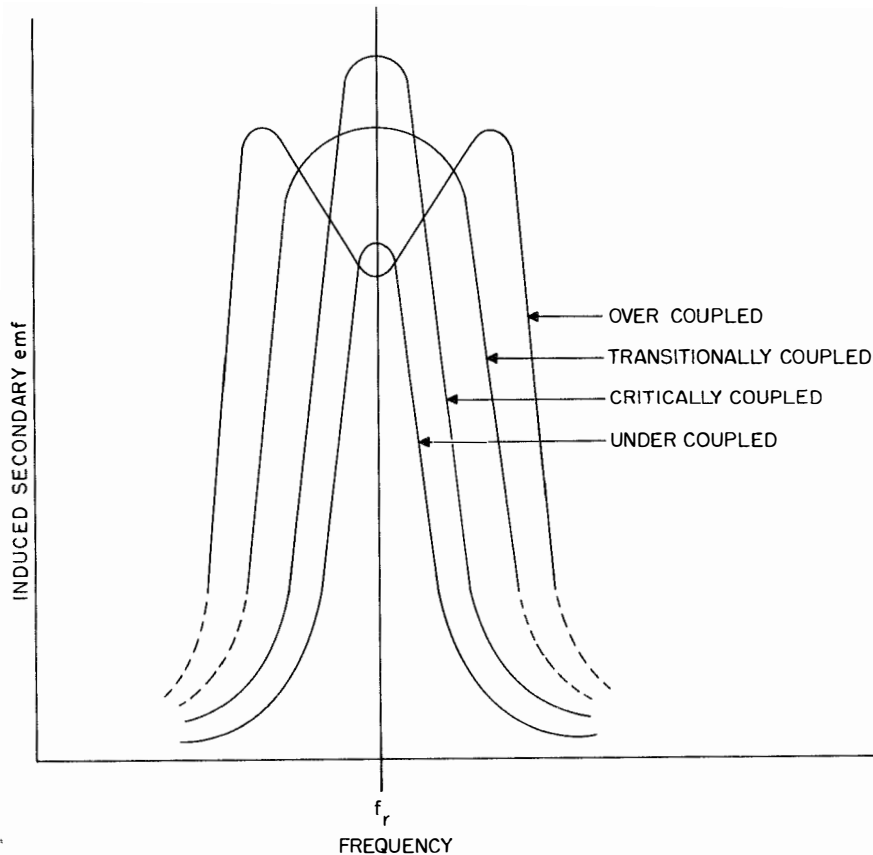


Figure 4-31

a thorough analysis is given in the Electronic Designer's Handbook by Landee, Davis and Albrecht, starting on Page 13-3.

C. Effect of Size of L and C on Selectivity

There are an unlimited number of combinations of L and C that will give resonance at any specified frequency. However, practicality limits the extremes of the L/C ratio. In a high Q series LCR circuit, where L/C is low, the tuning is broad and the selectivity curve is symmetrical about the resonant frequency. When the L/C ratio is large, the tuning is sharp and the circuit becomes more selective on the low frequency side of resonance. See Table II and III, and Figure 4-32.

TABLE II

($C=900 \mu\mu\text{f}$ and $L=.1128 \text{ mh}$, giving resonance at 500 kc)

Freq (kc)	X_c (ohms)	X_L (ohms)	X_{total} (ohms)	Z (ohms)
100	1768	71	-j 1697	R-j 1697
200	884	141	-j 743	R-j 743
300	589	212	-j 337	R-j 377
400	441	283	-j 158	R-j 158
500	353	353	0	R
600	294	424	+j 130	R+j 130
700	252	494	+j 242	R+j 242
800	221	565	+j 344	R+j 344
900	196	636	+j 440	R+j 440

TABLE III

($C=100 \mu\mu\text{f}$ and $L=1.012 \text{ mh}$, giving resonance at 500 kc)

Freq (kc)	X_c (ohms)	X_L (ohms)	X_{total} (ohms)	Z (ohms)
100	15900	636	-j 15264	R-j 15264
200	7950	1270	-j 5680	R-j 5680
300	5300	1910	-j 3390	R-j 3390
400	3980	2540	-j 1340	R-j 1340
500	3180	3180	0	R
600	2670	3820	+j 1150	R+j 1150
700	2270	4450	+j 2180	R+j 2180
800	1990	5090	+j 3100	R+j 3100
900	1770	5730	+j 3960	R+j 3960

By taking the reciprocals of the reactances in Tables II and III, and plotting the susceptances as in Figure 4-32, it will become apparent that a small L/C ratio in a high Q

parallel circuit will give very sharp tuning, and a selectivity curve that is sharper on the low frequency side of resonance.

D. Other Factors Affecting Selectivity

Among other factors affecting selectivity are avc, stage gain of amplifiers, cascading of stages, etc., which are too detailed for treatment here. Communications and Radio Engineering texts by Terman, Everitt, Henny, and Lande, Davis & Albrecht may be helpfully employed for amplified treatment of these items as well as other information contained in this chapter.

4.9. DESIGN CONSIDERATIONS

The special intended use for any circuit, including frequency range, selectivity, adaptability to varying modes of operation, power capabilities, etc., dictates special design considerations in each instance. Special considerations of FAA receiver inductively coupled circuits are of special interest in this chapter.

A. The RUQ and RVA Receivers

The use of the RUQ and RVA receivers indicates need for an RF input section operating in the 108-152 mc range, with enough selectivity to give attenuation $\cong 80$ db for all signals 36.6 mc away from the receiving frequency in order to furnish ample image frequency rejection.

In the relatively broad-band, high-frequency first intermediate frequency section of these receivers, enough selectivity is desired to provide for adequate signal-to-noise ratio. At the same time, the band-pass should be broad enough so that a minor shift in alignment of the section would not result in loss of the desired signal prior to conversion to the low intermediate frequency.

The second (low frequency) intermediate frequency section is designed to provide sufficient bandwidth that the signal of any air-ground transmitter operating within tolerance, may be received without undue loss of signal; also to provide sufficient selectivity that the nearest adjacent communication frequency is effectively rejected. Inasmuch as the transformers are transitionally coupled, provisions

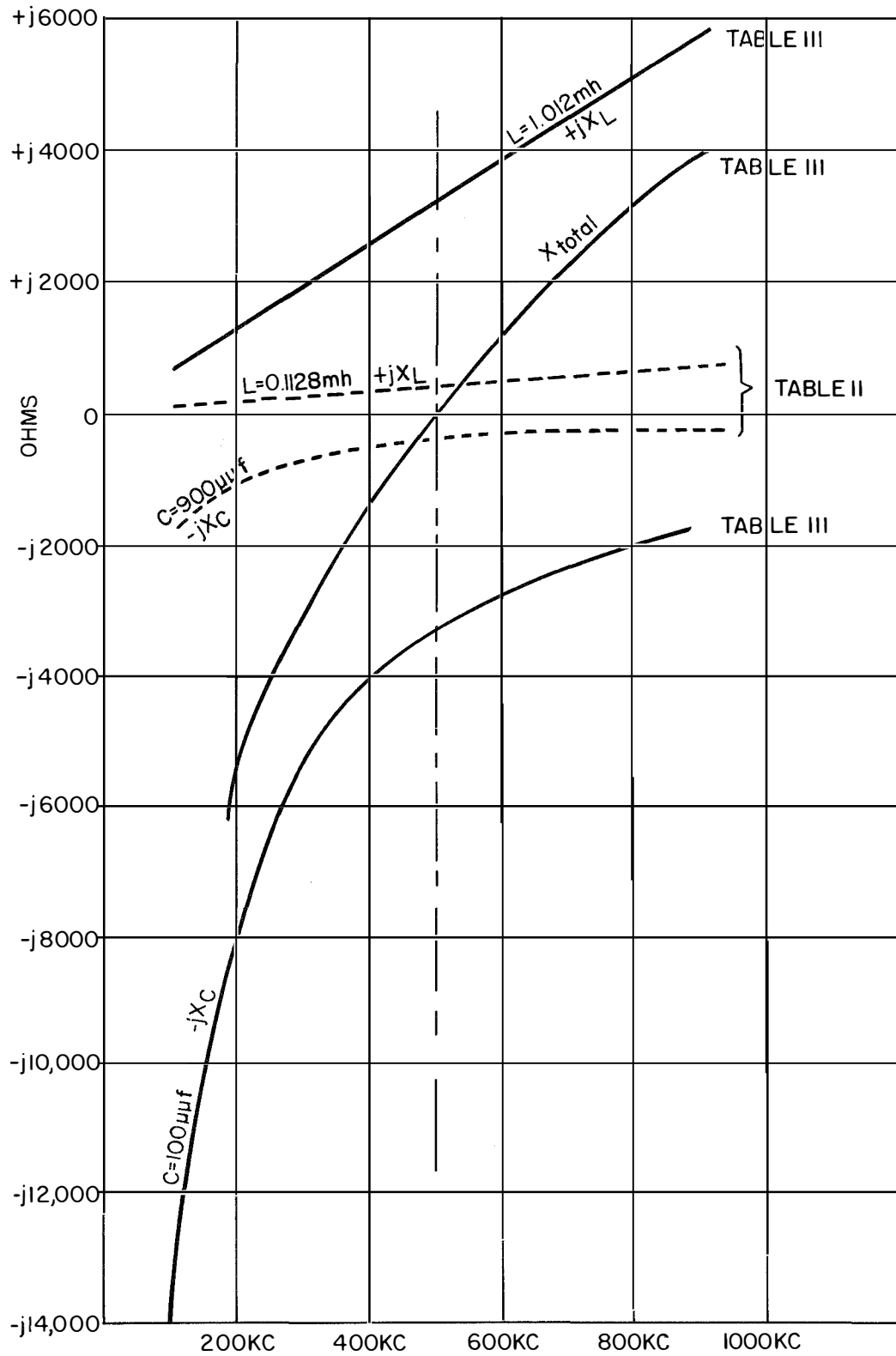


Figure 4-32

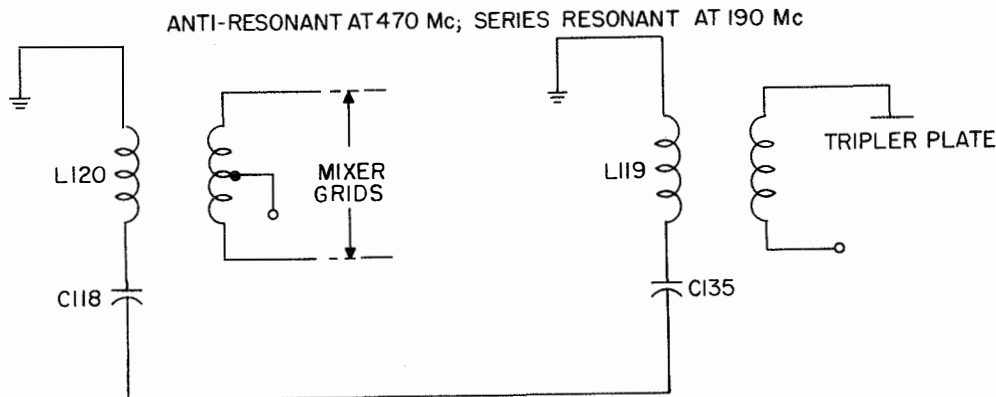


Figure 4-33

to detune one-half of each inductively coupled stage while tuning the other half is required and is indicated in the alignment instructions.

B. The RCP Receiver

The RCP and RCQ receivers were designed to operate on both voice and CW, and in fixed-tuned and tunable modes of operation. The intermediate frequency stage transformers were designed to have high Q to obtain selectivity. Because of the need in times past to receive AM signals of varying nature, and to compensate for some amount of transmitter drift, provisions were made to broadband the intermediate frequency stages by closing the circuit of the tertiary winding on the transformer coil forms. This causes a loading effect on the transformer, thus broadening the response curve.

C. The AN/URR-13 Receiver

In the AN/URR-13 receiver, deliberate use of coil shunt capacity, which is usually minimized as much as possible, was made to provide more nearly linear response of the receiver across the entire band. Energy fed from the plate of the tripler to the mixer grid is fed through an inductively coupled circuit consisting of L_{119} , L_{120} , C_{118} and C_{135} . This circuit is series resonant at about 190 mc (corresponding to a receive frequency of about 170 mc), and anti-resonant by virtue of coil

shunt capacitance at about 470 mc, thus essentially constant current exists in L_{119} and L_{120} at all frequencies within the receive band. See Figure 4-33.

4-10. SUMMARY

Inductively coupled circuits are used throughout the range of the ac spectrum; to isolate two or more circuits from transfer of dc energy while coupling them with ac energy; as voltage regulator components, to effect change in emf levels; as frequency-selective circuit components; as impedance matching or transforming devices; as frequency changing or phase shifting components; and as components with which to modify circuit characteristics.

In this chapter the reader has been given a quick review of general LCR circuit theory and circuit characteristics, and provided with a concept of the applications of inductive components and inductively coupled circuits to FAA communications receiver application. Although it is impossible to cover such an extensive subject any more than sketchily in a chapter such as this, the thoughts provided herein should furnish the reader with concepts that lead him to channels for further investigation even as it provides a fundamental working concept to facilitate his work on FAA equipment.

CHAPTER 5

NOISE

5-1. GENERAL

Intelligent maintenance of electronics equipment demands a sound understanding of the basic problems encountered in design. In no other way can some of the obscure malfunctionings of complex communications type equipment be found. Since communications receivers must accurately reproduce the intelligence conveyed by the originating signal, a general understanding of the effects of noise on receiver operation is necessary. Certain adjustments in the equipment are based on the existing noise level. In addition, certain malfunctionings of the receiver or associated equipment may occur which would be extremely difficult to trace if the general properties of noise and their effects are not understood.

Noise can be defined as any random or aperiodic signal which interferes with the extraction of intelligence from a desired signal. It would appear that by cascading amplifier stages, almost any degree of amplification is attainable. However the effect of noise is to limit the minimum signal amplitude which the receiver can demodulate satisfactorily since these noises mask small signal voltages received.

5-2. NOISE SOURCES

The source of noise contained in the output of a receiving system may be either external to, or located within the system. Sources which are external to the receiving system fall into two general classes i. e., natural disturbances and man-made static. Noise sources located within the receiving system may be sub-divided into those generated in conductors due to thermal agitation, and those originating

within the vacuum tube due to thermionic emission.

A. *Thermal Agitation*

Thermal agitation noises consisting of minute, unpredictable variations of voltage and current exist in all electrical circuits as a result of the random motion of free electrons. The frequency spectrum resulting from the motion of these electrons is almost continuous in range and thus uniformly distributed over the useful radio frequency spectrum. If a sufficiently sensitive measuring device could be attached across any two points of a conductor, a voltage fluctuation would be measured even though no voltage was applied to it. This voltage fluctuation would increase as the conductor was heated. These minute variations will have no appreciable effect on most electrical circuits. It is only in those circuits involving high gain, such as found in a communications receiver, that these voltages and currents will assume troublesome proportions. Not only the weak signal but also these minute aperiodic voltages and currents will be amplified by the receiver.

It is not the signal as such that determines the useful gain, but rather the ratio of the signal magnitude to the noise magnitude that determines the minimum usable signal. Thermal agitation is generated in the antenna conductors as well as in the receiver circuits.

Noise generated within the early stages of the receiver is the major offender in that it will be amplified by all of the stages that follow. If several stages are cascaded, the later stages have little effect on the overall noise.

Noise introduced into a circuit by thermal agitation may be calculated from the equation:

$$E_n = \sqrt{4KT\Delta fR} \quad (5-1)$$

Where; E_n is rms noise voltage developed.

K is 1.38 times 10^{-23}

T is absolute temperature in degrees Kelvin.

Δf is the bandwidth of the frequency under consideration.

R is the resistive component in ohms.

For example, a cold solder connection in the front end of a receiver would increase "R" and consequently the noise generated.

B. Vacuum Tube Noise

Noises generated within vacuum tubes stem from the following principal sources:

1. SHOT EFFECT

The number of electrons emitted by the cathode varies minutely from instant to instant and this random emission with respect to time introduces a voltage in the circuits through which these electrons pass. The space charge obtained in a vacuum tube under normal operating conditions smooths out the shot effect, due primarily to variations in the space charge currents which reduces the shot effect by about one-half.

2. FLICKER EFFECT

Variations in electron emission at a relatively low frequency, below 500 cps, are sometimes experienced from the entire cathode surface due primarily to impurities, diffusion, evaporation and ion bombardment. The magnitude may exceed shot-effect noise by several times at low audio frequencies.

Flicker effect can be reduced by careful cathode preparation and does not place a fundamental limitation on amplifier sensitivity.

3. ION EFFECT

Electrons passing to the plate may strike gas molecules causing short energy bursts that release free electrons contained in the molecule, forming positive ions. The ions

are attracted to the grid resulting in electron flow through the external grid circuit to neutralize the positive ions. Since the electron collision is aperiodic in nature, the grid and plate currents resulting constitute a source of noise. Collision ionization effects appear mostly below 10 mc.

4. INDUCED GRID NOISE

Ultra-high frequency components of the fluctuations in space charge induce variations in the grid charge, which in turn reacts to vary the current flow in the tube. This is not a result of electrons being intercepted by the grid but occurs when electrons comprising the space current approach the grid plane. These electrons induce a positive current in the grid circuit and a negative current upon leaving the grid plane and proceeding to the plate. Due to transit time in the tube, induced grid noise increases linearly with frequency and becomes a trouble source particularly above 30 mc.

5. INTERCEPT NOISE

Resulting from a random distribution of electrons between the plate and other positive tube elements, intercept noises are from three to five times greater in multi-electrode tubes than in triodes due to the variations in plate current as a result of the electron distribution between these multiple elements.

C. Natural Disturbances

Natural disturbances stem from the following principal sources:

1. ATMOSPHERICS OR STATIC

The energy generated by electrostatic discharges between clouds decreases as the frequency is increased, so that it is relatively unimportant at ultra high frequencies. Static is a natural radio signal and under suitable conditions this impulse type of energy is reflected from the ionosphere and may travel great distances to create noise in a receiver at a particular location. Night time propagation is usually greater than day time propagation, since there is greater attenuation of the sky wave during the daylight hours.

2. COSMIC RADIATIONS

Originating in regions beyond the earth's atmosphere, solar and galaxy radiation is believed due principally to celestial thermal noise radiated by thermal bodies. Cosmic radiations are quite weak, but may limit a receiver in the 10 to 300 mc portion of the spectrum in the absence of other noise sources.

D. Man-made Noise

Noise is generated by most electrical appliances and devices. Almost any item that produces an electrical spark can interfere with reception. Man-made noises are generally carried by power lines or radiated in the neighborhood of the antenna.

E. Breakdown Noise

This type of noise exists in addition to thermal noise in some types of resistors. Contact and breakdown noise is due to the breakdown of insulation or the interruption of minute paths within components. This is often an undesirable effect of corroded center conductors and together with insulator breakdown is responsible for leakage at antenna installations.

5-3. FIGURE OF MERIT

The noise components appearing in the output of a receiver are those at frequencies within the overall receiver bandwidth, thus the wider the bandwidth, the greater the amount of noise produced. Antenna radiation noise, which is a form of thermal agitation in objects located within the radiation pattern of the antenna, tube noise and thermal agitation are all similar in nature and distributed continuously in frequency. Since the noise at the input is a function of bandwidth, and the noise generated within the receiver is also a function of bandwidth as determined from Formula 5-1, the ratio of noise output to noise input is independent of bandwidth and describes the quantity of noise generated within the receiver circuits. This ratio is the noise figure. In a practical system it is the source noise as well as signal and consequently the noise figure is the ratio of the input signal-

to-noise ratio to the output signal-to-noise ratio.

$$F = \frac{\frac{S_i}{N_i}}{\frac{S_o}{N_o}} \quad (5-2)$$

Where; S_i/N_i is the input signal-to-noise ratio.

S_o/N_o is the output signal-to-noise ratio.

F is the noise figure, sometimes called the figure of merit of a receiver.

Since the noise with which a signal must compete in the output of the receiver determines the minimum signal which can be utilized, the noise figure is among the most important receiver characteristics.

5-4. NOISE REDUCTION

The best noise reduction method is to attack the noise at its source. This method is often practicable and can be accomplished through the use of line filters or through repositioning the offending device. The modern design of receiving equipment has reduced to a minimum the problem of noisy reception due to receiver design itself, primarily due to tube design, internal wiring, interstage noise suppression, the care exercised in shielding of radio frequency coils, mechanical rigidity and complete shielding of the receiver itself. If the receiver incorporates all of these, it will be improbable for it to pick up any stray electromagnetic fields produced by ignition system, contact interruption of switches, loose connections or any other external source of impulse excitation. However, when the antenna is connected to the receiver, all such electrical disturbances will be picked up and amplified through the receiver. Hence, great care to reduce radiation from undesirable sources must be exercised by shielding when possible.

A. Filters

Since noise is a function of bandwidth, it frequently becomes desirable to discriminate between frequency bands, accepting some fre-

quencies and rejecting others in a particular receiver. Combinations accomplishing this purpose are called "filters." They differ from simple resonant circuits by providing an essentially constant attenuation over the band they accept. An ideal filter would produce no attenuation in the band desired and would give infinite attenuation at all other frequencies. Actual filters fall short of this ideal, the undesired frequencies being attenuated by a finite amount. Filters are generally treated upon the basis of long lines theory. The effect of a filter inserted between any supply and receiving impedances is a combination of the effect of reflection and attenuation. The component parts of a filter designed for this purpose are lumped inductances and capacitances, which are selected for as low an effective resistance as possible so there will be a minimum of absorption in the desired band.

1. BAND SUPPRESSION FILTER

The cable filter assembly of the AN/URR-13 Receiver shown in Figure 5-1, contains RF filter circuits for the audio output and RF filter circuits for the power input circuits. The filter plates are mounted on a base plate which is attached to the rear wall of the receiver cabinet by means of snap fasteners. The filter cover, when attached to the base plate, constitutes an RF shield.

Inductances L-401, L-402, L-403 and L-404 reject radio frequencies while passing power at 60 cps. Capacitors C-401, C-402, C-403 and C-404 present a low impedance path to ground for radio frequencies while rejecting 60 cps. Inductances and capacitances associated with the audio filter perform the same function, allowing the audio frequencies to pass through the filter to the audio line while bypassing to ground radio frequencies that are picked up by the cable.

B. Noise Limiters

Superheterodyne receivers sometimes make use of noise reducing devices placed at the second detector. This type of device operates on the principle that since noise pulses from impulse sources are of short duration, no appreciable loss of intelligence will result if the receiver is made inoperative during the pulse. These pulses though of short duration are of extremely high amplitude and may have a peak value fifteen to twenty times as great as the incoming radio signal. Some noise limiters limit the maximum peak signal which reaches the output by presenting a high impedance path to the audio signals, while others punch a hole in the signal by removing the entire output for the duration of the noise pulse. The amount of limiting that can be obtained is a function of the audio distortion that can be tolerated.

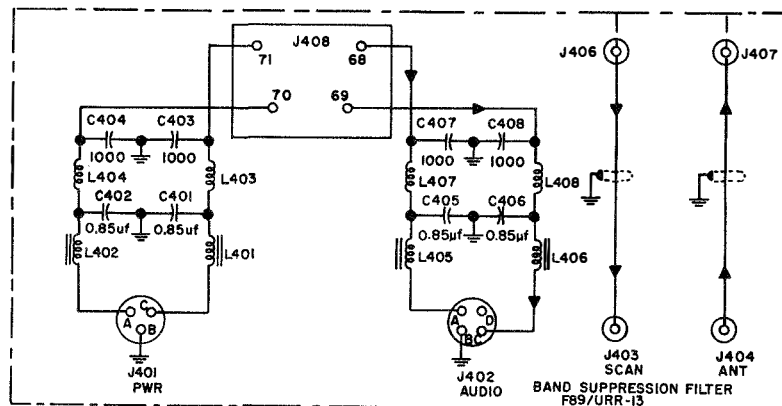


Figure 5-1. AN/URR-13 Receiver Band Suppression Filter

1. SERIES NOISE LIMITER

The series diode clips only positive peaks by limiting upward modulation to a percentage determined by circuit components. Downward modulation peaks in excess of 100% are automatically limited in the detector, inasmuch as the rectified voltage cannot be less than zero. The limiter is biased by dc voltage obtained from a voltage divider comprising the plate load, so that it no longer conducts when short duration pulse voltage exceeds the carrier voltage by a certain percentage. Thus a high impedance path is presented to the noise. This also clips modulation peaks but not enough to impair intelligibility.

2. SHUNT NOISE LIMITER

The shunt noise limiter works on the same principle as the series type, except that during conduction, a low impedance path is presented to the noise and signal, removing both the noise and intelligence for the duration of the noise pulse.

C. RUQ/RVA Detector-Noise Limiter

Referring to Figure 5-2, the diode detector V-9A, one-half 6SN7GT, operates as a rectifier; the diode load consists of R-38 and R-39 in series, with intermediate and radio frequencies bypassed by C-56. The output of the detector is a negative voltage of a nominal -8 volts, upon which is superimposed the audio modulation signal.

The series diode limiter circuit employs diode element V-11B, one-half 6AL5, in a circuit which limits the audio output signal to approximately 40% upward modulation. Since impulse noise of the type encountered at very high frequencies consists of very sharp pulses of several hundred percent modulation, this limiting action greatly reduces the amount of noise energy transmitted to the audio circuits without adversely affecting the intelligibility of voice signals. In the circuit employed, the plate of the diode is connected to the junction of R-38 and R-39 which form a voltage divider. R-40 and C-58 operate as a filter to establish a reference potential at point C. R-41 and R-42 in series form the diode load

resistor across which the output audio signal is developed.

The operation of this circuit can best be understood by assuming V-11B to be an ideal diode; that is, perfectly conducting in one direction and completely non-conducting in the other direction of current flow. With a carrier developed bias of about -8 volts dc at A, the current flow through the circuit will result in dc potentials at points B, C, and D. B and D are the plate and cathode respectively of diode V-11B. Note that C is bypassed to ground so that its instantaneous potential to ground is not appreciably affected by the modulation signal.

When the audio modulation is superimposed on the developed bias, the audio signal will be transmitted through the diode from point B to point D so long as the instantaneous potential of point B does not become more negative than that of point D. If this occurs, the electronic conduction of the diode will be cut off and point D will remain at the potential of point C until conduction is re-established. The upward modulation of the signal transmitted through the noise limiter circuit is limited to approximately 40% modulation, the amount required to instantaneously drive the potential at B from approximately -5 volts dc to about -7 volts dc. The series limiter is self-adjusting; the negative voltages established at A, B, C, and D depend on carrier level. However, the ratio remains unchanged and limiting occurs with modulation levels of approximately 40%. Downward modulation is not affected by the circuit.

The amount of noise energy remaining in the output of the series diode limiter circuit is further reduced by the introduction of noise pulses of about equal amplitude but of opposite polarity to cause noise cancellation. These pulses are generated by noise cancellation detector V-9B, one-half 6SN7GT. This tube operates as an "infinite impedance" type detector in which the intermediate frequency signal is applied to the grid of the tube and the detected output appears at the cathode. In this application, the tube is biased by means of voltage divider R-35, R-32 and R-33 in series to about plus 15 volts, so that only

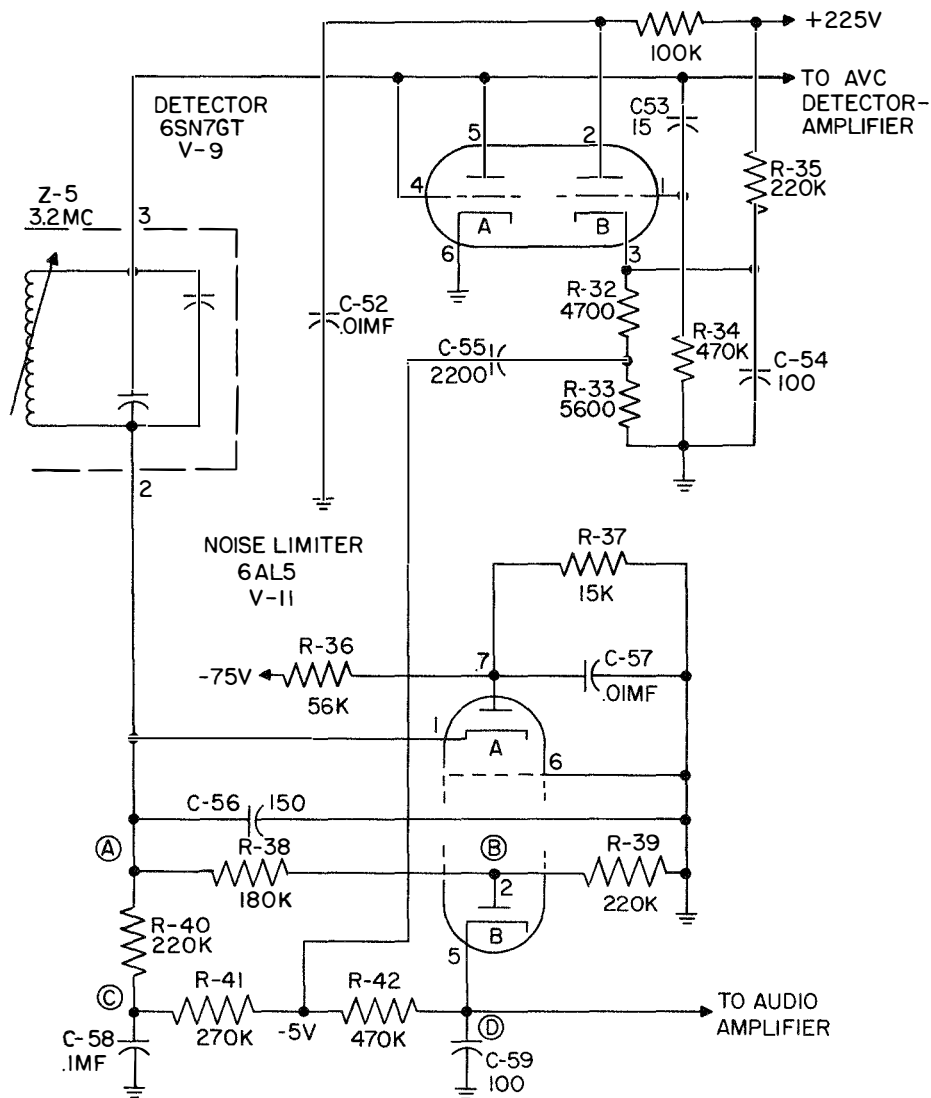


Figure 5-2. RUQ Detector and Noise Limiter Circuits

signals exceeding approximately 100% upward modulation will be detected. Noise pulses will appear as positive pulses in the cathode circuit and are coupled into the output of the series diode limiter circuit by C-55 and R-41. The values of R-32 and R-33 have been chosen so the amplitude of the cancelling pulses will approximately equal that of the original noise pulses.

A shunt type limiter has also been included

in this receiver for the purpose of protecting the avc and squelch circuits from the effects of noise pulses. The shunt type noise limiter works on the same principle as the series, except during conduction a low impedance path is presented to the noise and signal, removing both for the duration of the noise pulse. In this circuit, diode element V-11A, one-half 6AL5, is biased by voltage divider R-36 and R-37 to about -16 volts. When an

intermediate frequency of peak amplitude greater than -16 volts is applied to this circuit, the diode conducts and effectively connects a very low impedance path across the diode load resistors R-38 and R-39, thus limiting the intermediate frequency voltage which can appear across Z-5. This reduces the effect of noise impulses which ordinarily would generate appreciable avc control voltage, thus reducing the sensitivity of the receiver to desired signals. Since the silencer control voltage is also developed by the avc detector amplifier, this limiter also reduces the tendency of the silencer circuit to open in the presence of noise impulses.

D. Carrier Operated Noise Suppressor

Carrier operated noise suppressor circuits are designed to render the receiver inoperative and suppress noise during the absence of a carrier. Two examples of carrier operated noise suppressors will be discussed, although numerous variations are in use based on the same principles.

1. ELECTRO-MECHANICAL CARRIER OPERATED NOISE SUPPRESSOR

Referring to Figure 5-3, the CONS relay, A-12, is equipped with a double coil. The relay is normally controlled by the CONS tube plate current flowing through one of the coils.

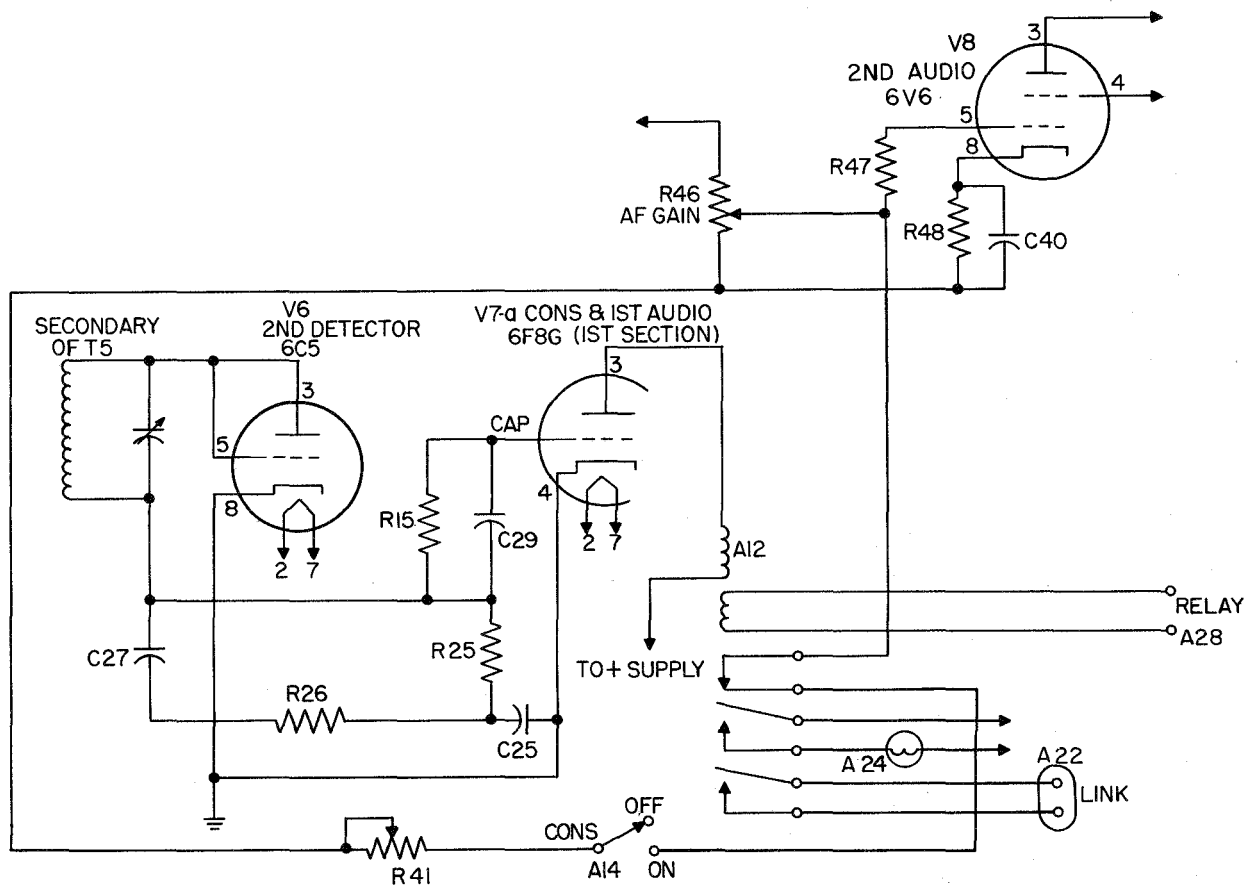


Figure 5-3. RCP Receiver CONS Circuit

The relay may also be controlled by an external 6 volt dc source and thereby mute the receiver even when a carrier is present on the channel to which the receiver is tuned. In Figure 5-3, the CONS relay is shown in the no signal condition (energized). Without an effective carrier there is not sufficient voltage developed across R-25 and R-26 by the second detector to bias the CONS tube V-7A to cut-off. The plate current of V-7A flowing through the high resistance coil of the CONS relay, A-12, energizes the relay and holds its contacts in the position shown. With the relay in this position and the CONS switch on, the output of the receiver is muted due to the shunting action of R-41 across the input to the second audio amplifier, V-8. Lamp A-24 will be out and the link circuit will open. When an effective carrier is received there will

be sufficient voltage developed across R-25 and R-26 to cut off the CONS tube plate current and thereby de-energize the CONS relay. Under this condition the shunting action of R-41 is eliminated and normal output will be obtained from the receiver. Lamp A-24 will be on and the link circuit will be closed.

2. ELECTRONIC CARRIER OPERATED NOISE SUPPRESSOR

The RHZ carrier operated noise suppressor, as shown in Figure 5-4, is one-half of a dual triode 6SL7GT, which performs the combined function of a noise suppressor and audio amplifier. For purposes of circuit description, the individual triodes in the 6SL7GT envelope will be considered as separate tubes.

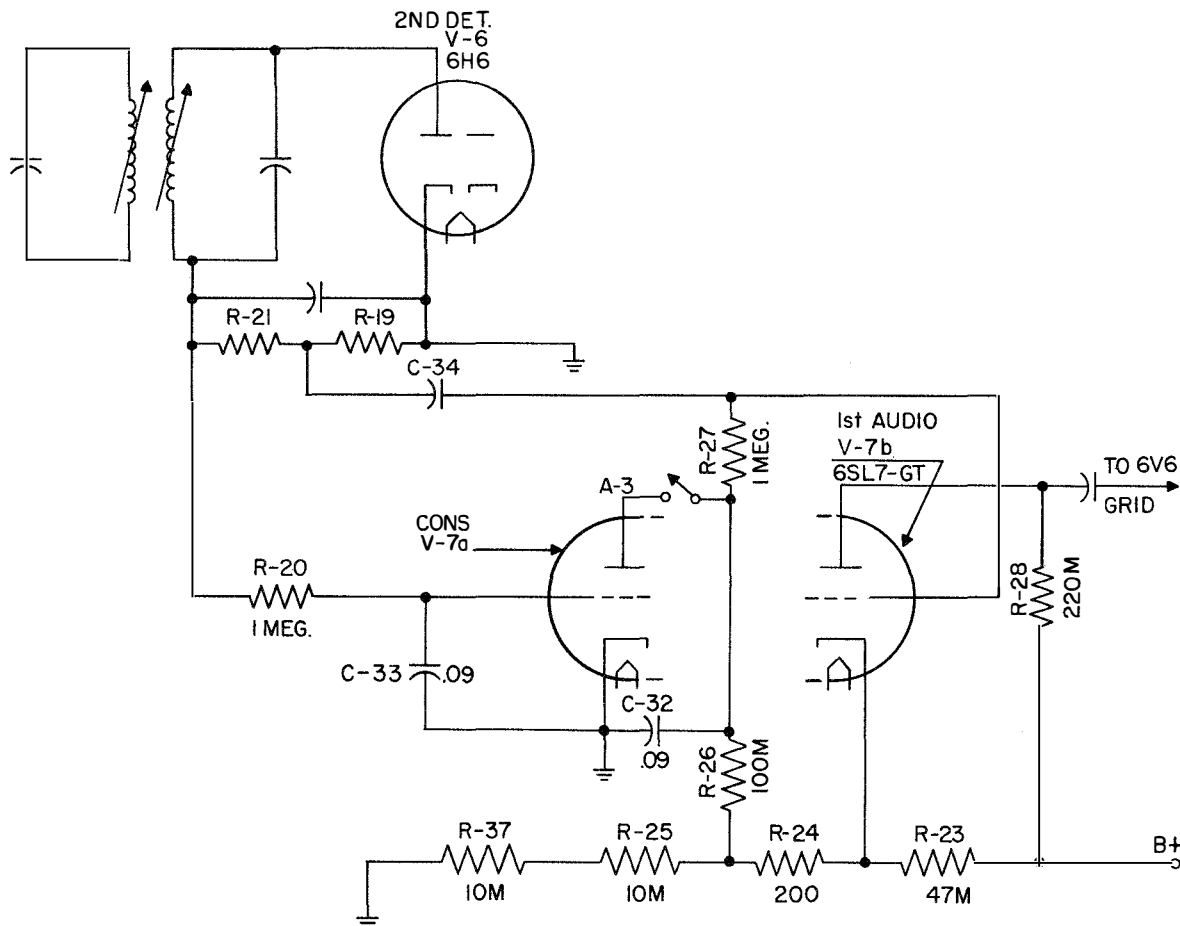


Figure 5-4. RHZ Receiver CONS Circuit

Since the actual silencing operation occurs in the first audio triode under the control of the CONS triode, the two triodes will be referred to as the "controlled tube," V-7B, and "controller tube," V-7A, respectively. The ability of the receiver to "squench" noise is dependent upon the biasing and plate potentials of both these triodes. V-7A receives its biasing potential from the voltage developed across the diode load, R-19 and R-21. Its plate voltage is obtained by the voltage drop across R-25 and R-37. Cathode and grid returns for V-7B are connected to the receiver voltage divider in such a manner that, although the grid is negative with respect to its cathode by the amount of voltage drop across R-24, it is positive with respect to ground by the potential developed across R-25 and R-37. It will be noted that the plate return of V-7A and grid return of V-7B are common. Resistor R-26 acts as a plate load resistor for V-7A. R-20 and C-33 are filters to provide an average dc biasing potential on V-7A, independent of varying percentage of modulation.

In normal operation with the controller tube V-7A disconnected by switch A-3, the controlled tube V-7B acts independently as a straight audio amplifier; however, with the CONS switch on and in the absence of a carrier, the tube passes current and offers a low resistance to ground for the grid return of V-7B. Since this grid return and V-7A plate are common, controlling bias will develop on V-7B grid depending upon the voltage drop across R-25 and R-37. In this manner V-7B is biased beyond plate current cut-off, "squenching" the receiver completely.

With the application of a carrier, grid bias will be developed across R-19 and R-21 for V-7A causing it to block and thereby release the controlling bias on the first audio tube, which will then function as an amplifier.

A variable threshold adjustment has not been provided. Resistors R-24, R-25 and R-37 have each been predetermined so an input signal within certain definite limits will automatically develop a blocking bias for V-7A across the detector load resistor. At lower values of receiver sensitivity, a higher signal input will be required to develop the blocking bias for V-7A.

E. Effect of AVC on Noise

Automatic volume control and noise are grouped together to show briefly how over-all receiver noise can vary when avc is applied to vacuum tubes at different stages in the receiver. Shot effect voltages, appearing at the grid of a vacuum tube, may be determined from the equation:

$$E = \frac{A\sqrt{I_b\Delta f}}{G_m} \quad (5-3)$$

Where; E is the equivalent shot effective voltage at the grid.

A is a variable dependent on the type of tube used. Triodes have the lowest value, multi-element tubes are higher.

I_b is direct plate current.

Δf is bandwidth.

G_m is mutual conductance or grid-plate transconductance.

From equation 5-3 it can be shown that for equivalent tube types, the lower the ratio of direct plate current I_b to grid-plate transconductance G_m , the lower will be the shot noise. Using a 6SK7 tube as an example, as bias for the remote cut-off type tube is increased beyond that used for maximum gain, shot effect E is increased. With five volts of bias applied, the ratio of direct plate current I_b to grid-plate transconductance G_m is 1.6. With ten volts bias the value rises to 3.5. Thus the shot effect is more than doubled. As the gain of the stage is reduced, the noise contribution from the following stage will be enhanced even in conjunction with a reduction in gain of the following stages where avc control is applied. From the viewpoint of a good signal to noise ratio, the last intermediate frequency amplifier is the preferred stage for avc application, however, this introduces the problem of distortion due to overloading of the stages not controlled by avc. A compromise has been achieved in some receivers wherein the use of delayed avc for the first stage has resulted in no avc bias application before a signal of a predetermined level is received. A smaller amount of delay or no delay in the application

of avc bias to the succeeding controlled stages results in a substantial increase in the signal to noise level compared to a receiver without delayed avc.

1. AUTOMATIC VOLUME CONTROL OF THE RUQ/RVA RECEIVER

The avc detector-amplifier circuit Figure 5-5, employs a type 6AQ6 dual-diode-triode in a circuit that produces a delayed and amplified avc control bias voltage, which maintains the output of the receiver very nearly constant over a wide range of input signal level. The circuit is designed to maintain a high signal to noise level or ratio.

The avc detector consists of one diode section, load resistors R-60 and R-61 in series, and coupling capacitor C-71. Part of the output of this detector is applied to the grid of the triode section, which functions as a cathode-loaded dc amplifier. Its load consists of R-62, R-63 and R-64 in series. The dc output of this amplifier is coupled through the other diode section to the avc time constant circuit, R-66 and C-74. The dc amplifier also provides a control voltage for the operation of the silencer circuit; this voltage is taken from the junction of R-63 and R-64.

Operation of this circuit is best understood by considering the action which takes place as a signal of increasing amplitude is applied to the receiver. With no signal applied, the triode section will be fully conducting and the cathode will assume a potential of about plus 40 volts with respect to ground. Because of the uni-directional characteristic of the output diode, no current will flow through it and the avc bias voltage will remain at the level determined by the setting of the radio frequency gain control, R-68. As signal is applied, the detector circuit develops a negative bias voltage with respect to the cathode; part of this voltage is applied to the grid of the triode, reducing the flow of current through the tube. When sufficient signal is applied, the current of the triode section will be cut off so that the cathode will become negative with respect to the plate of the output diode causing the generation of negative avc bias in the output circuit. This bias voltage is applied to the grid circuits of V-3, first intermediate frequency amplifier, and V-5 the first stage of the second intermediate frequency amplifier. A reduced or partial avc is applied to V-1, the RF amplifier stage, and a still further reduced avc bias voltage is applied to the grid of the first mixer, V-2, and the first audio amplifier, V-12B.

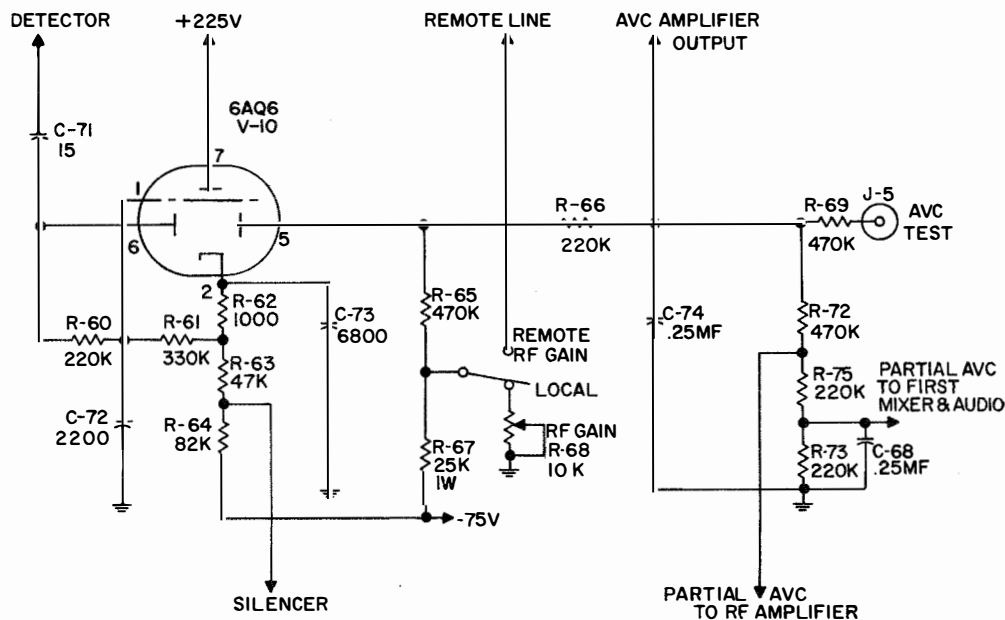
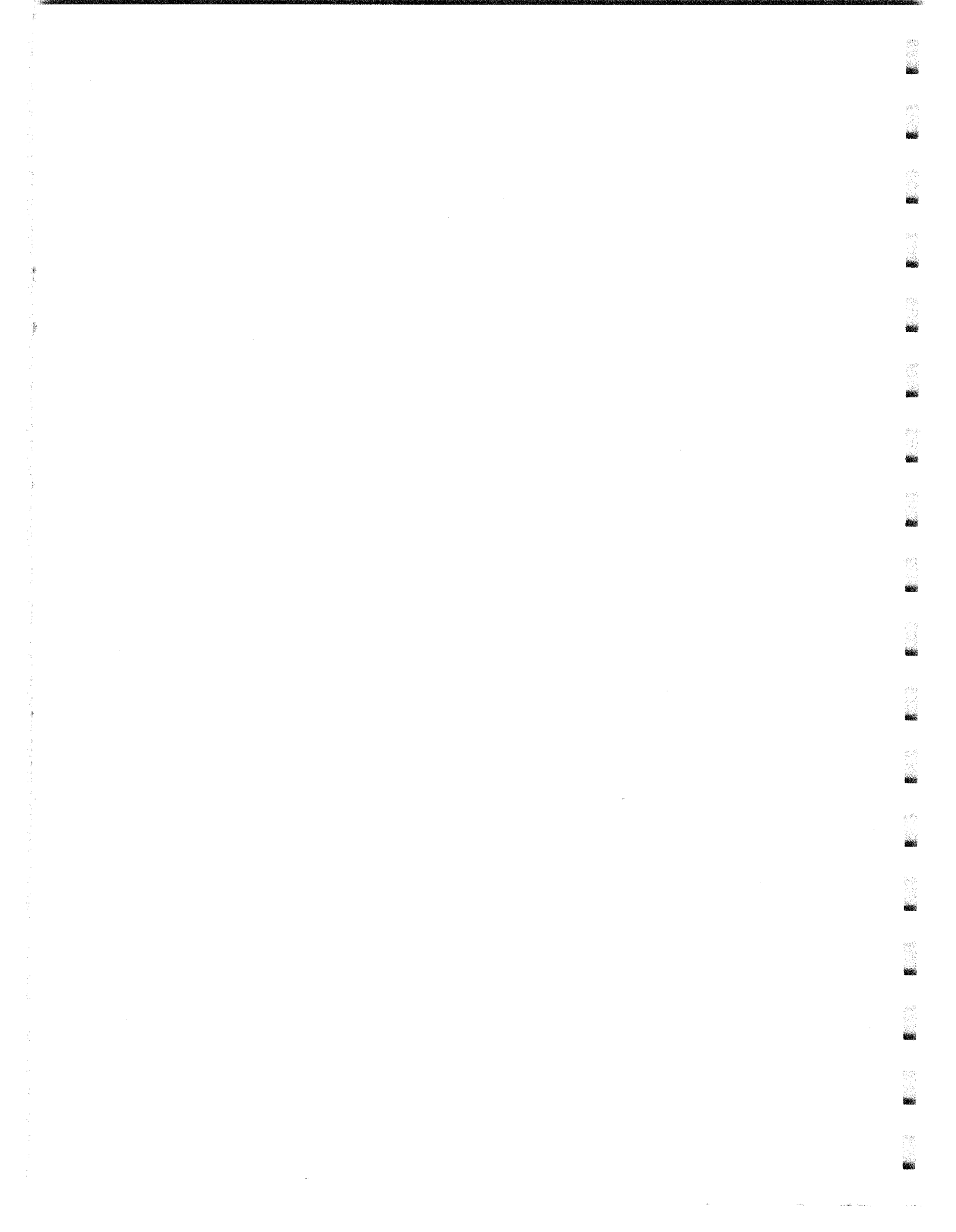


Figure 5-5. RUQ Receiver AVC Detector and Amplifier Circuits

5-5. SUMMARY

Noise exists throughout the frequency spectrum. It is generated both externally to and within the receiver and is the factor limiting receiver sensitivity. Many special cir-

cuits are designed to reduce noise. Research is continuing to develop new and better methods to combat it. Improved vacuum tubes, circuit components and circuit arrangements are needed to eliminate or reduce its ill effects.



CHAPTER 6

FREQUENCY MODULATION

6-1. INTRODUCTION

The FAA has installations to provide service over very large areas encompassing mountains, oceans and ice-capped regions. Landlines are not available to provide communications to some of these remote locations. FM link equipment is frequently used to overcome this deficiency. Frequency modulation is employed in this equipment.

Not all FAA stations require this equipment, but most field offices are responsible for a VHF Omni-directional Range which also employs frequency modulation to produce the 30 cps reference signal. Therefore, everyone should benefit from the general discussion of the characteristics of frequency modulation that is presented in this chapter.

6-2. DEFINITION

Modulation is the process of altering a radio wave in accordance with the intelligence to be transmitted. A radio wave is characterized by its amplitude, frequency and phase. This may be shown:

$$e = E_m \sin (\omega_c t + \phi) \quad (6-1)$$

Where: e is the instantaneous value of voltage.

E_m is the maximum carrier voltage.
 ω_c is 2π times the frequency of the carrier.

ϕ is the initial phase of the carrier voltage.

t is time in seconds.

Modulation may be achieved by altering any one of these characteristics. In amplitude modulation the amplitude of the resultant wave is varied in accordance with the intelligence while the frequency remains constant.

For frequency modulation the amplitude of the resultant wave remains constant while the instantaneous frequency is varied in accordance with the intelligence. In phase modulation the instantaneous phase angle of the carrier is varied in accordance with the modulating signal while the amplitude of the resultant wave remains constant.

Phase modulation and frequency modulation are essentially the same thing, the chief difference lies in the distribution of energy in the sidebands. Through the use of corrector networks and phase modulation, it is possible to dispense with self-controlled oscillators and to obtain directly crystal-controlled frequency modulation. Reception of FM, using limiters with discriminators for detection, is dependent upon the change in frequency of the radiated wave and not upon the distribution of sideband energy for a given frequency shift.

6-3. ADVANTAGES OF FM

Many of the problems associated with an amplitude modulated carrier can be overcome with frequency modulation. These problems include interchannel interference, fading and extraneous interference. Although the following discussion will be related to frequency modulation, the discussion would apply equally well to phase modulation.

A. Reactance Tube

The most common frequency modulator involves use of a reactance tube, wherein the frequency of an oscillator is made to vary in accordance with the amplitude of the modulating signal. A typical example is the CA-1390 Frequency Shift Exciter, which uses a 6SN7 as a balanced reactance tube modulator. The CA-1390 exciter provides a means for shifting a radio frequency carrier for the

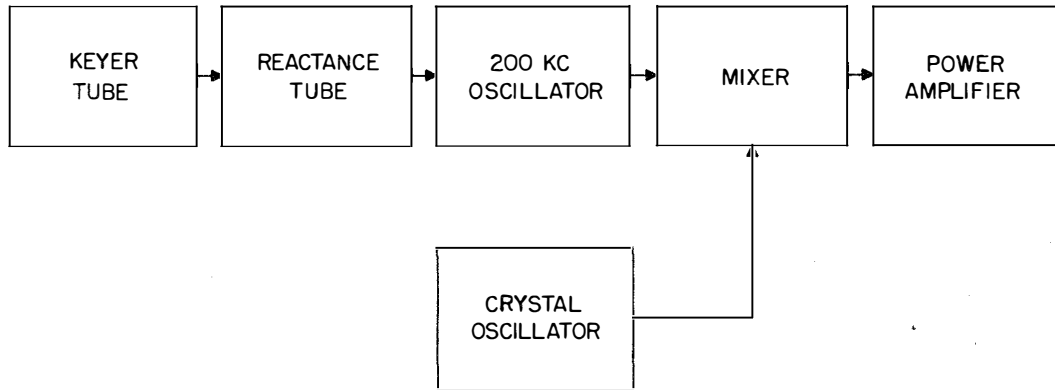


Figure 6-1. CA-1390 Frequency Shift Exciter

transmission of teletype or telegraph signals. A block diagram of this exciter is shown in Figure 6-1.

The reactance tube modulator, as shown in Figure 6-2, consists of a pentode tube connected across the resonant circuit of an oscillator. By means of an RC-phase-shift network the RF voltage between grid and cathode of the reactance tube is caused to be very nearly 90 degrees out of phase with the RF voltage across the oscillator resonant circuit. This quadrature voltage relationship as established by the reactance tube V-1 has the objective of injecting reactances into the

associated oscillator network. The associated oscillator V-2 is not stabilized at any one frequency as in phase modulation. The reflected reactance from V-1 changes the frequency of the generated oscillations, causing frequency deviations directly. The reactance can be made either capacitive or inductive by having the grid-cathode voltage lead or lag the oscillator voltage by 90 degrees. In Figure 6-2, grid-cathode voltage of V-1 lags the oscillator voltage by 90 degrees, producing an effective inductive reactance in parallel with the oscillator resonant circuit. If R and C are reversed the tube acts as a variable capacitance. The

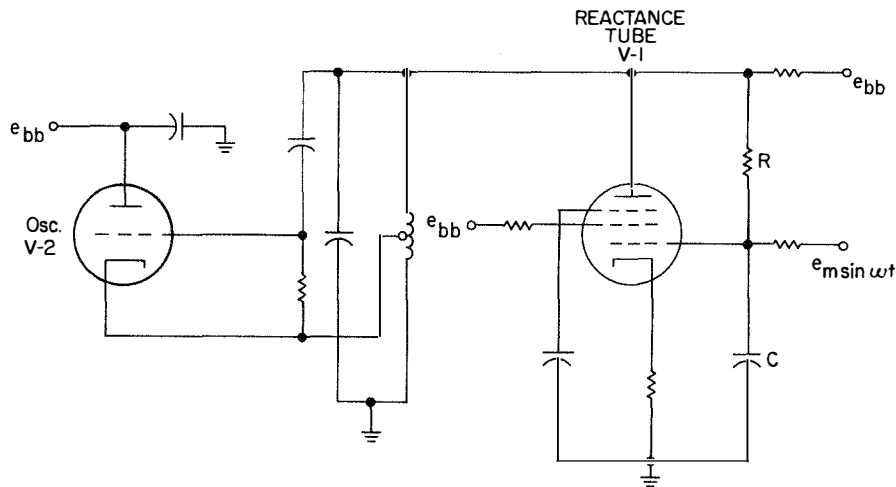


Figure 6-2. Reactance Tube Modulator

admittance presented to the oscillator by the RC phase shifting network and the reactance tube V-1 is very nearly

$$Y = \frac{1}{r_p} + \frac{j\omega C}{1 + j\omega RC} + \frac{g_m}{1 + j\omega RC} \quad (6-2)$$

Where: g_m is grid-plate transconductance of V-1.

r_p is the dynamic plate resistance.

In direct FM applications the modulating voltage is applied as an instantaneous bias change to the reactance tube, thereby changing the g_m of the tube. The resultant is a change in tube reactance, which in turn changes the oscillator resonant frequency. The approximate frequency of operation may be expressed as

$$f_r = \sqrt{CL \left(1 - \frac{g_m L}{R_{gp} C_{gk}} \right)} \quad (6-3)$$

Where: C_{gk} is grid-to-cathode interelectrode capacitance of the tube.

R_{gp} is dynamic grid-to-plate resistance.

It is to be noted g_m is the primary variable

and it is dependent upon the amplitude of the modulating signal and not its frequency.

In the absence of a modulating signal voltage, the oscillator generates a rest or center frequency. Assume the center frequency to be 50 megacycles and that it is desired to transmit a 1000 cycle note. When the oscillator frequency is shifted to 50 kilocycles below center frequency then back through center to 50 kilocycles above center frequency and back again to center frequency at a rate of 1000 of these operations per second, the output of the oscillator is then an FM wave modulated with an audio note of 1000 cycles per second. The number of these excursions from center frequency is dependent upon the frequency of the modulating signal employed.

If the oscillator swing is increased to wider limits of ± 75 kilocycles, a receiver tuned to this FM wave will receive the same 1000 cycle tone, see Figure 6-3; however, the amplitude of the 1000 cycle tone will be greater than the amplitude of the same tone recovered from an FM wave with a 50 kilocycle deviation. Change from a maximum deviation of ± 50 kilocycles to a maximum deviation of ± 75

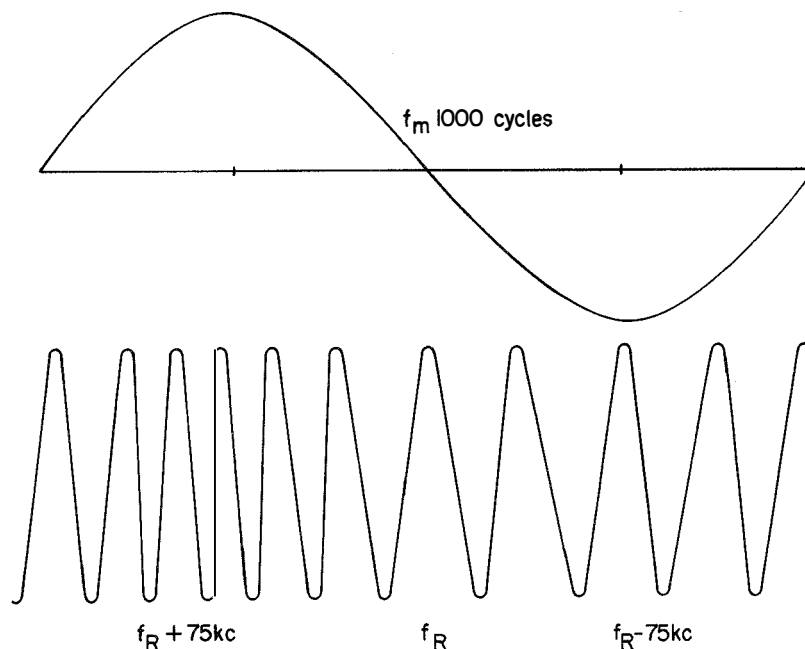


Figure 6-3. Frequency Modulated Wave

kilocycles is caused by a change in the amplitude of the modulating signal voltage. Wide-band FM is defined as one-hundred-per-cent modulated when the carrier is deviated from center frequency ± 75 kilocycles for a total swing of 150 kc.

B. Phase Modulation

Even though the main purpose of this chapter is to describe frequency modulation, it must be recalled that in indirect FM, phase modulation is instrumental in bringing about all of the features of direct FM. If the frequency of the oscillator associated with a reactance tube modulator is stabilized as in a crystal oscillator, the reflected reactance will cause a corresponding phase shift. The subsequent plate-current variations during a certain period of time are advanced in phase with respect to the plate-current variations affecting the grid potential of the oscillator by feedback action. The reason is that a stabilized frequency source will not allow a phase balance, and the resultant oscillator current will yield to periodic phase shifts. The oscillator current is advanced and retarded without changing the number of cycles within the current wave. The entire current wave shifts back and forth along the time axis, see Figure 6-4, and instantaneous current may be computed using

Eq. 6-4.

It is true, equivalent frequency variations exist. This is analogous to the relationship between velocity and distance traversed. Velocity is the rate of change of distance with respect to time. If phase variation is desired, a corresponding amount of frequency deviation must be produced.

$$i = I_m \sin(\omega_c t + \Delta\phi \sin \omega_m t) \quad (6-4)$$

Where: ω_m is 2π times the modulating frequency.

One way of causing phase modulation would be to apply a voltage of a stabilized frequency "f" to the grid of a pentode tube and to tune the plate-output tank circuit to the output carrier frequency "f". A reactance tube network is in parallel with the oscillator plate-tank circuit. Since the signal voltage " $e_m \cos \omega t$ " is also present, the plate tank circuit will be modulated in phase because the frequency "f" is stable. It should be noted " $e_m \cos \omega t$ " is used because of the 90 degree phase shift between grid and plate. A coupling circuit connected to a load experiences a phase modulated current. The phase-angle variation is proportional to the modulating signal amplitude. The frequency deviation will increase directly in proportion to the modulation frequency.

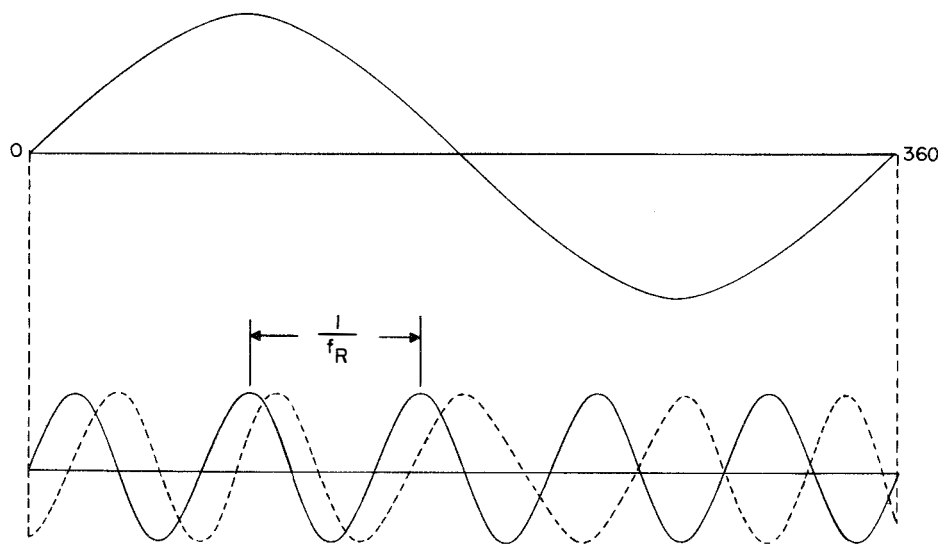


Figure 6-4. Phase Modulation

For example, a two thousand cycle note will produce two times the frequency deviation that a one thousand cycle note will produce. Recalling the principles of FM, where the frequency deviation must represent the magnitude of the modulating signal, and the higher the modulating frequency the smaller the phase-angle variation, a practical comparison between the two systems can be made. Briefly, if the frequency deviation is proportional to the modulation frequency, then the system is PM. If the frequency deviation is proportional to the modulation amplitude, then the system is FM.

The TFE very-high-frequency transmitter employs phase modulation to bring about a frequency modulated wave; see block diagram Figure 6-5.

This equipment produces frequency modulation through the use of pulse-shaping and pulse-positioning circuits which phase modulate the output of a crystal oscillator. Since the frequency deviation produced by a phase modulator is directly proportional to the modulating signal frequency, a capacitor equalizer is shunted across the audio input circuit to attenuate the higher frequencies.

Phase modulation is produced by shaping the output of the crystal oscillator to a saw-tooth. The saw-tooth is then passed through

a gating diode which retards the starting point of conduction in accordance with the instantaneous value of the audio signal. Thus the point of conduction is delayed in time and phase as the audio signal is changed in amplitude. The output of the gating diode is applied across an L-C ringing circuit which forms single pulses representing the point of saw-tooth conduction and of return. The pulse corresponding to the point of conduction is amplified and shaped in two successive stages and is then used to drive the grid of the first of the chain of frequency multipliers. This method of modulation is employed because it is inherently linear and permits the use of a crystal-controlled oscillator.

6.4. SIDEBANDS

Amplitude modulation develops two sidebands of a magnitude dependent upon the modulation factor "m." When modulation factor "m" becomes 1, for 100% modulation, the amplitude of each sideband becomes

$$\frac{mI}{2}$$

or 1/2 the magnitude of the carrier current. Each sideband contains 1/4 the carrier power. The resultant increase in power is 50%.

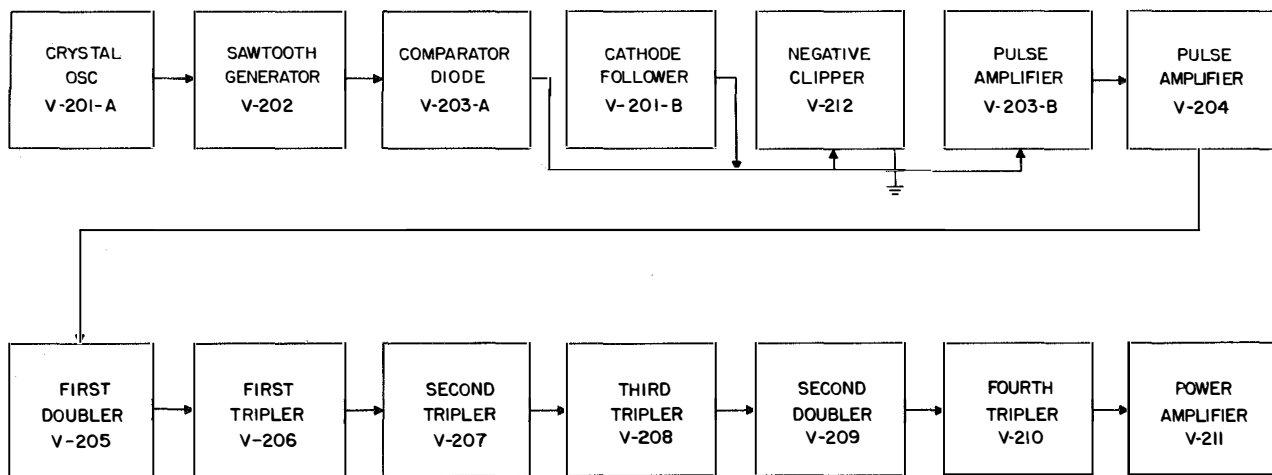


Figure 6-5. Block Diagram of TFE Transmitter

$$i = I \sin \omega t + \frac{mI}{2} \cos (\omega_c - \omega_m)t$$

$$- \frac{mI}{2} \cos(\omega_c + \omega_m)t \quad (6-5)$$

Where: i is instantaneous current.

I is average current amplitude.

ω_c is the carrier frequency.

ω_m is the modulating frequency.

In the Eq. 6-5, the first term of the equation refers to the carrier, the second term to the lower sideband and the third to the upper sideband.

Amplitude modulation can be clearly demonstrated by the use of vectors if the carrier vector can be assumed to be stationary. The upper sideband will then rotate counterclockwise with respect to the carrier at f_m revolutions per second. The lower sideband will rotate clockwise at the same rate. The upper sideband and the lower sideband generate equal angles with respect to the carrier.

The resultant R adds or subtracts in line with ω_c to form the overall resultant $\omega_c + R$. Figure 6-6 shows a vector presentation of amplitude modulation for various instants of time.

Frequency modulation, unlike amplitude modulation, contains many sidebands or side frequencies which are dependent upon the modulation index. Modulation index m_p is similar to the term defined as modulation factor in amplitude modulation. It is described by the following equation:

$$m_p = \frac{\Delta f}{f_m} \quad (6-6)$$

Where: m_p is the modulation index.

Δf is the instantaneous frequency deviation from center.

f_m is the instantaneous frequency of the modulating signal.

For better illustration, assume a maximum frequency deviation of 75 kilocycles and a modulating frequency of 15,000 cycles. Substitution in Eq. 6-8 shows the deviation ratio to be five. Although FM is composed of an infinite number of sidebands, the majority are of little significance. The sidebands that remain carry most of the intelligence. In the example just cited, sidebands above the fifth order are negligible, see Figure 6-7. The band width remains essentially constant over the audio range for a given frequency deviation regardless of the frequency of the modulating signal. This can be seen from Eq. 6-6.

If the modulating frequency is 50 cycles, for 100 per cent modulation the value of m_p is 1500, which indicates the upper and lower sidebands are significant at low signal frequencies and can not be neglected. These sidebands are spaced only 50 cycles apart and occupy 2 times 50 times 1500, or 150 kilocycles of bandwidth. With a modulating signal frequency of 15,000 cycles, approximately five upper and five lower sidebands are of significance. These sidebands are 15,000

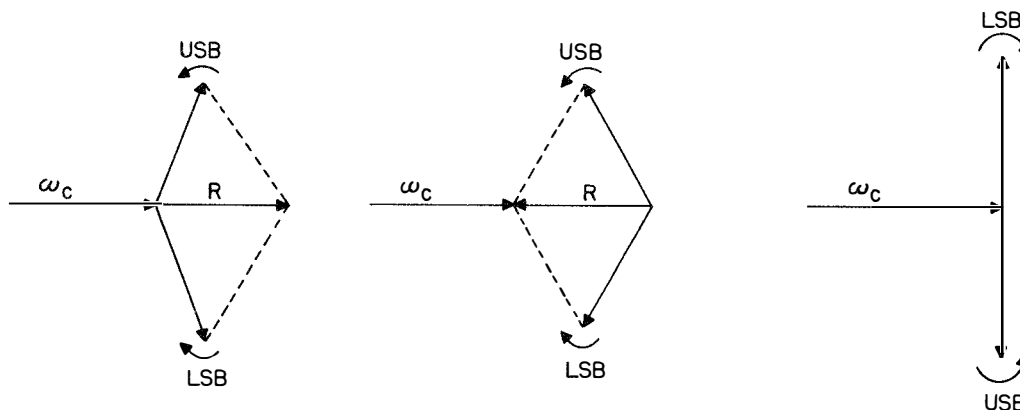


Figure 6-6. Carrier and Sideband Vectors in AM

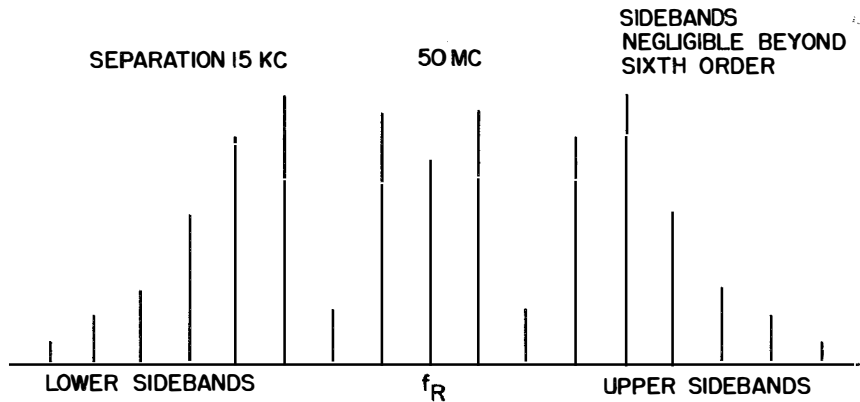


Figure 6-7. Spectrum of Frequency Modulated Wave

cycles apart and occupy 2 times 5 times 15,000 or 150 kilocycles, which is the same bandwidth as before when the modulating frequency was 50 cycles.

Consider next Eq. 6-7, for a frequency modulated wave.

$$i = I \sin(\omega_c t + m \sin \omega_m t) \quad (6-7)$$

This basic equation can be expanded into a series of terms as follows:

$$\begin{aligned}
 i &= I J_0 m \sin \omega_c t && \text{carrier frequency} \\
 + J_1 m \sin(\omega_c + \omega_m)t - \sin(\omega_c - \omega_m)t &&& \text{1st order sidebands} \\
 + J_2 m \sin(\omega_c + 2\omega_m)t - \sin(\omega_c - 2\omega_m)t &&& \text{2nd order sidebands} \\
 + \dots
 \end{aligned}$$

Where: J_0 with I represents the amplitude of the carrier.

J_1 represents the first order of sidebands.

J_2 represents the second order of sidebands.

$J_3, J_4,$ etc., represent higher order sidebands.

" J_0 " is a Bessel function of the zero order, " J_1 " is the first order and so on. As " m " varies, each " J " function will vary in its individual manner. These variations are shown in Figure 6-8.

Frequency modulation can be demonstrated using vectors as was done for amplitude modulation. However, in FM the first-order

sidebands involve $\pm \sin(\omega_c \pm \omega_m)$ instead of the cosine functions used in AM. The lower sideband is thus reversed in direction and the resultant appears at right angles to the carrier vector. As the sideband vectors rotate, they maintain equal angles with a line perpendicular to ω_c .

As the sideband vectors rotate they cause the overall resultant $\omega_c + R$ to swing to and fro with respect to ω_c , oscillating about the carrier vector at a rate determined by the modulating frequency. The extreme limits of its oscillation are $\pm \phi$. True frequency modulation does not permit any change in amplitude. Figure 6-9 clearly indicates the overall magnitude will vary as R swings between the limits of $\pm \phi$. However, addition of all the infinite order of sidebands produces an overall resultant of constant magnitude with varying phase angle. Figure 6-10 shows the addition of two pairs of sidebands with the carrier which indicates the overall resultant is approaching a constant magnitude represented by the dotted curve.

The total amount of energy in an FM wave is constant because of its constant amplitude; the relative energy in each of the sidebands and in the carrier varies, however, with the degree of modulation and also with the frequency of the modulating signal. The amplitude of the different frequency components depends upon m , the modulation index and can be calculated with the aid of a table of Bessel functions, see Figure 6-11. With zero

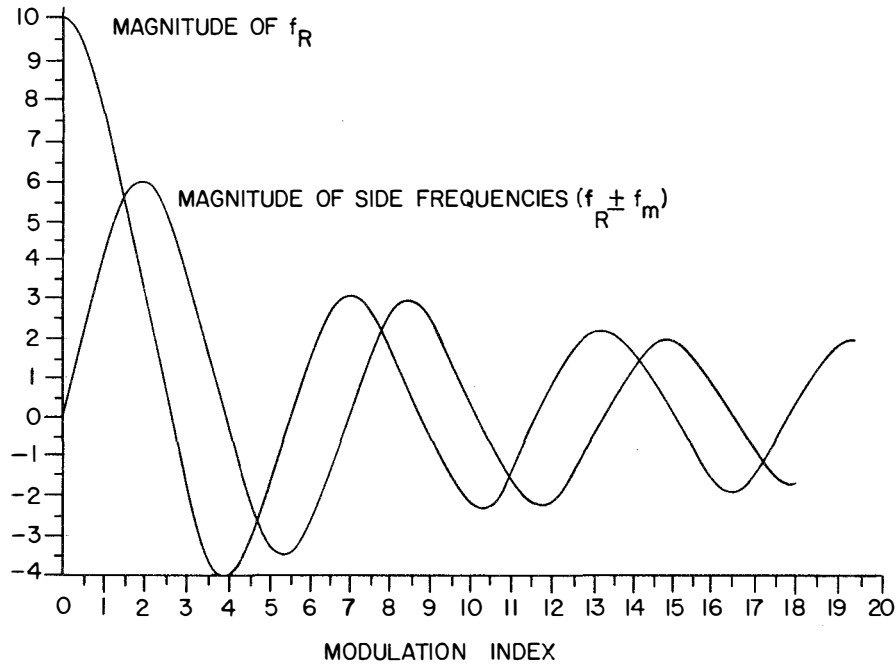


Figure 6-8. Plot of Bessel Functions

modulation, all energy is confined to the carrier. At other degrees of modulation, energy of the carrier may vary down to zero, and this former carrier energy is distributed among the sidebands. A spectrum analysis for a modulating signal frequency of 50 cycles has not been presented. The 1500 significant sidebands would crowd the presentation until it was meaningless.

Deviation ratio is another term often as-

sociated with frequency modulation. Deviation ratio is determined from the maximum frequency deviation under conditions of full modulation and the maximum modulating frequency to be used. Deviation ratio is described further by Eq. 6-8.

$$\text{Deviation Ratio} = \frac{\max \Delta f}{\max f_m} \quad (6-8)$$

Where: Δf and f_m are maximum values.

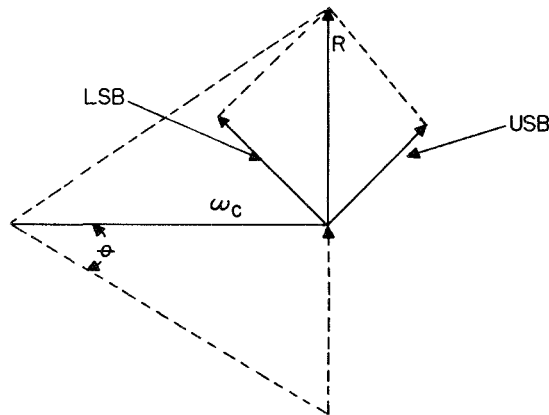


Figure 6-9. Carrier and First-Order Sideband Vectors

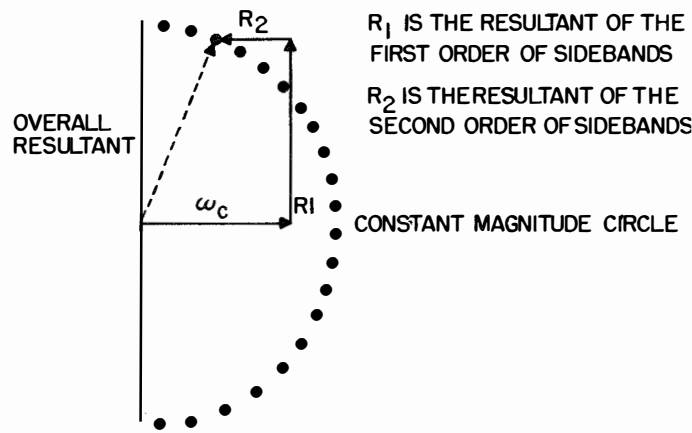


Figure 6-10. Carrier, First and Second-Order Sidebands Only

Order n of spectrum current	Modulation index $m_p = \Delta F/f_m = \Delta \theta$						
	$m_p = 1$	$m_p = 2$	$m_p = 3$	$m_p = 4$	$m_p = 5$	$m_p = 6$	$m_p = 7$
0 (center freq.)	0.7652	0.2239	-0.2601	-0.3971	-0.1776	0.1506	0.3001
1 (for $f_r \pm f_m$)	0.4401	0.5767	0.3391	-0.06604	-0.3276	-0.2767	-0.004683
2 (for $f_r \pm 2f_m$)	0.1149	0.3528	0.4861	0.3641	0.04657	-0.2429	-0.3014
3	0.01956	0.1289	0.3091	0.4302	0.3648	0.1148	-0.1676
4	0.00248	0.034	0.1320	0.2811	0.3912	0.3576	+0.1578
5		0.00704	0.04303	0.1321	0.2611	0.3621	0.3479
6		0.001202	0.01139	0.04909	0.131	0.2458	0.3392
7			0.00255	0.01518	0.05338	0.1296	0.2336
8				0.004029	0.01841	0.05653	0.1280
9					0.00552	0.02117	0.05892
10					0.001468	0.006964	0.02354
11						0.002048	0.008335
12							0.002656
13							0.0007702

Figure 6-11. Table of Bessel Functions

6.5. POLARIZATION

The horizontally polarized antenna is favored for FM because it has certain advantages over the vertically polarized antenna. The wave front of a radio wave at considerable distance from the antenna may be considered a plane at right angles to the direction of energy propagation. This wave front is composed of electric and magnetic lines of force. These lines of force are at right angles to each other and to the direction of energy propagation. The electric field has been chosen as a reference for defining vertical and horizontal polarization, primarily because the field intensity of a radiated wave is usually measured in terms of electric field intensity; millivolts or microvolts per meter. A vertically polarized wave is one in which the electric lines of force lie in a vertical direction, that is at right angles to the surface of the earth. The magnetic field is horizontal. A horizontally polarized wave is one in which the electric lines of force lie in a horizontal direction with respect to the earth's surface and the magnetic field is vertical as can be seen in Figure 6-12.

If a single wire antenna is used to extract energy from a radio wave, maximum pickup will result when the antenna lies in the same direction as the electric field. A vertical antenna should be used for efficient reception of

vertically polarized waves and a horizontal antenna for the reception of horizontally polarized waves. It is important to remember that whichever polarization is employed at the transmitter, this same polarization must be employed at the receiver for maximum signal pickup. Since ignition and most man-made noises are vertically polarized, see Chapter 5, the horizontal antenna receives a minimum of this type of noise.

Some fundamental differences between horizontal and vertically polarized dipole antennas are:

Horizontal Polarization

1. Directional.
2. Receives noise or interference from one line of direction only.

Vertical Polarization

1. Omni-directional. Receives signals from all directions.
2. Receives interference from all directions except up and down.

6.6. FM RECEIVERS

A frequency modulated wave should be constant in amplitude as it leaves the transmitting antenna. Any amplitude variation is produced by extraneous noise and unwanted signal. It is of paramount importance that some limiting device be incorporated to com-

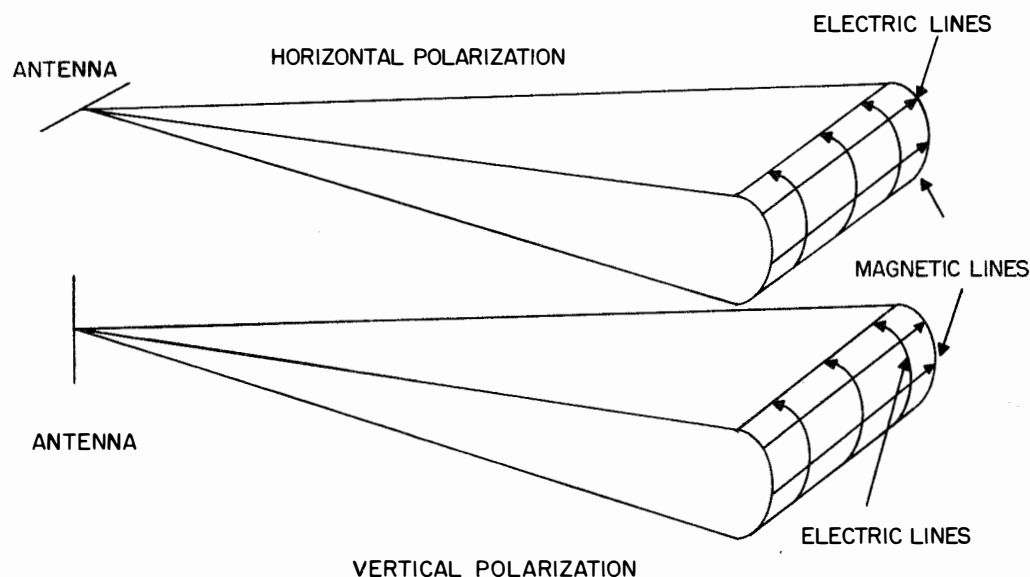


Figure 6-12. Wave Polarization

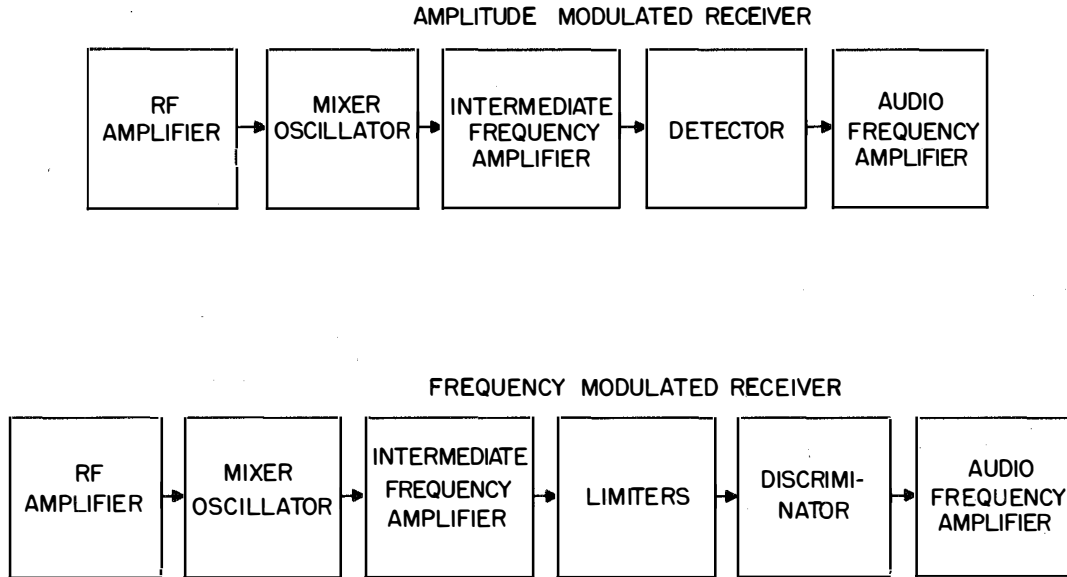


Figure 6-13. Comparison of AM and FM Receivers

but these unwanted amplitude changes and to make full use of the chief advantage FM has over amplitude modulation, greatly reduced interference. This limiting device, generally one or more limiting stages, is located ahead of the discriminator, shown in Figure 6-13.

The main substitution made in converting an AM receiver to an FM receiver is the method of detection. Some of the more common FM detectors are the slope detector, discriminator, ratio detector and the gated-beam detector. Figure 6-13 employs the discriminator in a typical FM receiver.

Broad-banded intermediate frequency circuits and a high quality audio system must be included if full advantage is to be taken of the high fidelity feature inherent in frequency modulation.

A. Limiters

The limiter is used to remove interfering amplitude variations and to insure a resultant carrier of constant amplitude. This is obtained by operating the limiter tubes at a low plate potential so that the input signal will drive the tube to saturation on the positive peaks and to cut-off on the negative peaks. In Figure 6-14, a low Q coil is employed in the plate circuit to broaden the resonance curve which enables it to accept the intermediate frequency at its widest deviation from center. The voltage developed across this tank is not constant over the bandwidth, but most of this voltage variation is balanced out in the discriminator. Signal voltage at the limiter grid should be large enough to cause plate saturation on positive peaks and plate current cut-

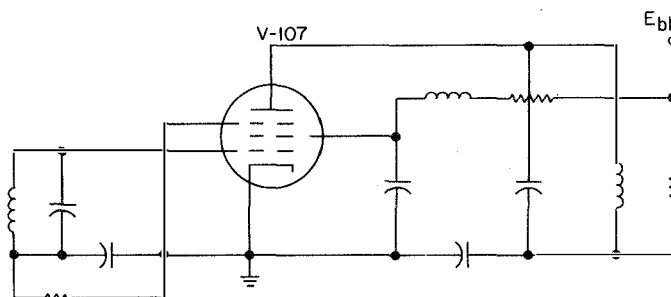


Figure 6-14. RFE Receiver, Third Limiter

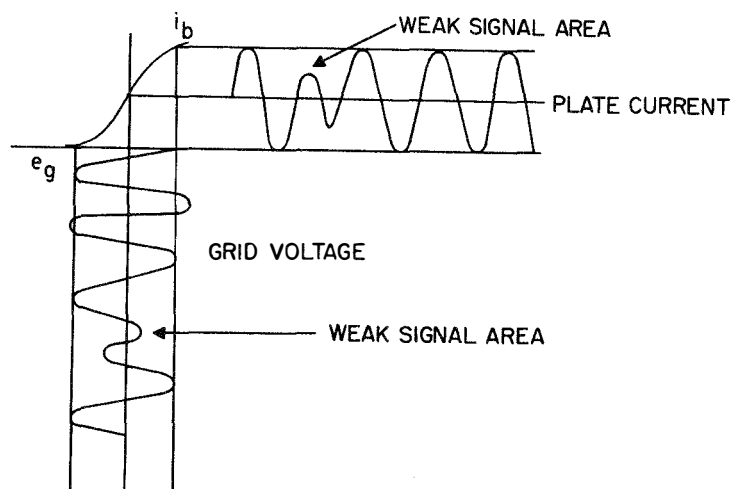


Figure 6-15. Typical Limiter Output

off on negative peaks. Thus the resultant plate voltage swings between definite limits regardless of the signal amplitude at the limiter grid. The frequency variations, representing the intelligence, remains (see Figure 6-15) since the frequency passed by the tube is independent of the tube characteristic curve.

When signal voltages fall below an acceptable minimum, the limiter does not eliminate amplitude variations and noise discrimination is lost. This suggests that multiple limiters are better than one, and three limiters are employed ahead of the Foster-Seeley discriminator in the RFE receiver. Sharp-cutoff vacuum tubes make ideal limiters and require only small grid voltages to effect limiting action.

B. Demodulators

The function of the demodulator is to detect an FM wave and recover the audio signal which was originally superimposed on the carrier. The demodulated signal must be in the form of an amplitude variation.

1. SLOPE DETECTOR

The simplest device for converting frequency variations to amplitude variations is an off-tune resonant circuit with sufficient "Q" that frequency variations will be detected on one side of resonance. With the carrier tuned in at A in Figure 6-16, an amount of RF voltage is developed across the resonant circuit.

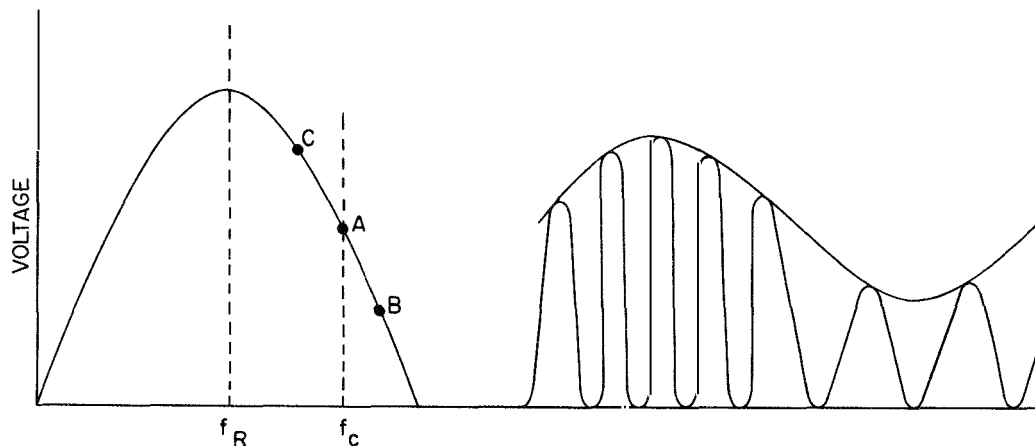


Figure 6-16. Slope Detector Output Voltage

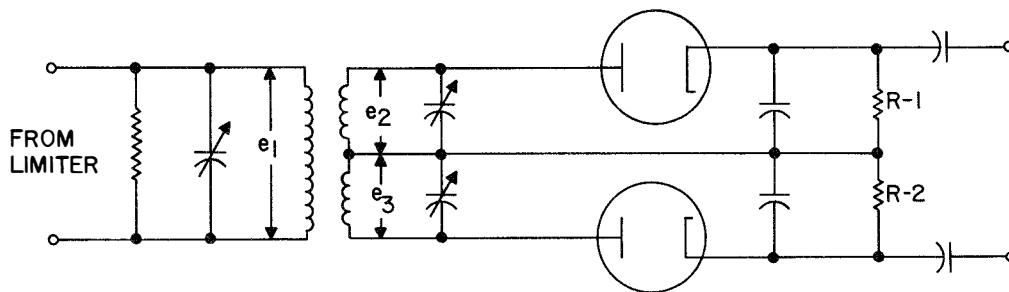


Figure 6-17. Travis Discriminator

The RF voltage amplitude increases or decreases as the frequency is varied either side of this carrier center frequency in accordance with the modulation. This resultant amplitude variation is then rectified by an AM detector. Only a small linear portion of the resonance curve is usable for conversion from FM to amplitude variations. Any frequency variation beyond this linear portion results in distortion. No noise limiting occurs and any amplitude increases at either the lower or higher portion of the resonance curve will be reflected in the output. Slope detection is employed in the CA-1277 Monitor, where f_c is 8.6 kilocycles and f_m is 9.96 kilocycles.

2. TRAVIS DISCRIMINATOR

Two tuned circuits are used in this discriminator secondary. One tank circuit is tuned above the intermediate frequency and the other tuned below. One resonant frequency should be slightly above the expected total frequency deviation and the other slightly below. Their individual outputs are combined

across series load resistors R-1 and R-2 in Figure 6-17. When a signal at rest frequency is received, the currents through R-1 and R-2 are equal in magnitude but opposite in phase; the sum of the voltages developed is zero. As the RF varies from center, these voltages become unequal. Above center the diode connected to the resonant circuit tuned to the upper frequency conducts more heavily, the other diode conducts less heavily, disturbing the voltage balance established across the load resistors at rest. As these individual voltages become unequal, a voltage having the polarity of the larger and equal to the difference between the two voltages is developed across the series resistors and applied through coupling capacitors to the audio amplifiers.

3. FOSTER-SEELEY DISCRIMINATOR

The Foster-Seeley discriminator requires only two tuned circuits instead of three as in the Travis discriminator, and it is much simpler to align. See Figure 6-18.

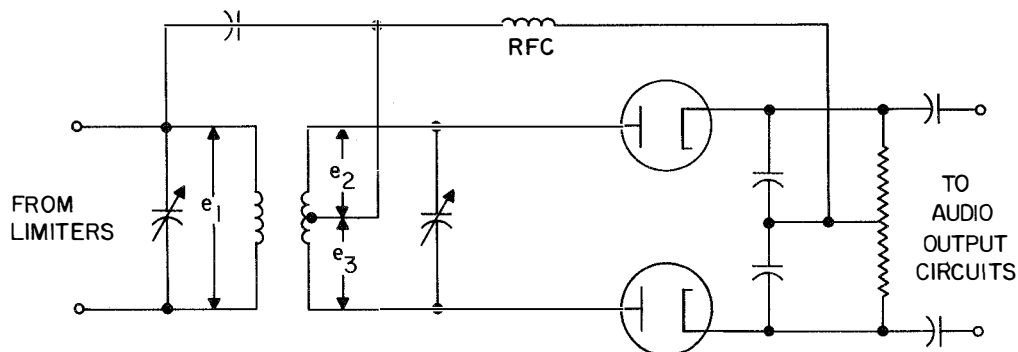


Figure 6-18. Foster-Seeley Discriminator

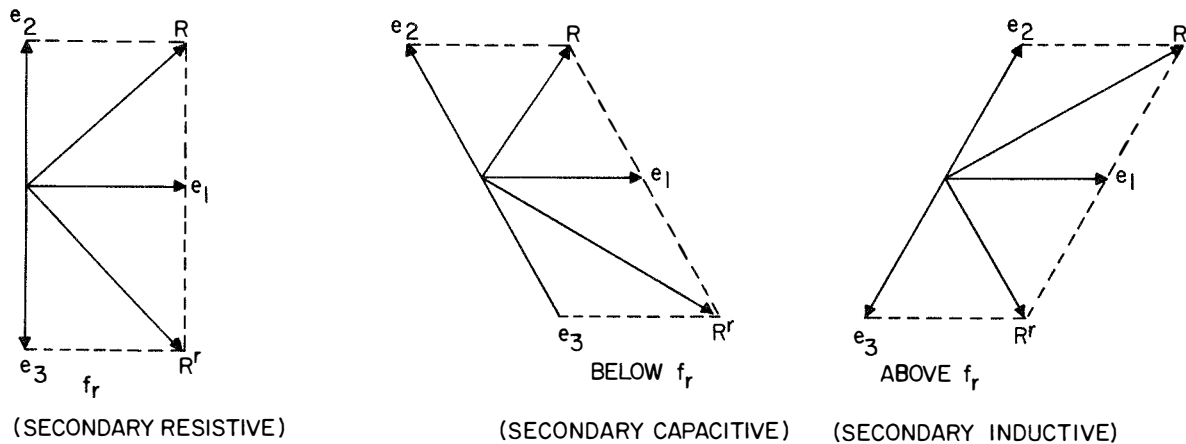


Figure 6-19. Discriminator Vectors

The tuned circuits are resonant to the center frequency and inductively coupled at less than critical value. The center of the secondary is connected to the primary by a coupling capacitor. At rest frequency the voltage taken from the primary and applied to the center tap of the secondary is 90 degrees out of phase with the secondary voltage. Phase relations existing in the discriminator are shown in Figure 6-19. The phase relations exist because of the impedance of the primary at resonance and the impedance of the secondary at resonance is purely resistive. At frequencies off resonance the secondary voltages are reactive and no longer in quadrature with the capacitor voltage taken from the primary. The magnitudes of the voltages applied to the diodes are equal at resonance, but at frequencies above resonance, one resultant voltage is increased in magnitude and the other is decreased; below resonance this condition is reversed. The diode outputs are combined in a differential rectifier, and the voltage developed across the load resistors is equal to the algebraic sum of the individual output voltages of each diode rectifier as developed across its individual load resistor. This difference voltage, caused by frequency excursions at an audio rate, is coupled through capacitors to the following audio stages. At frequencies encountered off resonance, phase shift in the primary is negligible. This may be attributed to low Q in the primary circuit

because of the low resistance of the limiter tube during conduction.

Figure 6-20 shows a frequency versus output voltage curve. For values of mutual coupling between primary and secondary less than critical, the frequency separation of the peaks of the discriminator characteristic is approximately

$$f = \frac{f_r}{Q} \quad (6-9)$$

Where: f is peak separation.

f_r is the center frequency.

Q is the Q of secondary circuit including diode loading.

At center frequency a zero output voltage is developed. On either side of this frequency a voltage is developed, the magnitude and polarity of which depends upon the direction and amount of frequency shift. If the frequency shift is greater than the limits shown in Figure 6-20, the output voltage will drop to zero. Discriminator adjustments should be made where the voltage is changing most rapidly. This zero voltage indication will be on the linear portion of the curve. One common alignment error is to adjust the discriminator for a false zero indication on either side of the linear area. A version of the Foster-Seeley discriminator is employed in the RFE receiver.

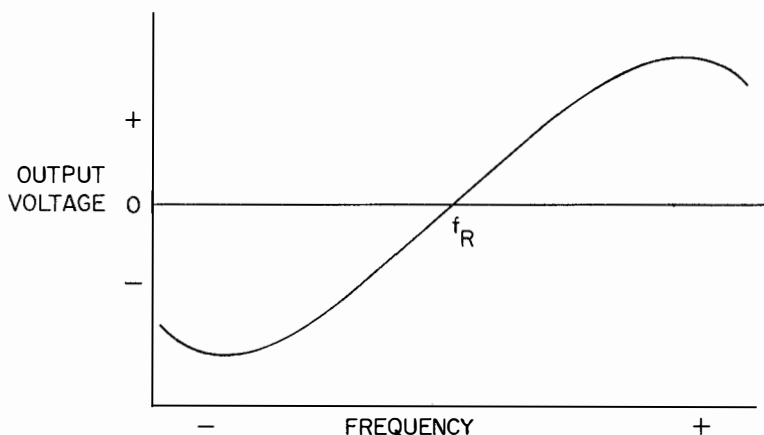


Figure 6-20. Discriminator Voltage vs. Frequency Curve

4. RATIO DETECTOR

The Ratio Detector circuit is in some respects similar to the Foster-Seeley discriminator. The vector presentation of Figure 6-19 applies equally well to the ratio detector. The two diode polarities are such that their output voltages add instead of subtracting as in the Foster-Seeley and Travis discriminators. At rest frequency the voltage at the avc terminal is twice that developed across the potentiometer R-1; the contribution of each diode is the same. As the input frequency deviates above or below the center frequency,

the relative contributions differ in accordance with the direction and amount of the deviation. Voltage across the potentiometer increases for frequency deviations in one direction and decreases in the other. The audio output voltage is equal to the ratio of the relative contributions of the two diodes. Ratio Detectors are essentially not responsive to amplitude variations. The total output voltage of the detector remains relatively constant, and is stabilized by a large capacitor, C₃ of Figure 6-21. Rapid variations in the amplitude such as those produced by noise are filtered out by this capacitor. The Ratio

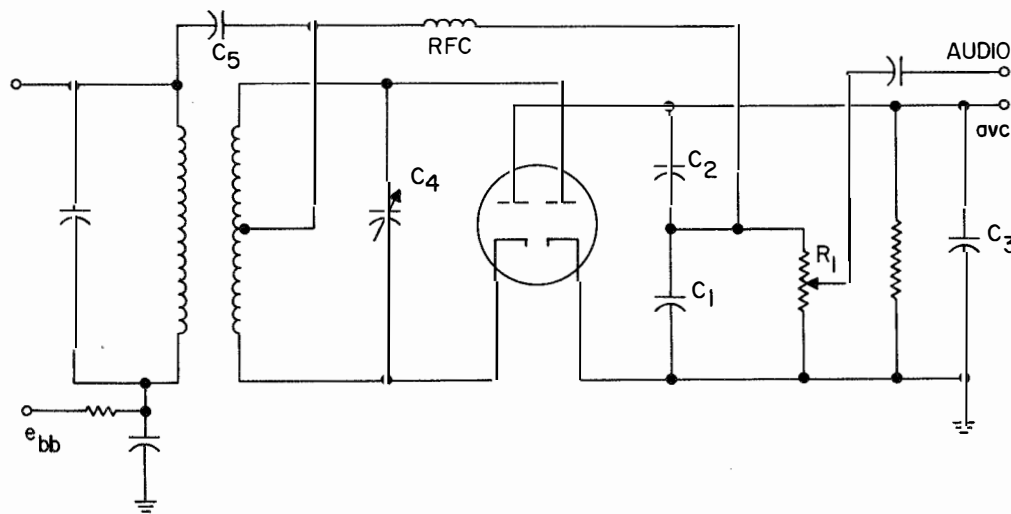


Figure 6-21. Ratio Detector

COMMUNICATIONS RECEIVERS

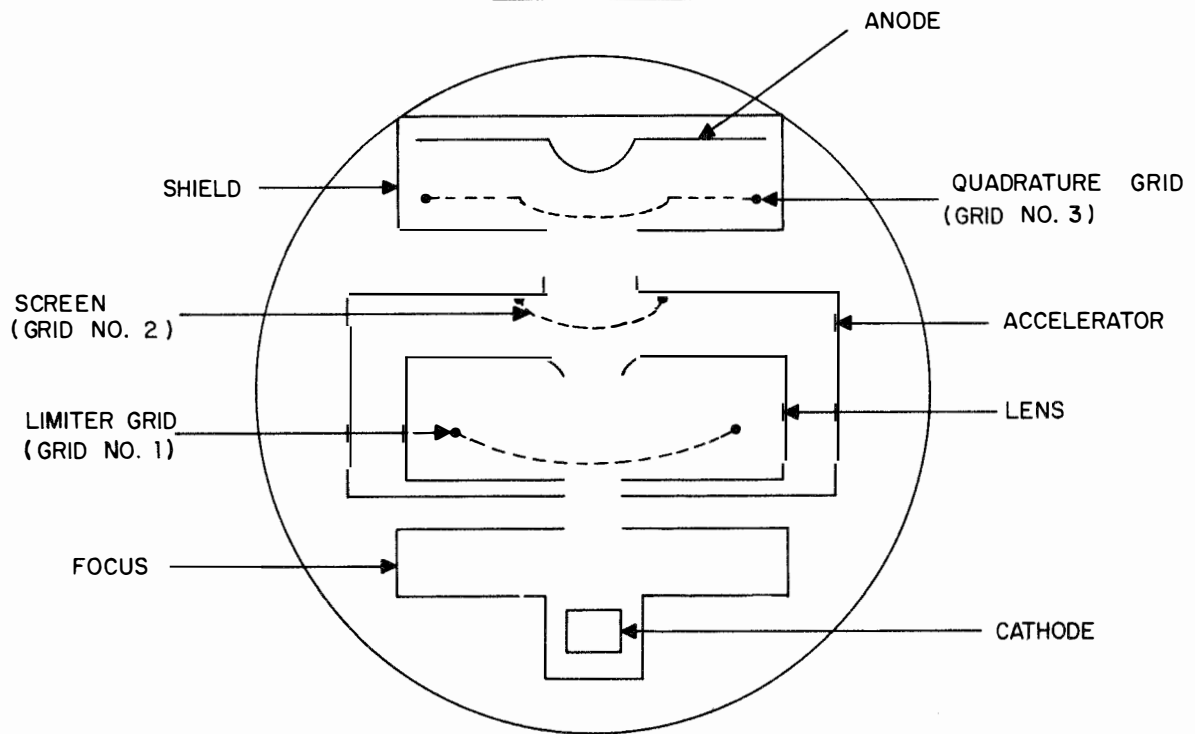
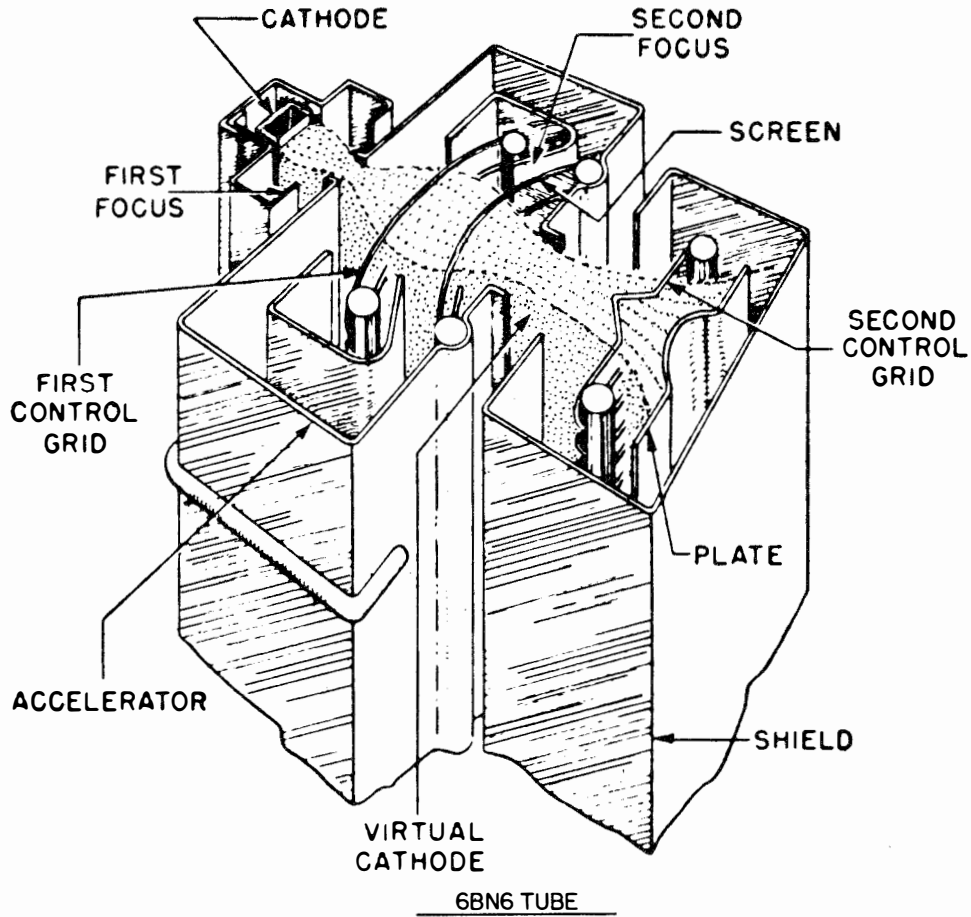


FIGURE 6-22. Gated-Beam Discriminator Tube and Schematic Diagram.

Detector usually does not employ a limiter preceding it. A Ratio Detector is employed in the CA-1621/4 Frequency Shift Receiver.

5. GATED-BEAM DISCRIMINATOR

Another approach to a combined limiter-discriminator combination, one that differs considerably from any of the previous circuits, is provided by the 6BN6 gated-beam tube. This tube, designed by Dr. Robert Adler of the Zenith Radio Corporation, possesses a characteristic such that, when the grid voltage changes from negative to positive values, the plate current rises rapidly from zero to a sharply defined maximum level. This same maximum value of plate current remains, no matter how positive the grid voltage is made. Current cutoff is achieved when the grid voltage goes about 2 volts negative.

The reason for this particular behavior of the tube stems from its construction. See Fig. 6-22. The focus-electrode, together with the first accelerator slot, forms an electron gun which projects a thin-sheet electron stream upon grid 1. The curved screen-grid, together with the grounded lens slot and aided by the slight curvature of grid 1, refocuses the beam and projects it through the second accelerator slot upon the second control grid. This grid and the anode which follows are enclosed in a shield box. Internally, the focus, lens, and shield electrodes are connected to the cathode. The accelerator and the screen grid receive the same positive voltage because both are connected internally.

The foregoing design is such that the electrons approaching the first grid do so head-on. Hence, when grid 1 is at zero potential or slightly positive, all approaching electrons pass through the grid. Making the grid more positive, therefore, cannot increase the plate current further. When, however, grid 1 is made negative, those electrons that are stopped and repelled back toward the cathode do so along the same path followed in their approach to the grid.

Because of the narrowness of the electron beam and its path of travel, electrons repelled by the grid form a sufficiently large space charge directly in the path of other approach-

ing electrons, thus causing an immediate cessation of current flow throughout the tube. In conventionally constructed tubes, the opposite is true.

These differences between tubes may be compared to the differences between the flow of traffic along narrow and along wide roads. Along the narrow road, failure of one car can slow down traffic considerably; along the wide road, more room is available and the breakdown of one car has less effect. The electron beam leaving the second slot of the accelerator approaches grid 3 also in the form of a thin sheet. Thus, this section of the tube may also serve as a gated-beam system. If this second grid is made strongly negative, the plate current of the tube is cutoff no matter how positive grid 1 may be. Over a narrow range of potentials in the vicinity of zero, the third grid can control the maximum amount of current flowing through the tube. However, if the third grid is made strongly positive, it also loses control over the plate current, which can never rise beyond a predetermined maximum level.

A typical gated-beam FM detector system is shown in Fig. 6-23, utilizing a 6BN6 gated-beam tube. The accelerator grid structures of the tube are in reality in the form of plates, which help shape the electrons into a narrow beam. The positive voltage of the accelerator grid structure increases the electron beam velocity and forces the beam through a narrow slot in the accelerator electrode. The electron beam then encounters the limiter grid, which, in conjunction with the quadrature grid, acts to control the electron flow. As with other tubes, the anode is made positive to attract the electrons emitted by the cathode.

The limiter grid has sufficient control over the electron beam to produce cutoff for any negative voltage. If the limiter grid has zero voltage or a positive voltage applied, however, it will permit current flow within the tube. The quadrature grid, if slightly negative, will also cause plate current cutoff. Thus, both grids are influential in preventing or permitting current flow within the tube.

With a small value of fixed bias, such as one volt, an incoming signal has sufficient amplitude to cause the tube to be operated at satura-

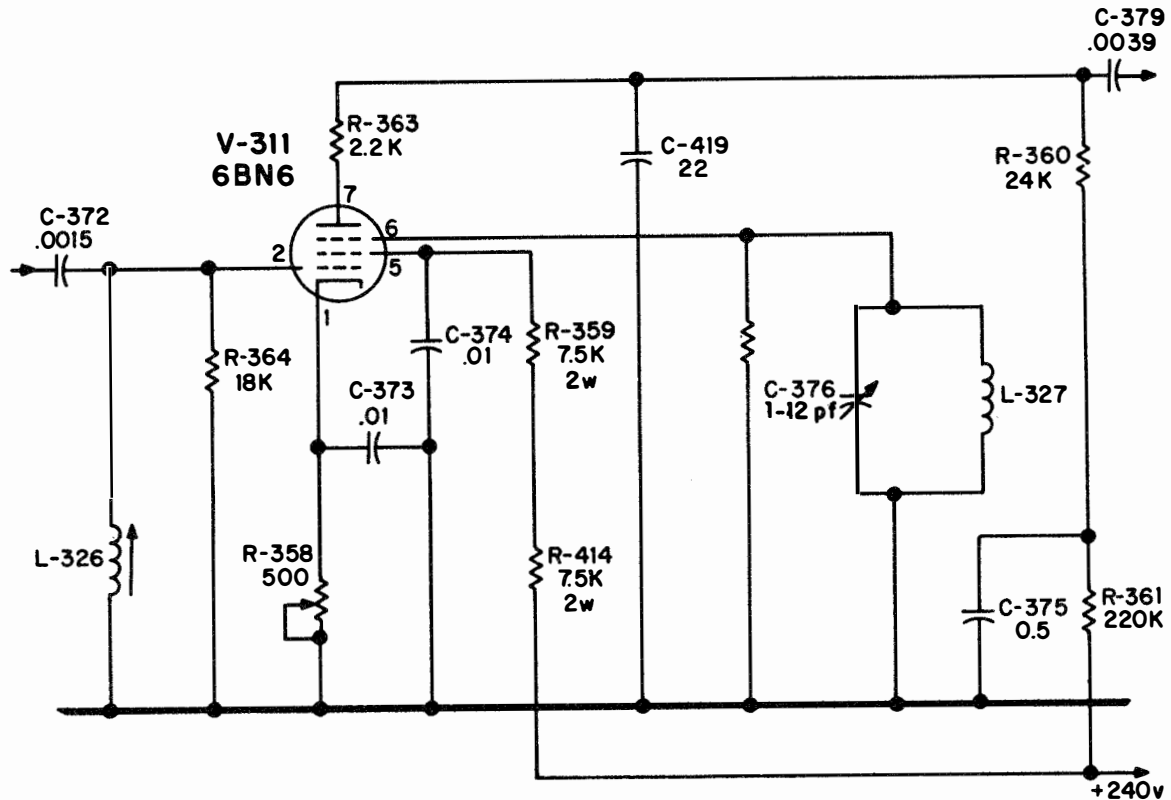


FIGURE 6-23. Typical Gated-Beam FM Detector System.

tion for the positive peaks of the grid signal, or at cutoff for the negative peaks of the incoming signal. Because the grid structure releases current flow rather suddenly, and also stops current flow quickly, a square wave of beam current occurs within the tube in the region beyond the input grid. Thus, the tube acts as a self limiting device, and will eliminate AM variations of incoming signal. A parallel resonant circuit is connected to the quadrature grid. The quadrature resonant circuit is tuned to the center carrier frequency of the incoming FM signal. During signal input, the cloud of electrons (space-charge) around the cathode varies and the quadrature grid is also affected by the electron beam because of space-charge coupling. Hence, the square wave type of signal generated within the tube is also present at the quadrature grid, and will pulse the quadrature circuit into a resonant flywheel condition.

The signal voltage which appears across the quadrature circuit, however, lags the input signal by approximately 90 degrees. The phase lag occurs because of the nature of the space-

charge coupling. With a 90 degree lag between the signal at the quadrature grid and that at the limiter grid, the plate current of the tube is cutoff for a greater period of time than would otherwise be the case.

This can be seen from an inspection of Fig. 6-24, which shows that the plate current can only flow when neither the limiter grid nor the quadrature grid is negative. Thus, only about one-half of each square-wave alternation reaches the anode during the time the carrier is at center frequency. When the incoming FM carrier shifts to a higher frequency, the quadrature circuit will be off resonance with respect to the shifted carrier frequency. The quadrature circuit becomes predominantly capacitive, because the higher frequency impressed on it increases the inductive reactance and decreases the capacitive reactance. Since the capacitive reactance is low, the current through the capacitive reactance is higher than that in the inductive reactance. Because a parallel resonant circuit with the resonant frequency impressed on it, exhibits a high imped-

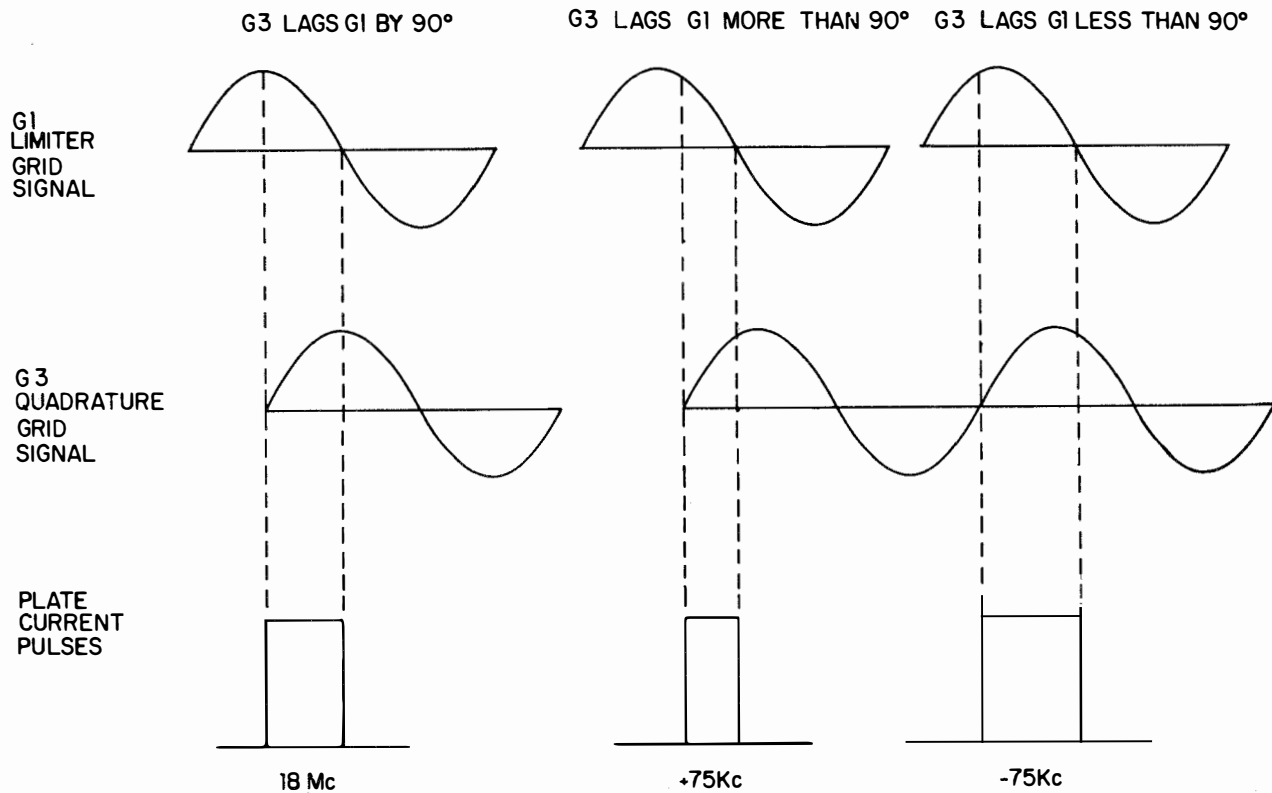


FIGURE 6-24. Gated-Beam Discriminator Waveforms.

ance, the reduction of such impedance through decreased capacitive reactance causes the circuit to be predominantly capacitive. The capacitive characteristics of the quadrature circuit will now cause the signal voltage at the quadrature grid to lag the signal at the limiter grid by more than 90 degrees, which is the lag at center carrier-frequency. Because of the increased phase difference between the two current-controlling voltages, less than one-half of each square-wave alternation arrives at the anode of the tube. Hence, the average value of plate current decreases. When the carrier signal at the limiter grid shifts lower in frequency, the quadrature circuit becomes predominantly inductive and the voltage tends to lead. As shown in Fig. 6-24, more than one-half of each square-wave alternation reaches the anode and thus the average value of plate current increases.

In the circuit of Fig. 6-23 a 2.2 K ohm resistor, R363 is inserted between the load R360 and the plate of the tube. By-passing of the IF voltage is accomplished by C419 but since this capacitor is placed beyond the 2.2 K ohm

resistor, a small IF voltage appears at the anode of the tube. Through the interelectrode capacitance that exists between the anode and grid 3, the IF voltage developed across R363 is coupled into L327-C376. The phase relationship existing in this circuit is such that this feedback voltage aids in driving the tuned circuit. Capacitor C419 and resistor R363 in the anode circuit also form an integration circuit. An integration circuit has the ability to produce an average value from a series of pulses having various widths. Resistor R360 is the conventional load resistor across which the audio signal voltages develop. Capacitor 419 has a low shunt reactance for the high carrier frequency, and thus eliminates the latter from the output circuit. Bias for grids 1 and 3 is obtained by placing a resistor R358 in the cathode leg of the tube. Since AM rejection, especially at low input signals near the limiting level, is a function of the correct bias, the cathode resistor R358 is made variable. This permits adjustments to be made in the field in order to compensate for tube or component changes.



FIGURE 6-25. Vector Resultants of Two Angular Modulated Waves.

C. PM and FM in a Receiver

It is interesting to note that frequency and phase modulated waves are detected in the same manner since there is no difference between them for a given modulation index. However, modulation index does vary differently with modulating frequency and should be taken care of by a transmitter compensating network. If an actual phase modulated wave is received on an FM receiver, an *RC* network should be placed in the receiver output to attenuate the developed audio signal in inverse proportion to modulation frequency if normally acceptable fidelity is to be maintained.

D. Acceptable Fidelity

Normal conversational speech frequencies contained in spoken language may vary from about 80 to 10,000 cycles. More than half of the speech energy is contained in frequency range below 500 cycles, but these frequencies contribute only five per cent to the interpretation of speech sounds. At the other extreme, eliminating frequencies above 1500 cycles reduces the speech energy by only about ten per cent, but the articulation is reduced more than 30 per cent. In practical applications a frequency bandwidth between 250 and 2500 cycles will usually provide a good voice circuit.

Carrier equipment used with FM links is often designed to provide an over-all channel bandwidth falling within these limits. Several audio channels can be utilized simultaneously to frequency modulate one transmitter, thus multiple sources and types of information may be carried on one FM link.

6-7. INTERCHANNEL INTERFERENCE

Frequency modulation results in greatly reduced interchannel interference since an FM wave has the desirable characteristic of excluding all except the strongest signal. When two FM waves of the same carrier frequency are added, each carrier must be computed individually and the results combined to determine the magnitude and phase of the resulting undesirable signal. This can be accomplished graphically as shown in Fig. 6-25, or by using Eq. 6-1 to determine magnitude and phase of the individual waves.

Assume two carriers A and B, with A twice the magnitude of B. With A unmodulated, line R is the vector sum of the two carriers. Whatever the variation of B, total angular change between A and resultant R can not exceed an angle only slightly greater than the angle whose tangent is B/A. The maximum undesired angular variation of R from A is no more than 29 degrees, whereas the desirable

modulation of A may be many hundreds of degrees; see Eq. 6-10. Thus the possible variation caused by the undesirable carrier B is insignificant, although B's amplitude is as great as 50% of the amplitude of A. In FM an undesirable signal can be as great as fifty per cent of the magnitude of the desired signal without causing objectionable interference. In amplitude modulation by comparison, an undesirable signal as little as one per cent of the desired signal can cause objectionable interference.

Captures sometimes occur in threshold areas when an undesired FM carrier stronger than the desired carrier takes over the FM link. This problem may originate from a transmitter hundreds or even thousands of miles distant if prevailing freak ionospheric conditions are favorable for directing the energy to the receiving station.

6-8. FM MEASUREMENTS

In a frequency modulated wave, amplitude of the desired intelligence is developed according to the instantaneous angular velocity or deviation and is computed by the relationship shown in Eq. 6-10.

$$\omega_i = \omega_c + 2\pi \Delta f \cos \omega_m t \quad (6-10)$$

Where: ω_i is instantaneous angular velocity.

ω_c is average angular velocity or 2π times center frequency.

ω_m is 2π times modulating signal frequency.

Δf is maximum deviation from center frequency.

When single frequency modulating signal voltages are used with an FM transmitter, relative magnitudes of sidebands and carrier vary widely with deviation. Changing the amount of modulation by increasing or decreasing the magnitude of modulating signal voltages has a definite relationship to magnitude of sidebands and carrier. This relationship results in a simple method for measuring deviation of a frequency modulated wave based upon Fourier's tables for repeating waveforms.

A. Measuring Frequency Deviation

1. NULL METHOD

An accurate method for measuring frequency deviation of an FM wave utilizes an AM receiver to detect carrier nulls at a point where all of the transmitted energy is contained in the sidebands. The frequency deviation may be readily computed using Eq. 6-2 and a carrier extinction chart, Fig. 6-26.

MODULATION INDEX	NULL POINT No.
2.405	1
5.520	2
8.654	3
11.792	4
14.931	5
18.071	6
21.212	7
24.353	8
27.494	9
30.635	10

FIGURE 6-26. Carrier Extinction Chart.

A known modulating signal frequency is used at the transmitter. Modulating signal voltage is increased from zero at the transmitter until the carrier component detected at the receiver is zero and all of the energy is contained in the sidebands. The first null or zero carrier is obtained when the modulation index is 2.405. From Eq. 6-6, this means the deviation is equal to 2.405 times the modulating signal frequency. For example, if a modulating signal frequency of 2,000 cycles is used, the first carrier null is obtained when the frequency deviation is 2.405 times 2,000 cps or 4.81 Kc. The second null occurs where the modulation index is 5.520 and the frequency deviation is 5.520 times 2,000 cps or 11.04 Kc. Null separation at lower modulating frequencies is sometimes difficult to detect and an accurate determination of deviation is simpler using higher modulating frequencies.

2. SUBSTITUTION METHOD

An alternate method for determining frequency deviation utilizes substitution, wherein

an FM signal generator resonated to the center or carrier frequency is applied to an FM receiver ahead of the limiter. A signal generator modulating frequency which is the same frequency as the transmitter modulating signal is selected, and the signal generator deviation is increased until the value of audio output developed equals the audio measured when the transmitter is used as the originating signal source. The deviation is read directly from the signal generator deviation meter, normally in kilocycles.

B. Sensitivity Measurements

1. AUDIO OUTPUT METHOD

Over-all sensitivity measurements of an FM type receiver differ somewhat from those of a standard AM type of communications receiver. The same test equipment, a signal generator and a meter to measure the audio output, is needed. However, unlike AM, FM sensitivity is determined by measuring the decrease in audio output level. With the signal generator connected to the antenna input terminals and the generator carrier off, the audio output noise level is noted. An unmodulated carrier from the generator is applied at the antenna terminals and the carrier magnitude increased until 20 db of quieting has been achieved as measured across the audio output. For example, if 50 milliwatts of audio output power is delivered to a power meter with no signal from the signal generator, then the generator carrier at center frequency must be increased in magnitude until the audio output power drops to 1/100 of its original value or 0.5 milliwatts. The microvolt reading indicated by the signal generator is then a measure of receiver sensitivity. FAA performance standards specify the sensitivity shall be 5 microvolts or less.

2. LIMITER SATURATION METHOD

An alternate method used for measuring FM receiver sensitivity requires only the use of a signal generator and the receiver microammeter. With the generator connected to the antenna input, the generator is adjusted to center frequency as determined by the micro-

ammeter in the discriminator circuit. The receiver meter is then switched into the last limiter stage and observed for grid current saturation as the signal generator output level is increased from zero. The point where a linear increase in input no longer produces a linear increase in limiter current is considered to be saturation. The microvolt reading from the signal generator is then a measure of the receiver sensitivity. This alternate method normally produces sensitivity measurements that vary not more than 1 microvolt from those measured by the method outlined in the previous paragraph.

6-9. SUMMARY

For years the major problem of radio communication has been man-made and natural interference. This noise is difficult to discriminate against because it contains nearly all frequencies in the usable spectrum. Early FM systems using very small frequency deviations to maintain narrow band width showed little improvements. Wide frequency deviation, however, provides for great discrimination against noise. Frequency modulation and the allied system of phase modulation is becoming of increasing importance in communications. Frequency modulation offers an improved signal-to-noise ratio for high field intensities available in the coverage area since noise is chiefly an amplitude variation.

6-10. REVIEW QUESTIONS

1. (a) What is the frequency deviation of a wideband FM wave?
 $f_r = 175 \text{ Mc}$ $f_m = 25 \text{ Kc}$ $m_p = 3$
 - (b) What is the deviation ratio?
 - (c) What is the effective bandwidth?
 - (d) Are deviation ratio and modulation index m_p always identical?
2. (a) What is the modulation percentage for wideband FM? When:
 $m_p = 4$ $f_m = 15 \text{ Kc}$ $f_r = 175 \text{ Mc}$
 - (b) What is the modulation percentage When Δf is $\frac{1}{2}$ that found for (a)?

3. An unmodulated FM transmitter produces 10 watts of RF output. f_m voltage is increased until $m_p = 2.405$.
- (a) An oscilloscope shows what percent of the total power is in the carrier?
 - (b) What percentage of the power is in the sidebands?
 - (c) What is the ratio of carrier power to sideband power?
 - (d) What is the ratio of modulated to unmodulated total power output?
 - (e) How much power is radiated during 100% modulation?
4. An FM generator is modulated by a 1,000 cycle tone at 30%. The modulation percentage is tripled:
- (a) What is the over-all frequency swing $2\Delta f$?
 - (b) What is the modulation frequency f_m ?
 - (c) What is the modulation index m_p ?

6-11. ANSWERS TO FM REVIEW QUESTIONS

- 1. (a) 75 Kc
(b) 3
(c) 150 Kc
(b) no
- 2. (a) 80%
(b) 40%
- 3. (a) Zero
(b) 100%
(c) 0:1
(d) 1:1
(e) 10 watts
- 4. (a) 135 Kc
(b) 1000 cycles
(c) 67.5



CHAPTER 7

DIVERSITY RECEPTION

7-1. GENERAL

The reliability in the reception of signals for point-to-point communications is dependent upon the factor called FADING. According to the standards of the I. R. E., fading "is the variation in intensity of radio signals resulting from changes in the transmission medium." Of the two major types of fading, which can be classified as either single-path or multipath transmission, the minimizing of the latter is of most importance to us. This is due to the fact that in multipath transmission the fading that occurs is much more rapid, and the extent of such fading is much greater than that which occurs in single-path transmission. In the majority of cases, fading that occurs in single-path transmission is of small enough magnitude that a good AVC circuit within a receiver can compensate for it. The rapidity of fading in multipath transmission, however, depends upon the rate at which the different transmission path lengths change, and a change in the transmission of as little as one-half wave length may produce a change in the received signal that will go from a maximum to a minimum which cannot be counteracted by AVC circuits in the receiver.

By the use of two or more receivers on the same frequency, and by feeding these receivers from spaced directional receiving antennas, the fading effects of multipath transmission may be reduced to a negligible factor. When this is accomplished, and the outputs of the separate receivers are combined in such a manner that a fairly constant output is achieved, we have what is termed "DIVERSITY RECEPTION."

It is to be noted that one of the requirements listed for DIVERSITY RECEPTION was that

of SPACED DIRECTIONAL RECEIVING ANTENNAS. It is these aforementioned antennas that makes reception possible in spite of fading present in multipath transmission. It is a well established fact that if two or more antennas are spaced a number of wave lengths apart, that due to the different path lengths of a given signal arriving at these separate antennas and the time factor involved, that a fade will not occur simultaneously on these antennas. Experimentally, it has been observed that optimum reception is possible when the antennas are spaced from five or six to ten wave lengths apart. However, it is known that there is a limit to the spacing, for after reaching a distance of approximately ten wave-lengths, or one thousand feet which ever is attained first, any gain received from further spacing is lost in the form of losses in the added length of transmission line necessary to carry the signal to the receivers. So a compromise normally has to be made where the spacing between the antennas, and the length and type of the transmission lines are both carefully considered.

It can undoubtedly be said from the foregoing that DIVERSITY RECEPTION is due primarily to the spaced antennas, but there is something else necessary, and that is properly designed receivers and accessory equipment for the combining of the signals after they are picked up by the antennas, in order that the desired signal may be heard.

7-2. ANTENNAS

It is true that the individual antennas that go to make up a diversity antenna system may be of any type, however, there are certain antennas that lend themselves especially well to this type of work. It has been found from

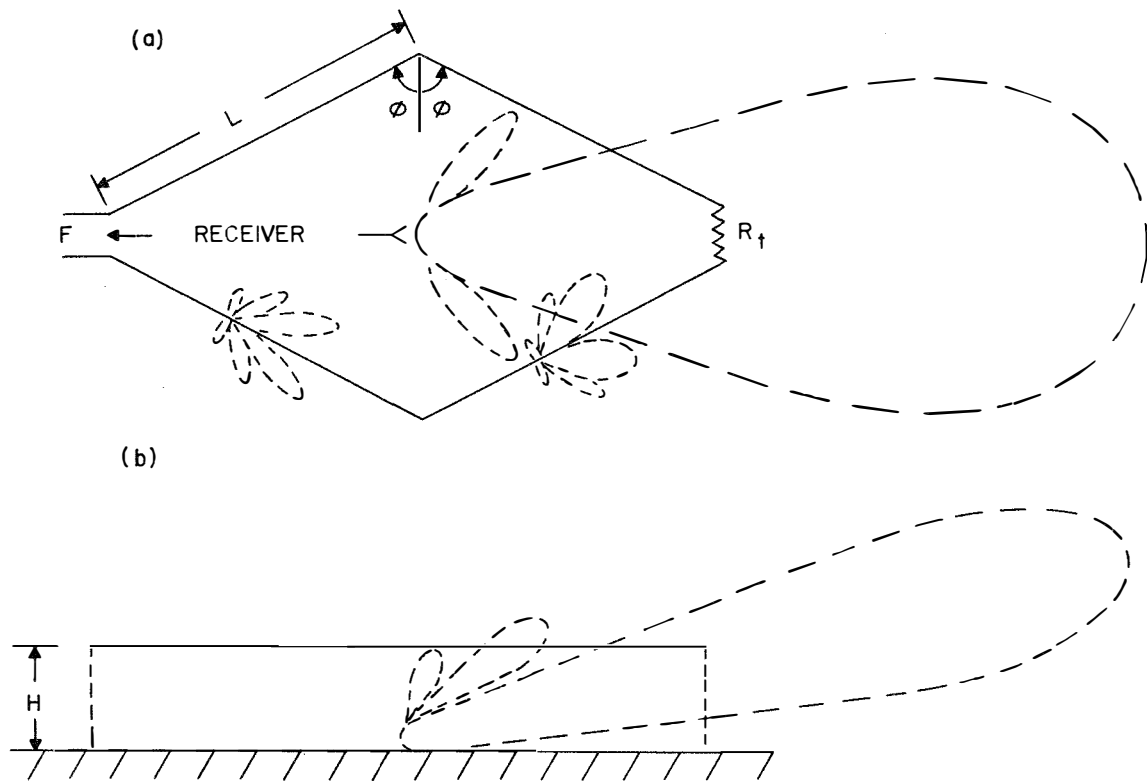


Figure 7-1. Single Wire Rhombic Antenna.

experience in long distance point-to-point communications during the past years that horizontally polarized antennas possessing good directional characteristics will give the best results. Some of the most universally used types giving excellent results are (1) rhombic, (2) horizontal doublets, (3) horizontal doublets with reflectors, (4) V and inclined V, and (5) the Beverage, or Wave Antenna. Of these five types, the first four lend themselves equally well to the transmission of signals as they do to reception of signals.

A. Rhombic Antenna

This antenna consists of four nonresonant wires arranged to form a diamond as illustrated in Figure 7-1. This arrangement has a very high directivity along a line drawn through the apexes, and the signals appearing on each leg of the diamond tend to be in phase, and are additive. This antenna has good gain characteristics, the magnitude of which is dependent upon the length of the legs as measured in wave lengths of the frequency being used. With the length of the legs of the antenna two or more wave lengths for the particular frequency being used, a gain of approximately 10 db can be expected when compared with a half wave dipole at the same elevation. The amount of gain secured from such an antenna varies considerably with the vertical angle used. Much higher gains can be secured by increasing the length of the legs where space will permit.

The rhombic has several variables inherent in it that to a large extent makes it a very flexible antenna. For instance, the height of the antenna above ground, the length of the legs and the tilt angle ϕ are all variables. It is quite possible, and often convenient to control the angle of the main lobe of the pattern of the antenna above the horizontal for maximum response by changing the tilt angle ϕ while the length of the legs and the height above ground are both left constant. The height of the antenna has considerable bearing on the directional characteristics, but lack of height can be compensated for by increasing the length of the legs of the antenna and modifying the tilt angle a small amount. The optimum tilt angle is not very critical as long

as the length of the legs is maintained at a length of two or more wave lengths for the main frequency to be used with the antenna.

There is a definite advantage when constructing a rhombic in making the antenna legs consist of two or three spaced wires. See Figure 7-2. This procedure reduces the characteristic impedance of the antenna making it easier to match the antenna to the transmission line, and also to help compensate for the fact that the different spacing between the sides of the legs tend to cause a varying characteristic impedance at different points in the antenna. This method of construction gives a slightly higher gain than the single wire type rhombic, and improves the frequency response of the antenna. Indications are that this method substantially reduces precipitation static intensity, which is an aid in certain geographical locations. A rhombic constructed of two spaced wires provides a good impedance match between the antenna and the open wire type of transmission line, which is generally used. Experience has shown that the increased performance secured from adding the third wire to such a system is so small compared with the cost that it is not common practice to do so. As a result of this, the rhombic can be used over a band of frequencies without requiring readjustment, or tuning. The main effect of changing the frequency used is to change the vertical angle of the main lobe of the pattern, and to change the sharpness of the directivity of the antenna.

With its distant end open the rhombic is bidirectional, but when the distant end (the apex opposite the feed point) is terminated with the correct resistance the antenna then becomes unidirectional with the terminal resistance absorbing the energy of the lobe for one direction. In this latter case the frequency of the system is broadened, thus making it desirable for many applications, especially reception. It is quite possible to modify the minor lobes to the rear of a terminated rhombic, and to obtain a desired null in any desired backward direction by modifying the terminating resistance a slight amount as is indicated by experiment. In order to obtain an ideal unidirectional characteristic of the antenna by use of a terminating resistor, that resistance must

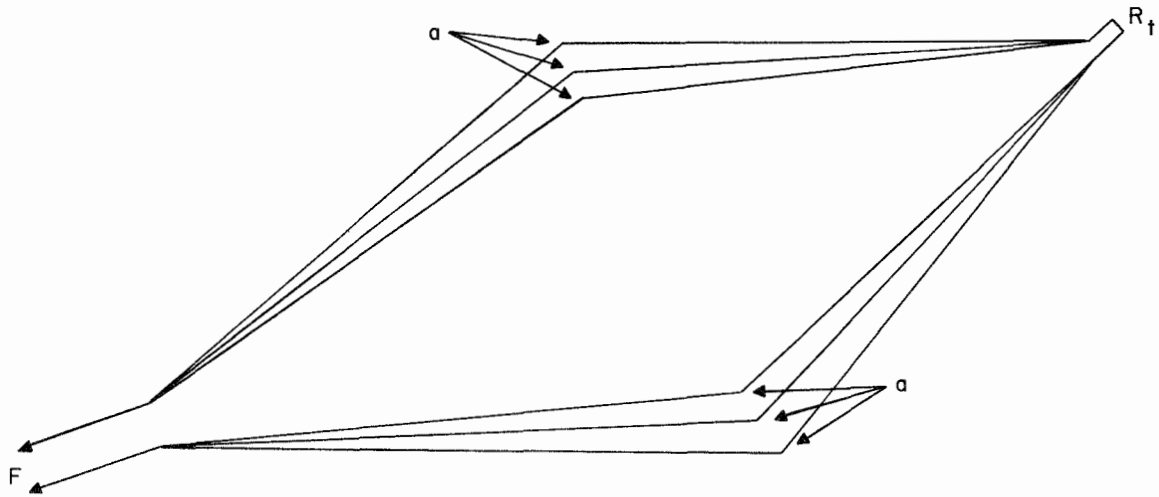


Figure 7-2. Multiple-Wire Rhombic Antenna.

equal the characteristic impedance of the feed line.

When the plane of the rhombic antenna is maintained parallel to the ground, the basic radiation characteristic of the antenna is that of a horizontally polarized antenna.

The critical parts of the rhombic antenna are: (1) the length L , which is the length of the legs of the antenna, (2) the tilt angle ϕ , which is sometimes referred to as the phase angle, (3) the height of the antenna above ground, and (4) in the case of a terminated rhombic, the value of the terminating resistor. However, as was previously mentioned, any two of these items can be modified to compensate for a lack of optimum conditions in the other items.

B. Doublet Antenna

For comparatively short and medium distance point-to-point communications work, the DOUBLET antenna is often used. This type of antenna has marked advantages and disadvantages over other types of antennas. Some of the advantages are: (1) Convenient size where lack of space prevents use of a

larger directional array; (2) Small cost as compared with larger directional arrays; (3) Ease of construction, and the rapidity with which this type of antenna may be lowered, repaired, and put back in service as compared with the more extensive arrays; (4) Its suitability for short distance communications where the high gain derived from higher directive arrays is not a necessity. On the other hand, some of its disadvantages are: (1) Definite lack of gain as compared with larger directive arrays for which it is possible to derive a 20 db gain as compared with the doublet; (2) In all compass directions, except perpendicular to the radiator, the radiation pattern has a vertically polarized component, and at all vertical angles directly off the ends, ONLY the vertical component exists. This is an important consideration when such an antenna is used for reception, for when the antenna is located over either average or poor ground, the antenna will have considerable response to vertically-polarized man-made noise waves arriving end-on at oblique angles; (3) This type of antenna is influenced much more by ground conditions than the larger directive

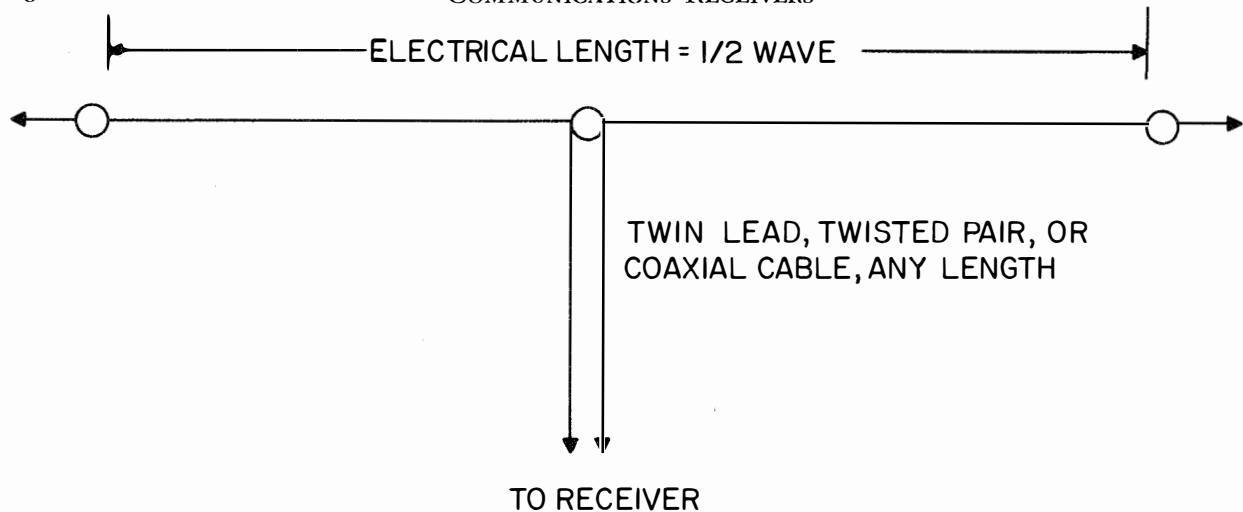


Figure 7-4. Half-Wave Doublet Antenna.

cable appears to be a better impedance match for the radiator than the higher value of the twin-lead or twisted pair types.

Many times it is desirable to increase the response of a signal in one direction or to reduce the response to an interfering signal in the opposite direction from that which is being worked. The term applied to this desired characteristic is **FRONT-TO-BACK RATIO**, or **FRONT-TO-BACK GAIN**. An increase in the ratio of the desired signal to all other signals is often achieved with the doublet by use of a reflector as a tuned but unfed parasitic element, which in the case of reception gathers the energy from a wave coming from the desired direction and focuses that energy on the receiving antenna. Reflectors performing this function are usually constructed of metal in order to have as perfect a reflection as possible for radio waves. In parabolic types of reflectors the antenna is located at the focus and the concentrating action is very much like that of a concave mirror. The plane surface type of reflector is used primarily to secure a unidirectional characteristic of the radiation pattern, and replaces any other reflecting components of the antenna system, but to be effective it must have large dimensions with respect to the diameter of the radiator.

C. "V" Antenna

One form of the long wire antenna used extensively in the past, and which is still found in use in many locations, is the V antenna.

There are many variations of this type of antenna, but only the basic unit (also known as the RCA Model D) will be considered here. This antenna is found operating under either resonant or non-resonant conditions. It can be constructed in a plane parallel to the ground, or it can be tilted with either the open ends, or the apex, at an angle. See Figure 7-5.

Basically, the V antenna consists of two long wires arranged to form a V with the wires in phase opposition. If the legs are two or more wave-lengths long the angle between the two long wires is normally twice the angle that would be formed between the first radiation lobe and the radiator of a single long wire antenna. This reinforces the radiation or response of the V in the plane of the antenna, and tends to cancel in other directions.

The V when operated as a resonant antenna is bidirectional, but by the use of any one of several methods it can be made unidirectional. One method of securing a unidirectional pattern is to erect an array of two or more of the basic V units wherein the added units are spaced, and are either fed in such a phase as to give the desired radiation pattern, or unfed and operating as reflectors. Another method of securing a unidirectional characteristic is to terminate the ends of each leg to ground with a noninductive resistor of half the value of the transmission line impedance. See Figures 7-6 and 7-7. This method is seldom used due to the difficulty in obtaining a good termination. It is generally believed that the rhombic is

arrays, which tends to increase the end-on response to noise; (4) Atmospheric noise, existing in the lower portion of the high frequency range, originating in equatorial regions, is pronounced off the ends of this type of antenna. Therefore, if a horizontal antenna of this type is being used for medium distance high-angle reception, the antenna should be orientated so that it is broadside to the distant noise sources, or erected in an east-west direction. When this is impossible, and it is necessary to run the antenna north and south, end-on response to noise may be minimized to a certain extent by tilting the antenna at an angle of approximately 10 degrees, with the high end of the antenna towards the noise area. When possible, the height of the antenna can be increased in order to increase the directivity and gain of the horizontal response, but doing this has little effect on the end-on response to vertically polarized noise.

A half-wave radiator center-fed with a resonant feed line having a characteristic impedance of 500-600 ohms as illustrated in Figure 7-3 is one of the most common applications of the doublet type of antenna. The main advantage of this type over other doublets is that it may be operated off its natural resonant frequency without much line loss, or without introducing undesirable antenna effect, if the radiator is symmetrically balanced to ground and nearby objects. Its main disadvantage is that it is critical as to transmission line length and the SWR is hard to control. However, when a good impedance matching system is used, this antenna may be used at any frequency between one-half and

up to twice the natural resonant frequency of the radiating portion of the antenna proper, without affecting the efficiency of the radiator, or the horizontal radiation pattern. Truthfully speaking, this antenna is not a half-wave antenna except when operated at one frequency in its band of operation, because the length of the radiator varies from less than $\frac{1}{4}$ to more than one wave-length. For all practical purposes, however, it may be considered a half-wave doublet, because the radiation pattern is consistent over the entire frequency range.

Another type commonly used is the half-wave doublet with symmetrical feed as illustrated in Figure 7-4. The radiation resistance, which also becomes the feed point resistance if the antenna is center fed, differs considerably from theoretical values, and is approximately 30 to 85 ohms.

This value is dependent upon the constants of the ground, the effective height of the antenna, nearby objects, and the ratio of the effective length to the diameter of the radiating conductor. The transmission line for use with this type of doublet can be of the twin-lead, or twisted pair types which are available with a characteristic impedance of 70-75 ohms. This type of line is suitable for direct connection, making it unnecessary to use any form of impedance matching unit at the antenna feed point. In most cases this type of line will also match the input of the majority of receivers. Solid dielectric coaxial cable can be used with this antenna with even better results than the twisted pair transmission line. This is due mainly to the fact that the 52 ohm coaxial

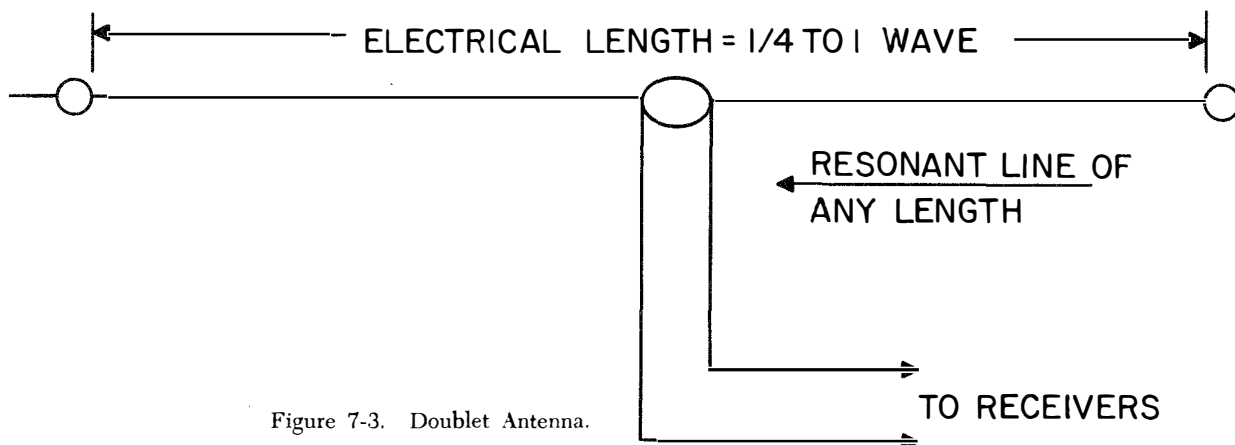


Figure 7-3. Doublet Antenna.

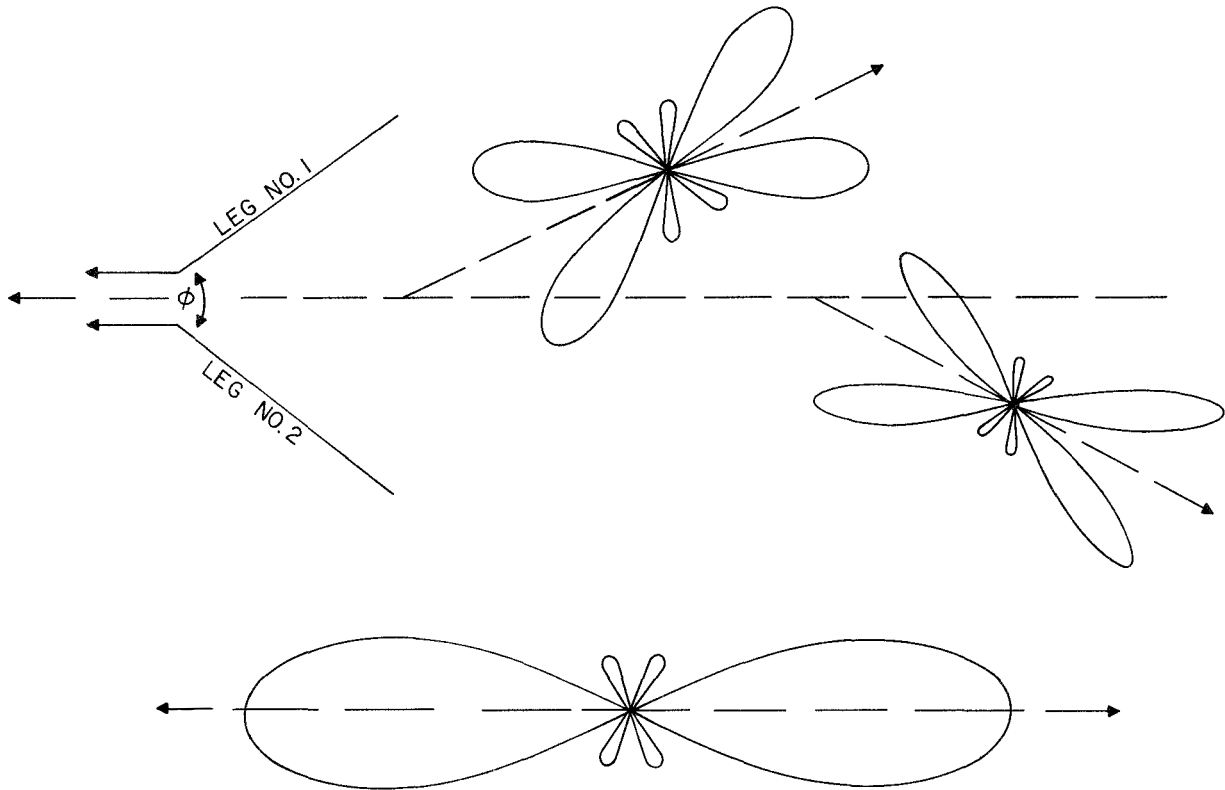


Figure 7-5. Basic "V" Antenna.

more satisfactory for securing unidirectional patterns by the termination method.

When the legs of the V antenna are several wavelengths long, good gain and high directivity may be achieved. However, the feed point resistance of the antenna varies a great deal with frequency, thus necessitating a very

flexible antenna coupling unit for operation over a band of frequencies. This last mentioned characteristic, along with the high SWR present on the transmission lines, is undesirable for reception over a band of frequencies. The SWR makes it necessary to use one or more matching stubs to flatten the transmission line. See Figure 7-8.

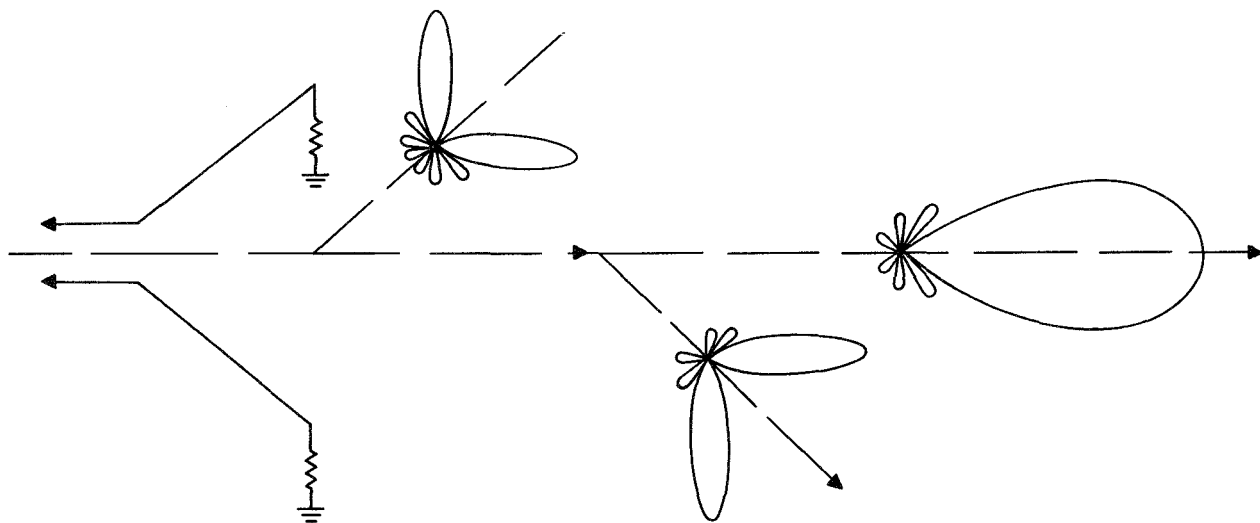


Figure 7-6. Non-Resonant "V" Antenna.

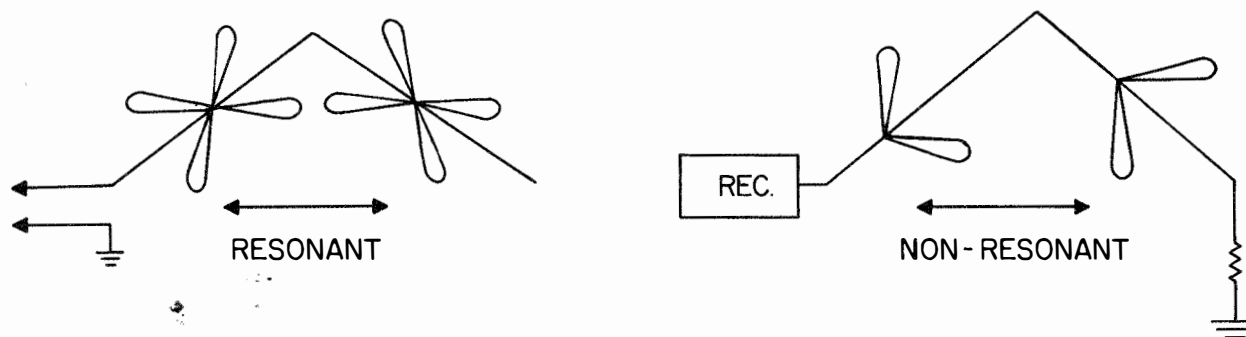


Figure 7-7. Other Types of "V" Antennas.

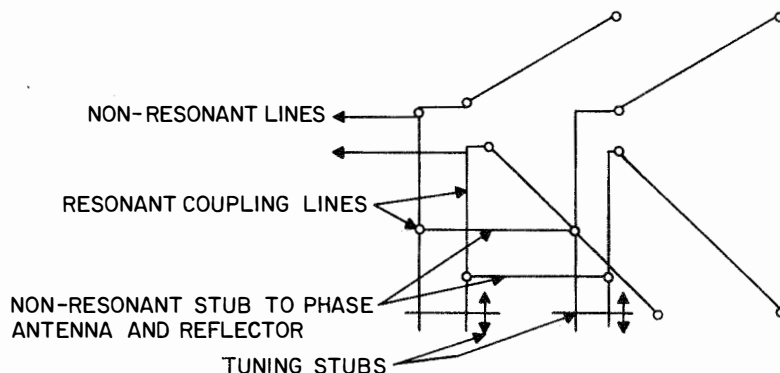


Figure 7-8. "V" Antenna and Reflector with Feed and Tuning System.

The radiation from a V array along the line of maximum directivity is horizontally polarized due to the cancellation of the vertical components from the two wires. In all other directions there are both horizontal and vertical components present. When it is desirable to broaden the vertical radiation or response along the line of the maximum directivity, without affecting the horizontal width, it is possible to do so by tilting the entire array a few degrees; this can also be accomplished by the use of another V used as a reflector.

D. Beverage Or "Wave" Antenna

One of the first long distance point-to-point communications receiving antennas, designed as such, was the BEVERAGE, (sometimes known as the "WAVE" antenna). The BEVERAGE antenna consists of a long non-resonant wire pointing in the direction of the transmitting station from which it is desired to receive signals. Ordinarily, this type of antenna is several wave lengths long. The end of the antenna pointing toward the trans-

mitting station is grounded with a resistor having a value that approximates the characteristic impedance of the antenna when the antenna is considered as a grounded one-wire transmission line.

The wave tilt produced by earth losses in polarized low-frequency waves traveling along the ground's surface makes it possible for the BEVERAGE antenna to abstract energy from the passing wave. When the wave is traveling in the direction of the receiver there are currents induced in different parts of the antenna which are additive due to the phase relationship of the currents.

When the direction of the wave is towards the transmitter the energy built up on the antenna is absorbed by the terminating resistance, and there is practically no effect on the receiver.

Due to the non-resonant operating conditions of the BEVERAGE antenna, it may be used for reception over a band of frequencies as long as all of the desired signals are in the same direction. Because of the high earth losses compared with the radiated energy, pro-

ducing low efficiency, the BEVERAGE antenna is not suitable for transmitting purposes.

7-3. TRANSMISSION LINES

In order to transfer energy from antenna to receiver, it is necessary to have a medium for so doing. Such a medium is known as a transmission line (or lines). While the study of transmission lines is a complex matter, we will be concerned only with those applicable to FAA installations. Specifically, we are concerned with both resonant and non-resonant two-wire lines, and concentric conductors.

Transmission lines may be operated in either a resonant or non-resonant manner. The manner in which they are operated is determined by the load impedance at the receiving end of the line. If the load impedance is a resistance, the value of which is the same as the characteristic impedance of the line, then, and then only, we have a non-resonant line. When operated in a non-resonant condition the voltage and current in the line decrease uniformly as one moves from the source toward the load end of the line. A non-resonant line does not have resonance points regardless of the length of the line; the voltage and current throughout the line are in phase, and the phase changes at a uniform rate of 360° per wave length of transmission line.

Non-resonant lines transmit energy at unity power factor and are thus more efficient than resonant lines; and, in proportion to the power handled, there is less stress on the insulation material.

On the other hand, if the load impedance terminating the line is not the same as the characteristic impedance of the line, then the line is operating as a resonant line. When operated as a resonant line, the voltage and current distribution in the line is such that there are definite resonance points repeating every half wave length. There is a marked phase difference between the voltage and current except where the voltage and current pass through their minima points. The terms "balanced" and "unbalanced" are employed to indicate how a line is functioning with respect to the presence of "antenna effect." "Antenna effect" is the term used to describe a transmission line that either radiates a por-

tion of a transmitted signal sent over it, or that receives a certain amount of signal from another station. Either of these two conditions would destroy the directivity characteristic of a directive antenna array. "Antenna effect" is therefore, very undesirable in transmission lines.

Line unbalance can be caused by an unbalanced condition existing at either the load or the source, or by lack of symmetry of the two wires with respect to ground or surrounding objects.

The TWO-WIRE OPEN LINE is probably the most common type of transmission line in use for long distance point-to-point communications work. One of the main reasons for this is that this type of line can be designed to have the lowest losses of any practicable transmission line for operation in the band of frequencies used in point-to-point communication work. Other reasons for its use are as follows: (1) it is symmetrical with respect to ground, thereby eliminating "antenna effect," (2) it is inexpensive.

Normally the Two-Wire Open Line consists of two unshielded, spaced, round conductors constructed in a horizontal plane at a definite height above ground. When it is necessary to have an extremely long line, the two wires are sometimes transposed periodically to minimize unbalance.

For practically all high frequency (above 50 mc) work, and for some specific cases in low and medium frequencies, concentric cable is used for transmission lines. This type of line can take the form of two spaced (one within the other) copper tubes, or the later designed and manufactured coaxial cable in which the braided shielding on the outside of the dielectric forms one of the conductors.

Two very important constants of any transmission line, that to a major degree determines the usage of that transmission line under a given set of circumstances, are (1) the CHARACTERISTIC IMPEDANCE (Z_0), and (2) the ATTENUATION factor α .

The characteristic impedance (Z_0) of any transmission line is a function of (1) the properties of the material that is used in constructing the line, and (2) the physical construction of the line. Specifically, Z_0 is de-

pendent upon (1) the series impedance per unit length of the conductor used, which is determined by the resistance (including skin effect), frequency used, and the inductance of the conductor, (2) the shunt admittance per unit length of the conductor used, determined by the leakage (in the form of dielectric losses), capacity of and between the conductors, and the frequency used. Ground effect (presence of ground and nearby objects), and the spacing between conductors must also be taken into consideration, for they also play an important part in determining the characteristic impedance of transmission lines.

The aforementioned data can be taken into consideration and necessary details can be determined mathematically, or in almost all cases can be secured from prepared tables which are given in practically all good engineering hand-books. For practical application of radio frequencies in the field, the characteristic impedance of a two-wire open transmission line can be determined by:

$$Z_0 = 276 \log_{10} \frac{2D}{d}$$

Where D=spacing between conductors
(center to center).
d=diameter of the conductors.

The characteristic impedance of a concentric conductor can be determined by:

$$Z_0 = 138 \log_{10} \frac{D}{d}$$

Where D=inner radius of the outer conductor.

d=outer radius of the inner conductor.

The attenuation factor of any transmission line determines the energy that will be dissipated by that line. This factor is determined by the resistance, leakage, frequency, and the extent to which the ratio of the resistance to the inductance differs from the ratio of the leakage to the capacity. For practical application of radio frequencies in the field the attenuation factor can be determined by:

$$\alpha = \frac{R}{2Z_0} + \frac{GZ_0}{2}$$

Where R=resistance of the line.

G=leakage of the line.

Z₀=characteristic impedance of the line.

CHAPTER 8

VHF FM RECEIVER TYPE RF-2

8-1. GENERAL DESCRIPTION

The Type TF-2 Transmitter, TF-2/3 Modulator and RF-2 Receiver have been designed to operate together to provide a high quality wide band VHF communication system suitable for use with multi-channel link equipment. Particular emphasis has been placed on design features which contribute to low over-all system distortion and flat audio frequency response. The audio input and output impedances are closely maintained over the operating frequency range to insure satisfactory performance when used in connection with suitable link terminating equipment. In addition, this equipment has been designed to provide maximum freedom from the generation of, and susceptibility to, spurious radiations. Standard door type panel rack construction has been employed, with all tuning controls accessible from the rear.

The TF-2 Transmitter requires 12-7/32 inches of vertical mounting space and the TF-2/3 Modulator requires 8-23/32 inches. The transmitter and modulator may be installed in the same rack or separated by a distance of as much as fifty feet. The only interconnection necessary between the units is a suitable length of type RG-8/U cable.

The RF-2 Receiver requires 13-12/32 inches of vertical space.

A. *Basic Principles of Operation*

The RF-2 receiver employs a total of 23 tubes and includes a self-contained regulated power supply. Two stages of RF amplification are provided ahead of the mixer which is followed by a three stage 18 Mc intermediate frequency amplifier, two limiters and the discriminator. The local oscillator is crystal controlled by a

series resonant harmonic mode crystal operating at one-third of the injection frequency. An electron coupled circuit is employed and the crystal frequency is tripled in the plate circuit of the oscillator. An injection amplifier follows the oscillator and amplifies the injection frequency to the proper level for application to the mixer. Low side injection is used.

Following the discriminator, a cathode follower reduces the source impedance of the signal to a low level for application to a frequency compensation network which provides nominal 10 db of boost at 300 cps, 5 db at 650 cps and 1 db at 50,000 cps. The response of this network is essentially flat between approximately 3,000 and 30,000 cps. Its over-all response is designed to compensate for the audio response of the Type TF-2 transmitter and result in a system response which is flat within ± 0.5 db between 300 and 50,000 cps.

The output of the frequency compensation network is fed to a three stage audio amplifier which provides 100 milliwatts output at 600 or 150 ohms impedance. The gain of the audio amplifier is adjusted by varying the amount of negative feedback used, and in addition, two fixed, strap-attenuators are provided. One of these has a loss of 6 db and the other a loss of 12 db. They may be connected when needed under certain circumstances to extend the range of audio level control.

Two indicating instruments are provided. A front panel-mounted meter is connected in the grid circuit of the 3rd IF amplifier and a chassis-mounted meter and switch permit metering crystal voltage, IF amplifier and limiter grid currents, injection amplifier grid current and audio output ahead of the frequency compensating network.

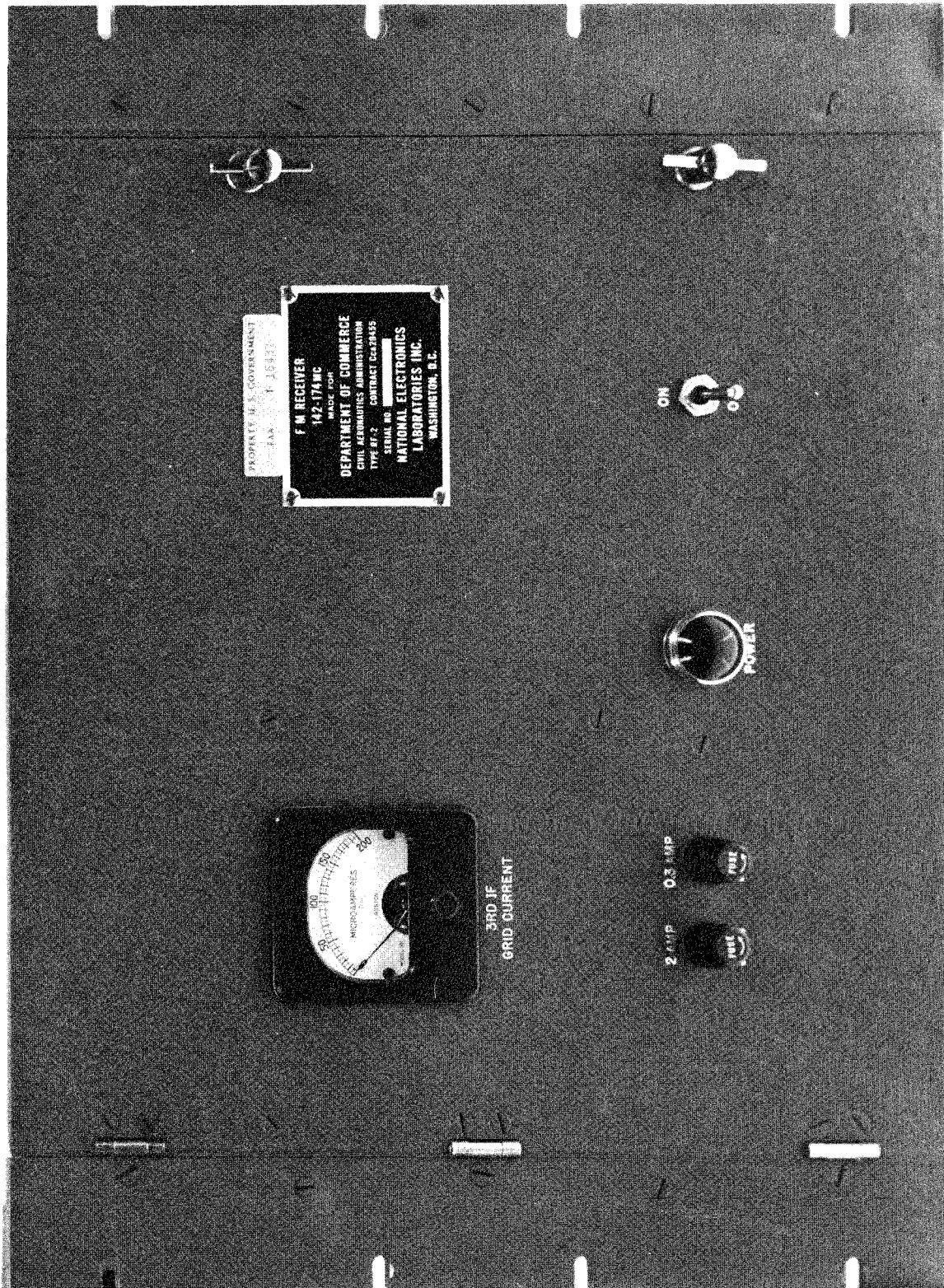


FIGURE 8-1. Type RF-2 Receiver.

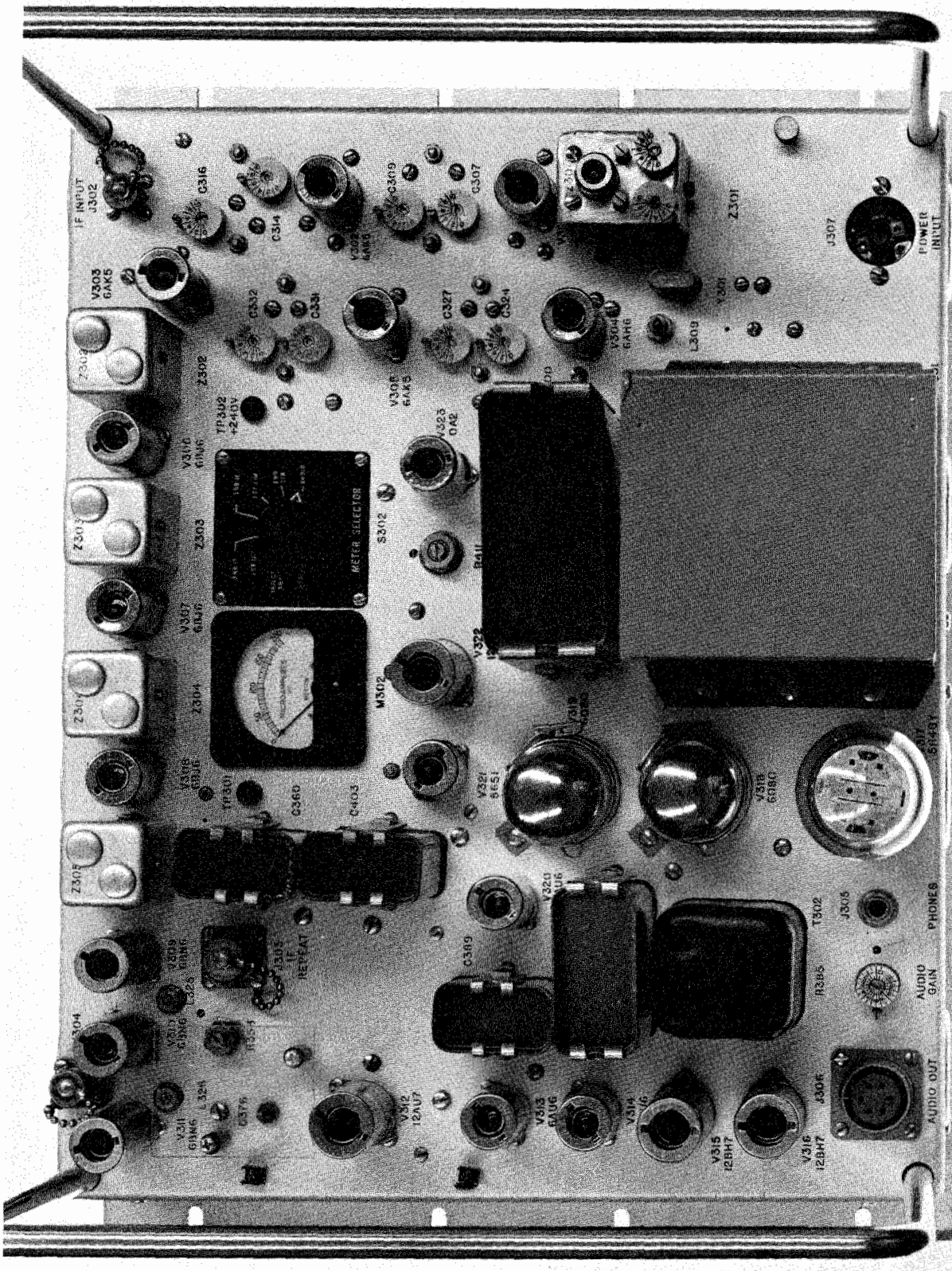


FIGURE 8-1 (a). Type RF-2 Receiver. (Back View)

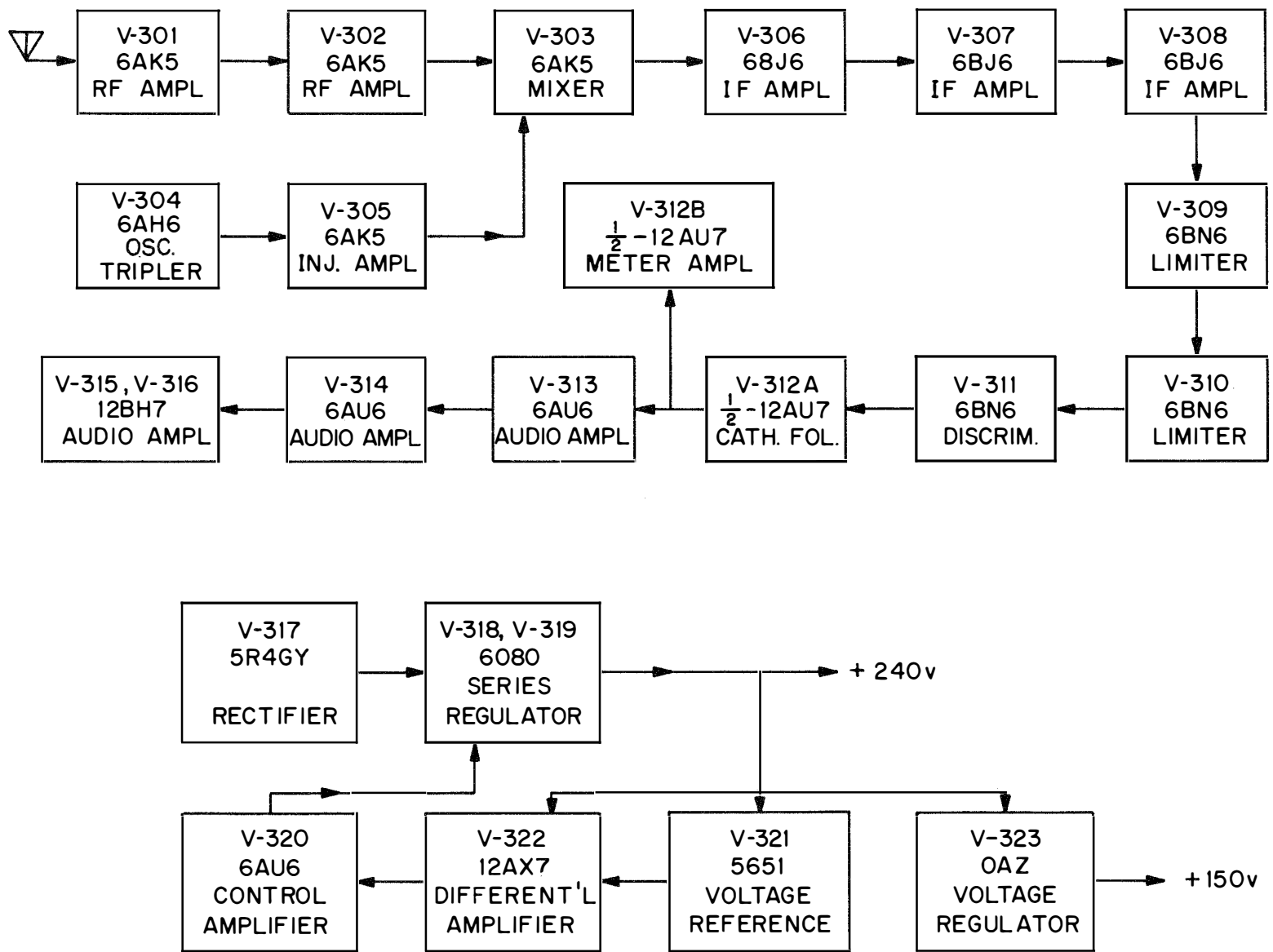


FIGURE 8-2. Block Diagram of RF-2 Receiver.

8-2. THEORY OF OPERATION

A. General

Figure 8-2 shows a functional block diagram of the RF-2 Receiver. This section will discuss in detail the circuit theory of the equipment following the general sequence of the block diagram.

B. RF Amplifier

Reference is made to the schematic diagram. Figure 8-12 shows the RF amplifier uses two type 6AK5 pentodes in a conventional two-stage circuit. The antenna is coupled to the grid of V-301 by an RF transformer Z-301 which is a critically coupled double tuned circuit. The primary and secondary are tapped to provide an accurate impedance match between a 51-ohm transmission line and the input admittance of V-301. Tuning is accomplished by means of variable capacitors C-301 and C-302.

The plate circuit of V-301 is coupled to the grid of V-302 by inductors L-304 and L-305 which are tuned by variable capacitors C-307 and C-309. The plate circuit of V-302 is likewise coupled to the grid of the mixer by inductors L-307 and L-308 which are tuned by capacitors C-314 and C-316.

The midband gain of the RF amplifier from antenna terminal to the grid of the mixer is approximately 175. The 3-db bandwidth is slightly over 2 Mc and the 50-db bandwidth is approximately 8 Mc.

The entire RF amplifier is constructed in a shielded compartment and decoupling is employed in both plate and filament supply leads.

C. Local Oscillator

The local oscillator employs an electron-coupled modification of the impedance-inverting Pierce circuit and uses a type CR-52/U crystal unit in the frequency range of from 41.333 to 52.000 Mc.

Impedance inverting oscillators employ a network similar to that shown in Fig. 8-3, to permit conventional lower frequency oscillators to be operated with crystal control in the VHF range. The impedance inverting network is designed to behave as quarter-wave line having

a characteristic impedance:

$$Z_o = \omega L_1 = \frac{1}{\omega C_n} = \frac{1}{\omega C_o} \quad (8-1)$$

With this design, the network always appears as an inverted Z_s equal to $Z_n = Z_o^2/Z_s$ where Z_s is in the series-arm impedance of the crystal. If $Z_s = 0$, C_o is shorted out and Z_n is infinite (assuming zero losses). If Z_s is infinite, L_1 is series resonant with C_o , and $Z_n = 0$. If $Z_s = Z_o$, the network appears as an infinite line with $Z_n = Z_o$. When Z_s is a small inductive reactance, Z_n is a large capacitive reactance and vice versa. With Z_o much greater than the resistance, R , of the crystal unit, the network serves to invert the crystal resistance to a high impedance equal to Z_o^2/R . Figure 8-4 shows the actual circuit of the local oscillator. The impedance inverting network consists of inductor L_1 (L-309), the capacitance (C_n) of crystal unit Y-301 and the input capacitance (C_o) of the 6AH6 V-304. Since the input capacitance of a 6AH6 is somewhat higher than the capacity of the crystal unit, its value is effectively reduced by the use of a shunt inductor L-303. Capacitors C-320 and C-322 are for dc blocking. Resistor R-308 across the crystal unit damps out a second anti-resonant frequency—the square of which is $\omega^2 = (C_n + C_o)/(C_n C_o L_1)$. In addition, R-308 serves to reduce the effective series resistance of crystals which approach the upper limit of resistance permitted for the type CR-52/U. Diode CR-301 is provided to permit metering of the crystal voltage to provide a means of properly adjusting L-309. Even

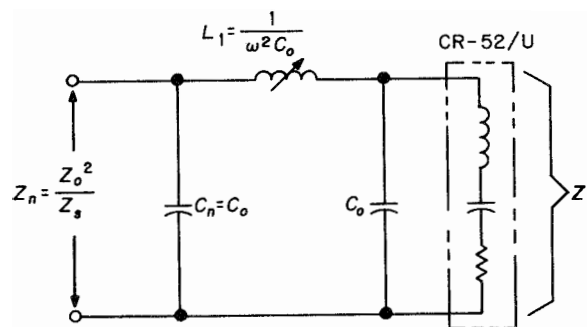


FIGURE 8-3. Basic Impedance Inverting Network.

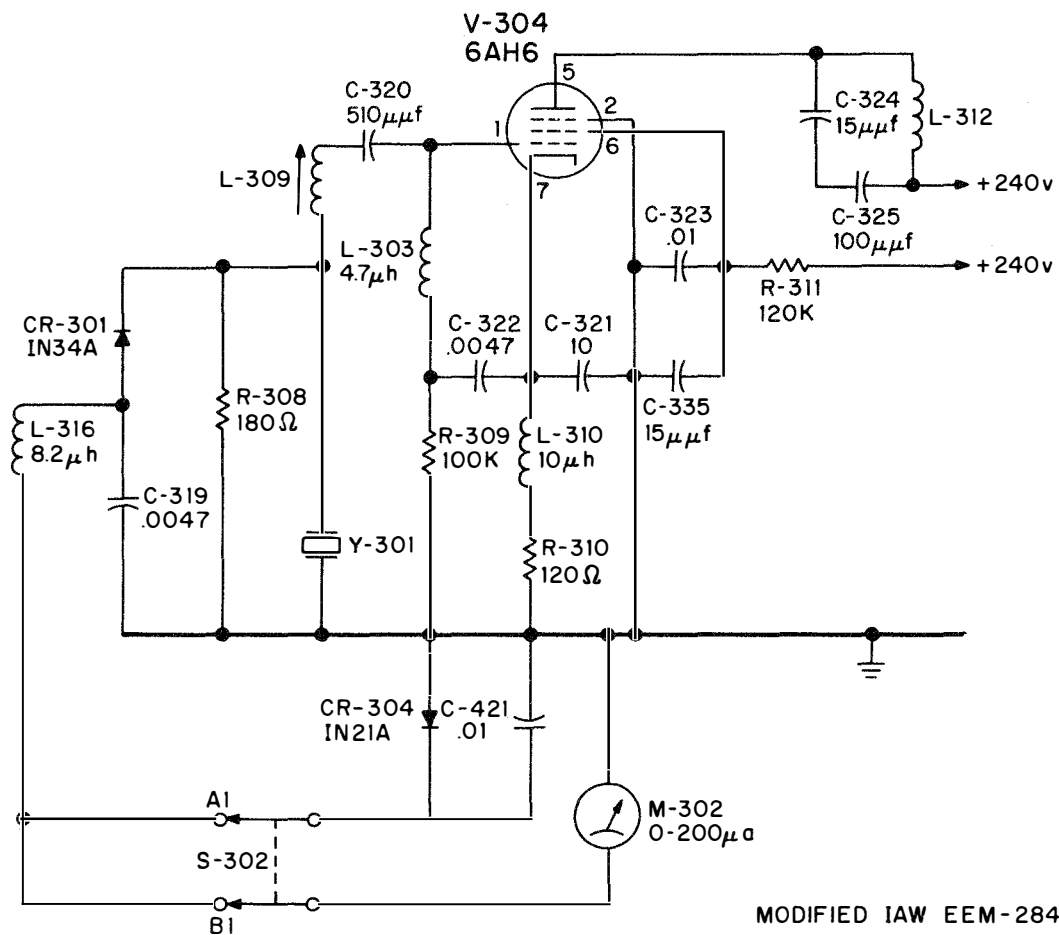


FIGURE 8-4. Partial Schematic, Local Oscillator.

though the equivalent impedance inverting network is designed to be anti-resonant at approximately the crystal frequency, the operating frequency may well require that the crystal network facing the actual terminal connections be reactive if the necessary phase reversal is to be accomplished. As a result of this, proper adjustment of L-309 is obtained when a minimum RF voltage appears across the crystal. This minimum represents the condition when the crystal network is adjusted for proper phase reversal causing the reactive components of the crystal impedance to cancel out (series resonance) leaving only the effective series resistance. If inductor L-309 is detuned, the reactive elements of the crystal impedance increase, causing an increase in the voltage appearing

across the crystal together with a shift in operating frequency and a decrease in frequency stability. If it is detuned far enough, oscillation will cease and the voltage across the crystal will drop to zero. Figure 8-5 shows a plot of meter indication versus adjustment of inductor L-309. It should be noted that since the impedance of the crystal network changes over the operating band and because of the wide variation in the resistance of individual crystal units, both the height of the reactive voltage peaks and the depth of the resonant voltage minimum are subject to wide variations. Diode CR-304 is connected across meter M-302 so that its forward resistance shunts the meter and limits the indication obtained on the reactive peaks to approximately 3/4 of full scale.

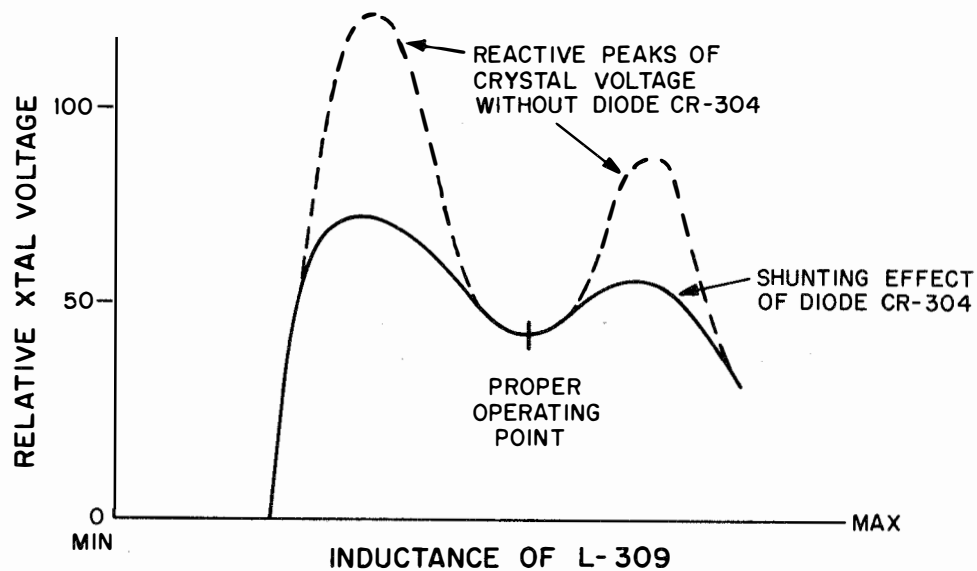


FIGURE 8-5. Crystal Tuning Characteristics.

The cathode circuit of the oscillator is broadly tuned by inductor L-310 and capacitor C-321 to a frequency somewhat below the lowest crystal frequency and the plate circuit is tuned by inductor L-312 and capacitor C-324 to the third harmonic of the crystal frequency. Almost complete isolation between the plate circuit and the crystal circuit is obtained with the configuration, and tuning of the plate circuit has essentially no effect on the output frequency.

The following formula should be employed to determine the crystal frequency (f_x) required for a given operating frequency (f_o)

$$f_x = \frac{f_o - 18 \text{ Mc}}{3} \quad (8-2)$$

D. Injection Amplifier and Mixer

The output of the crystal oscillator stage is coupled to V-305—a 6AK5 operated as a Class A injection amplifier. Since the output of the oscillator varies widely with different frequencies and different crystals, a very high value of screen dropping resistor R-315 is employed for V-305 so that the resulting screen limiting will stabilize the output voltage level of the injection amplifier. In addition, diode CR-303 is used as a meter shunt in the grid return. The forward resistance of CR-303 in parallel with that

of diode CR-304 limits maximum meter deflection to approximately 3/4 of full scale when the meter selector switch S-302 is in the INJ. AMP. position.

The output of the injection amplifier is inductively and capacitively coupled to the grid of the mixer V-303, which is also a type 6AK5 pentode. It is mixed in the grid circuit of V-303 with the output of the second RF amplifier V-302. The plate circuit of the mixer is tuned to the 18 Mc difference frequency of these two signals by the primary of the first IF transformer Z-302. The screen of V-303 is operated at a somewhat lower than normal voltage in order to minimize noise resulting from partition effect in the mixer.

A type BNC Jack (J-302) is provided in the grid circuit of the mixer as a convenience in applying an 18 Mc signal to the mixer for aligning the IF amplifier.

E. IF Amplifier

Three stages of IF amplification are employed using type 6BJ6 pentodes, V-306, V-307 and V-308. Grid limiting is incorporated in each and the grid current of the third IF amplifier, V-308 is metered by the front panel-mounted meter, M-301. Grid current of all three stages may be metered by the chassis-mounted meter,

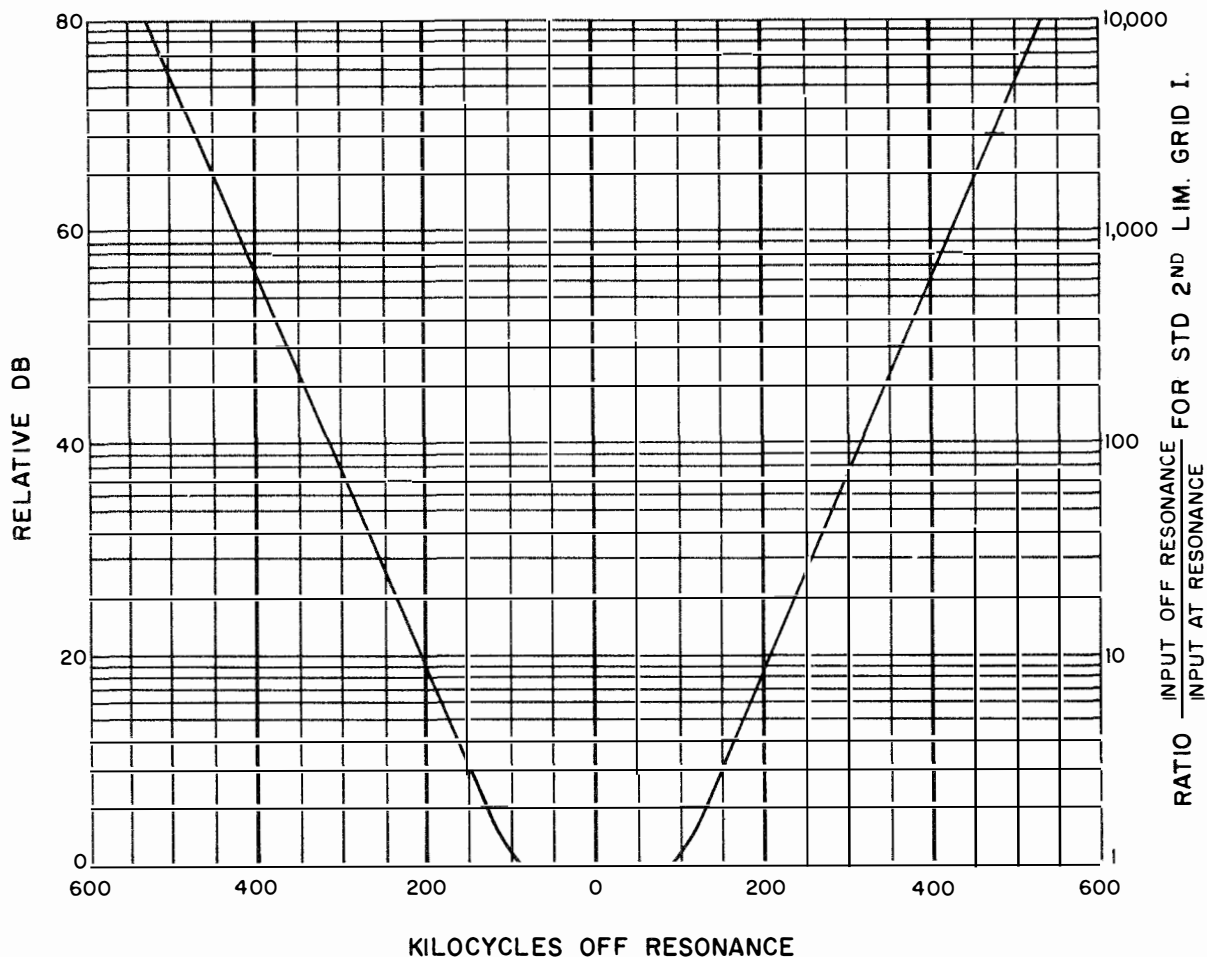


FIGURE 8-6. IF Selectivity Curve.

M-302 and meter selector switch S-302. The intermediate frequency is 18 Mc and the IF transformers, Z-302, Z-303, Z-304 and Z-305 are each double-tuned by means of the iron cores in their respective primaries and secondaries. The first three IF transformers are critically-coupled but the fourth, Z-305, is over-coupled and test point TP-301 is provided as a convenience in tuning.

A small capacitor C-418, diode CR-305 and resistor R-317 are connected to the primary of Z-305. With a dc vacuum tube voltmeter connected to TP-301, transformer Z-305 is properly tuned by first *detuning the secondary* and then adjusting the primary for maximum indication on the VTVM. The secondary is then tuned for maximum dip in the VTVM indication without further readjustment of the primary. This procedure is a simple yet quite accurate method

of aligning over-coupled transformers. The first three IF transformers, being only critically-coupled, may be aligned by tuning for maximum indication on meter M-302 with the meter selection switch in the appropriate position. Because of the flat top response curve of these transformers, it is necessary to adjust the tuning carefully to the center of the slug position which gives maximum response. This is necessary to avoid tilting the nose of the response curve and to ensure that the IF is centered on the desired pass-band.

Figure 8-6 shows the IF response curve characteristics as measured from the IF input jack J-302 (grid of the mixer to the grid of the second limiter) as obtained by increasing the off resonance input to the level necessary to maintain constant second-limiter grid current as indicated on meter M-302.

F. Limiters

Although the IF amplifier incorporates grid limiting, it is only operative on relatively strong signals; therefore two stages of supplementary limiting are provided following the IF amplifier. The first and second limiters, V-309 and V-310 employ the type 6BN6 gated-beam pentode. The 6BN6 has characteristics such that essentially step-function control of plate current is realized by application of relatively small signal voltages to either the number one or number three grid. In both stages the number one grid is employed as the signal grid and the number three grid is maintained at zero bias by being directly connected to the cathode.

The output of the first limiter V-309 is capacitively coupled to the grid of the second limiter and also to the IF repeater jack J-303 where a minimum voltage of 10 millivolts is available for excitation of a heterodyne type repeater transmitter. Connection of any type load, from open circuit to short circuit, to J-303 will have no significant effect on the operation of the receiver.

The grid of V-310 is tuned by inductor L-325 which is a miniature slug-tuned shielded coil. L-325 tunes quite broadly because of the loading effect of resistor R-349. A type BNC connector, J-304, is provided to permit ready connection of an external oscilloscope to the grid circuit of the second limiter in the event it is desired to employ sweep-alignment techniques in adjusting the IF amplifier and limiter tuning.

The second limiter provides limiting on noise alone and is saturated with input signals of 2 microvolts or more at the antenna terminals. Its plate circuit is capacitively coupled to the discriminator V-311.

G. Discriminator

The 6BN6 gated-beam pentode functions as a discriminator by virtue of the fact that its two control grids are operated in a quadrature phase relationship at center frequency. As the input frequency is varied, the phase of one grid is caused to vary with respect to the other about the quadrature point. The resultant plate current when properly integrated contains the modulation content of the incoming fre-

quency modulated voltage.

Since each of the control grids (No. 1 and No. 3) are capable of exercising essentially step function control of plate current, the action at center frequency may be visualized by considering that a plate current pulse is initiated by a positive excursion of the signal voltage on the No. 1 grid. The No. 3 grid is excited at quadrature phase relationship by space charge coupling and it swings negative during a part of the cycle when the No. 1 grid is still positive and thus terminates the plate current pulse. The result is a square wave of plate current which occurs at a repetition rate equal to the signal frequency. If the signal frequency is shifted to one side of center, the phase relation with respect to the tuned quadrature grid is shifted and the duration of the plate current pulse is increased. If the signal frequency is shifted to the other side of center, the duration of the plate current pulse is reduced. The integrated sum of these plate current pulses when developed across a load resistor yields a voltage which varies in accordance with the frequency modulation of the input signal. In order to obtain proper operation it is essential that the following conditions be met:

1. The voltages applied to the signal grid and space charge coupled to the quadrature grid must both be of sufficient amplitude (above 3 v) to control the plate current between the limits of cut-off and saturation.
2. The cathode bias and the voltage on the No. 2 grid must be such that total cathode current is essentially independent of the voltage applied to the control grids.
3. The quadrature grid must be tuned exactly to center frequency and its tuned circuit must exhibit linear phase shift characteristics either side of center frequency for a distance adequate to accommodate the farthest significant sideband currents associated with the frequency modulated signal.
4. The average plate current must remain constant with and without signal.

Figure 8-7 shows a partial schematic of the discriminator circuit employed in the type RF-2 Receiver. Output of the second limiter is coupled through capacitor C-372 to the signal grid of V-311 which is broadly tuned by the shielded miniature inductor L-326. Cathode bias is established by variable resistor R-358.

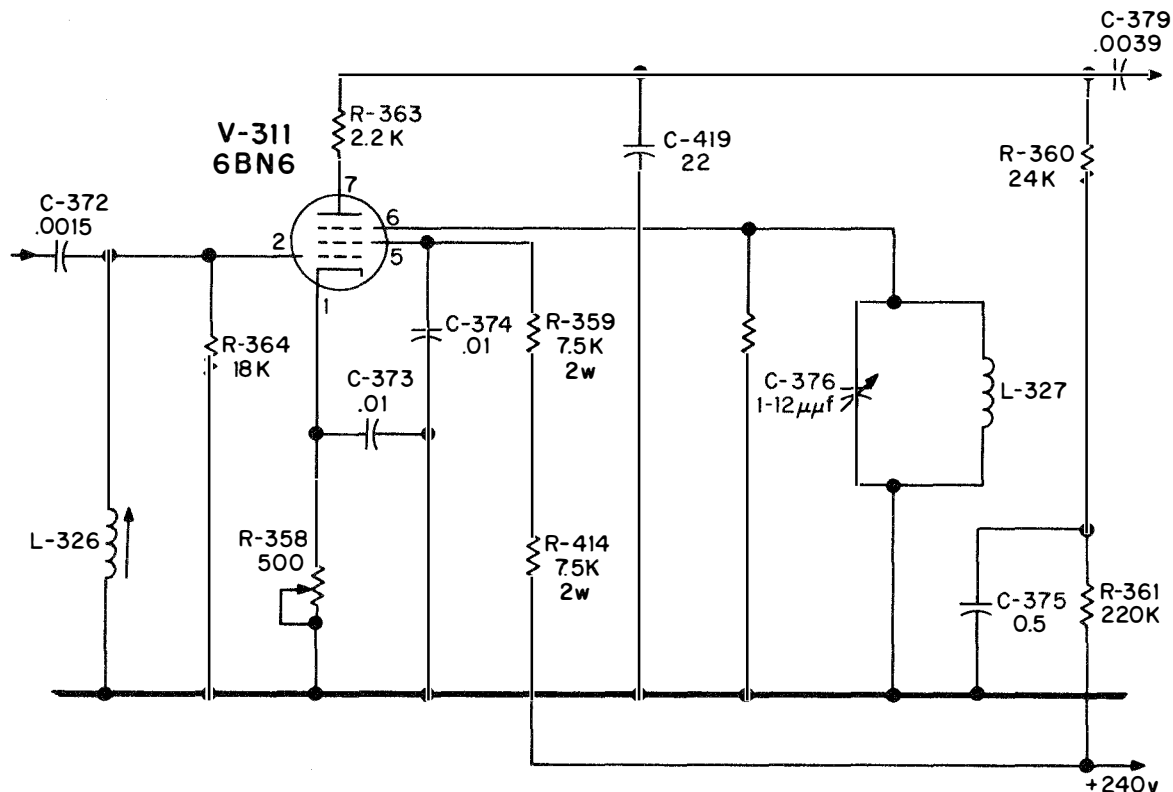


FIGURE 8-7. Partial Schematic, Discriminator.

Resistor R-363 together with capacitor C-419 and resistor R-360 form the plate circuit integrating network; resistor R-361 and capacitor C-375 provide plate decoupling. The quadrature grid is tuned by variable capacitor C-376 and inductor L-327 which are shunted by resistor R-362 to establish proper operating Q . Tuning of the discriminator is accomplished by adjusting C-376 and R-358 to provide maximum audio output from a modulated incoming signal. Use of a distortion analyzer when making these adjustments will enable the total harmonic distortion of the audio output to be reduced to less than 1/2%, however, less than one per cent distortion will result if these controls are simply adjusted carefully to yield maximum audio.

The tuned circuit in the quadrature grid has been carefully designed to provide maximum frequency stability through the temperature range encountered over the service conditions. Capacitor C-376 employs glass and invar construction and has a temperature coefficient of zero ± 50 ppm/ $^{\circ}$ C. Inductor L-327 is wound on a grooved form made from material

that has a coefficient of expansion such as to cause the distributed capacity of the coil to vary in such a way as to largely offset the change in inductance of the coil over the temperature range.

H. Cathode Follower and Meter Amplifier

The audio output of the discriminator is coupled through capacitor C-379 to the grid of 1/2 of a type 12AU7 dual triode, V-312A which is operated as a cathode follower. V-312A serves to reduce the source impedance of the audio signal to a low and constant value for application to the audio de-emphasis network. This network has frequency response characteristics which are the complement of those of the audio pre-emphasis network employed in the type TF-2 Transmitter equipment; *i.e.*, response at 300 cps is up 10 db, and 650 cps is up 5 db over response at a mid-frequency of say 10,000 cps. In addition the response at 50,000 cps is boosted by from 0.75 to 1.25 db, depending on temperature, over mid-frequency response. Figure 8-8 is a partial schematic of the cathode follower and de-emphasis network.

Operation of the de-emphasis network is best understood by examination of equivalent circuits. Figure 8-9 shows the equivalent circuit at mid-frequencies. The cathode follower has been replaced with a constant voltage generator having an internal impedance of 500 ohms which is loaded with a voltage divider consisting of R-373 and R-374 and the output voltage is seen to be slightly more than one-fifth of the generator voltage.

Figure 8-10 shows the equivalent circuit at 50,000 cps. In this case, capacitors C-380, C-384 and shunting resistor R-373 cause the output voltage to rise slightly with increasing frequency. Thermistor RT-301 in series with capacitor C-384 modifies its effect depending on temperature. At -10°C , RT-301 has a resistance of approximately 100,000 ohms and capacitor C-384 is relatively ineffective. At $+60^{\circ}\text{C}$, RT-301 has a resistance of approximately 3,000 ohms and the effect of capacitor C-384 is to raise the output voltage about 0.5 db at 50,000 cps. This characteristic compensates for an inverse temperature dependent response characteristic which exists in the type TF-2 transmitting equipment.

Figure 8-11 shows the equivalent circuit at 300 cps. In this case the impedance of L-329 and C-387, which are parallel resonant at 300 cps, is high and since it is in series with resistor R-374, the output voltage is over three-fifths of the generator voltage. Below 300 cps the increasing reactance of capacitor C-385 together with the decreasing impedance of the tuned circuit causes the output voltage to fall rapidly.

Output of the cathode follower (V-312A) is coupled to meter amplifier (V-312B) from a convenient point in the de-emphasis network. V-312A is operated as a Class A amplifier with degeneration due to its unbypassed cathode resistor. Its output is rectified by diode CR-302 and is metered by the rear chassis-mounted meter (M-302) when the METER SELECTOR switch is in the AUDIO position.

I. Audio Amplifier

The audio amplifier consists of three stages in cascade. The first two stages employ type 6AU6 pentodes (V-313 and V-314) and the third stage uses a pair of dual triodes, type

12BH7, with all four sections in parallel. Overall current feedback is employed from the cathode of the third stage to the cathode of the first stage. The variable feedback resistor R-385 serves as the gain control for the amplifier and provides a range of 30 db control by varying the feedback from 20 db to 50 db.

Output of the de-emphasis network is coupled through capacitor C-386 to the grid of V-313. Resistor R-375 is the grid load resistor and its value is selected to cause audio response to fall off rapidly below 300 cps. Together with R-421 it forms the grid return for V-313. Capacitor C-411 provides audio decoupling so that the feedback voltage appearing at the junction of resistors R-389 and R-376 is prevented from reaching the grid. R-389 is the cathode bias resistor and R-376 establishes, together with R-385, the minimum fraction of voltage which is fed back when the AUDIO GAIN control is set for maximum. It also establishes the dc voltage at the junction of R-389 and R-376 to a level approximately equal to the dc voltage appearing on the cathodes of V-314 and V-315. This prevents the gain control (R-385) from conducting appreciable dc at the minimum resistance setting and provides smoother operation of the control. Capacitor C-412 and resistor R-372 control gain and phase shift characteristics of V-313 at frequencies well removed from the desired response-band and prevent an oscillation from developing at a frequency of about 2 Mc. Output of V-313 is coupled to V-314 which is operated as a pentode with localized feedback because of its unbypassed cathode resistor R-381.

The second stage output drives the third stage consisting of the parallel triodes of V-315 and V-316. Resistor R-390 serves as the plate load resistor for this stage. Since at least 20 db of current feedback is employed at all times, the effective plate resistance of this stage is raised by a factor of 10 or more and this high resistance in parallel with the 650 ohm plate load resistor therefore appears as a 600 ohm source impedance to the output transformer T-302. Capacitor C-395 provides coupling to this source impedance from the output transformer. Capacitor C-394 corrects a portion of the reactive component of the output impedance at the higher audio frequencies.

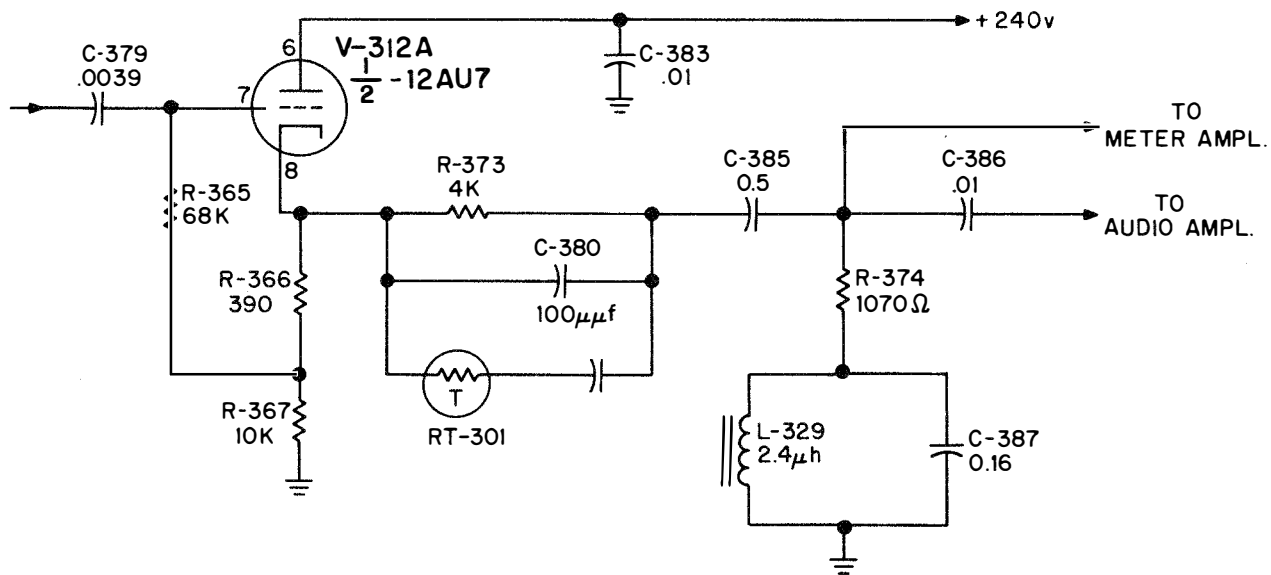


FIGURE 8-8. Partial Schematic, Cathode Follower and De-emphasis Network.

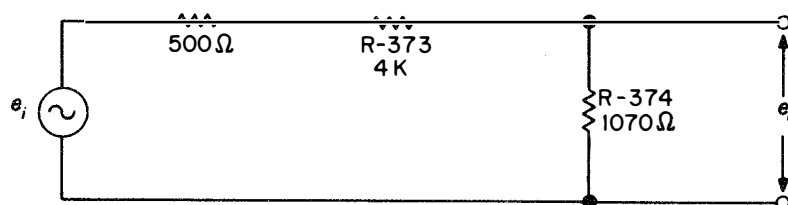


FIGURE 8-9. Equivalent Circuit at Mid-Frequencies.

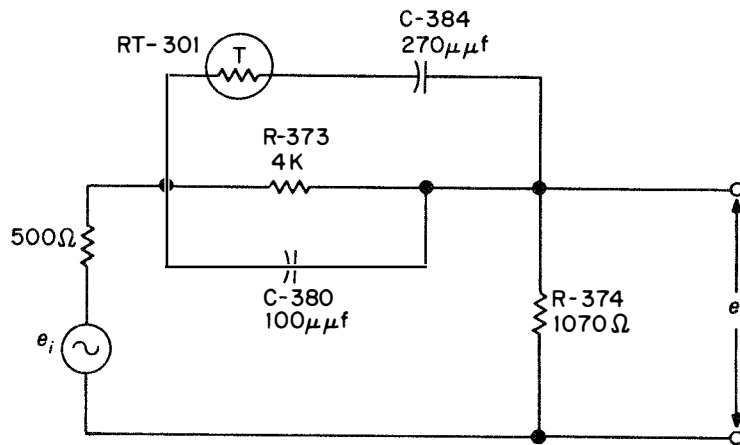


FIGURE 8-10. Equivalent Circuit at 50,000 cps.

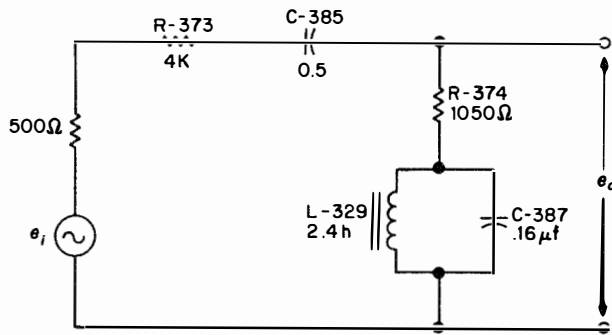


FIGURE 8-11. Equivalent Circuit at 300 cps.

A two-section, 600-ohm strap-type T pad is provided between C-395 and T-302. If terminals A and C only are strapped together, the pads are cut out of the circuit entirely. If A is strapped to B and E to F, 12 db attenuation is provided. If C is strapped to B and D to E, then 6 db attenuation is provided. If D, E and F are strapped together, then 18 db of attenuation results. The 18 db attenuation provided by the pads together with the 30 db in the gain control permits a total gain variation range of 48 db in the amplifier.

The audio amplifier is designed to provide a maximum power output of only slightly over 100 milliwatts; therefore if the pads are inserted when more than +10 dbm output level is required from the receiver, the amplifier will be called on to deliver more than 100 milliwatts power level to the pads and distortion will result. The use of pads is required only when output levels below 0 dbm are desired.

The secondary of T-302 consists of two balanced 300-ohm windings which are connected in series-aiding for a 600-ohm output impedance. They may be connected in parallel-aiding for 300 ohms output impedance.

An auxiliary output jack, J-305 is provided for monitoring the audio output with a 2,000 ohm headset. The output jack is connected to the cathode of the third audio stage through resistor R-387 and capacitor C-393 and is arranged so that insertion of the headset plug will remove load resistor R-388 and prevent any appreciable change in output level.

J. Power Supply

The power supply employs a conventional full-wave rectifier followed by an electronic regulator to minimize ripple and effect of line voltage variations.

The ac input from receptacle J-307, is brought through Z-306, a shielded lowpass RF filter which is mounted directly on the rear of the power receptacle, through the fuse F-301 and switch S-301 to the primary of power transformer T-301. T-301 provides plate and filament power for all stages in the receiver. The high voltage output is rectified by a type 5R4GY full-wave rectifier V-317. The output of the rectifier is partially filtered by resistor R-392 and capacitor C-400 and applied to the plates of a pair of type 6080 series regulator tubes – V-318 and V-319 connected in parallel. The grids of the series regulators are controlled by amplifier V-320 and the cathodes of the regulator tubes supply the positive 240 v dc bus.

Operation of the circuit is best understood as follows:

Assume the positive 240 v dc bus tends to go positive. This positive voltage excursion is applied to the grid of a differential amplifier V-322 through the voltage divider composed of resistor R-410, R-411 and R-412. This causes the cathode of V-322 to go positive and since the opposite grid of V-322 is held at a constant voltage by reference tube V-321, the increase in cathode voltage increases the bias on the stage reducing plate current through R-406 and causing the voltage appearing at the grid of the control amplifier (V-320) to increase. This in turn increases plate current through V-320 causing grid voltage applied to the regulator tubes (V-318 and V-319) to be reduced thus increasing their plate resistance and reducing the voltage appearing on the 240 v dc bus.

Screen voltage for the control amplifier V-320 is taken from the unregulated output of the rectifier. Resistors R-393 and R-413 together with capacitor C-416 set the amplitude and phase of the ripple component of the voltage applied to the screen of V-320 so as to obtain maximum ripple reduction through taking advantage of the screen to plate transconductance of this stage. The total ripple voltage appearing on the 240v dc bus is less than 200 millivolts

and regulation is approximately ± 3 volts over the range of service conditions.

Regulator tube V-323 together with resistor R-409 are used to provide a 150 v dc plate and screen supply source for the stages using type 6AK5 tubes and a screen voltage supply source for the IF amplifier stages.

Resistor R-411 is used to adjust the output of the regulator to provide 240 v dc on the main plate supply bus.

8-3. SUMMARY OF CHARACTERISTICS

TYPE OF EQUIPMENT

VHF FM Receiver Type RF-2

FREQUENCY RANGE

142 to 174 Mc

TYPE OF FREQUENCY CONTROL

Crystal

DEVIATION CAPABILITY

± 75 Kc nominal

OVER-ALL SYSTEM AUDIO FREQUENCY RESPONSE

± 0.5 db between 300 and 50,000 cps

OVER-ALL SYSTEM HARMONIC DISTORTION

Less than 1½%

OVER-ALL SYSTEM INTERMODULATION

Less than 1%

TYPE OF RECEIVER

Superheterodyne

INTERMEDIATE FREQUENCY

18 Mc

IF SELECTIVITY

6 db BANDWIDTH ± 120 Kc Min.
60 db BANDWIDTH ± 500 Kc Max.
80 db BANDWIDTH ± 600 Kc Max.

CRYSTAL REQUIREMENTS

Type CR-52/U; 41.3333 to 52.0000 Mc

FREQUENCY STABILITY

$\pm .005\%$

OUTPUT

- 15 to +20 dbm into 600 or 150 ohm balanced load

INPUT

52 ohm unbalanced coaxial line

TYPE OF RECEPTION

F 3

SENSITIVITY

Less than 5 μ v input required for 25 db of quieting

POWER SUPPLY

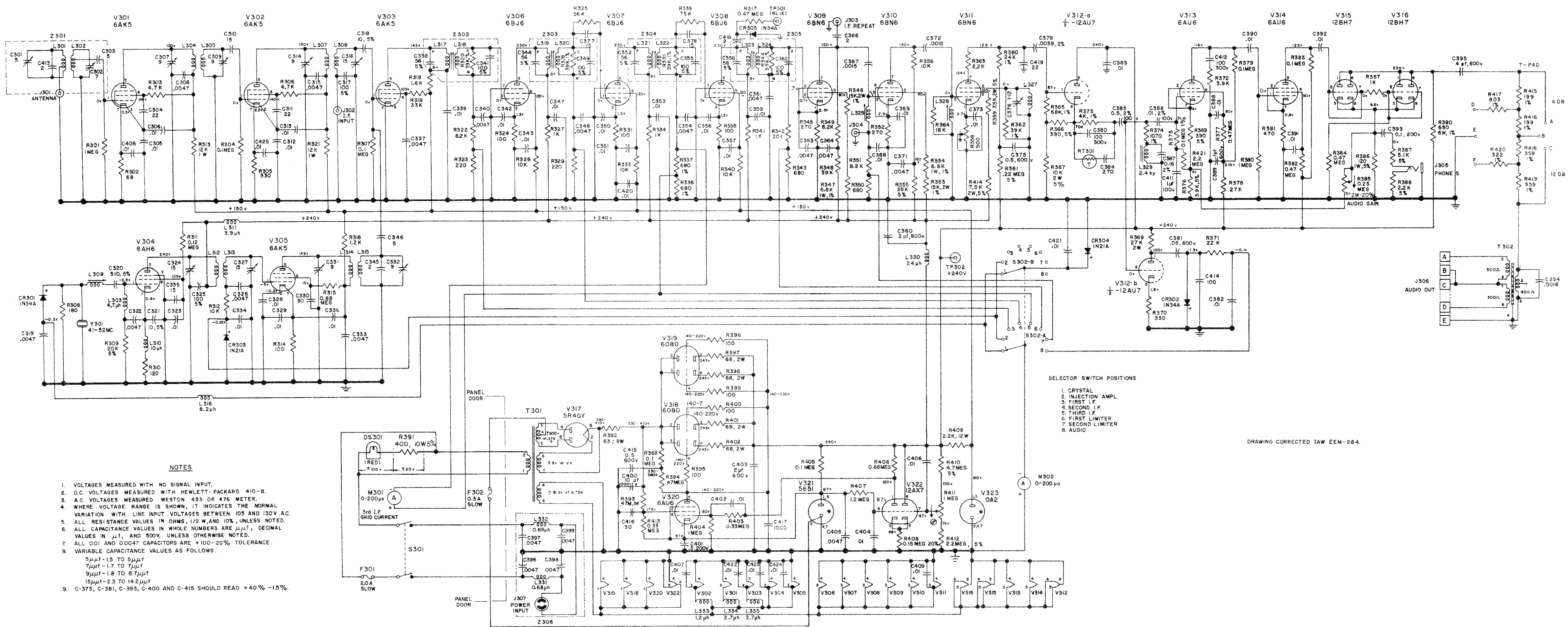
Self contained, operating from 105 to 130 v ac single phase 58.5 to 61.5 cps source; requires 1.81 amps at .905 PF for normal operation

WEIGHT

Crated 46 lb Uncrated 40 lb

ELECTRON TUBE AND FUSE COMPLEMENT

<i>Type</i>	<i>Quantity</i>
OA2	1
5R4GY	1
6AH6	1
6AK5	4
6AU6	3
6BJ6	3
6BN6	3
12AU7	1
12AX7	1
12BH7	2
5651	1
6080	2
MDL 2 amp	1
MDL 3/10 amp	1



CHAPTER 9

AN/URR-13 RECEIVER

9-1. INTRODUCTION

Radio receiving set AN/URR-13 is designed to provide means for reception of amplitude-modulated and MCW transmission in the 255-400 mc range. The AN/URR-13 is a VHF/UHF type of receiver, designed primarily for operation as a pretuned, single-channel, crystal-controlled unit. Provisions are made for variable manual tuning. A single tuning control is employed for either crystal controlled or manual operation.

The receiver has a sensitivity of better than 8 microvolts for a 10-db signal-to-noise ratio. The balanced push-pull circuit arrangement of the RF amplifier and oscillator-multiplier circuits provides for stable operation and freedom from spurious antenna radiation.

The front panel is fitted with handles to permit withdrawal of the panel and chassis assembly from the cabinet. A spring stop mechanism on each side permits nearly complete withdrawal, but prevents its falling out. When these stop mechanisms are pressed upward by the fingers through holes in the bottom of the chassis frame, the assembly may be completely withdrawn.

A built in power supply is provided which can be adjusted to operate from a 110-115-120 volt, 50-60 cps, single-phase source. The audio and power source connections to the receiver are filtered to limit radio frequency interference.

9-2. DESCRIPTION OF SUB-ASSEMBLIES

A. Preselector Section

The preselector section is mounted along the left side of the chassis and comprises all parts of the RF amplifier-converter and oscillator-multiplier sections. The RF amplifier-converter section is above the oscillator-multiplier section,

and each section consists of an aluminum casting with removable covers. The two RF amplifier stages and the mixer, or first detector, are mounted in the RF amplifier-converter section. The basic oscillator, two frequency-doubler stages and a frequency-tripler stage are mounted in the oscillator-multiplier section. Partitions in the casting provide RF shielding between stages.

B. Intermediate-Frequency/Audio-Frequency Section

The intermediate frequency and audio-frequency section is located on the right-hand side of the chassis and mounts the circuit parts of the five intermediate-frequency stages, the second detector, the automatic volume control circuit, the silencer and the silencer amplifier circuits, the noise limiter circuits and the three stages of audio amplification.

C. Cable Filter Assembly

The cable filter assembly consists of RF noise filter circuits for the audio output and power input circuits and provides connections from the receiver proper to the antenna input and scan channel output circuit connectors. The filter parts are mounted on a base plate, which is attached to the rear wall of the receiver cabinet by means of snap-slide fasteners located on the inside of the cabinet. The filter cover, when attached to the base plate, constitutes an RF shield. The ac power input, audio output, scan channel output and antenna transmission line input connectors, P-401-P-404, to which all external connections except headphones are made, are mounted on the underside of the filter assembly on an angle bracket attached to the base plate. When the receiver assembly is slid into the cabinet,

three connectors on the rear of the receiver engage mating connectors on the rear of the filter base plate, establishing connections between the external and internal receiver circuits.

9-3. ASSOCIATED EQUIPMENT

A. Scan Circuit

The scan connector, J-403, on the cable filter at the rear of the cabinet can be connected to a panoramic radio frequency scanning adapter to provide visual indication of the signals being picked up by the receiver. The scanning channel has a frequency response which is flat within 6-db over a bandwidth of 600 kilocycles. If scanning equipment is used, a 52-ohm coaxial transmission line cable should be employed. However, no such equipment is required for satisfactory operation as a receiver.

B. Antenna

The antenna to be used with this receiver must be designed to have an impedance of approximately 52 ohms with characteristics that result in good matching with the transmission line over the frequency range of 225-400 mcs. A coaxial transmission line having a characteristic impedance of about 52 ohms is required for connection between the antenna and the receiver.

C. Phones and Speaker

The audio output circuit at the audio receptacle, J-402, on the rear of the equipment is designed to operate into any load impedance between 60 and 600 ohms and to maintain its output voltage constant within 3 db into an impedance within this range. Audio output is also wired to a phone jack on the front panel, J-501. A 600-ohm headphone set can be connected into this jack.

9-4. THEORY OF OPERATION

Radio receiver AN/URR-13 is of the superheterodyne type and is designed for either manual tuning or crystal-controlled operation

over a frequency range of 225-400 megacycles. The receiver will provide audio output to headphones, to an external speaker or intercommunication system. Output provisions are also included for an external panoramic adapter. The receiver is basically conventional in most respects. Two stages of RF amplification precede the mixer stage. The local injected signal is obtained from an oscillator which is followed by three stages of frequency multiplication. The oscillator functions as either a crystal-controlled or self-excited circuit, depending on the position of the CRYSTAL-MANUAL switch. The five-gang capacitor which tunes the RF and mixer stages is geared to the four-gang capacitor in the oscillator-multiplier section to provide single control tuning.

The received signal is converted to an intermediate frequency of 18.6 megacycles and is amplified by five stages of intermediate-frequency amplification. The signal is then rectified by the second detector and by the avc rectifier. The developed avc voltage is applied to the grids of both RF amplifiers, the first four intermediate-frequency amplifiers and the first AF amplifier. The audio output of the second detector passes through a series noise limiter, which may be bypassed, into the first AF voltage amplifier. The output of this amplifier passes through an AF bandpass filter, Z-201, which has a bandpass of 350-3500 cps, and through a silencer diode into the second AF amplifier. The amplified audio signal is then applied to the AF output stage, the output of which drives the headset or external speaker. Output is also monitored by the output meter, M-502.

The silencer diode, between the bandpass filter and second audio amplifier, is controlled by the silencer amplifier, which permits the diode to conduct when a signal is present at the second detector. If no signal is present, the diode cuts off, preventing any noise from reaching the second audio stage. The noise silencer may be cut in or out by action of the silencer switch.

9-5. DETAILED CIRCUIT ANALYSIS

A. First RF Amplifier

As shown in Figure 9-2 and the schematic diagram, Figure 9-11, the antenna circuit is coupled to the grid input circuit of this stage. The grid circuit consists of tank inductance L-102, trimmer inductance L-103, a section of capacitor C-101, and trimmer capacitor C-140. Critical coupling is employed at the high end of the band to maintain uniform sensitivity over the entire frequency range. Two type 9003 tubes, V-101 and V-102, are operated in push-pull. These tubes are remote cut-off pentodes to permit employment of avc. The push-pull arrangement serves to reduce the length of connecting leads to a minimum and to cancel out spurious radiation of unwanted

signals into the antenna. The avc voltage is applied to the grid circuit of this stage through decoupling resistor R-101 and the center tap on coil L-102. A feed-through type RF bypass capacitor, C-105, functions to bypass the avc line at this point. This capacitor is located in the casting wall. The screen voltage is applied through decoupling resistor R-104 and through screen voltage dropping resistors R-102 and R-103. The 180 volt supply is bypassed to ground at this point by capacitor C-112, while resistors R-102 and R-103 are bypassed by the capacitors C-104 and C-103, respectively. The plate circuit is similar in design to the grid circuit and consists of basic inductance L-105, a section of C-101, trimmer inductance L-104 and trimmer capacitor C-141. Plate voltage is applied through resistor R-104 to the center tap on coil L-105. No appreci-

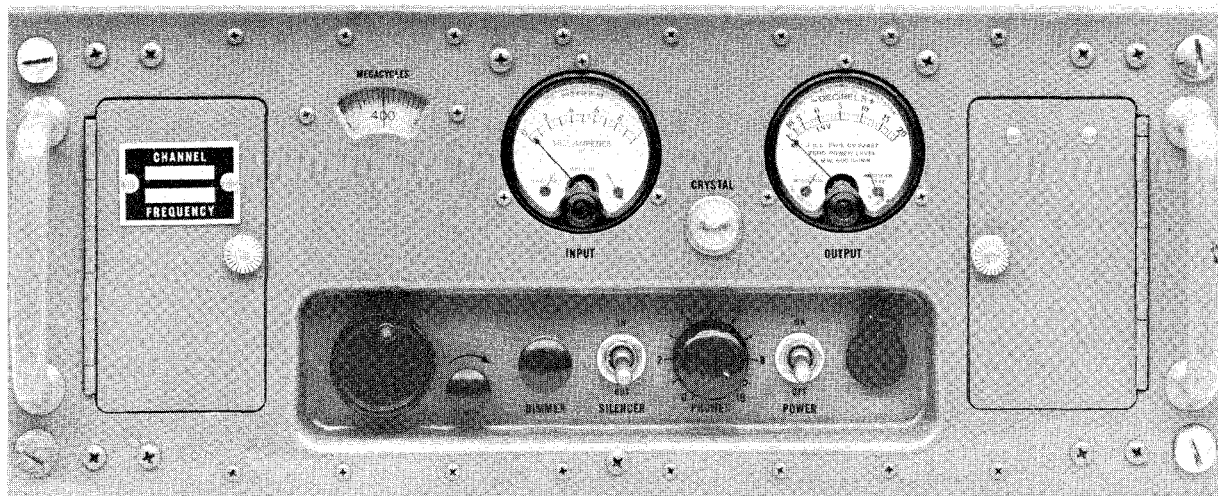


Figure 9-1. AN/URR-13 Receiver

able inductive coupling exists between the plate circuit of the first RF stage and the grid circuit of the second RF stage because coils L-105 and L-106 are shielded from each other by the wall of the casting.

Coupling between these two stages is accomplished by means of capacitors C-108 and C-109, which are connected directly from the plates of the tubes in the first stage to the grids of the tubes in the second stage. With tubes having a nominal value of transconductance, the gain of the first RF amplifier is approximately 7 db.

B. Second RF Amplifier

This stage, employing tubes V-103 and V-104 in push-pull, is identical in design and in circuit constants to that of the first RF stage.

C. Mixer

The mixer, or first detector, stage employs a type 6J6 dual triode, V-105, in a push-push arrangement. The grid circuits are connected in push-pull through the resonant circuit consisting of a section of capacitor C-101, tank inductor L-110, trimmer capacitor C-144 and trimmer inductor L-111. The plates of the tubes are connected in parallel to plate transformer L-122. Capacitor C-119, connected between the plate of the mixer tube and ground, serves to tune L-122 to resonance at the intermediate frequency. This capacitor is connected directly to ground to provide a low impedance path for other than intermediate-frequency RF currents. With the push-push arrangement of the mixer a high order of conversion gain is obtained because the push-push conversion transconductance is approximately twice that of a single converter tube.

The output of the second RF stage is coupled to the grid circuit of the mixer stage by capacitors C-111 and C-110, while the output of the oscillator-multiplier circuit is inductively coupled to the grid circuit of the mixer through a coupling loop consisting of coil L-119 and capacitor C-135 in the oscillator-multiplier unit and coil L-120 and capacitor C-118 in the RF amplifier unit. This coupling

arrangement maintains a more uniform mixer injection voltage over the frequency range of operation. The values of capacitors C-118 and C-135 are so chosen that, together with the shunt capacity of the short coaxial connecting line and the coupling coils, the loop is series resonant at 190 mc, a frequency below the low end of the band, and parallel resonant at 470 mc, a frequency above the high end of the band. Consequently, at the low-frequency end the transfer voltage induced in the converter input tank is increased while at the high frequency end of the band the induced transfer voltage is reduced. If the coupling were not resonant, as described above, the induced voltage would tend to drop off at the low end and increase at the high end of the band due to the change in "Q" of the oscillator-multiplier tank circuits.

The output of the mixer stage is link coupled to the grid circuit of the first intermediate-frequency stage. This type of coupling is made necessary by the physical layout of the equipment. The plate coil of the mixer, L-122 is contained in the mixer compartment of the RF amplifier casting, while the intermediate-frequency input transformer, T-201, is located on the intermediate and audio frequency chassis. A length of dual conductor RF cable joins the secondary winding of transformer L-122 on the RF amplifier chassis and the primary winding of transformer T-201 on the intermediate and audio frequency chassis, thus establishing low impedance inductive coupling between the two circuits.

Plate voltage is supplied to the primary of L-122 through resistor R-110. Capacitors C-115 and C-117 bypass the B supply line and R-110 respectively. The grids of V-105 are biased by the voltage developed across grid leak resistor R-109, and capacitor C-116.

D. Oscillator-Multiplier Section

The oscillator-multiplier section generates a local injection signal which is 18.6 mc higher in frequency than the received signal. The basic oscillator frequency is generated in tube V-106 and multiplied twelve times in the two doublers and one tripler which follow, as

shown in the schematic diagram. Also see Figure 9-3. Tuning of the various stages is accomplished by capacitor C-102, which is a four-section capacitor, each section being of the split-stator type. The use of the split-stator capacitors in the balanced tank circuits permits the use of a grounded rotor to reduce intersectional capacitance. Rotor grounding is accomplished by wiping contacts located between the capacitor sections. Since no appreciable RF currents flow through these contacts, the inherent noise associated with wiping contacts is not present. Capacitor C-102 is geared to the five-section capacitor C-101 to provide single control tuning of the receiver. The trimmer inductances and capacitors utilized in the oscillator-multiplier stages are integral parts of the tuning capacitor, similar to those previously described.

E. Oscillator-First Doubler

The oscillator and first doubler stages utilize a type GL-5670 dual triode. One half of the tube V-106A functions as a grounded-grid oscillator. The second half V-106B is arranged as a split-load cathode follower and serves as a frequency doubler and a source of feedback to the oscillator cathode. The oscillator functions as a crystal-controlled circuit when switch S-203 is in the crystal position. The crystal, Y-201, is a harmonic-mode type, which establishes the feedback from the cathode of V-106B to the cathode of V-106A. Capacitor C-159, in series with the crystal, is utilized to resonate the inductance of the crystal leads so that zero phase shift exists between the two cathodes.

In MANUAL position, for manual tuning, the crystal is shorted out by switch S-203. V-106A then functions as a free-running oscillator, the frequency of which is determined by the setting of tuning capacitor C-102A. Since the feedback path between the cathodes of the two triode sections is not frequency selective, the stability of the free-running oscillator is not as great as the crystal-controlled circuit.

On manual tuning, the receiver may be operated with or without a crystal in the crystal socket. However, since the original

factory adjustment of the receiver was made with a crystal in the socket, it follows that dial calibration will be more accurate, and the reserve gain greater, if the receiver is operated in the same manner during manual operation in the field. In the neighborhood of 400 mc the resonant frequency of the receiver increases approximately 0.1%; near 225 mc the corresponding increase is 0.04%. This effect is attributable to the fact that there is some capacity between the crystal and ground, which also exists effectively between the cathodes of the oscillator tubes and ground. Removing the crystal from its socket removes this capacity and so causes a shift in the frequency of the oscillator.

The oscillator tank circuit consists of variable capacitor C-102A, trimmer capacitor C-148 and coil L-114. The tank circuit is kept balanced by the use of capacitor C-121, which equalizes the output capacitance of the tube. Resistor R-112 tends to suppress spurious oscillation. Inductance L-123 counteracts the heater to cathode capacitive reactance so that the phase shift of the signal applied from the cathode of the first doubler tube to the oscillator cathode is as small as possible. Capacitor C-160 keeps L-123 from shorting the bias developed across resistor R-111 to ground. Plate voltage is fed to the center tap of coil L-114 through resistor R-113 from a 150 volt regulated source; the B+ supply line is bypassed for RF by capacitor C-120.

The output of the oscillator tube is coupled to the grid of the first doubler through capacitor C-137. Grid bias for V-106B is provided by the voltage drop across cathode resistor R-116 and by the drop across grid resistor R-114. Plate voltage is applied through the untuned primary of transformer L-115; voltage is obtained from a 150 volt regulated source through decoupling resistor R-125, which is bypassed for RF by capacitor C-123.

The first doubler cathode circuit is not bypassed, so the RF voltage drop across resistor R-116 may be fed back to cathode resistor R-111. The values of R-111 and R-116 are such that the feedback is limited; therefore, oscillations which might occur due to the capacity across the crystal holder are prevented.

A test point is provided at the grid of the first doubler to measure the dc bias on the tube. The measured voltage is indicative of the amount of drive from the oscillator. Resistor R-115 and capacitor C-122 decouple the grid circuit from the point of measurement.

F. Second Doubler

The second doubler stage employs the two sections of the type 6J6 dual triode, V-107, in a push-push circuit arrangement. The grid circuit is balanced and is tuned to the second harmonic of the oscillator frequency. The tuned circuit consists of the center-tapped secondary winding of transformer L-115, a section of the ganged capacitor C-102, and trimmer capacitor C-147. Capacitor C-126 compensates for the capacitance unbalance to ground in the secondary winding of transformer L-115.

The grids are returned through the center tap on the secondary of L-115 and through resistor R-117 and R-118 to a -3 volt terminal on the power supply. This bias holding voltage functions to prevent excessive rise in plate current if driving voltage is removed. Capacitors C-124 and C-125 function to bypass RF currents around this circuit.

The plates are tied in parallel; untuned coil L-116 constitutes the plate load. This coil is center-tapped to provide a balanced load for coupling to the grids of the tripler stage. Capacitor C-128 at the ground side of coil L-116 functions to balance the output capacity of the second doubler tubes at the other side of the coil. Plate voltage is applied through decoupling resistor R-119 to the center tap of coil L-116. C-127 functions as an RF bypass capacitor. The output of the second doubler stage is coupled to the grid circuit of the tripler stage, V-108, by means of capacitors C-129 and C-130. No appreciable inductive coupling exists between L-116 and L-117 since the two coils are isolated by the preselector casting.

The push-push circuit arrangement reduces the generation of odd harmonics. This eliminates the possibility of resultant spurious signals which might otherwise be amplified and radiated by the receiver antenna.

G. Tripler

The tripler stage employs two type 6AK5 pentodes, V-108 and V-109, in push-pull. The parallel-resonant grid circuit consists of coil L-117, a section of capacitor C-102, and trimmer C-145, and the circuit is tuned to the fourth harmonic of the oscillator frequency. The plate circuit is tuned to resonance at a frequency three times its grid input frequency, which results in a plate circuit output frequency 18.6 mc higher than the receiver incoming signal frequency. The tripler plate circuit is a parallel-resonant combination of coils L-112 and L-113, and capacitors C-102-D and C-146. Plate and screen voltages are applied through resistor R-124 to the center tap of L-113 and the junction of screen grid dropping resistors R-122 and R-123 respectively.

The output of the tripler stage is inductively coupled through coil L-119 and capacitor C-135 on the oscillator-multiplier chassis and through capacitor C-118 and coil L-120 on the RF amplifier chassis to the grid circuit of the mixer stage.

9-6. INTERMEDIATE FREQUENCY STAGES

The mixer output is link-coupled to the receiver intermediate-frequency section, which consists of five stages of amplification tuned to 18.6 mc.

A. First Intermediate Frequency Amplifier

A type 6BA6 remote cut-off pentode, V-201, is arranged as a split-load amplifier; the tube functions as both the first intermediate frequency amplifier and as a cathode follower to provide output for a scanning channel indicator, such as a panoramic adapter. The cathode resistor R-202, is not bypassed so that the RF voltage developed across it may be utilized as the scanning channel output.

The signal from the mixer is applied to the grid of V-201 through transformer T-201. The primary is an untuned, low-impedance winding; the secondary is tuned and is resistance loaded to provide a scanning bandwidth of 600

kc which is flat to within 6 db. An avc voltage is applied to the grid circuit of V-201 through decoupling resistor R-201, which is bypassed by capacitor C-201A. Screen voltage is obtained from the 105-volt source through decoupling resistor R-203, while the 180-volt plate supply is brought through decoupling resistor R-204. Resistors R-203 and R-204 are each bypassed by a separate section of capacitor C-202. The output of V-201 is coupled to the second intermediate-frequency amplifier by means of transformer T-202.

B. Second, Third, Fourth and Fifth Intermediate Frequency Amplifiers

Each amplifier utilizes a type 9003 remote cutoff pentode, with double-tuned transformers used as their interstage coupling device. All transformers used in the intermediate-frequency section are tuned by means of iron

cores (powdered). Screen voltage for all tubes is obtained from the regulated 105-volt tap on the power supply, and plate voltage is obtained from the 180-volt tap. Both screen and plate voltages are applied through suitably bypassed decoupling resistors. The grid circuits of the second, third and fourth amplifiers, V-202, V-203 and V-204, are returned to the avc line through a resistance-capacitance decoupling network. The avc is not applied to the fifth amplifier V-205; instead, terminal No. 6 of transformer T-205 is grounded. Grid bias for V-205 is developed across cathode resistor R-217, which is bypassed by capacitor C-213A.

C. Input Meter Circuit

Input meter M-501 is provided to indicate the approximate incoming signal strength. It also serves as an alignment indicator for the

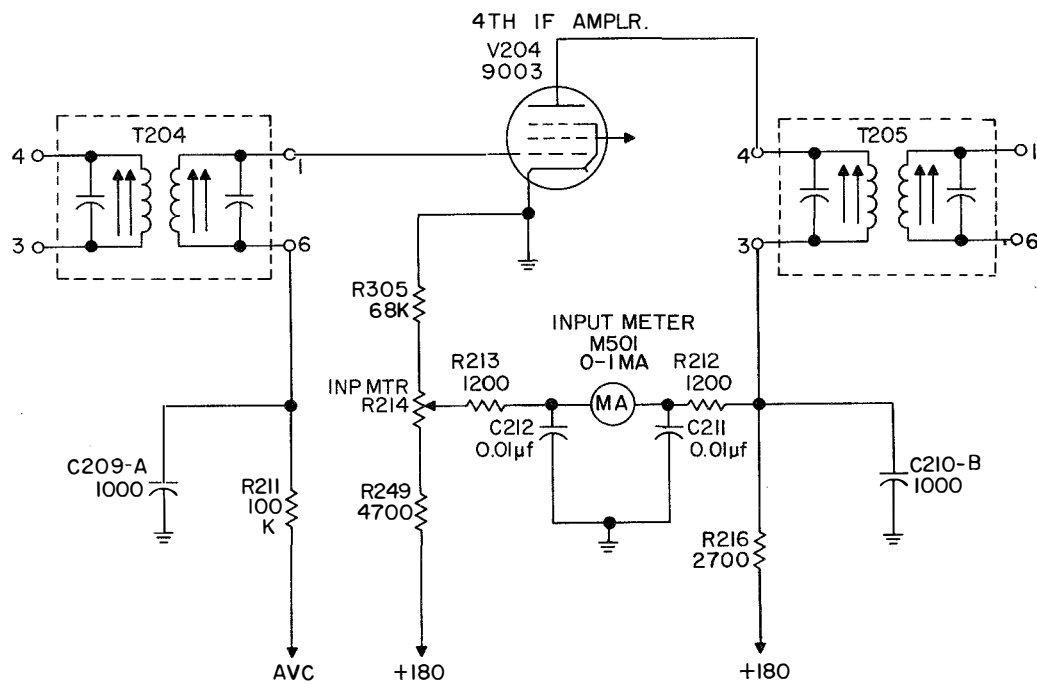


Figure 9-4. Simplified Schematic—Input Meter (M501) Circuit

oscillator-multiplier section when switch S-201 is placed in the ALIGN position. See Figure 9-4. One side of meter M-501 is connected to terminal No. 3 of transformer T-205 through a low-pass filter consisting of R-212 and C-211. The other side of the meter is connected through a similar filter, R-213 and C-212, to the arm of input meter control R-214. Potentiometer R-214, along with resistors R-249 and R-305, is in a voltage-divider network connected between the 180-volt supply and ground.

With no signal present in the receiver, pentode V-204 conducts heavily, causing a voltage drop across resistor R-216. The input meter control, R-214, is then adjusted so that the center arm potential is equal to the potential at terminal No. 3 of transformer T-205. This results in equal potentials at each side of meter M-501 and, consequently, no meter indication. When a signal is received,

the developed avc voltage increases the bias on V-204. The reduction in plate current decreases the voltage drop across resistor R-216. Since a voltage difference then exists across the meter terminals, an indication is obtained on the meter.

The input meter functions in the same manner when used for alignment purposes. However, the avc voltage is replaced by the tripler grid-leak bias.

D. Second Detector

One-half of a type 6AL5 dual diode V-206A, is used in a conventional diode circuit. Intermediate-frequency transformer T-206 couples the signal from the fifth intermediate-frequency stage, V-205, to the detector. See Figure 9-5. Resistors R-224, R-225 and R-226 constitute the diode load, which is bypassed for radio frequency by capacitors C-216

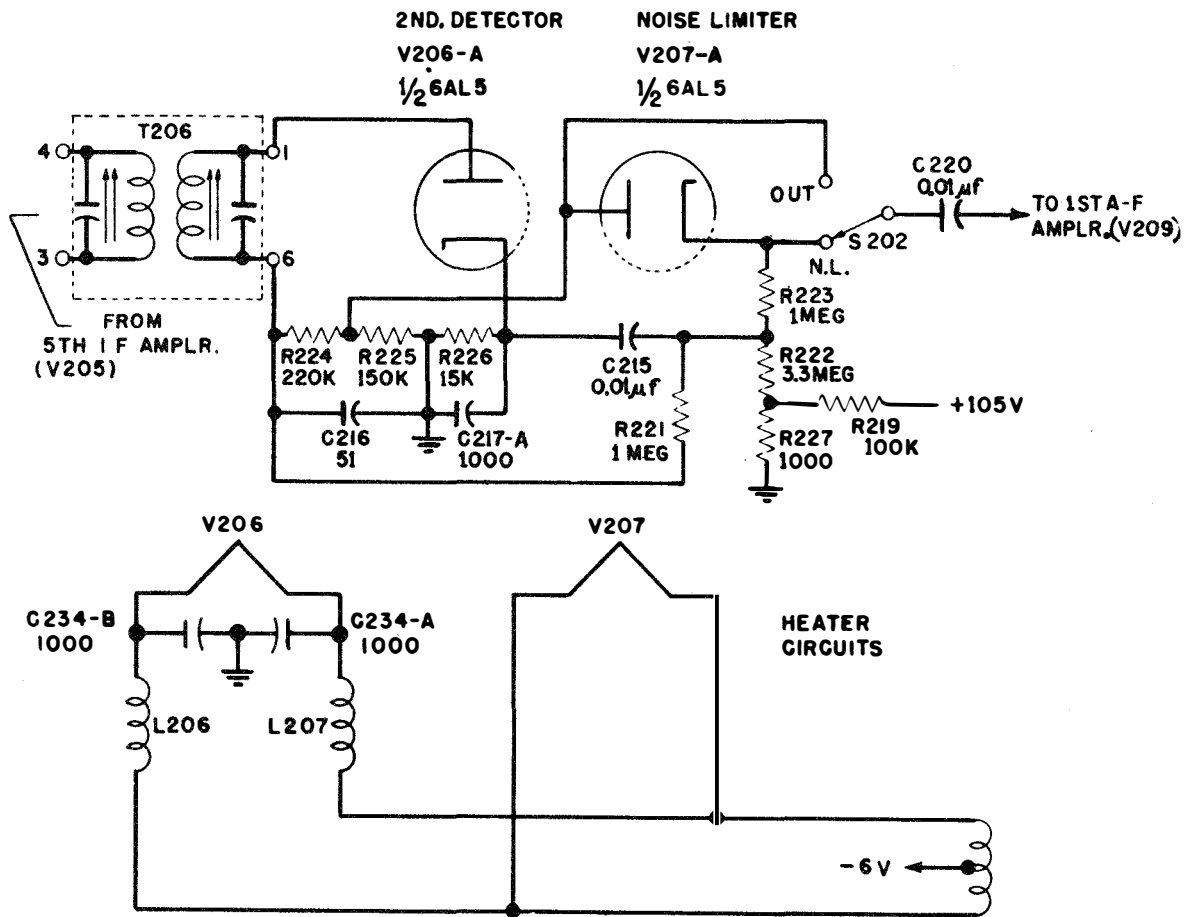


Figure 9-5. Simplified Schematic—Second Detector and Noise Limiter

and C-217A. The audio frequency output, obtained at the junction of resistors R-224 and R-225, is coupled through capacitor C-220 to the first audio-frequency amplifier grid when noise-limiter switch S-202 is in the OUT position.

9-7. SPECIAL NOISE LIMITER, AVC AND SILENCER CIRCUITS

A. Noise Limiter

When noise-limiter switch S-202 is in the IN position, a series noise limiter is placed in the circuit between the second detector and first audio stage. The limiter, V-207A, is one-half of a 6AL5 dual diode which functions as follows:

The negative voltage, developed across second detector resistors R-224 and R-225, is applied through resistor R-221 to capacitor C-215, building up on this capacitor a negative potential equal to the total average rectified dc voltage as measured between terminal No. 6 of T-206 and ground. The AF component of the rectified voltage is taken from the detector diode circuit at the junction of resistors R-224 and R-225. The AF path is then from plate to cathode of V-207A, across switch S-202 and through capacitor C-220 to the grid of the first audio amplifier tube, V-209. It will be noticed that the cathode of V-207A is at the potential of terminal No. 6 of T-206, which is more negative than the plate potential of V-207A because of the voltage divider action of R-224 and R-225. Since the diode cathode is at negative potential with respect to the plate, electrons flow from cathode to plate and an AF path is established through the tube.

In the event that a sharp pulse of noise is received, the long time constant of R-221 and C-215 does not permit capacitor C-215 to become charged to the high transient voltage. However, terminal No. 6 of T-206 rapidly follows the change, placing the plate of V-207A at a more negative potential than the cathode, thereby, cutting off electron flow in the tube for the duration of the noise pulse. Consequently, the noise pulse does not enter the audio-frequency amplifier circuit. Resistor R-226 in the cathode of the second detector

acts as an accelerating circuit to bring the noise limiter diode V-207A to the condition of non-conduction more quickly when a noise pulse enters the receiver. A positive pulse from the cathode end of R-226 is coupled to the limiter diode cathode through capacitor C-215 and resistor R-223. Thus, an additional positive voltage is initially present at the cathode, which aids in cutting off tube V-207A. This action is similar to the operation of the noise-cancellation detector in the RUQ receiver.

Resistors R-219 and R-227 form a voltage divider across the 105 volt regulated supply producing a positive voltage of approximately one volt. This voltage is applied through resistor R-222 and R-223 to the cathode of V-207A. The positive bias is required to balance out the contact potential of this tube, permitting operation of the diode as a noise limiter on lower levels of noise. By normally operating the cathode of V-207A considerably more negative than the plate, clipping does not occur on modulation peaks below a certain level.

B. AVC Rectifier

One-half of a 6AL5 dual diode, V-206B, is utilized in the avc circuit, which is shown in simplified form in Figure 9-6. The diode is connected as a shunt rectifier across the secondary of transformer T-206; coupling is accomplished through capacitor C-218 and capacitors C-216 and C-217B in series. With switch S-201 in the RECEIVE position, the diode load consists of resistors R-229 and R-230. The junction of these two resistors is at a potential of three volts negative, which places the diode plate at this same voltage. The three-volt negative potential is developed across resistors R-230 and R-248 by the return plate current which flows through them. The cathode is connected to the junction of resistors R-230 and R-248, placing it at a potential of approximately two volts negative. Thus, avc delay is provided by the resultant one-volt bias. The avc voltage is applied to the RF amplifiers, the first four intermediate-frequency amplifiers and the first RF amplifier through resistor R-228. Suitable RC decoupling networks are incorporated in the grid

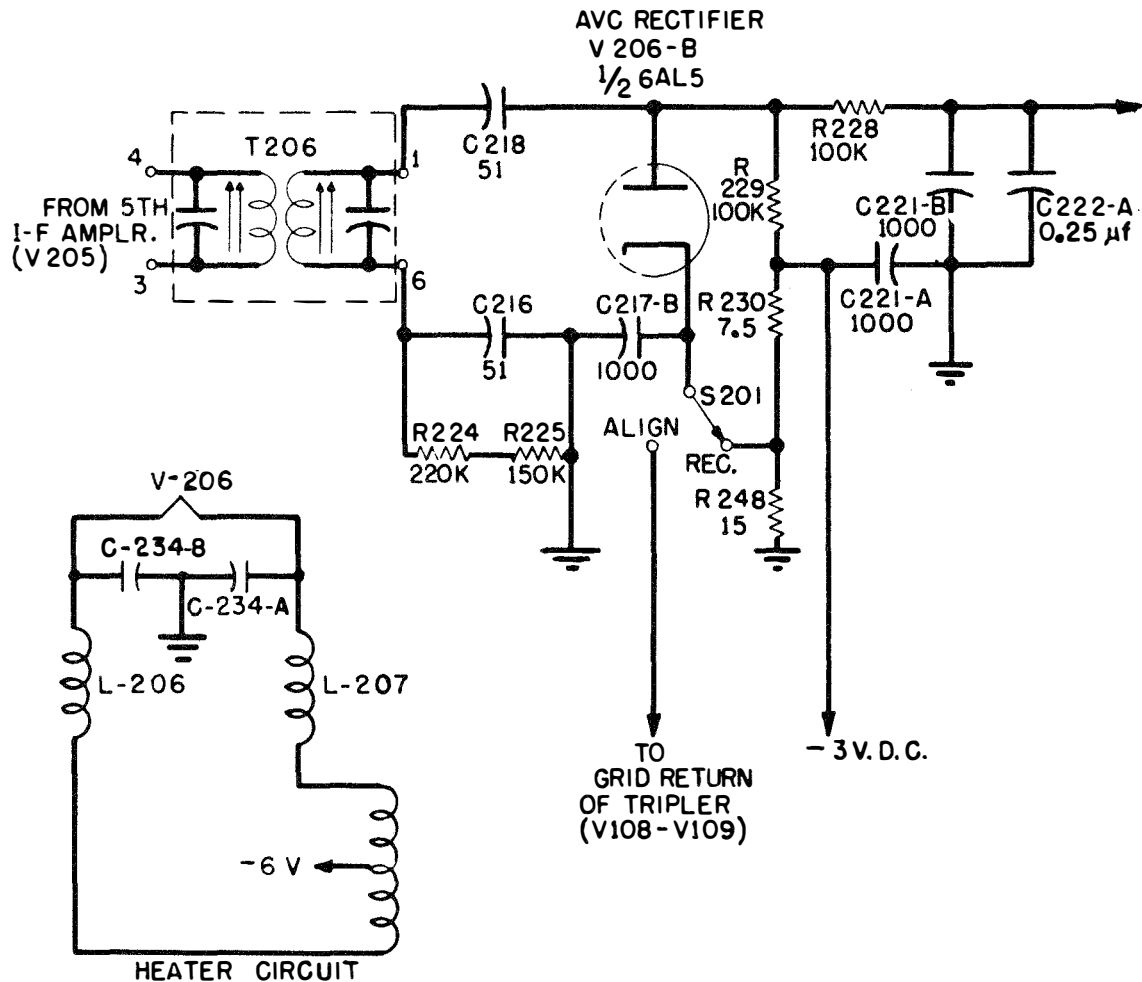


Figure 9-6. Simplified Schematic—AVC Rectifier Circuit

circuit of each of the controlled stages. The long time constant of R-228 and C-222A prevents the avc voltage from following rapid variations in carrier level. Capacitor C-221 is a dual RF bypass unit.

The avc circuit also provides means for indicating the alignment of the oscillator-multiplier section of the receiver. When switch S-201 is placed in the ALIGN position, the cathode of V-206B is removed from the voltage divider across the three-volt bias source and is connected to the grid return of the tripler stage, V-108 and V-109. Under this condition, the diode load consists of resistors R-229, R-230 and R-248 in the diode circuit and resistor R-121 in the tripler circuit. Both cathode and plate are at negative potential of three volts. As the tuned circuits of the oscillator, first and second doublers, and tripler

grid are tuned to resonance, the drive to the tripler circuit will increase. The portion of the resultant tripler grid-leak bias developed across resistor R-121 is applied to the avc diode cathode. Diode current will then flow causing a negative voltage to appear on the avc bus. This voltage in turn, will cause a reading on input meter M-501 as explained previously. The greater the meter deflection, the greater the drive and the grid-leak bias at the tripler grid; hence, an indication of alignment is obtained.

C. Silencer

The noise silencer (squelch) circuit is used to prevent noise from reaching the audio section of the receiver in the absence of an incoming signal of some predetermined mini-

imum level. A controlled diode V-207B between the first and second audio stages permits the audio signal to pass during conduction, and cuts off the audio signal when the diode is not conducting. The diode V-207B is one-half of a type 6AL5 dual diode, and tube V-208, the silencer amplifier, is a dc amplifier which controls the diode. See Figure 9-7.

The audio signal is fed to the cathode of V-207B from filter Z-201 through audio gain control R-238 and capacitor C-225. When the diode conducts, the audio signal reaches the second AF amplifier through capacitor C-226. The silencer circuit functions when the silencer switch S-501 is in the IN position. During reception of a signal, a negative voltage is developed at terminal No. 6 of the intermediate-frequency transformer T-206. This voltage is applied through resistor R-231 to the grid of the type 6AK5 silencer amplifier tube, V-208, as negative bias.

An additional bias voltage is applied to the grid of tube V-208 through resistor R-251 for the purpose of establishing the threshold of operation of the silencer tube. This voltage, which is positive in potential, is obtained from the regulated 105-volt supply through the voltage divider action of resistors R-233 and R-253.

It can be adjusted from the front panel by means of silencer potentiometer R-233 located in the right-hand compartment.

When no signal is being received, the negative bias developed at the second detector is quite low, and without silencing, some noise would be present in the audio output of the receiver. The silencer control potentiometer acts to silence this no-signal noise output by increasing the positive bias on the grid of the tube V-208. This positive bias causes tube V-208 to draw more plate current and increases the voltage drop across resistor R-235, thereby, making the voltage applied through resistor R-236 to the plate of the silencer diode, V-207B, lower than the diode cathode voltage. This stops current flow in the diode and prevents the conduction of the audio signal through it.

When an input signal appears, the negative bias previously mentioned will increase, thereby reducing conduction in the silencer amplifier, V-208, and raising the plate voltage. This makes the plate of the control diode, V-207B, sufficiently positive for conduction, and the output from the first audio stage is then allowed to pass to the second audio

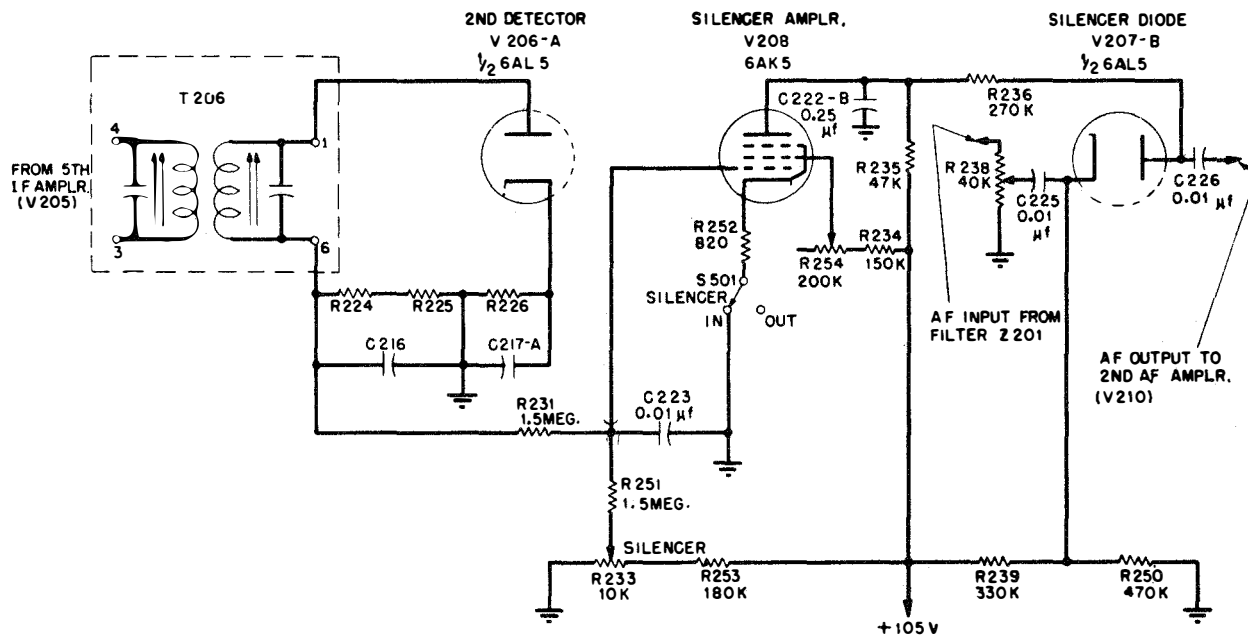


Figure 9-7. Simplified Schematic—Noise Silencer Circuit

amplifier via the diode. In this manner noise is prevented from reaching the second audio amplifier when no useful signal is being received.

The level at which the silencer tube, V-208, responds, can be adjusted with potentiometer R-254, the setting of which determines the screen voltage applied to that tube. This adjustment is used to compensate for possible changes in circuit constants occurring when V-208 is replaced with a new tube. The silencer switch, S-501, when thrown to its OUT position, opens the cathode circuit of V-208, thereby, stopping conduction in the silencer amplifier tube and rendering the silencer circuit inoperative.

9-8. AUDIO FREQUENCY CIRCUITS

The three audio amplifier stages are generally conventional. See Figure 9-8. The first stage is a triode-connected type 6AK5 pentode, V-209, and incorporates avc. Its output is fed through the 350-3500 cycle bandpass filter, Z-201, which discriminates against undesired frequency components but transmits the required voice frequencies. The signal passes through the AF level control, R-238, and the control diode (silencer), to the grid of the second audio stage, V-210. This stage is a conventional resistance-coupled pentode type 6AK5 voltage amplifier with an unbypassed cathode bias resistor. Its output drives the final type 6AK6 pentode power stage, V-211, which also operates with an unbypassed cathode resistor. Feedback is used from the plate of the output stage, V-211, to the cathode of the preceding stage, V-210, in order to maintain a constant output voltage characteristic with a variation of output load impedance, such as would result from the plugging in or withdrawing of headphones.

Audio signal from the amplifier is fed through the electrostatically-shielded output transformer, T-207, to the required output circuits. The impedance step-down of the transformer is 10,000 to 60 and the secondary provides balanced output. Signal from the secondary is fed via the RF filter to the output meter, M-502. Similarly, it is fed via the phones gain control, R-502, and associated RF filter to the headphones jack, J-501. A third

circuit feeds audio output via the connector P-301/J-405 for external use through the output jack, J-402, located at the rear of the cabinet. This line is also filtered against external RF fields by a combination of chokes and capacitors. This filter is a part of Band Suppression Filter F-89/URR-13. See Figure 9-10.

For headphone use, the front-panel phones volume control provides an audio level adjustment auxiliary to the main audio level control, R-238.

9-9. POWER SUPPLY

A single power transformer, T-301, supplies heater power for all tubes and the rectified dc voltage for plates, screens and negative bias. See Figure 9-9.

Filament power is obtained from three separate secondaries of the power transformer, namely: 5 volts for the type 5U4 rectifier, V-301; 6.3 volts for the detector, avc, noise-limiter and noise-silencer diodes, V-206 and V-207; and 6.3 volts from the third winding for all other tubes in the receiver. The winding for the diode heaters is center-tapped to -6 volts to minimize hum.

The fourth secondary winding provides power for full-wave rectification for the +180-volt plate supply, the +150-volt screen supply, and the separately filtered and regulated +150-volt supply for the local oscillator and first doubler. Capacitor C-302 and the portion of inductor L-301 between terminals 1 and 2 constitute a series resonant path to ground at the power supply ripple frequency, while capacitors C-301 and C-303 and the remainder of inductor L-301 form a conventional capacitor-input power supply filter for the plate power supply which is used for other than the oscillator and first doubler. The screen power supply is regulated at +105 volts by means of V-302 and resistors R-302, R-303, R-309, R-230, R-301 and R-248. The oscillator-doubler B+ supply is regulated at +150 volts by V-303 and resistors R-307, R-308 and R-310.

The dc bias voltages of -6 volts for minimizing hum in tubes V-206 and V-207, and of -3 volts for holding bias, are developed by resistors R-248, R-230 and R-301 which are connected between the negative side of the rectifier output and ground.

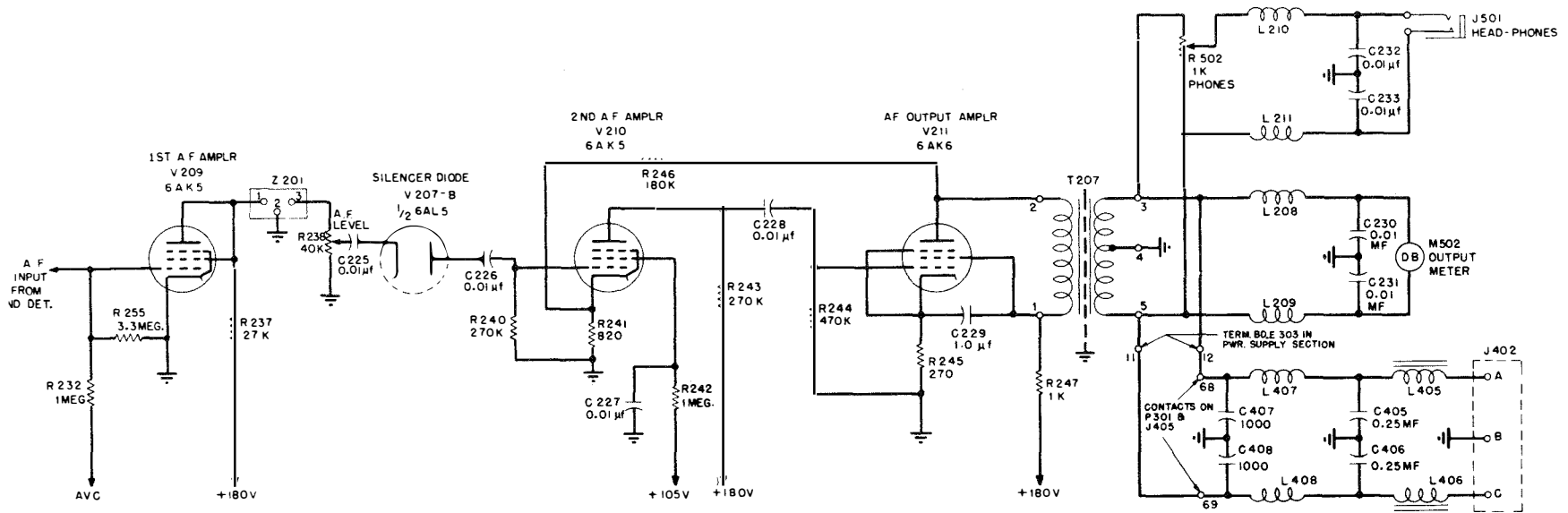


Figure 9-8. Simplified Schematic—Audio Frequency Circuits

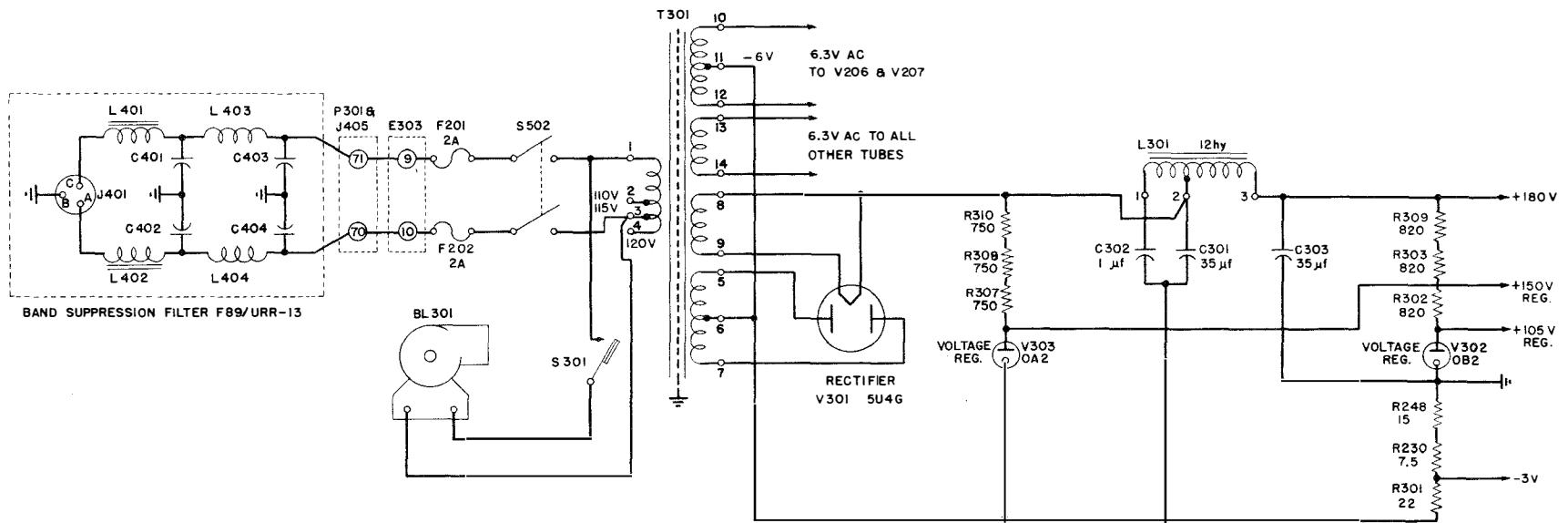


Figure 9-9. Simplified Schematic—Power Supply Section

The power transformer primary is tapped for operation on 110 v, 115 v, or 120 v at 50/60 cycles per second. Power application is controlled by power switch S-502 on the receiver front panel. The switch controls both sides of the power line. Both sides of the power line input are fused by 2 amp. fuses F-201 and F-202. Power input is through connector P-301/J-405 through the Band Suppression Filter F-89/URR-13 and input receptacle J-401. L-401, L-402, L-403 and L-404 in conjunction with capacitors C-401, C-402, C-403 and C-404 comprise the Band Suppression Filter designed to keep out external RF fields. See Figure 9-10.

9-10. BLOWER AND THERMOSTAT

In the original design of the receiver, the thermostat S-301 turned on the blower BL-301 whenever the ambient temperature reached 60° C. Because of the high blower failure incidence, and because external ventilation facilities have been improved, the receivers have generally been modified by disconnecting the blowers.

9-11. SUMMARY

Frequency range: 225-400 mc, crystal or manual operation.

Receiver type: Superheterodyne.

Intermediate frequency: 18.6 mc plus or minus 2 kc.

Power output: 60 mw into a 600-ohm load, or 600 mw into 60-ohm load, with 7% distortion. Phone jack; 60 mw into a 600-ohm load.

Scanning channel output: 10 microvolts across a 50-ohm load for a signal input voltage of 25 microvolts.

Silencing range: 0-15,000 microvolts input.

Antenna input impedance: Nominal 51 ohms.

Scan channel impedance: 51-ohm output.

Audio output impedance: 60 ohms, phone jack 600 ohms.

Sensitivity: Better than eight microvolts for a ten decibel signal-to-noise ratio.

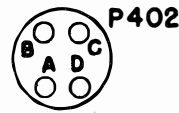
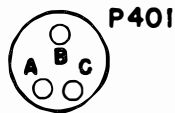
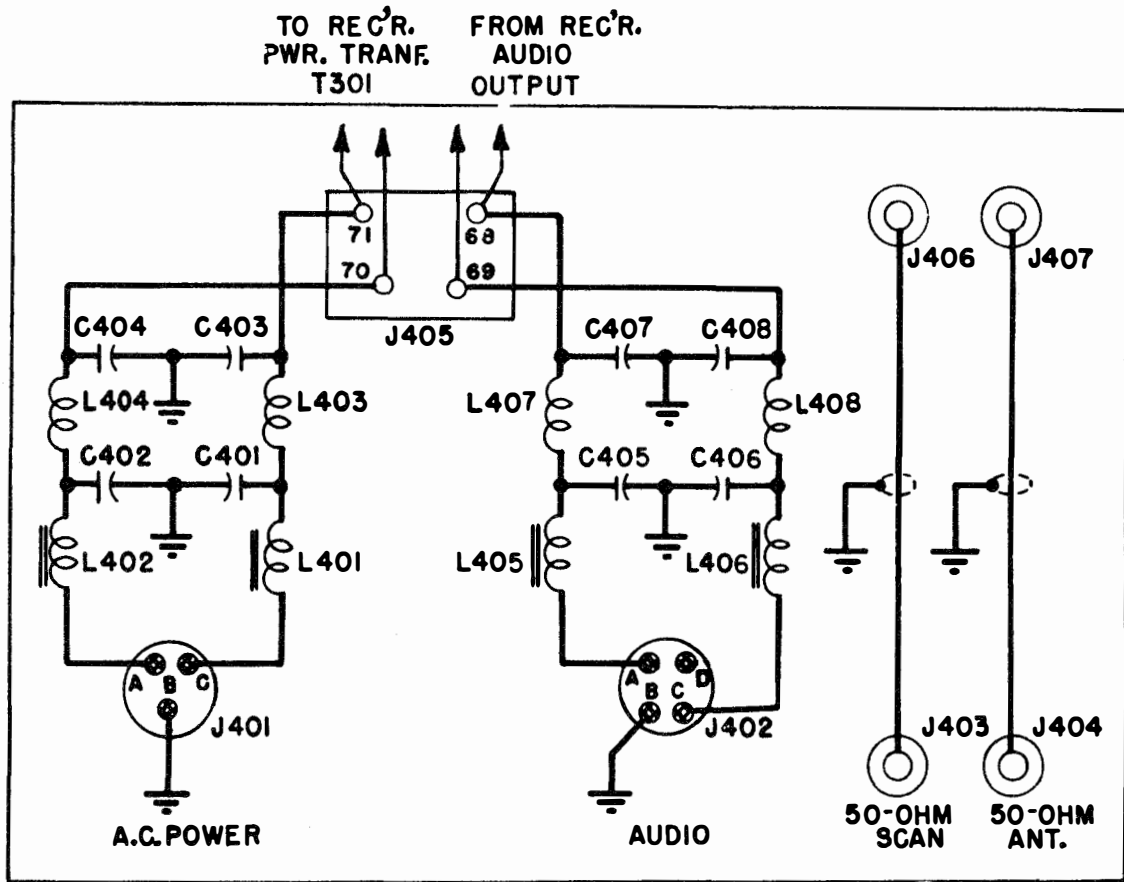
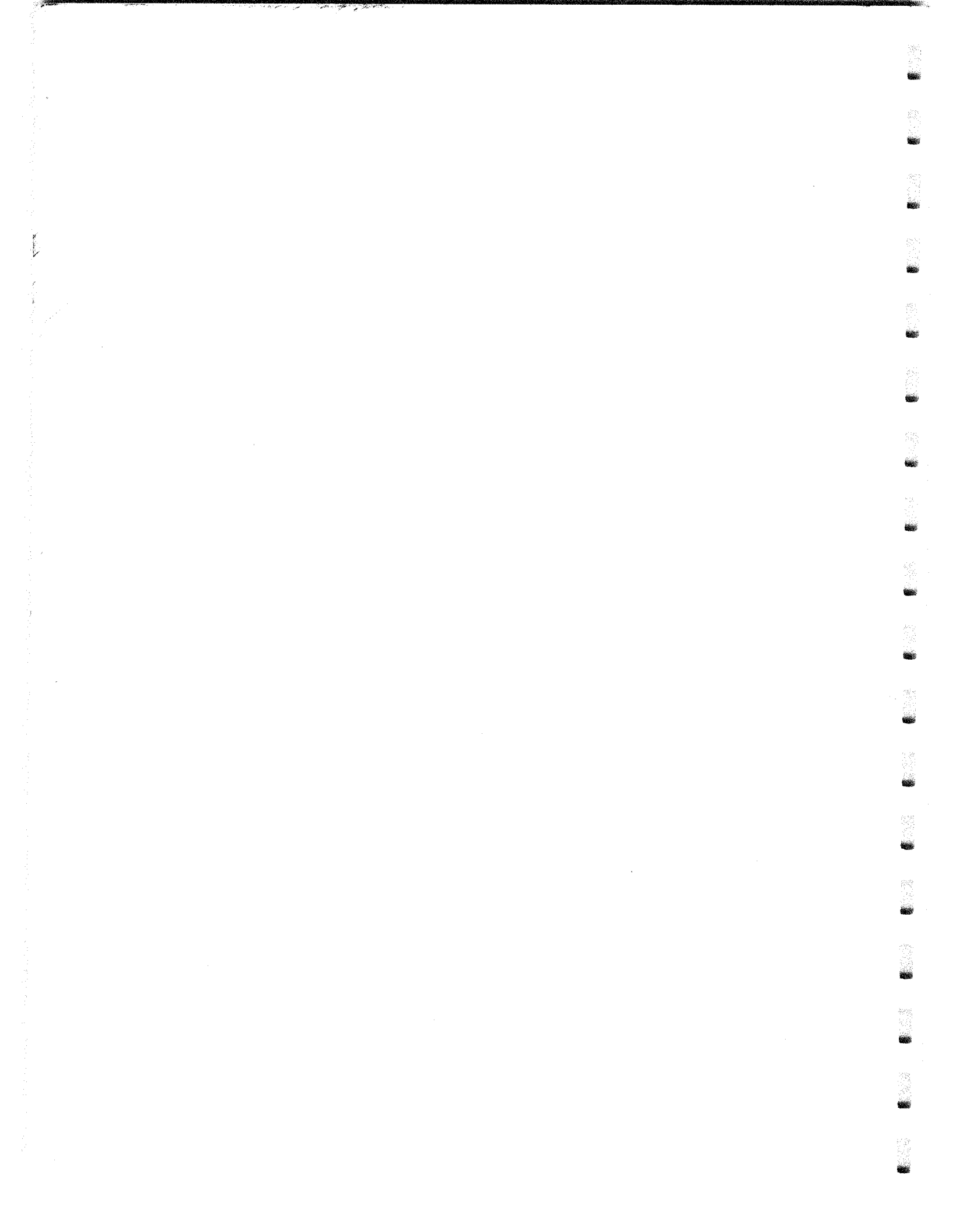


Figure 9-10. Simplified Schematic—Band Suppression Filter F-89/URR-13



CHAPTER 10

VHF TUNABLE RECEIVERS TYPES RV-5 AND RV-8

10-1. INTRODUCTION

Due to the similarity of the RV-5 and RV-8 receivers, they will be discussed in general in this chapter. Differences between the two receivers, other than those mentioned in the text, may be found by studying their respective schematic diagrams.

10-2. GENERAL DESCRIPTION

The RV-5 and RV-8 receivers are dual-conversion superheterodyne receivers. They are continuously tunable over a frequency range of 108 mc to 152 mc and are designed for aeronautical ground station reception of AM signals. They are equipped with automatic noise limiting features, delayed and amplified avc, and carrier operated squelch circuits. They are designed for mounting in a standard relay rack and will operate satisfactorily in ambient temperatures from -10° to 60° C and relative humidity up to 95%.

A. *Technical Summary of Receiver Characteristics*

Frequency Range
108-152 mc in one band.

Type of Reception
AM.

Antenna Input
50 Ohm unbalanced.

Intermediate Frequencies
30 mc and 4.5 mc.

Sensitivity
5 μ v, 30% modulated over entire frequency range for a minimum of 10 db signal to noise ratio.

Selectivity (RV-5)
Bandwidth at 6 db 100 kc minimum.
60 db 360 kc maximum.

Selectivity (RV-8)
Bandwidth at 6 db 50 kc minimum.
60 db 150 kc maximum.

Asymmetry of Selectivity
Less than 15%.

AF Output Impedance
For 600 and 20,000 ohm loads, balanced and ungrounded.

Frequency Stability
 $\pm 0.08\%$.

Audio Power Output
1 watt into each load.

Audio Response
 ± 3 db from 200 to 3,000 cps.

Harmonic Distortion
Less than 8% for 1 watt of audio output.

Hum and noise
At least 30 db below 1 watt output.

Noise Limiters
Series and shunt type.

Squelch
Threshold 1 μ v at maximum RF gain. More than 50 μ v at minimum RF gain.

AVC
Audio output constant within 2 db with an input level change of 80 db.

Power Requirements
Approximately 60 watts, 115 v ac, 50-60 cps.

Tube Complement
One 6AF4A, two 6AK5, three 6BA6, one 12AT7, (6201)*, and one 6AQ5.

* (6201 is an improved 12AT7 for the second mixer in the RV-8)

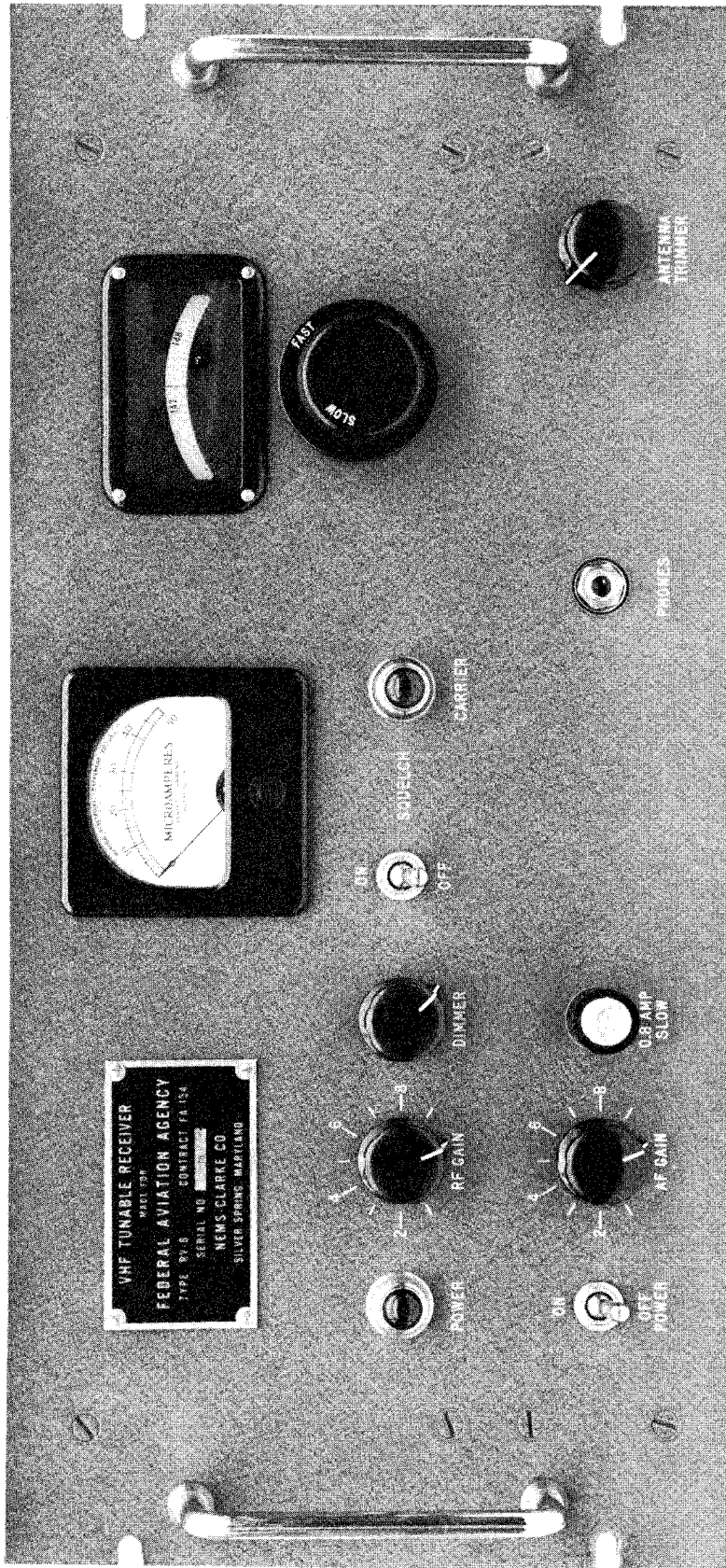


Figure 10-1. RV-8 Receiver.

NOTE: From the technical summary, the student should take cognizance of the sharper selectivity characteristics of the RV-8 receiver as compared to the RV-5 receiver.

10-3. THEORY OF OPERATION

The RV-5 and RV-8 receivers are dual-conversion superheterodyne receivers employing a RF amplifier, four stages of IF amplification in two sections, and an audio section.

The RF section consists of a RF amplifier, a local oscillator, and a mixer. The frequency of the local oscillator is 30 mc above the frequency of the incoming signal. The two signals are applied to the mixer and the difference frequency of 30 mc is supplied to the first IF section.

The first IF section employs a 30 mc double-tuned circuit. The output is fed to the second mixer V-302A. V-302B is a 12.75 mc crystal-controlled oscillator. This signal, together with its second harmonic of 25.5 mc, is coupled to the cathode of the second mixer. The second mixer then combines the 30 mc signal on the control grid with the 25.5 mc oscillator injection on the cathode to produce a 4.5 mc IF output in the plate circuit. This 4.5 mc signal is then amplified by the second IF section. The second IF section consists of three stages which are conventional double-tuned 4.5 mc amplifiers.

Both a crystal diode audio detector and a separate avc detector-amplifier are connected to the output of the second IF section. The audio from the AF detector is fed through a series noise limiter, a squelch circuit and three stages of audio amplification.

A. *Input Circuit and Radio Frequency Amplifier*

The RF signal from the antenna is applied through J-201 to the ANTENNA TRIMMER, C-201. The signal is then coupled to one section of the four section, Mallory Type, inductuner which tunes the grid circuit of the first RF amplifier.

Bias voltage is applied to the grid of V-201 to maintain operation of the stage at its most linear operating point. This bias also prevents modulation of the desired signal by nearby

undesired high level signals. In the RV-5, a partial avc bias is applied to the grid of V-201 through R-202. In the RV-8, a fixed bias (obtained from the -75 volt supply and voltage divider network, R-130 and R-131) is applied to the grid of V-201. R-130 is a variable screwdriver control, located on the RV-8 chassis, and requires resetting whenever V-201 is replaced.

The plate circuit of V-201 is coupled to the grid of V-202 by a double-tuned band-pass filter. A capacitive "T" network provides coupling between the primary and secondary tuned circuits.

B. *First Oscillator and Mixer*

A 6AF4A triode is used as the first local oscillator which is tuned by a section of the inductuner. The oscillator, a modified Colpitts, operates 30 mc above the incoming signal and its output is injected into the mixer grid of V-202.

An added feature incorporated in the RV-8, that is not in the RV-5, is a biased mixer stage. In the RV-8 receiver, a combination of fixed and avc bias is applied to the grid of V-202 to prevent the mixer from overloading due to extremely strong signals (such as 1 volt of RF at the antenna). A fixed bias of -7 volts is obtained from the -75 v supply and voltage divider network, R-130 and R-131. This bias is applied to the grid of V-202 through the clamping diode, CR-104. AVC is applied to the grid of V-202 through the clamping diodes, CR-105 and CR-106, and their associated resistors. Thus, when a weak signal is being received, the grid of the mixer remains biased at -7 volts. When the avc increases to a value exceeding -7 volts, the avc takes over and biases the grid of the mixer to compensate for the strong input signal and thus prevent overloading of the mixer stage.

The plate of the mixer is a double-tuned circuit consisting of T-201 and T-202 which are both tuned to 30 mc. The output of T-202 is coupled to the grid of the first IF amplifier, V-301, through C-229.

C. *First Intermediate Frequency Section*

The output of the first mixer is amplified by a single 30 mc IF stage, V-301, and coupled

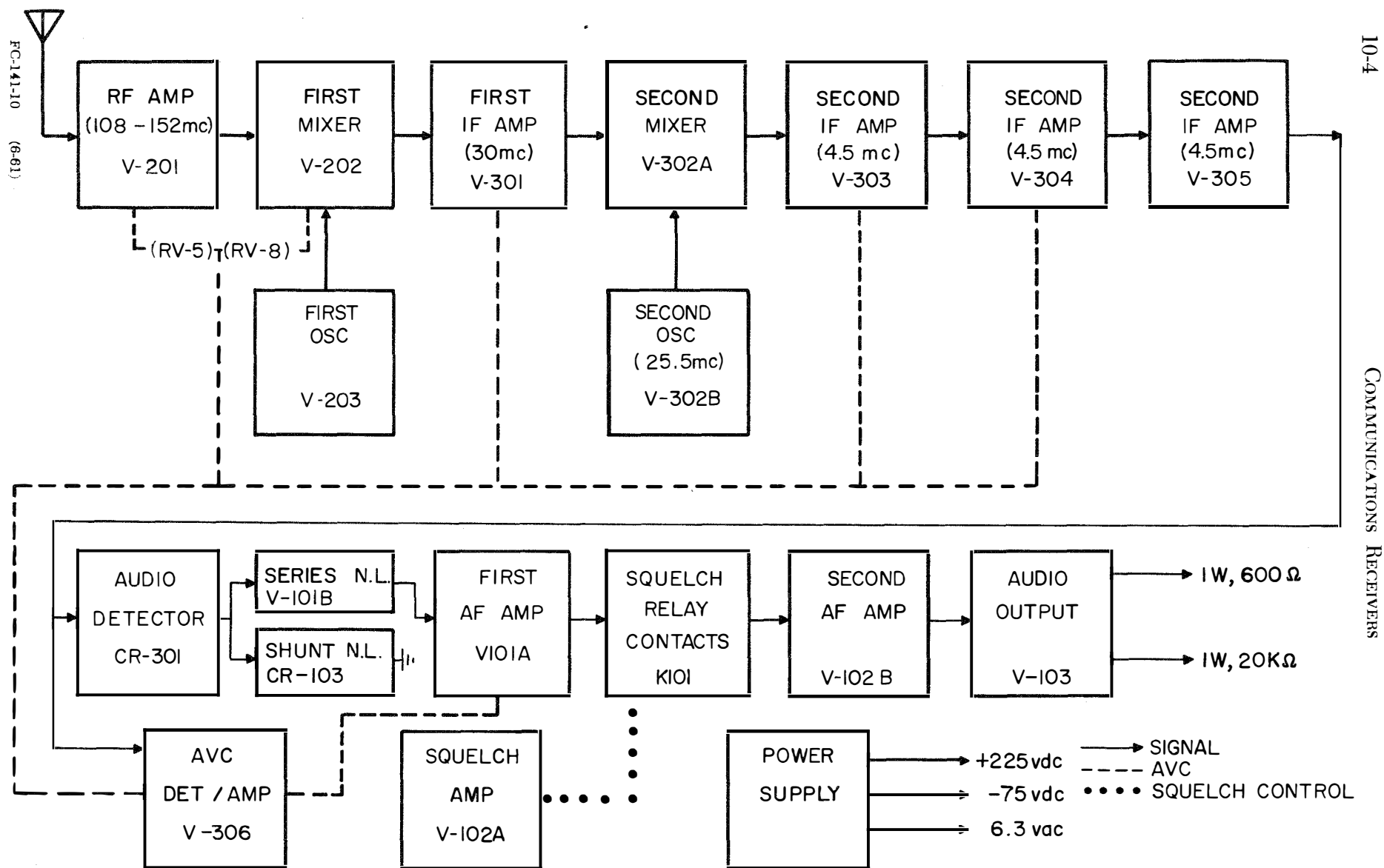


Figure 10-2. Type RV-5 and RV-8 Receiver Block Diagram.

to the second mixer by a double-tuned circuit T-301 and T-302.

D. Second Oscillator and Mixer

Half of V-302 is the second local oscillator, operating at 12.75 mc, and is controlled by crystal Y-301. The second harmonic of the crystal frequency is coupled to the mixer half of V-302, where it is combined with the 30 mc input signal to produce a 4.5 mc output signal. A double-tuned circuit, T-303 and T-304, couples the output of the mixer to the grid of V-303.

E. Second Intermediate Frequency Section

The 4.5 mc second IF section consists of three stages of conventional double-tuned IF amplifiers. The first two IF amplifiers, V-303 and V-304, use 6BA6 semi-remote cut-off pentodes (instead of sharp cut-off tubes) since avc bias is applied to the control grids of these amplifiers. The third stage, V-305, uses a 6AU6 sharp cut-off pentode since avc is not applied to this stage. The output of the final IF tuned circuit is coupled to the audio and avc detectors.

F. Automatic Volume Control

The avc detector-amplifier circuit employs a dual-diode-triode in a circuit that holds the output nearly constant over a wide range of input levels. The avc detector consists of one section of V-306 and associated load resistors (R-335 and R-336 for the RV-5; R-324 and R-325 for the RV-8). Part of the output of this detector is applied to the grid of the triode section, a cathode loaded dc amplifier. The dc output of this amplifier is coupled through the other diode section to the avc time constant circuit (R-338 and C-117 for the RV-5; R-327 and C-117 for the RV-8). The amplifier also provides the control voltage for operation of the squelch circuit.

The RF gain function of the receiver introduces a variable amount of negative bias to the avc buss. This bias is obtained by means of a voltage divider network (R-124 and R-125) and is coupled to the avc time constant circuit by R-339 in the RV-5, and by R-328 in the RV-8.

G. Detector and Noise Limiters

The detector-noise limiter section of the receiver includes an audio diode detector, a shunt diode noise limiter and a series diode noise limiter. The amplitude modulation detector is a conventional crystal diode operating as a signal rectifier. The output of the detector is a negative voltage proportional to the carrier, upon which is superimposed the audio modulation signal.

Both noise limiters have been designed to remove signals exceeding the level expected from normal modulation. In the shunt limiter, (CR-106 in the RV-5; CR-103 in the RV-8), a negative bias of 16 volts is applied to the anode and therefore delays clamping action until the detector output exceeds -16 volts. The normal output, due to the signal carrier, is approximately -5 volts.

In the case of 100% modulation the peak output is -10 volts. Since the bias on the diode exceeds the signal voltage normally applied with 100% modulation, no clamping action occurs on a normal signal. When noise spikes substantially stronger than the signal are received, the diode will clamp them at a -16 volt level. Since this clamping action cannot occur on a normal signal it is not necessary to provide switching to remove it from the circuit.

The series limiter has been designed to clip the audio signal to a value slightly less than 100% modulation in the RV-5 and to slightly greater than 100% modulation in the RV-8. This circuit is self adjusting by using the average value of the rectified signal carrier as a reference and permitting a peak audio voltage approximately equal to this value to pass through. For purposes of illustration, assume that a detector output of -5 volts dc is obtained. Approximately one-half of this is applied to the plate of V-101B through the voltage divider R-114 and R-113. Modulation is removed from the detector output by the action of filter R-115 and C-116. This dc voltage is proportional to the carrier level and is applied to the cathode of V-101B through R-116. During a modulation cycle V-101B will conduct and therefore couple the audio signal as long as the peak swing on its plate does not drop below the voltage across C-116.

Since the series limiter in the RV-5 operates at levels below 100% modulation, a switch is provided to remove it from the circuit when a noise free signal is being received. This permits full output on carriers modulated 100%.

H. Audio Circuit and Squelch

The output of the noise limiter circuit is coupled to the AF GAIN control, R-103, by C-105. From the AF GAIN control the signal is amplified by the first audio amplifier, V-101A, and then coupled to the MUTING control, R-106, and the silencer (Squelch) relay, K-101. This relay is actuated by V-102A which is in turn controlled by a signal developed in the avc amplifier circuit.

With no carrier being received, V-102A operates at zero bias and sufficient current flows through the tube to operate relay K-101. This couples the grid of the second audio amplifier, V-102B, to the center tap on the MUTING control. This control is adjustable from the rear of the chassis to provide any desired degree of audio silencing.

When a signal is received, a negative bias is developed on the grid of V-102A and current flow is reduced to the point where the relay will open. This transfers the connection of the second audio amplifier grid so that full audio output is obtained. The SQUELCH switch, S-102, is used to disable the silencer circuit, by opening the cathode of V-102A.

I. Power Supply

1. RV-5 POWER SUPPLY

The high voltage power supply utilizes one secondary of power transformer T-101, supplying a voltage of 240 vac to four silicon diodes, CR-101 through CR-104, connected in a full wave bridge rectifier circuit. The bridge rectifier circuit supplies pulsating dc to a capacitive input LC filter. The output of this filter is +225 vdc, with 0.22 v peak ripple voltage.

A low voltage 73 vac is supplied from another secondary of T-101 to the silicon diode, CR-105, which is connected as a half-wave rectifier feeding pulsating dc to a capacitive input RC filter, the output of which is -75 vdc with 0.92 v peak ripple voltage. This

bias voltage is used for operation of the avc and noise limiter circuits.

2. RV-8 POWER SUPPLY

The high voltage power supply utilizes one secondary of power transformer T-101, supplying a voltage of 240 volts ac to a dual section silicon diode, CR-101, connected as a full-wave rectifier and operates with a capacitive input LC filter to provide a stable supply of +225 volts dc.

The low voltage bias supply uses another secondary of T-101, supplying a voltage of 73 volts ac to the silicon diode CR-102. This diode is connected as a half-wave rectifier and operates with a capacitive input RC filter to provide a bias supply of -75 volts dc. The bias voltage is used for operation of the avc and noise limiter circuits, and also supplies a fixed bias for the RF amplifier and first mixer tubes.

10-4. OPERATING CONTROLS

The functions of the Type RV-5 and RV-8 VHF tunable Receiver is to receive Type A-3 signals at any channel frequency within the frequency range of 108 to 152 mc. All controls necessary for operation of the receiver are located on the front panel. Functions of the various controls and indicators are described in the following paragraphs.

The POWER ON-OFF control, S-101, is a SPST switch controlling the application of power to the receiver. On the RV-5, this switch is an integral part of the AF GAIN control and is OFF whenever the AF GAIN control is completely CCW. On the RV-8, the POWER ON-OFF control is a distinct SPST toggle switch.

The AF GAIN control, R-103, adjusts the audio output of either receiver in a conventional manner. Maximum audio output is obtained with full CW rotation of the control.

The RF GAIN control, R-125, is a continuously variable control which adjusts the RF gain and hence the squelch opening level of the receiver. By means of this control, the squelch opening level can be adjusted from approximately 1 μ v at maximum CW rotation of the control to approximately 100 μ v at maximum CCW rotation.

On the RV-5, a control that is not on the RV-8, is a MANUAL-AVC toggle switch S-104. This switch is open in the manual position but shorts out the RF GAIN control when in the AVC position.

The RF TUNING control adjusts the four section, Mallory Type, inductuner, L-201A, B, C, D to any channel frequency within the frequency range of the receiver. This control has both a coarse and a fine tuning adjustment.

An ANTENNA TRIMMER control, C-201, is adjusted so that maximum power from the antenna is fed to the RF amplifier.

The SQUELCH ON-OFF switch, S-102, permits the muting (squelch) circuit to be disabled. In the OFF position, the receiver remains unmuted; whether a carrier is being received or not.

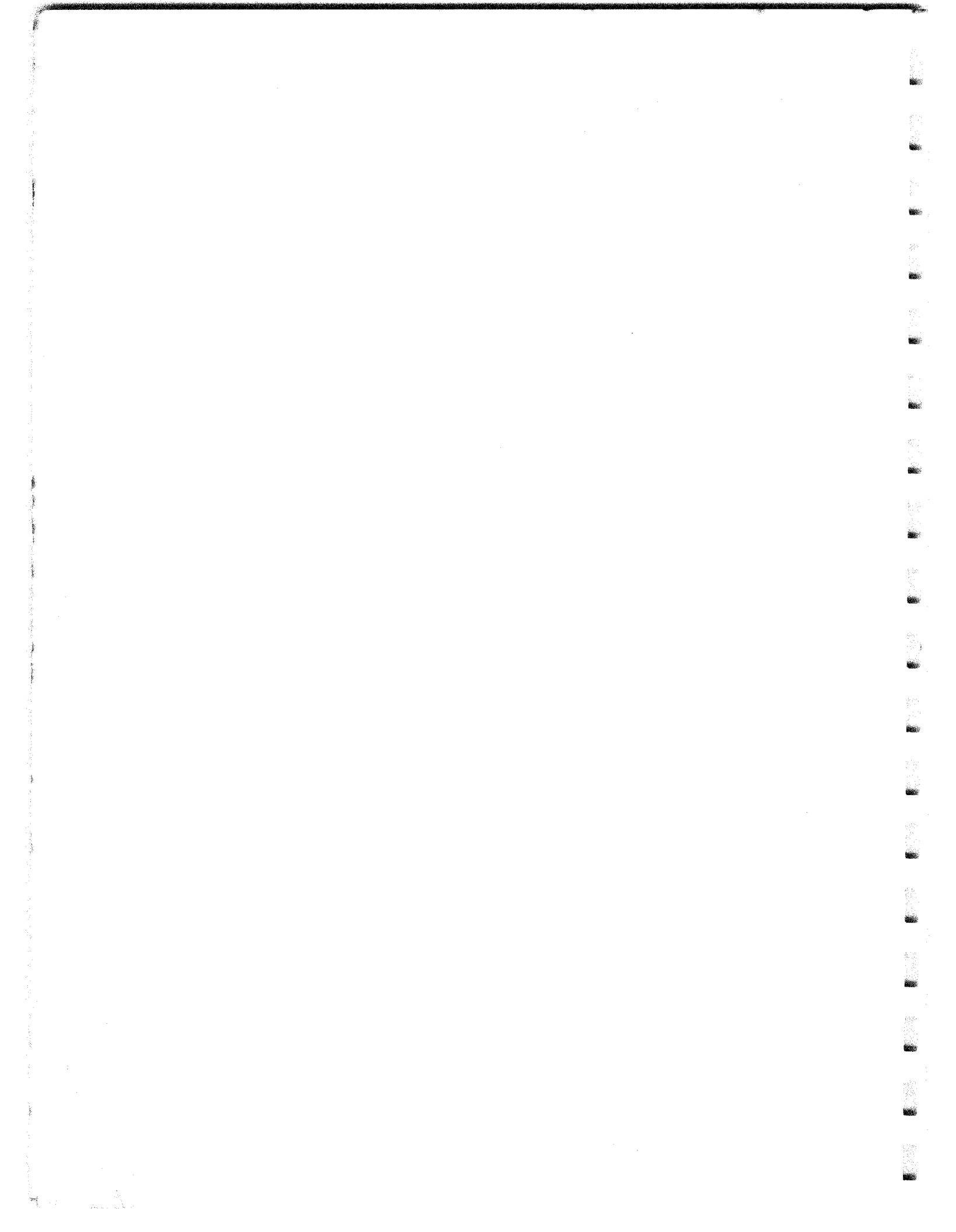
A control, which exists only on the RV-5, is a NOISE LIMITER ON-OFF spst switch S-103. In the ON position, the series noise limiter is in the circuit. In the OFF position, the series noise limiter is bypassed.

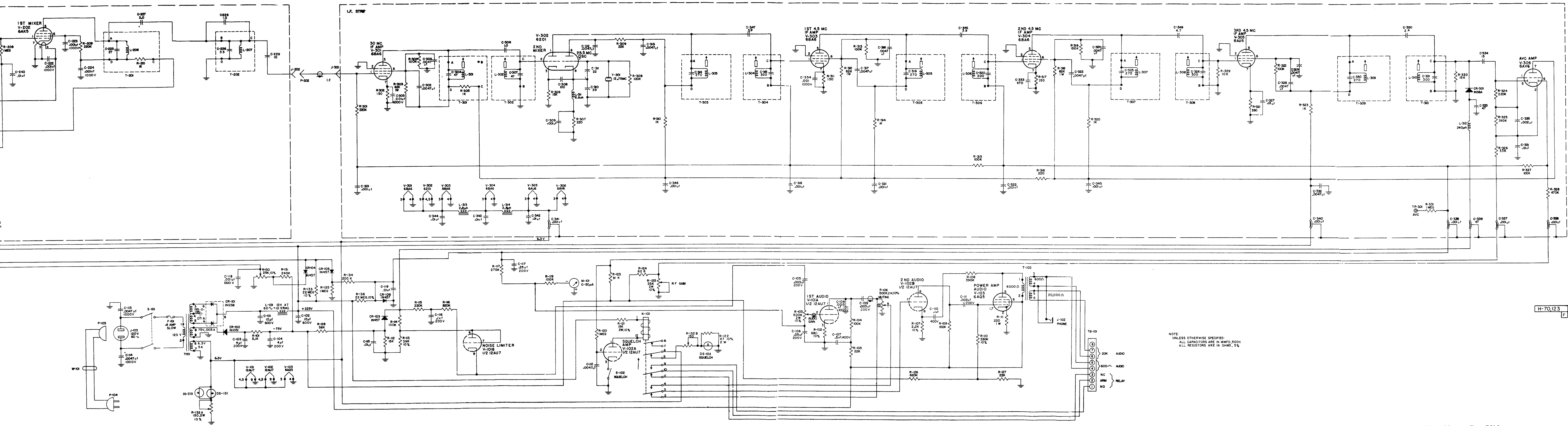
A control, which exists only on the RV-8, is a DIMMER CONTROL R-132. Use of this control permits the operator to dim the lamps (CARRIER AND POWER) to suit his convenience.

10-5. OPERATING PROCEDURE

Tune the RV-5 or RV-8 receiver to the desired signal. Proper tuning will be indicated by maximum reading of the signal strength meter. Also adjust the ANTENNA TRIMMER for maximum reading of the signal strength meter.

Adjust the AF GAIN control for desired audio output level. With the RF GAIN control set to minimum, advance the control until the ambient noise signal level just opens the squelch. Next, reduce the RF GAIN until the squelch just mutes the receiver. This will insure that the receiver is operating at the maximum sensitivity and still permit muting between transmissions.

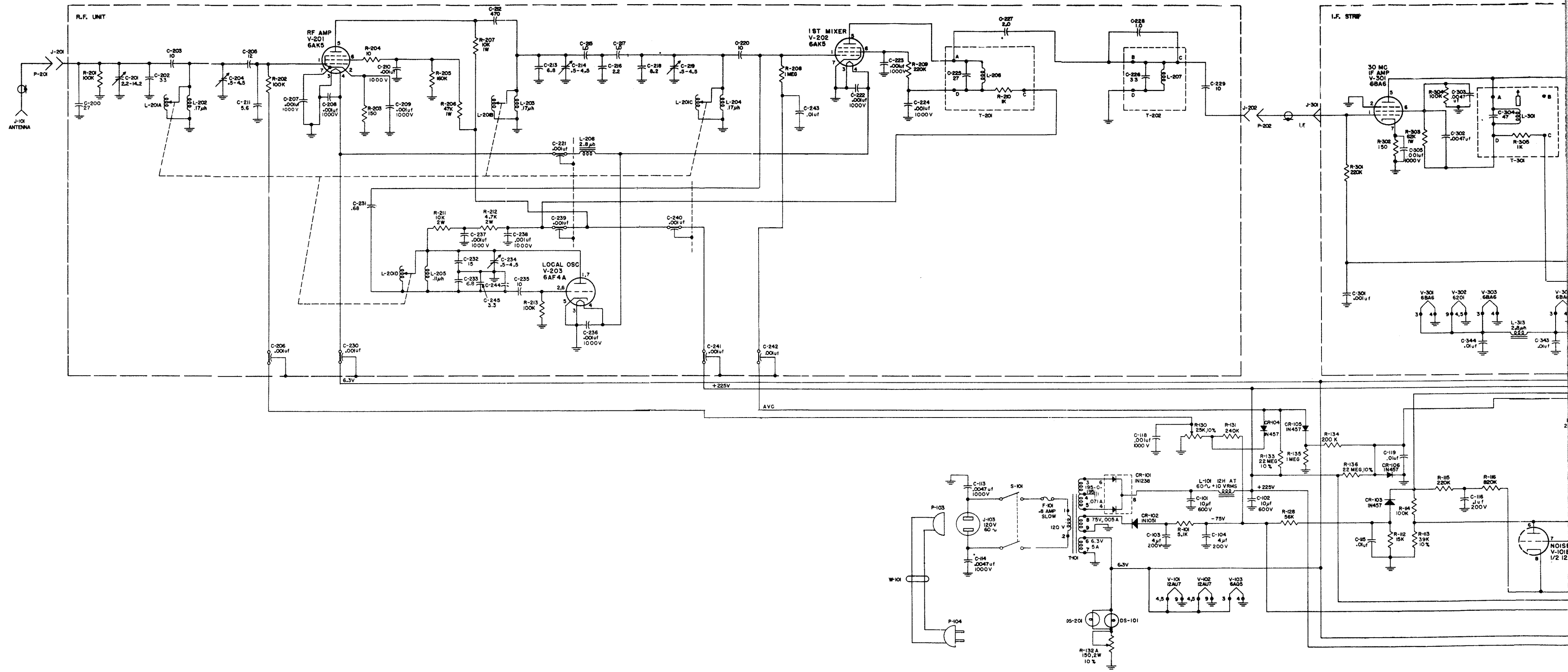


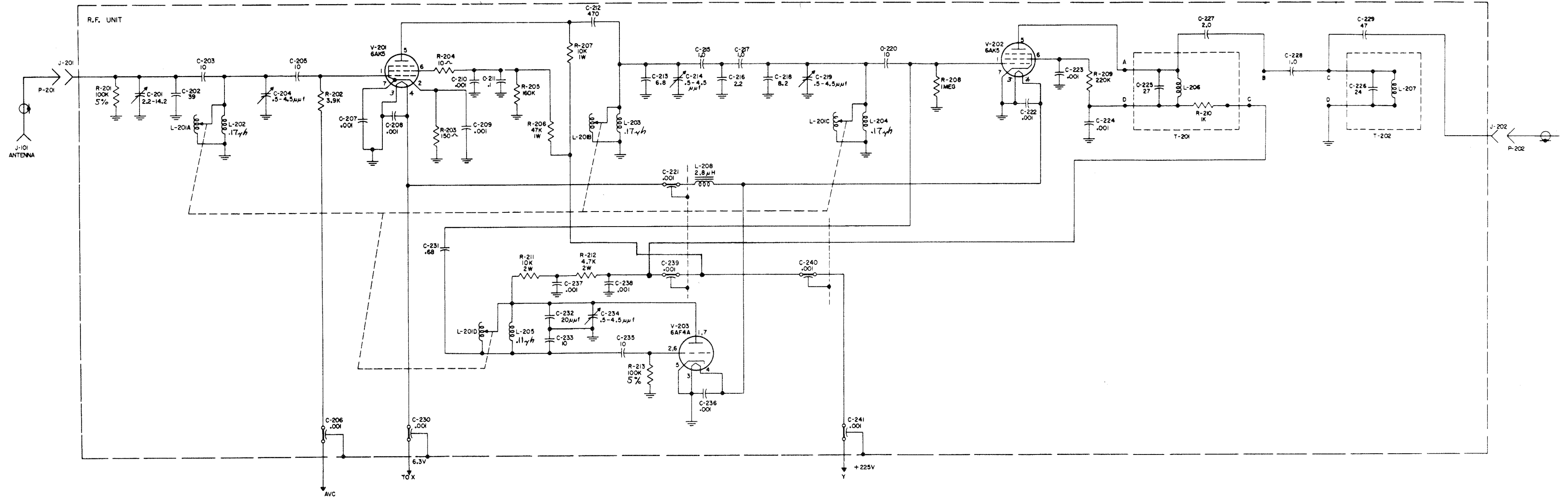


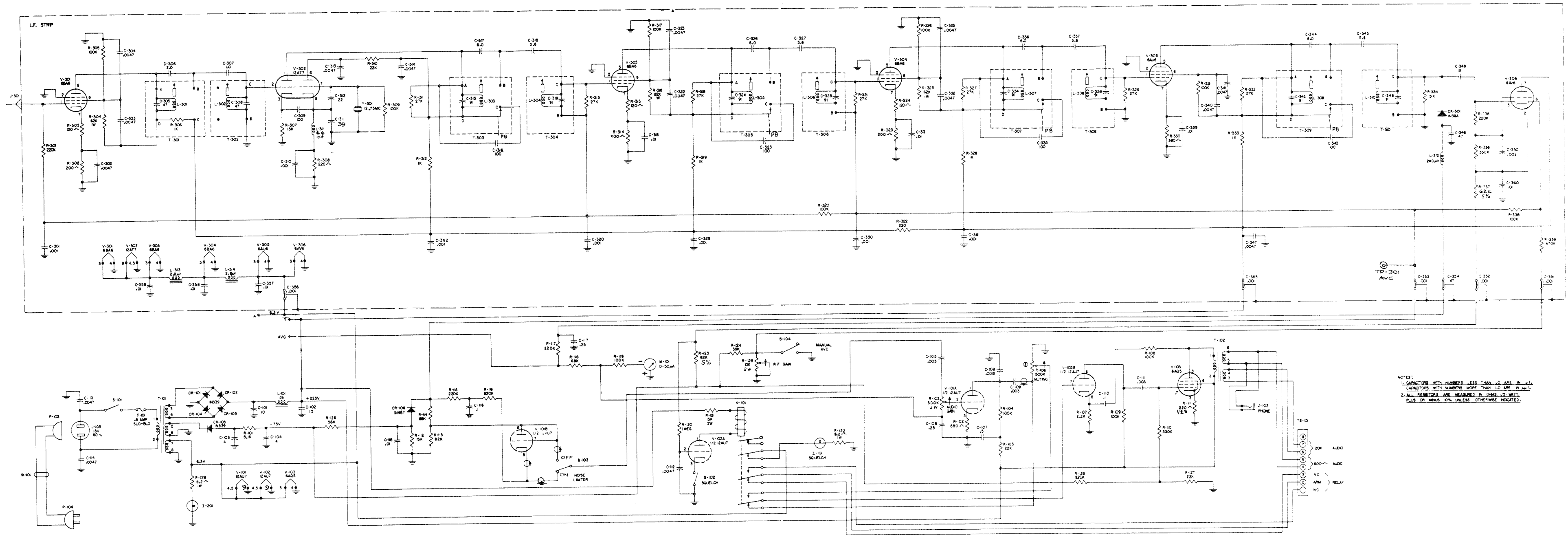
NOTE:
UNLESS OTHERWISE SPECIFIED:
ALL CAPACITORS ARE IN MMFD, 500V.
ALL RESISTORS ARE IN OHMS, 5%.

H-70,123 F

FIGURE 10-3 Schematic Wiring Diagram, Type RV-8
VHF Tunable Receiver







NOTES:
 1. CAPACITORS WITH NUMBERS LESS THAN 10 ARE IN P.P.
 CAPACITORS WITH NUMBERS MORE THAN 10 ARE IN P.P.
 2. ALL RESISTORS ARE MEASURED IN OHMS UNLESS
 PREFIX OR SUFFIX IS UNLESS OTHERWISE INDICATED.

Figure 10-4. RV-5 Receiver (Sheet 2 of 2)

CHAPTER 11

VHF FIXED TUNED RECEIVERS TYPES RUQ—RVA—RV-1—RV-3—RV-4 AND RV-6

11-1. INTRODUCTION

To the radio technician a good basic understanding of the theory, adjustment, operation, preventive and corrective maintenance of a particular piece of equipment is an absolute necessity, not only so that he may have a true, clear conception of how it works, but also that he may understand the effects of troubles common to its various circuits.

The information for this chapter was taken primarily from the instruction book for the Type RV-6 receiver. Differences among the various receivers, other than those mentioned in the text, may be found by studying the accompanying technical summary sheets, their schematic diagrams and pictures.

11-2. SAFETY NOTICE

The attention of operating and maintenance personnel is directed to Federal Airways MANOP IV-A-4 for instructions on the subject of radio-safety precautions to be observed.

This equipment employs voltages which are dangerous and may be fatal if contacted by operating personnel. Extreme caution should be exercised when working with the equipment.

While every practicable safety precaution has been incorporated in this equipment, the following rules must be strictly observed:

KEEP AWAY FROM CIRCUITS. Operating and maintenance personnel must at all times observe all safety regulations. Do not change tubes or make adjustments inside equipment with high voltage supply on. Under certain conditions dangerous potentials may exist in circuits with power controls in the off position due to charges retained by capacitors. To avoid casualties always remove

power, then discharge and ground circuits prior to touching them, by use of the grounding stick.

DON'T TAMPER WITH INTERLOCKS. Do not depend upon door switches or interlocks for protection but always shut down high-voltage rectifiers and other power equipment. Under no circumstances should any access gate, door, or safety interlock switch be removed, short-circuited, or tampered with in any way, except by authorized maintenance personnel when considered unavoidable, nor should reliance be placed upon the interlock switches for removing voltages from the equipment. The grounding stick should always be employed (after shutting off all power) to insure that circuits are electrically "dead."

RESUSCITATION. Maintenance personnel should familiarize themselves with the technique for resuscitation found in the Manual of First Aid Instructions.

11-3. BRIEF DESCRIPTION OF EQUIPMENT

These types of fixed-tuned receivers are communication receivers having a single preset crystal-controlled channel in the range 108 to 152 mc. (The RUQ has a range of 118 to 136 mc.) They are designed for aeronautical ground station reception of amplitude-modulated radio-telephone signals. These receivers employ a double-conversion super-heterodyne circuit and are equipped with automatic noise limiter, delayed and amplified automatic volume control, and carrier-operated squelch circuits. They have been designed in accordance with Federal Aviation Agency specifications.

The receivers have vertical chassis enclosed in a cabinet with a front door panel for mount-

ing in a standard 19-inch width rack. The height of the panel is 8- $\frac{3}{4}$ inches. The equipment has been designed for satisfactory operation from -10 degrees C to plus 60 degrees C (14 degrees to plus 140 degrees F) ambient temperature and for relative humidity to 95 percent plus or minus five percent.

11-4. GENERAL CIRCUIT THEORY

The function of these fixed-tuned receivers is to receive amplitude-modulated radio telephone signals on a single preset channel. The functional diagram, Fig. 11-1, presents a broad picture of the operation of the circuit elements. The complete schematic diagrams for the RUQ and RV-6 are shown at the end of this chapter.

These receivers employ a double-conversion superheterodyne circuit. Referring to Fig. 11-1, the signal to be received is amplified by a single stage RF amplifier, V-1. It is then mixed in the first mixer, V-2, with an injection signal generated by the crystal-controlled oscillator-multiplier section to produce an intermediate frequency of 18.3 mc. This 18.3 mc signal is amplified by the first IF amplifier, V-3, and applied to the second mixer-oscillator, V-4, where it is mixed with the oscillator signal (16.45 mc for the RV-3, RV-4 and RV-6; 15.1 mc for the RUQ, RVA and RV-1) to produce the second intermediate-frequency of 1.85 mc

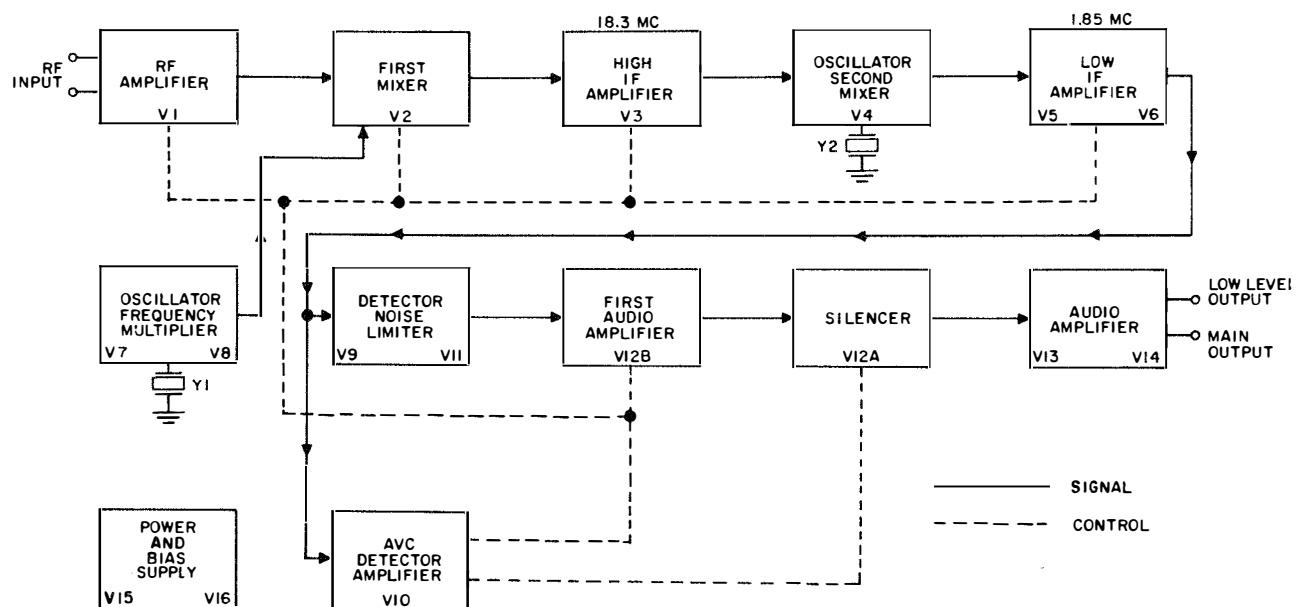
for the RV-3, RV-4 and RV-6. (3.2 mc for the RUQ, RVA and RV-1).

This intermediate-frequency signal is then amplified in the two-stage second IF amplifier, V-5, V-6, and applied to the detector-noise limiter, V-9, V-11, and the avc detector amplifier, V-10, which produces an avc voltage that is applied to amplifier tubes in order to maintain the audio output level very nearly constant for wide variations in RF input signal strength. This circuit also produces a control voltage for operation of the silencer (squelch circuit).

The audio signal developed at the detector is then amplified by the first audio amplifier, V-12B. From this stage the signal is transmitted through a filter section into a carrier-operated silencer (squelch) circuit which silences the receiver when no signal is being received. This circuit includes V-12A and Relay K-1, which is operated by a dc control signal developed by the avc detector amplifier. The audio signal is then amplified by V-13 and V-14. This section provides two separate audio output circuits.

The power and bias supply, V-15 and V-16, provides the positive and negative voltages required for operation of the various receiver circuits.

To prevent the introduction of undesirable signals into the receiver, the power input leads



and all output leads terminating in connector J-4 are filtered.

11.5. DETAILED CIRCUIT THEORY

A. RF Amplifier.

Referring to Fig. 11-2, the RF amplifier consists of V-1, a Type 6AK5 pentode, operating in conjunction with three capacitance-tuned circuits. The RF signal is coupled to the input tuned circuits, L-2, C-1, C-2, by means of a coupling link, L-1. This coupling is connected to both RF input connectors, J-1 and J-2, so as to permit the series connection of two receivers to a common antenna. The second tuned circuit, L-4, C-10, is coupled to the third tuned circuit, L-5, C-11, C-12, by means of a wire link capacitor. This circuit is coupled to the input of the first mixer by means of C-13. AVC voltage is applied to the grid of V-1 through a resistance-capacitance filter, R-1, C-4.

B. Oscillator-frequency multiplier.

The oscillator-frequency multiplier section includes V-7, a type 6AK5 pentode, oscillator-first multiplier and V-8, a type 6AK5 pentode, second multiplier. The crystal oscillator employs Y-1, a type CR-27/U quartz crystal, (frequency range 5606.25 to 8356.25 kc for the RV-3, RV-4 and RV-6) or a type CR-18/U for RUQ, RVA and RV-1, (frequency dependent on the channel frequency). The RV-3, RV-4 and RV-6 have a temperature controlled oven OV-1 (70 degrees C.) but the RUQ, RVA and RV-1 do not have an oven.

It operates in a modified Colpitts circuit in which no tuned circuit at crystal frequency is required. In this circuit, the crystal presents an inductive impedance which resonates with capacitors C-39 and C-40. V-7 also acts as an electron-coupled frequency quadrupler, the tuned circuit L-9, C-42, C-43 being tuned to a frequency four times the crystal frequency. This circuit is tuned by means of a powered iron slug which varies the inductance of L-9. The frequency range of this circuit is from 22.425 to 33.425 mc in the RVA, RV-1, RV-3, RV-4 and RV-6. In the RUQ the range is

from 24.9 to 29.4 mc. Note that the screen of V-7 is bypassed to ground to provide isolation. The frequency of oscillation is independent of the resonant frequency of the plate circuit.

The frequency of the crystal unit, Y-1, is determined from the channel frequency of the receiver by means of the formula:

$$f_{\text{xtal}} = \frac{f_{\text{channel}} - 18.3}{16} \text{ times } 1000$$

Where: f_{xtal} = crystal frequency in kc
 f_{channel} = channel frequency in mc
 likewise, the channel frequency can be determined from the xtal frequency by means of the formula:

$$f_{\text{channel}} = \frac{16 \text{ times } f_{\text{xtal}} \text{ plus } 18.3}{1000}$$

The second frequency multiplier, V-8, operates as a frequency quadrupler, its output tuned circuits being tuned to four times the frequency of the input signal. Two inductively coupled capacitance-tuned circuits, L-11, C-49 and L-12, C-50, are employed to reduce the amplitude of undesired crystal harmonic frequency signals to a minimum. (Note: Values of certain capacitors and resistors are different in the RF, mixer and multiplier circuits of the RUQ receiver due to its narrower frequency range.) The output signal is coupled to the first mixer by means of C-14.

C. First mixer.

The first mixer circuit (Fig. 11-2) consists of V-2, a type 6AK5 pentode, in a circuit which mixes the amplified RF signal from the RF amplifier with the injection signal provided by the oscillator-frequency multiplier section to produce the first intermediate frequency signal of 18.3 mc. Both the RF and injection signals are impressed on the grid of V-2. The plate load of V-2 is the primary circuit of double-tuned transformer T-1, (Z-1 in the RUQ/RVA) which is tuned to 18.3 mc. The secondary circuit of this transformer is coupled to the grid of the first intermediate frequency amplifier, V-3, through C-19. AVC voltage is applied to grid of V-2 through a resistance-capacitance filter, R-5, C-25.

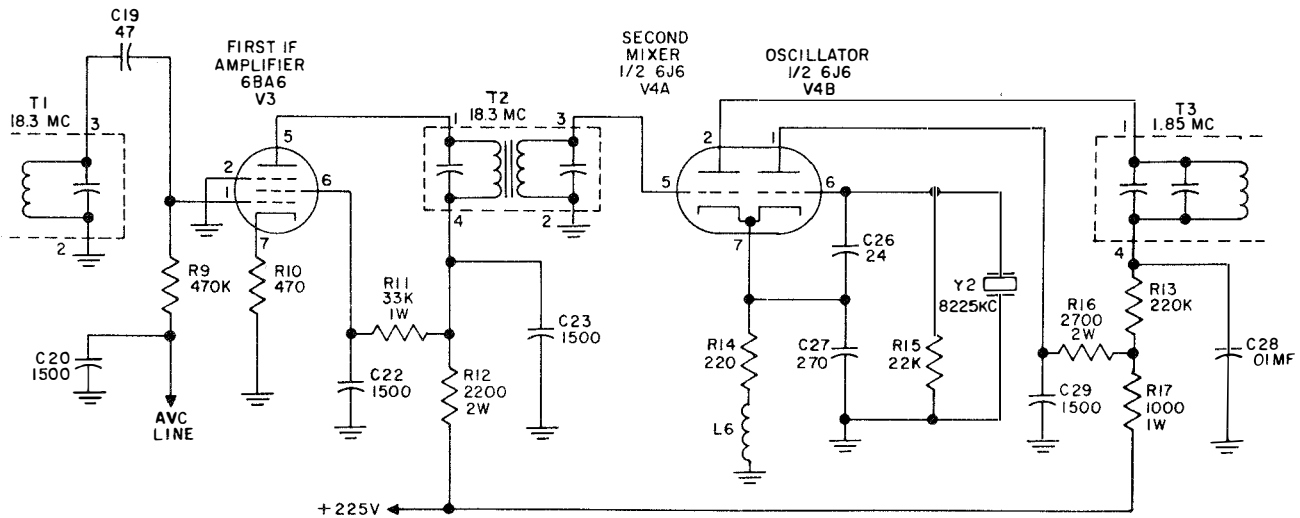


Figure 11-3. 1st IF Amplifier, Oscillator and 2nd Mixer Circuits

D. First IF amplifier.

Referring to Fig. 11-3, the 18.3 mc signal produced by the first mixer is then amplified by V-3, Type 6BA6. The plate load for this tube is the primary circuit of another double-tuned transformer, T-2 (or Z-2) which is also tuned to 18.3 mc. The avc voltage is applied to the grid of this tube, connection being made to grid resistor R-9 and filter capacitor C-20.

E. Second mixer-oscillator.

Referring again to Fig. 11-3 the second mixer-oscillator circuit employs V-4, Type 6J6, a double triode, one section of which acts as a

crystal oscillator, the other section as a triode mixer. The oscillator section, V-4B, employs a quartz crystal, Y-2, in a circuit similar to that of V-7. This oscillator operates at 8.225 mc for the RV-3, RV-4 and RV-6; while in the RUQ, RVA and RV-1 it operates at 7.550 mc. The second harmonic of the oscillator signal, 16.45 mc for the RV-3, RV-4 and RV-6, and 15.1 mc for the RUQ, RVA and RV-1 is mixed with the 18.3 mc first IF signal in the mixer section, V-4A, to produce the second IF signal of 1.85 mc for the RV-3, RV-4 and RV-6, and 3.2 mc for the RUQ, RVA and RV-1.

The mixer receives its injection signal from the oscillator by virtue of the common cathode

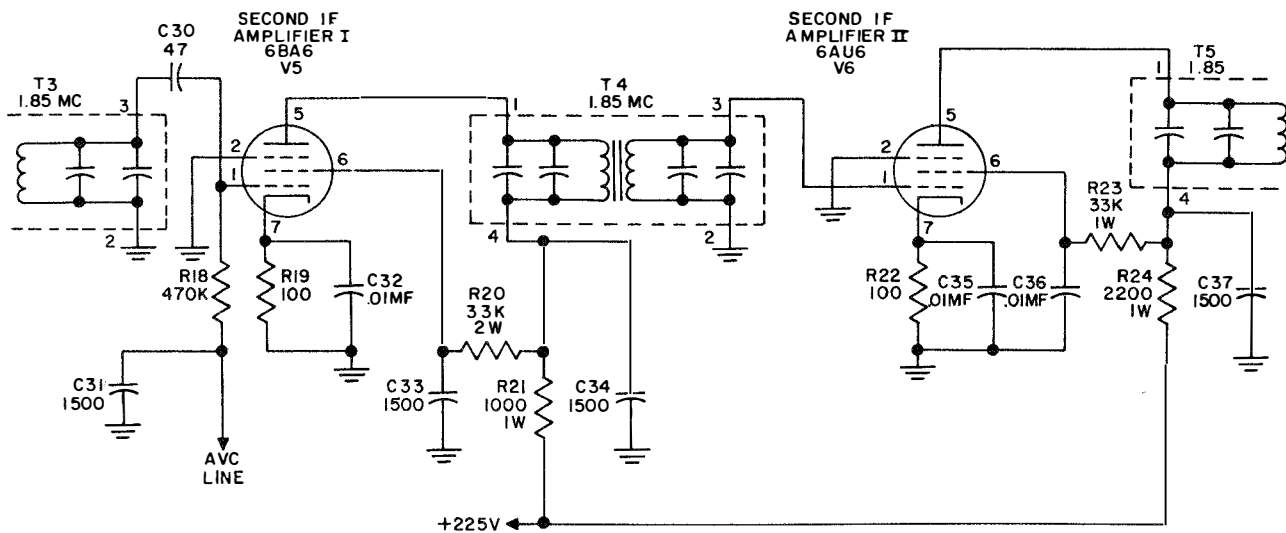


Figure 11-4. 2nd IF Amplifier Circuit

connection of the two triode sections. The plate load of the mixer is the primary circuit of a double-tuned transformer, T-3 (or Z-3), which is tuned to the appropriate frequency of 1.85 mc for the RV-3, RV-4 and RV-6, and 3.2 mc for the RUQ, RVA and RV-1. The secondary circuit of T-3 (or Z-3) is coupled to the grid of V-5 through capacitor C-30.

F. Second IF amplifier.

Referring to Fig. 11-4, the second IF amplifier includes V-5, Type 6BA6, and V-6, Type 6AU6, pentodes operating in conjunction with double-tuned transformers T-4 and T-5 (or Z-4 and Z-5). AVC voltage is applied to the grid of V-5 through grid resistor R-18. V-5 and V-6 operate as conventional, class "A" amplifiers. The selectivity of the receiver is obtained by virtue of the band-pass characteristics of T-3, T-4 and T-5 (or Z-3, Z-4 and Z-5). The secondary circuit of T-5 (Z-5) is coupled

to the detector-noise limiter circuit and also to the avc detector-amplifier.

G. Detector-noise limiter.

Referring to Fig. 11-5, the detector-noise limiter section of the receiver includes a diode detector, V-9A, 1/2 Type 6SN7GT, a series diode noise limiter, V-11B, 1/2 Type 6AL5, a shunt diode noise limiter, V-11A, 1/2 Type 6AL5.

The diode detector circuit is a conventional circuit in which V-9A, 1/2 Type 6SN7GT, operates as a rectifier; the diode load consists of R-38 and R-39 in series, bypassed by C-56. The output of the detector is a negative voltage (normally about -8 volts) upon which is superimposed the audio modulation signal.

The series diode noise limiter circuit employs diode element V-11B, Type 6AL5, in a circuit which limits the audio output signal to ap-

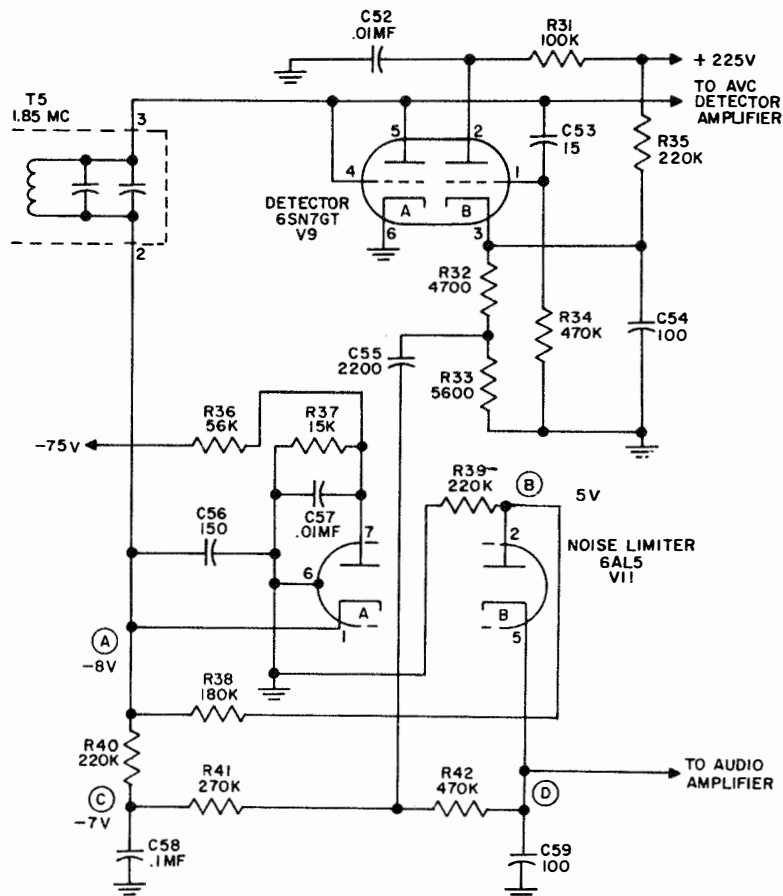


Figure 11-5. Detector and Noise Limiter Circuits

proximately 40 percent upward modulation. Since impulse noise of the type encountered at VHF consists of very sharp pulses, this limiting action greatly reduces the amount of noise energy transmitted to the audio-amplifier circuits without adversely affecting the intelligibility of voice signals. In the circuit employed, the plate of the diode is connected to the junction of R-38 and R-39, which act as a voltage divider. R-40 and C-58 act as a filter to establish a reference potential at point C, the junction of R-40, R-41 and C-58. R-41 and R-42 in series act as a diode load resistor across which the output audio signal is developed.

The operation of this circuit can best be understood by assuming V-11B to be an ideal diode; that is, perfectly conducting in one direction and completely nonconducting in the other direction of current flow. With a dc developed bias of about -8.0 volts at point A (the output of the detector) the current flow through the circuit will result in dc potential at points B, C and D as shown. Note that point C is bypassed to ground so that its instantaneous potential to ground is not affected by the modulation signal.

When the audio modulation is superimposed on the developed bias, the audio signal will be transmitted through the diode from point B to point D so long as the instantaneous potential of point B does not become more negative than that of point D. If this occurs, the electron conduction of the diode will be cut off and point D will remain at the potential of point C until conduction is re-established.

From this it can be seen that the upward modulation of the signal transmitted through the noise limiter circuit is limited to about 40 percent modulation (2 volts peak audio signal with respect to -5 volts dc bias). Downward modulation is not affected by the circuit.

The amount of noise energy remaining in the output of the series diode noise limiter circuit is further reduced by the introduction of noise pulses of about equal amplitude but of opposite polarity to cause noise cancellation. These pulses are generated by the noise cancellation detector, V9B, 1/2 Type 6SN7GT. This tube operates as an infinite-impedance type detector in which the IF signal is applied to the

grid of the tube and the detected output appears at the cathode. In this application, the tube is cathode biased by means of voltage divider R-35, with R-32 and R-33 in series, to about plus 15 volts so that only signals exceeding about 100 percent upward modulation will be detected. Noise pulses will appear as positive pulses in the cathode circuit and are coupled into the output of the series diode noise limiter circuit by C-55 and R-41. The values of R-32 and R-33 have been chosen so that the amplitude of the cancelling pulses will approximately equal that of the original pulses.

A shunt-type noise limiter has also been included in this receiver to protect avc and squelch circuits from the effects of noise pulses. In this circuit, diode element V-11A, 1/2 Type 6AL5, is biased by means of voltage divider R-36 and R-37 to about minus 16 volts. When an IF signal of peak amplitude greater than minus 16 volts is applied to this circuit, the diode conducts and effectively connects a very low impedance across the diode load resistors, R-38 and R-39, limiting the IF voltage that appears across T-5. This reduces the effect of noise impulses which ordinarily would generate appreciable avc control voltage, thus reducing the sensitivity of the receiver to desired signals. Since the silencer control voltage is also developed by the avc detector-amplifier, this limiter reduces the tendency of the silencer circuit to open in the presence of noise impulses.

H. AVC detector amplifier.

The avc detector-amplifier circuit (Fig. 11-6) employs V-10, a Type 6AV6 dual-diode-triode, in a circuit which produces a delayed and amplified avc control voltage which maintains the output of the receiver very nearly constant over a wide range of input signal levels.

The avc detector consists of "A" diode section of V-10, load resistors R-60 and R-61 in series, and coupling capacitor C-71. Part of the output of this detector is applied to the grid of the triode section which functions as a cathode loaded dc amplifier; its load consists of R-63 and R-64 in series. The dc output of this amplifier is coupled through diode section

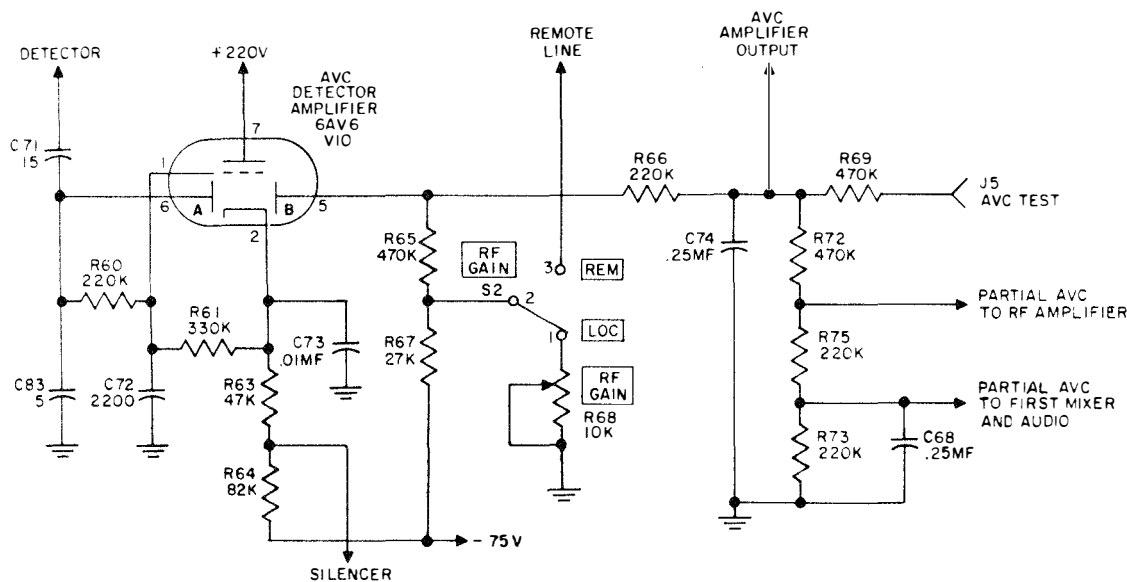


Figure 11-6. AVC Detector Amplifier

"B" to the avc time constant circuit R-66 and C-74. The dc amplifier also provides a control voltage for the operation of the silencer circuit, V-12A; this voltage is taken from the junction of R-63 and R-64.

The RF gain control function of the receiver is performed by the introduction of a variable amount of negative bias in series with the output of the avc detector-amplifier. This bias is obtained by means of voltage divider R-67 and R-68 (RF gain control) and is coupled to the avc time constant circuit by R-65. Switch S-2, the RF GAIN switch, is provided to permit the connection of an RF gain control potentiometer external to the receiver.

The operation of this circuit is best understood by considering the action which takes place as a signal of increasing amplitude is applied to the receiver. With no signal applied, the triode section of V-10 is fully conducting and the cathode assumes a potential of about plus 40 volts with respect to ground. Because of the unidirectional characteristic of the output diode, (plate 5), no current flows through it and the avc bias voltage remains at the level determined by the setting of the RF GAIN control, R-68. As signal is applied, the detector circuit develops a negative bias voltage with respect to the cathode; part of this voltage is applied to the grid of the triode, reducing the flow of current through the tube. When sufficient signal is applied, the current

of the triode section is reduced so that the cathode becomes negative with respect to the plate of the output diode, (plate 5), and current flows through the output diode, causing the generation of negative avc bias in the output circuits. This bias voltage is applied to the grid circuits of V-3 and V-5 to control their amplification thus maintaining the output of the receiver very nearly constant for a wide range of signal input level. A portion of the avc voltage is also applied to V-1, V-2 and V-12B.

I. First audio amplifier.

Referring to Fig. 11-7, the audio output of the noise limiter is coupled to the "AF GAIN" control potentiometer, R-43, by C-60. From the gain control, the audio signal proceeds to the grid of the first audio amplifier, V-12B, a 1/2 Type 6SN7GT. This amplifier operates as a conventional resistance-coupled amplifier. A small portion of the avc bias voltage is applied to the grid of this tube, connection being made through R-43.

The output of the first audio amplifier is transmitted through a low-pass filter consisting of L-13, C-63 and C-65. The function of this filter is to attenuate all signals above 4,000 cps. The attenuation of audio-frequency signals above 4,000 cps makes the normal receiver noise less annoying by filtering out the high-

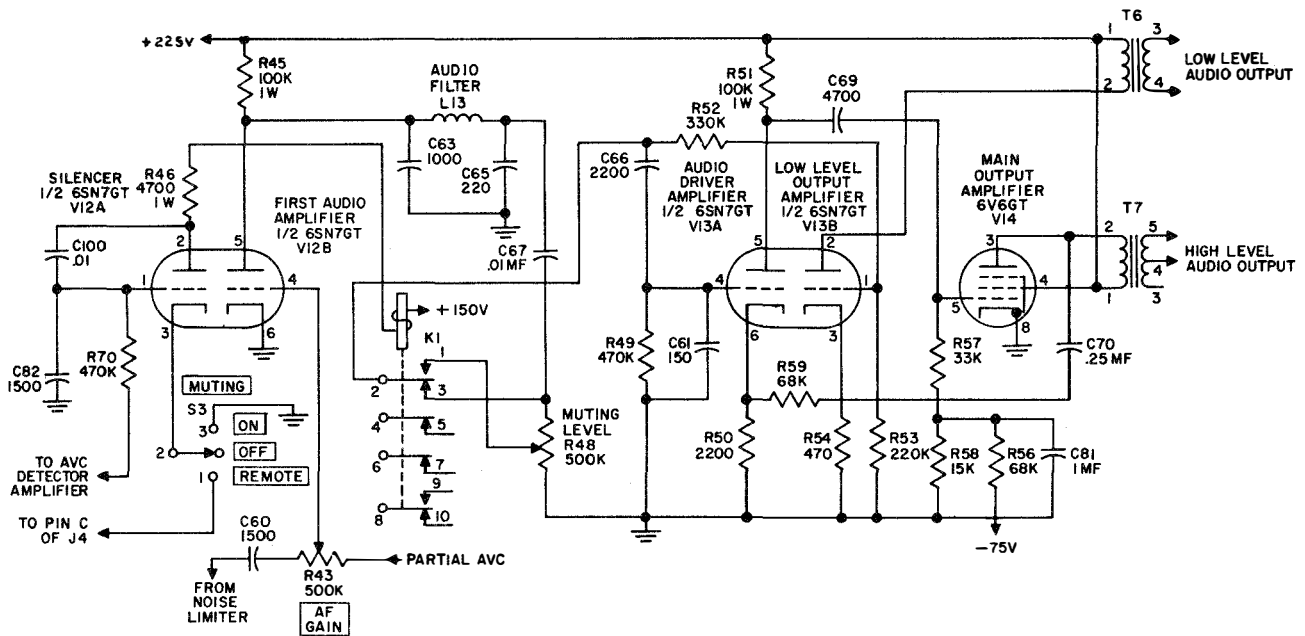


Figure 11-7. Silencer and Audio Amplifier Circuits

frequency energy of the noise pulses. The output of the filter is coupled to the "MUTING LEVEL" control and the silencer (squelch) relay by C-67.

J. Silencer.

The silencer, or squelch circuit, of the receiver (Fig. 11-7) includes relay K-1, V-12A, a 1/2 Type 6SN7GT, and muting level control potentiometer, R-48. The contacts of relay K-1 consist of one set of single-pole, double-throw contacts of the audio silencer, one set of single-pole double-throw contacts provided for operation of external equipment, (this set of contacts is single-pole, single-throw in the RUQ and RVA) and two sets of single-pole, single-throw contacts, one set for operation of the "CARRIER" indicator lamp, I-1, the other set for operation of external equipment. (Note: Terminal "M" on J-4 is ground in the RUQ and RVA). This relay is actuated by V-12A which is in turn controlled by a signal developed in the avc amplifier circuit.

When no carrier is being received, V-12A operates at zero bias, and sufficient current flows through the tube to cause the relay armature to operate when S-3 is on. This opens the normally-closed contacts and couples the grids of V-13A and B to the tap of the MUTING LEVEL control, R-48. This control is adjustable from the

rear of the chassis to provide any degree of audio silencing desired.

When a carrier of sufficient amplitude (as determined by the setting of the RF GAIN control) is received, a negative voltage developed in the avc amplifier cathode circuit is applied to the grid of V-12A, cutting off the flow of current through the tube and causing the relay armature to drop out. This causes the carrier indicator and external circuit contacts to close and transfers the connection of the audio amplifier grids so that full audio output is obtained. The MUTING switch S-3 is used to disable the silencer circuit, when required, by opening the cathode of V-12A, or for 40-percent remote control of the silencer circuit. Note: This remote control of the silencer circuit is found only on the RV-6 receiver.

K. Audio output amplifier.

The audio-output amplifier section of the receiver includes a low-level output amplifier, V-13B, audio-driver amplifier, V-13A, and main output amplifier V-14 (Fig. 11-7).

The student's attention is invited to a statement in the Technical Summary for the RUQ, RVA, RV-1, RV-3, RV-4 and RV-6 receivers which says in part, "Main audio output 1.0 watt into 20,000 ohm load impedance,"

implying thereby that the output tube plate impedance is matched to a 20,000 ohm load. However, a comparison of the specifications for the particular output transformer in each case reveals that the output tube plate for one model of receiver may be matched to 5,000 ohms, while in another model it may be matched to 20,000 ohms.

Tests made on available RUQ, RVA and RV-6 receivers indicate that the original equipment transformer for each model matched the tube to 5,000 ohms impedance, which is at variance with the transformer specifications for the RV-6.

The above is for information only and does not imply that a 20,000 ohm load should not be used.

The low-level output amplifier consists of a triode, V-13B, 1/2 Type 6SN7GT, operating in conjunction with audio-output transformer T-6 (T-2 in RUQ and RVA) to provide a low-level, audio signal for the operation of headphones and for transmission over telephone lines. Grid circuit voltage divider R-52, R-53, has been designed to provide a maximum output of approximately 10 milliwatts into 600 ohms at J-3. Cathode bias is provided by R-54.

Audio-driver amplifier, V-13A, a 1/2 Type 6SN7GT, operates as a conventional resistance-coupled amplifier to provide a signal for operation of the main output amplifier. Cathode bias is provided by R-50.

The output of the audio-driver amplifier is coupled to the grid of the main output amplifier, V-14, a Type 6V6GT. This tube operates as the audio output stage and drives the loudspeaker. V-14 operates with fixed bias obtained from the bias supply through voltage divider R-56, R-58. The plate of V-14 is coupled back to the cathode of V-13A through C-70 and R-59 to provide about 12 db inverse feedback for transformer protection and to improve the regulation of the output voltage with varying output load. The output circuits are so designed that the output voltage will not vary by more than 5 db for a 5 to 1 reduction in load resistance.

L. Power and bias supply.

(Fig. 11-8) The power and bias supply includes power transformer T-8 (T-1 RUQ and RVA), power rectifier V-15, and bias rectifier V-16.

Power rectifier V-15, a Type 5Y3GT, functions as a conventional full-wave rectifier and operates with a capacitor input filter, C-78,

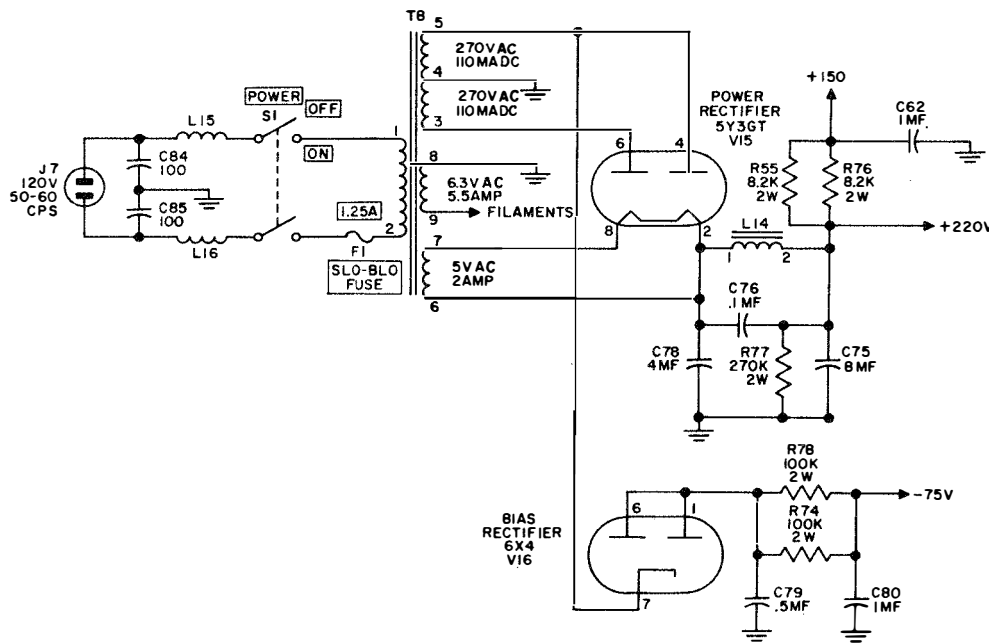


Figure 11-8. High Voltage and Bias Supply Circuits

L-14, C-76 and C-75, to provide a filtered high-voltage supply of approximately 220 volts. Voltage-dropping resistors R-76 and R-55 reduce the voltage applied to certain tubes, and C-62 provides additional filtering for operation of the RF amplifier, first mixer, and oscillator-multiplier sections.

Bias rectifier, V-16, a Type 6X4, functions as a half-wave rectifier and operates with resistance-capacitance filter C-79, R-74, R-78 and C-80, to provide a filtered bias supply of about 75 volts negative for operation of the AVC circuit and the fixed-bias audio-amplifier circuit.

Power input to the receiver is controlled by the POWER ON-OFF switch, S-1; the receiver is fused by a 1.25 ampere slow operating fuse, F-1 (This is a 1 amp fuse in the RUQ and is Slo-Blo). POWER indicator lamp, I-2, lights when power is applied to the receiver; the brightness of this lamp and the CARRIER indicator lamp, I-1, can be varied by means of the DIMMER control, R-71.

11-6. OPERATING CONTROLS

All controls necessary for operation of the receivers are located on the front panel. Functions of the various controls and indicators are described in the following paragraphs.

The POWER ON-OFF switch is a double-pole, single-throw switch controlling the application of power to the receiver.

The POWER lamp indicates when power is applied to the receiver.

The RF GAIN control is a continuously variable control which regulates the RF gain and hence the squelch-opening level of the receiver. By means of this control, the squelch opening level can be adjusted from approximately 2 microvolts at maximum clockwise rotation of the control (dial reading 10) to approximately 200 microvolts at maximum counterclockwise rotation (dial reading 0).

The RF GAIN switch determines whether the receiver is controlled by the local RF GAIN control immediately above the switch (switch set to LOC) or by a similar remotely located control (switch set to REM).

The AF GAIN control regulates the audio output of the receiver in conventional fashion. Maximum output is obtained at maximum

clockwise rotation of the control (dial reading 10).

The DIMMER control regulates the brightness of the POWER and CARRIER indicator lamps.

The CARRIER indicator lamp lights when the squelch circuit of the receiver opens. Its principal function is to indicate which frequency is being called when several receivers are being operated simultaneously.

The MUTING ON-OFF-REMOTE switch permits the muting (squelch) circuit to be cut out. In the ON position, the squelch circuit operates normally.

11-7. OPERATING PROCEDURE

Since this is a fixed-tuned receiver, its operation is comparatively simple. The only controls normally requiring adjustment are the RF GAIN and AF GAIN controls.

With the receiver installed and in operating condition, the AF GAIN control should be adjusted for desired audio output.

The RF GAIN control should be advanced to the point where ambient noise at the location will open the squelch circuit, then reduced to the point where the noise will not open the squelch. This will insure operation of the receiver at maximum sensitivity and still allow the muting circuit to silence the receiver between transmissions. All operations personnel should be instructed as to the proper adjustment of the RF GAIN control, because improper adjustment may result in missed calls.

The DIMMER control should be set to provide the desired brightness of the indicator lamps. The life of the pilot lamps will be increased if they are not operated continuously at maximum brilliance.

11-8. SUMMARY

The greater part of this chapter has been concerned with circuit theory. This theory being presented so that it will apply equally well to any one of these receivers that the student may be confronted with.

Further information such as alignment procedures, corrective maintenance data and installation notes should be taken from the instruction book on the particular receiver being worked on in the field. However, align-

ment and test procedures for the RUQ/RVA receivers used in the laboratory experiment will be adequately covered by the laboratory experiment sheets themselves.

11-9. TECHNICAL SUMMARY

The following information applies to type RUQ, RVA, RV-1, RV-3, RV-4, and RV-6 VHF fixed tuned receivers. Where there is a difference in the receivers, this difference is shown.

A. Frequency Range.

1. RUQ: 118-136 mc.
2. RVA, RV-1, RV-3, RV-4, RV-6: 108-152 mc.

B. Number of Pre-set Frequencies.

1. One.

C. Type of Frequency Control.

1. Quartz crystal.

D. Type of Receiver.

1. Double - conversion superheterodyne with automatic noise limiter, delayed and amplified automatic volume control, and carrier-operated squelch circuits.

E. Intermediate Frequencies.

1. RVA: First IF, 18.3 mc; Second IF, 3.2 mc. When operating this receiver on the 152 mc channel, the first IF is adjusted for 18.2 mc; and the second IF is adjusted for 3.1 mc.
2. RUQ, RV-1: First IF, 18.3 mc; Second IF, 3.2 mc.
3. RV-3, RV-4, RV-6: First IF, 18.3 mc; Second IF, 1.85 mc.

F. Receiver Output.

1. RUQ: Main audio output; 1.0 watt into 20,000-ohm load. Low level output; 10 milliwatts into 600-ohm load.
2. RVA, RV-1, RV-3, RV-4, RV-6: Main audio output; 1.0 watt into 20,000-ohm load (or 600 ohms by changing tap). Low level output; 10 milliwatts into 600-ohm load.

G. Type of Reception.

1. A-3 (Amplitude-modulation radiotelephone).

H. Crystals.

1. RUQ: channel crystal, type CR-18/U, (6.25 to 7.35 mc). Frequency dependent on channel frequency. Mixer crystal, type CR-18/U, (7.550 mc).
2. RVA, RV-1: channel crystal, type CA-1353, (5.60 to 8.35 mc). Frequency dependent on channel frequency. Mixer crystal, type CA-1353, (7.550 mc).
3. RV-3, RV-4, RV-6: channel crystal, type CR-27/U, (5.60 to 8.35 mc). Frequency dependent on channel frequency. Mixer crystal, type CR-18/U, (8.225 mc).

I. Frequency Stability.

1. RUQ, RVA, RV-1: over-all stability 0.005% from -10° C to $+60^{\circ}$ C ambient temperature.
2. RV-3, RV-4, RV-6: over-all stability 0.002% from -10° C to $+60^{\circ}$ C ambient temperature.

J. Receiver Sensitivity.

1. RUQ, RV-1, RV-3, RV-4, RV-6: at maximum setting of the RF GAIN control, a 5 microvolt signal, 30% amplitude modulated, will produce not less than 0.1 watt output at not less than 10 db signal-to-noise ratio.
2. RVA: at maximum setting of the RF GAIN control, a 2.5 microvolt signal, 30% amplitude modulated, will produce not less than 0.1 watt output at not less than 10 db signal-to-noise ratio.

K. Squelch Circuit.

1. RUQ, RVA: squelch threshold at maximum RF gain, approximately 2 microvolts; at minimum RF gain, approximately 100 microvolts. Muting adjustable from 0 to 30 db. Pro-

visions have been made for operation with a remote RF GAIN control.

2. RV-1, RV-3, RV-4, RV-6: squelch threshold at maximum RF gain, approximately 1 microvolt; at minimum RF gain, approximately 200 microvolts. Muting adjustable from 0 to 40 db. Provisions have been made for operation with a remote RF GAIN control.

L. Recommended Antenna.

1. Quarter-wave VHF ground plane or coaxial type antenna having 52-ohm output impedance. Receiver input circuit is designed for operation of two receivers from a single antenna. In this case, the operating frequencies of the receivers should differ by at least 1.0 mc and not more than 10 mc.

M. Receiver Selectivity.

1. RUQ, RVA, RV-1: bandwidth not less than 50 kc at 6 db attenuation; not more than 200 kc at 60 db attenuation.
2. RV-3, RV-4, RV-6: bandwidth not less than 36 kc at 6 db attenuation; not more than 108 kc at 60 db attenuation.

N. AVC Characteristic.

1. RUQ: audio output constant within 3 db for signal input from 10 microvolts to 1.0 volt.
2. RVA: audio output constant within 3 db for signal input from 5 microvolts to 0.5 volt.
3. RV-1, RV-3, RV-4, RV-6: audio output constant within 3 db for signal input from 15 microvolts to 1.0 volt

O. Power Requirements.

1. RUQ, RVA: 120 volts, 50 or 60 cycles AC, 90 watts. Receiver will operate satisfactorily with the line voltage from 105 to 130 volts.
2. RV-1, RV-3: 120 volts, 50 or 60 cycles AC, 100 watts. Receiver will operate satisfactorily with the line voltage from 105 to 130 volts.
3. RV-4: 120 volts, 50 or 60 cycles AC, 84 watts. Receiver will operate satisfactorily with the line voltage from 105 to 130 volts.
4. RV-6: 120 volts, 50 or 60 cycles AC, 96 watts. Receiver will operate satisfactorily with the line voltage from 105 to 130 volts.

P. Electron Tube Complement.

1. RUQ, RVA: 4 type 6AK5 pentode; 2 type 6BA6 pentode; 1 type 6AU6 pentode; 1 type 6J6 dual-triode; 3 type 6SN7GT dual-triode; 1 type 6AL5 dual-diode; 1 type 6AQ6 dual-diode triode; 1 type 6V6GT beam pentode; 1 type 6X4 rectifier; 1 type 5Y3GT rectifier.
2. RV-1, RV-3, RV-4, RV-6: 4 type 6AK5 pentode; 2 type 6BA6 pentode; 1 type 6AU6 pentode; 1 type 6J6 dual-triode; 3 type 6SN7GT dual-triode; 1 type 6AL5 dual-diode; 1 type 6AV6 dual-diode triode; 1 type 6V6GT beam pentode; 1 type 6X4 rectifier; 1 type 5Y3GT rectifier.

Q. Fuses.

1. RUQ: 1 3AG, 1 ampere, Slo-Blo.
2. RVA, RV-1, RV-3: 1 3AG, 1.25 ampere, Slo-Blo.
3. RV-4, RV-6: 1 MDL, 1.25 ampere, Slo-Blo.

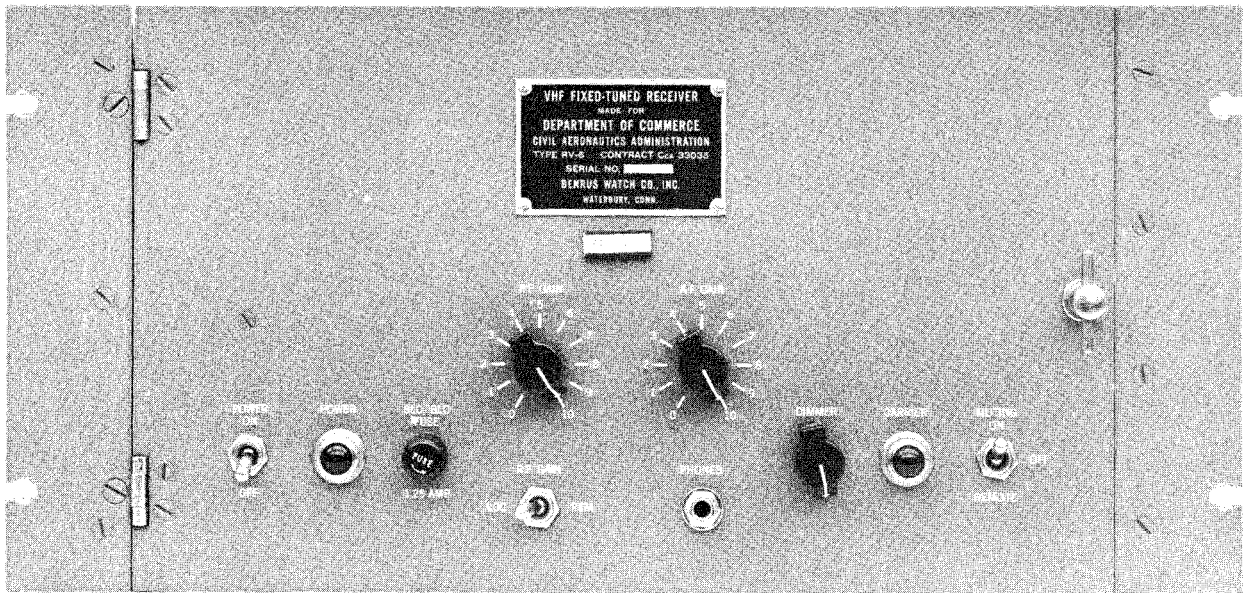


Figure 11-9. Type RV-6 Receiver, Front

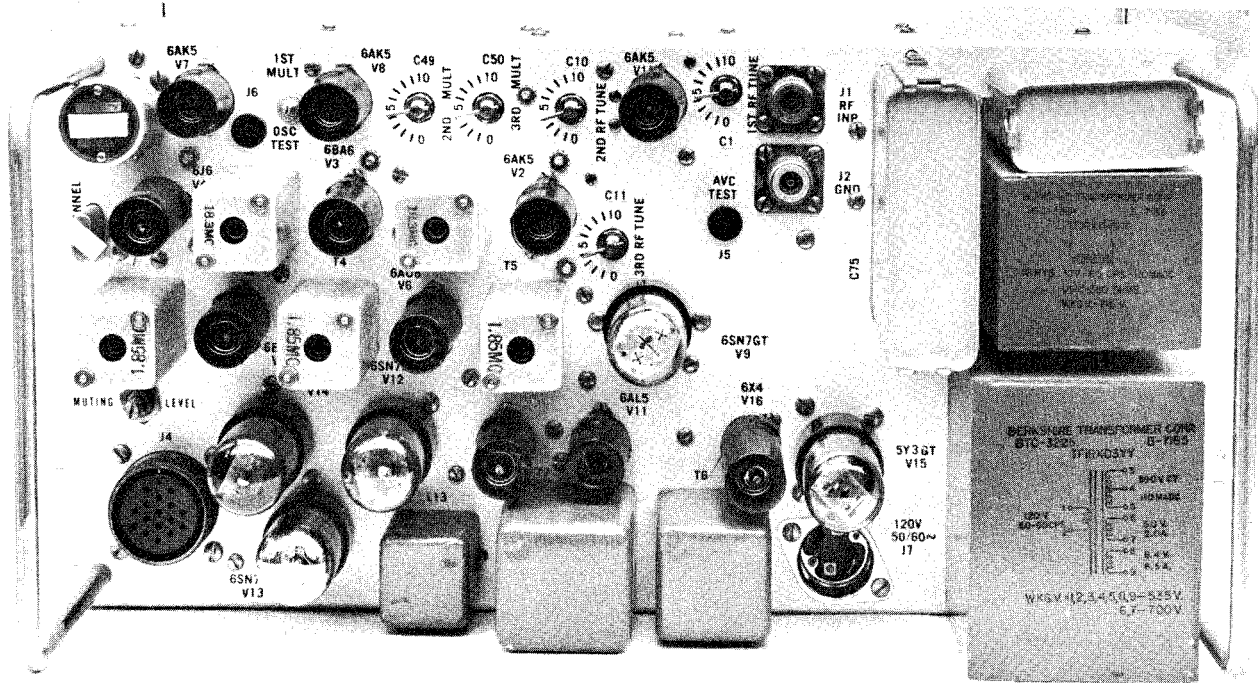
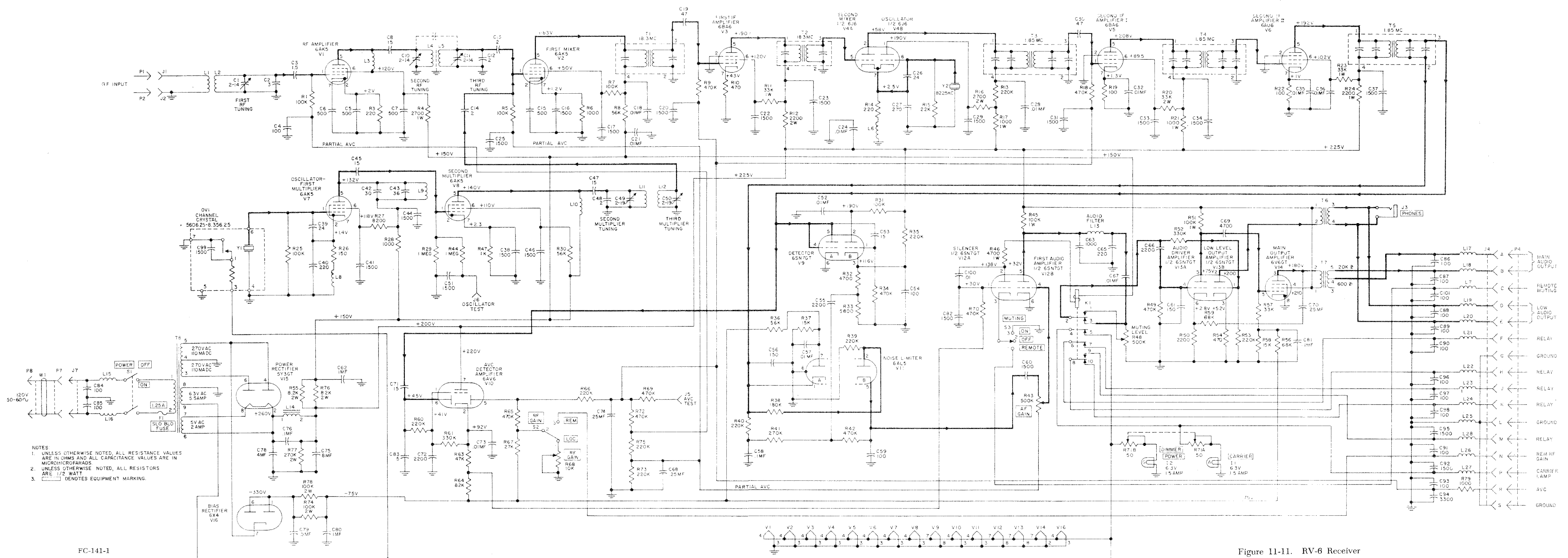


Figure 11-10. Type RV-6 Receiver, Rear



NOTES
 1. UNLESS OTHERWISE NOTED, ALL RESISTANCE VALUES ARE IN OHMS AND ALL CAPACITANCE VALUES ARE IN MICROMICROFARADS.
 2. UNLESS OTHERWISE NOTED, ALL RESISTORS ARE 1/2 WATT.
 3. [Symbol] DENOTES EQUIPMENT MARKING.

Figure 11-11. RV-6 Receiver

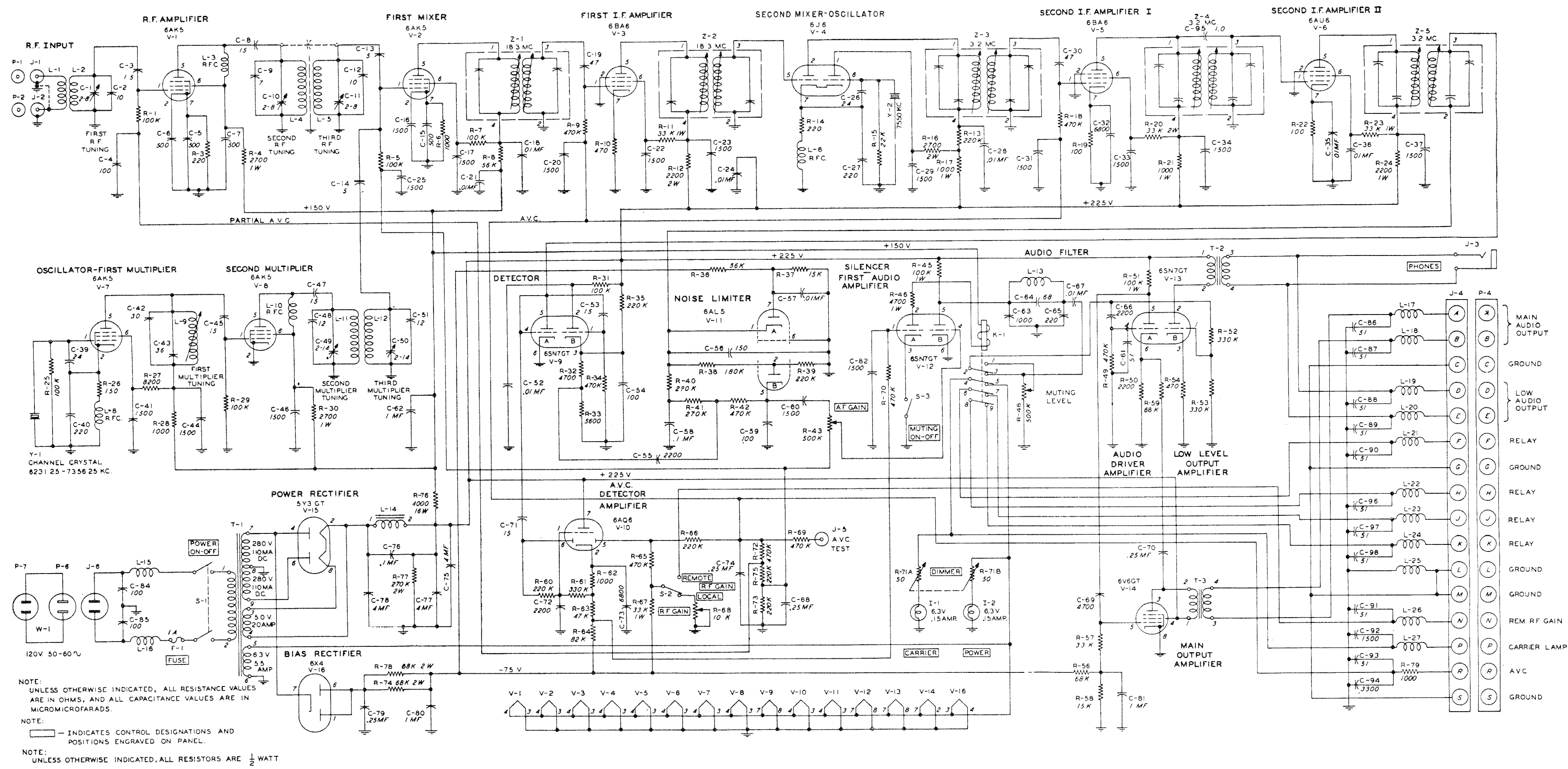
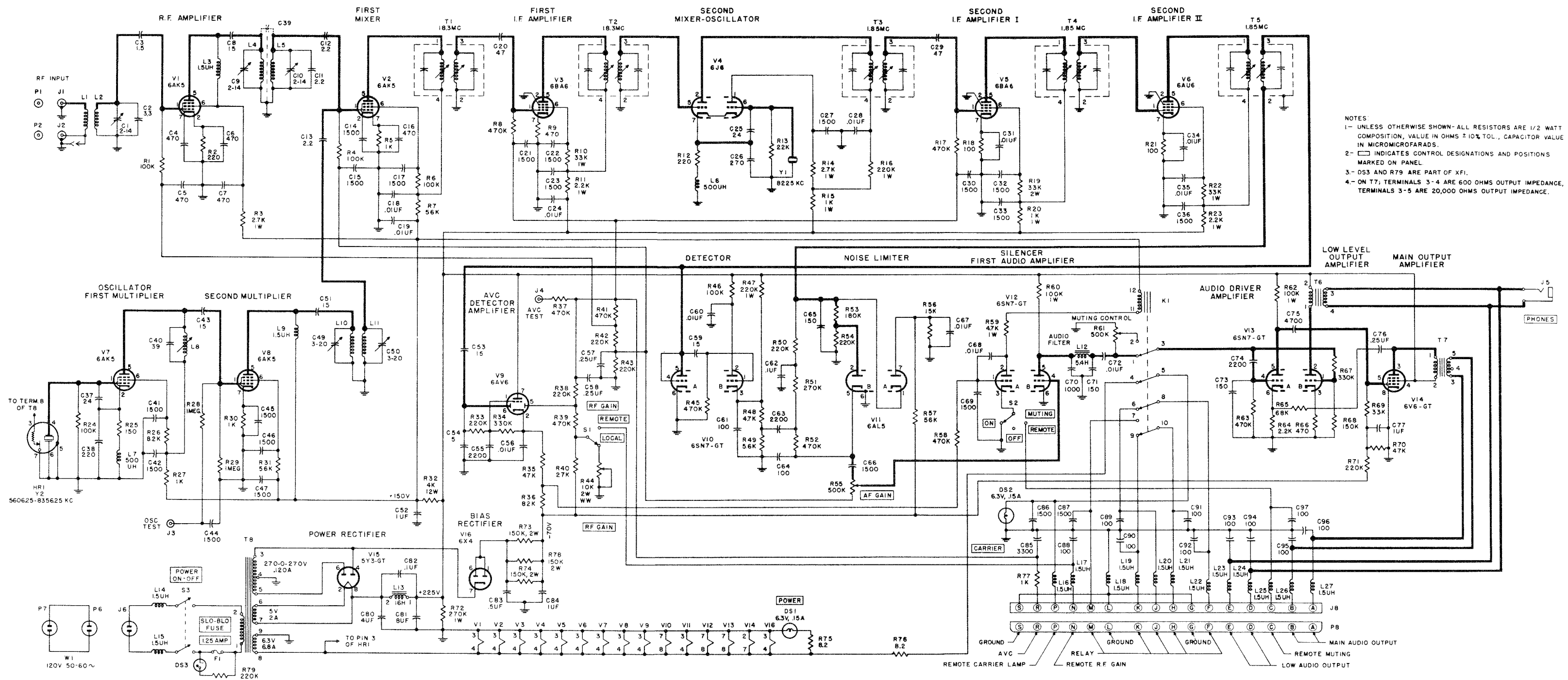


Figure 11-12. RUQ Receiver



NOTES:
 1- UNLESS OTHERWISE SHOWN-ALL RESISTORS ARE 1/2 WATT COMPOSITION, VALUE IN OHMS ± 10% TOL., CAPACITOR VALUE IN MICROMICROFARADS.
 2- □ INDICATES CONTROL DESIGNATIONS AND POSITIONS MARKED ON PANEL.
 3- DS3 AND R79 ARE PART OF XF1.
 4- ON T7, TERMINALS 3-4 ARE 600 OHMS OUTPUT IMPEDANCE, TERMINALS 3-5 ARE 20,000 OHMS OUTPUT IMPEDANCE.

CHAPTER 12

RCP RECEIVER

12-1. INTRODUCTION

The RCP receiver is designed for the reception of aural radio range, radio telephone and continuous wave signals. This receiver was originally a type RCK or RCL receiver. The type designation was changed to RCP at the time certain modifications were made to distinguish it from its original type.

A. *Technical Summary of Receiver Characteristics*

1. FREQUENCY RANGE: 200 to 400 kc and 1.3 to 30 mc.
2. INTERMEDIATE FREQUENCY: 455 kc.
3. OPERATING POWER: Approx. 70 w at 105 to 125 v, 50 to 60 cycles.
4. RF INPUT IMPEDANCE: 7 to 238 ohms.
5. OUTPUT IMPEDANCES: 20,000 ohms and 600 ohms
6. UNDISTORTED POWER PUTOUT: 1.5 watts.
7. CONS: Allows the receiver to operate at 5 μv or less.
8. SENSITIVITY: 5 μv or less for 1 watt, with signal-to-noise ratio of 13 db or better.
9. SELECTIVITY: Bandwidth at 6 db, 4 kc minimum; 20 db, 12 kc maximum; 40 db, 22 kc maximum; 60 db, 30 kc maximum.
10. AVC: +4 db and -0.5 db with input 10 to 10,000 μv .

12-2. THEORY OF OPERATION

As shown in the circuit diagram Figure 12-4, the receiver is a twelve tube superheterodyne covering the frequencies from 200 to 400 kc in one range and from 1.3 to 30 Mc in four ranges. The circuit employed on all

ranges consists of one RF amplifier stage, first detector, separate high frequency oscillator, two IF amplifier stages, a linear diode detector, a series noise limiter, a one-stage resistance coupled audio amplifier with a beam-power output stage. Maximum audio output is at least 1.5 watts. Two separate tubes are employed to provide amplified and delayed AVC action. A separate beat frequency oscillator is coupled to the second detector for CW reception. A carrier operated noise suppressor (or squelch) circuit is provided with an adjustment for controlling the muted level. A built-in power supply provides all voltages required. The receiver may be used with either a loudspeaker, headphones or a telephone line circuit.

A. *RF Amplifier and Oscillator*

The radio frequency amplifier and conversion oscillator circuits of the RCP Receiver have several features which deserve mention here. All tuned circuits are designed to have high Q and to be essentially free from the detuning effects of changing temperature and humidity. These characteristics are obtained by winding the coils on low loss forms of ceramic, or polystyrene material, by employing air dielectric trimmer capacitors in all circuits, and by incorporating temperature compensation in the conversion oscillator circuit.

The local oscillator may be operated as a variable frequency oscillator, or as a highly stable fixed frequency crystal-controlled unit by placing the TUNABLE-XTAL panel switch A-32 in the proper position. In either case the oscillator operates 455 kc above the incoming frequency. The receiver tuning dial should

be set at approximately the same position for XTAL or TUNABLE operation on the same frequency.

It is the function of the RF amplifier tube V-1 to amplify any RF signal to which the receiver is tuned. The amplified signal reaches the first detector V-2 where it combines with the output of the high frequency oscillator V-3 to produce a new signal of 455 kc, which is the intermediate frequency of the receiver. This 455 kc signal is amplified by the two IF stages V-4 and V-5, and is rectified or demodulated by the second detector V-6. The audio component passes through the series noise limiter, audio amplifiers V-7B and V-8, and to the speaker or phones. In the case of CW reception, a radio frequency signal is produced by a separate CW oscillator V-9, which is coupled by capacitor C-22 to the plate of the second detector V-6. The frequency of the audio signal is the difference between the IF signal and the CW oscillator signal.

B. Series Noise Limiter

A noise limiter modification has been incorporated in the RCP receiver. To accomplish this noise limiting, a type 6H6 duo-diode with circuit modifications was installed to perform the dual function of second detector and series noise limiter. Figure 12-2 shows the modified second detector circuit and the new limiter circuit in relation to the CONS and first audio stages.

The diode second detector circuit is conventional, using C-27, R-25 and C-25 as the RF bypass filter circuit. R-57 and R-56 serve as audio load and limiter tube cathode bias resistors, respectively. C-59 is an audio bypass capacitor to prevent audio from getting to the grid of the first audio amplifier through R-59 and R-58. The cathode of the N. L. portion of the 6H6 tube is biased negatively enough by the voltage drop across R-56 to allow the varying current, which is inversely proportional to the audio voltage impressed across the plate of the N. L. diode through capacitor C-58, to flow until the audio peaks exceed voltage representative of 30 percent modulation, at which

time this tube will be cut off from conduction. Because the tube no longer conducts after this specified audio voltage is reached, high level audio (noise) peaks will cause no plate current variation in the N. L. portion of the 6H6, and cannot, therefore, be coupled through C-28 to the grid of the first audio amplifier.

The noise limiter circuit is in series with the audio path from the diode demodulator to the audio amplifier, therefore, it is a series diode noise limiter. Because it presents a high (effectively infinite) impedance to the passage of high level noise voltages, it is a high impedance noise limiter.

C. CONS

The operation of the CONS circuit is fully explained in Chapter 5. The receiver remains silent on any pre-set frequency until an incoming signal of approximately $3 \mu\text{v}$ is received. At that time the audio stage of the receiver automatically becomes operative. With no signal being received the degree of muting or noise reduction is variable by screwdriver adjustment of R-41. The signal strength below which muting occurs is controlled by the setting of the RF GAIN control. With this control at maximum, individual receivers will respond to signals of less than $5 \mu\text{v}$. In extremely noisy areas, the setting of the RF GAIN control may have to be reduced to just below the point where the prevailing noise operates the CONS relay. Where the RCP is used as a monitor receiver, link A22 is removed and an alarm buzzer is connected in the circuit.

D. Selectivity

The selectivity characteristics of the IF amplifier depends largely upon the degree of coupling between the primary and secondary circuits of the individual IF transformers; bandwidth increases with the coefficient of coupling. The coupling of the various IF stages has been arranged to arrive at the most desirable IF selectivity curve. The tertiary windings shown on the schematic (Figure 12-4) have been removed leaving the receiver in the sharp tuned condition.

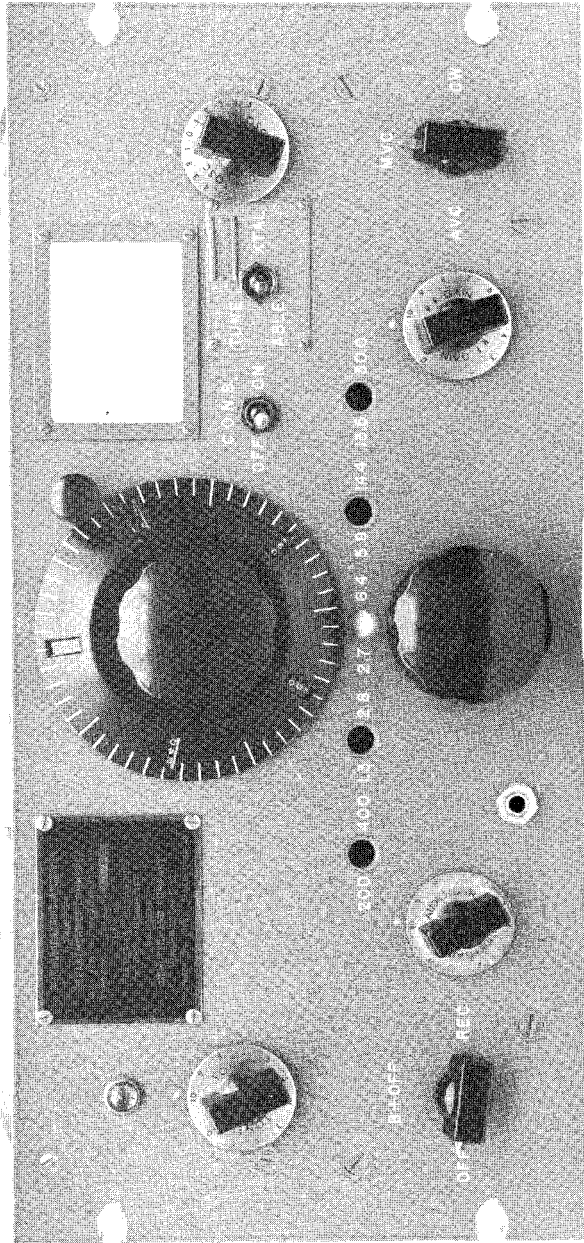


Figure 12-1. RCP Receiver - Front View

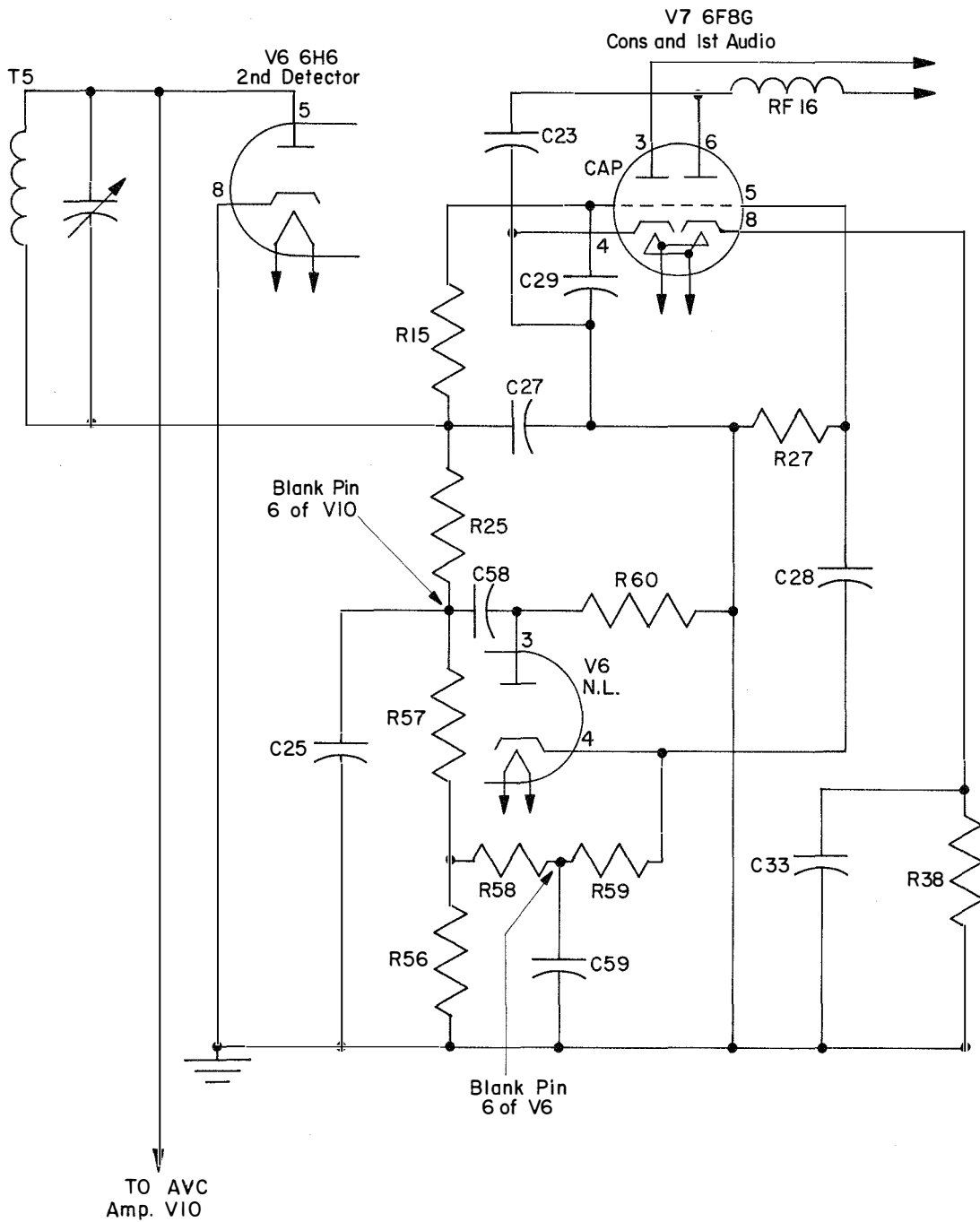


Figure 12-2. Series Noise Limiter Circuit

E. Audio

The audio amplifier circuits have been designed to give a frequency characteristic which is fairly flat from 100 to 2000 cycles and falls off slowly to give a response of more than 15 db down at 5000 cycles. With the tone control in full operation, the audio frequency response above 1000 cycles is sharply attenuated, with a maximum attenuation at 3000 cycles.

The second audio stage, V-8 using a 6V6, has several features worth mentioning. Cathode bypass capacitor C-40 was changed from 0.25 μ fd to 10 μ fd to prevent degeneration of low audio frequencies. The available audio input signal to this stage can exceed the self bias of the tube causing grid current flow. The grid current in flowing through R-47 (20,000 ohms) will limit this grid current by producing a grid leak bias across R-47. Capacitor C-39 (.0025 μ fd) is necessary to prevent a "fringe howl" in this stage. The maximum power output is at least 1.5 watts.

The output terminals on the terminal panel provide audio outputs as follows:

Between Terminals A29' and A29 (Gnd)—
20,000 ohms, full audio power, *or*

Between Terminals A30' and A29 (Gnd)—
600 ohms, full audio power, *or*

Between Terminals A30 and A29'—special high impedance output, used for lighting a neon light in A/G console installations, *and*

Between Terminals A31' and A29 (Gnd) and with the external jumper installed between A31 and A29 (Gnd)—special low impedance output, used for input to the receiver combining and muting unit of the A/G console.

If the original loudspeaker furnished with type RCK and RCL receivers is used, the input transformer of the speaker should be connected to the receiver as follows:

Terminals 3 and 5 of the speaker transformer—600 ohms

Terminals 4 and 5—20,000 ohms

A shielded lead should be used to connect the loudspeaker to the receiver output, the shield being connected to terminal A29 (Gnd).

If this receiver is used as a radio range "distant monitor" receiver, and it is desired to monitor by means of a meter as well as aurally, the monitor panel, CA-734, CA-760 or equal,

should be connected to terminals A30' and A29 (Gnd), while the 20,000-ohm output is simultaneously loaded with a speaker or other load connected across A29' and A29 (Gnd).

Normally, this type of receiver is not used by the FAA for remote receiver applications. If, however, this receiver is to be used at a remote site where the output works into a telephone company or other low level, ungrounded line, a suitable matching transformer should be mounted on a blank relay rack panel external to the receiver and connected to the receiver output. In addition, a fixed or variable T-pad attenuator should be used to achieve the necessary reduction in audio level, in order to avoid relative increase in hum level which would be caused by using the audio volume control alone to achieve the necessary low level.

F. AVC

Automatic volume control is obtained in the following manner. See Figure 12-3 for the schematic. The IF voltage across the secondary of the IF output transformer T-5 is capacity coupled to the grid of the AVC amplifier V-10 which amplifies and rectifies the IF voltage. This dc voltage appearing across the resistor R-30 is filtered with a resistance capacitance filter network, capacitor C-26 and resistor R-32, which removes the audio component. This dc voltage is applied to the grid of the AVC control tube V-11. A negative voltage, with respect to the plate of V-11, is applied to the cathode of V-11. A negative delay bias voltage developed across the delay bias resistor R-31 is applied to the grid. As the positive dc voltage from the AVC amplifier, developed across resistor R-30, reduces the effect of the delay bias voltage, the AVC control tube V-11 draws current through AVC resistor R-39. The voltage drop across R-39 is applied to the grids of V-1, V-4, and V-5 for AVC action in the receiver. Switch A-25 shorts out R-39 when AVC is not desired. The negative potential applied to the grid of V-11 is determined by the average carrier strength. The AVC therefore responds to the average carrier level, and not to the peak (noise) signal input. The circuit is so designed that for sig-

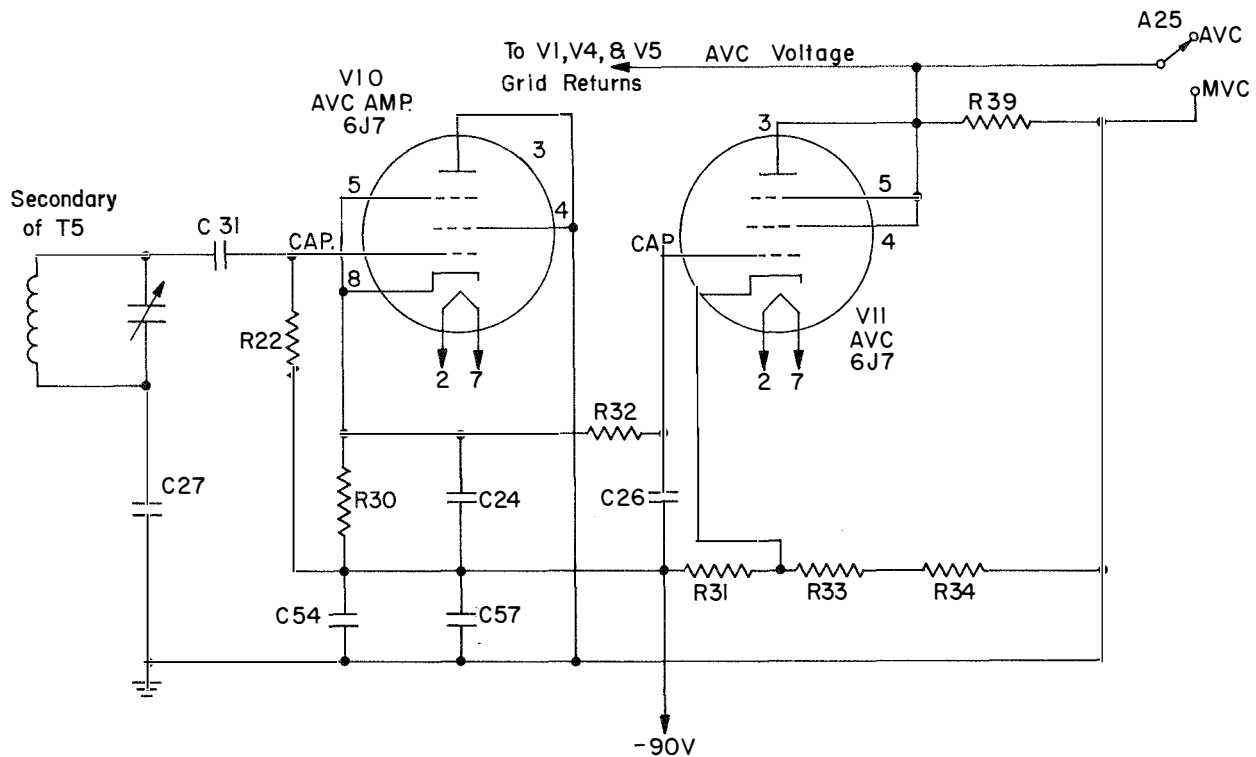


Figure 12-3. RCP Delayed AVC (Amplifier and Control)

nals of less than approximately $7 \mu\text{v}$ the AVC is essentially inoperative due to the fixed bias on V-11; hence, the term "delayed AVC."

G. Power Supply

The power supply uses a duo-diode 80 tube (V-12) as a full wave rectifier. The center tap of the power transformer T-2 is connected through R-31, R-33 and R-34 to ground. These three resistors develop a -90 volts.

The plate voltage for the RF amplifier, first detector, IF amplifiers, CONS and first audio, and both oscillators is the B+ to ground voltage. Plate voltage for the AVC tubes is ground to B-. The cathode of the second audio stage V-8 is connected to B- and the plate to B+. This applies the total or sum of the two voltages across the second audio tube. The primary of the power transformer T-2 is adjustable by A-3 to compensate for differences in line voltage.

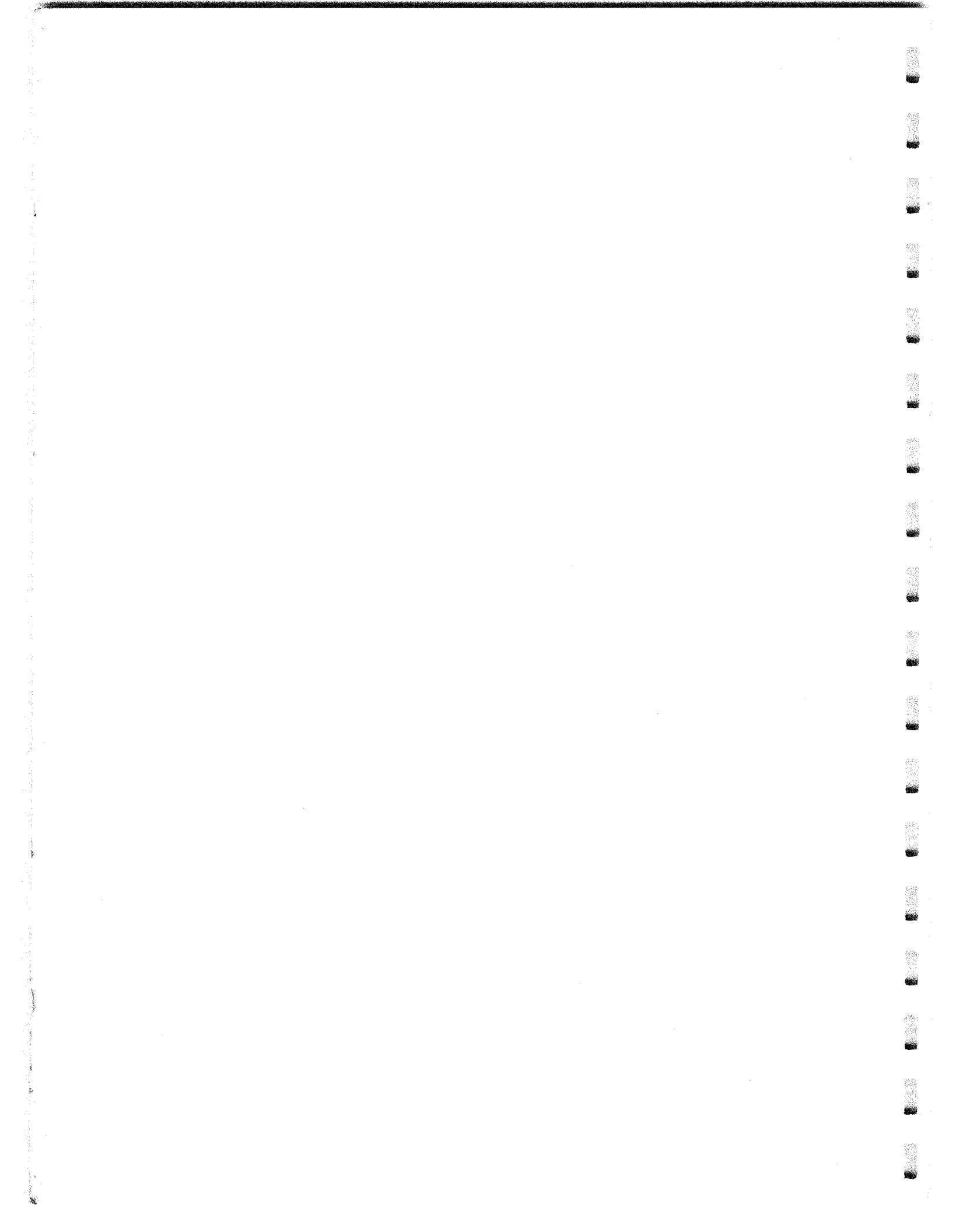
12-3. MECHANICAL DETAILS

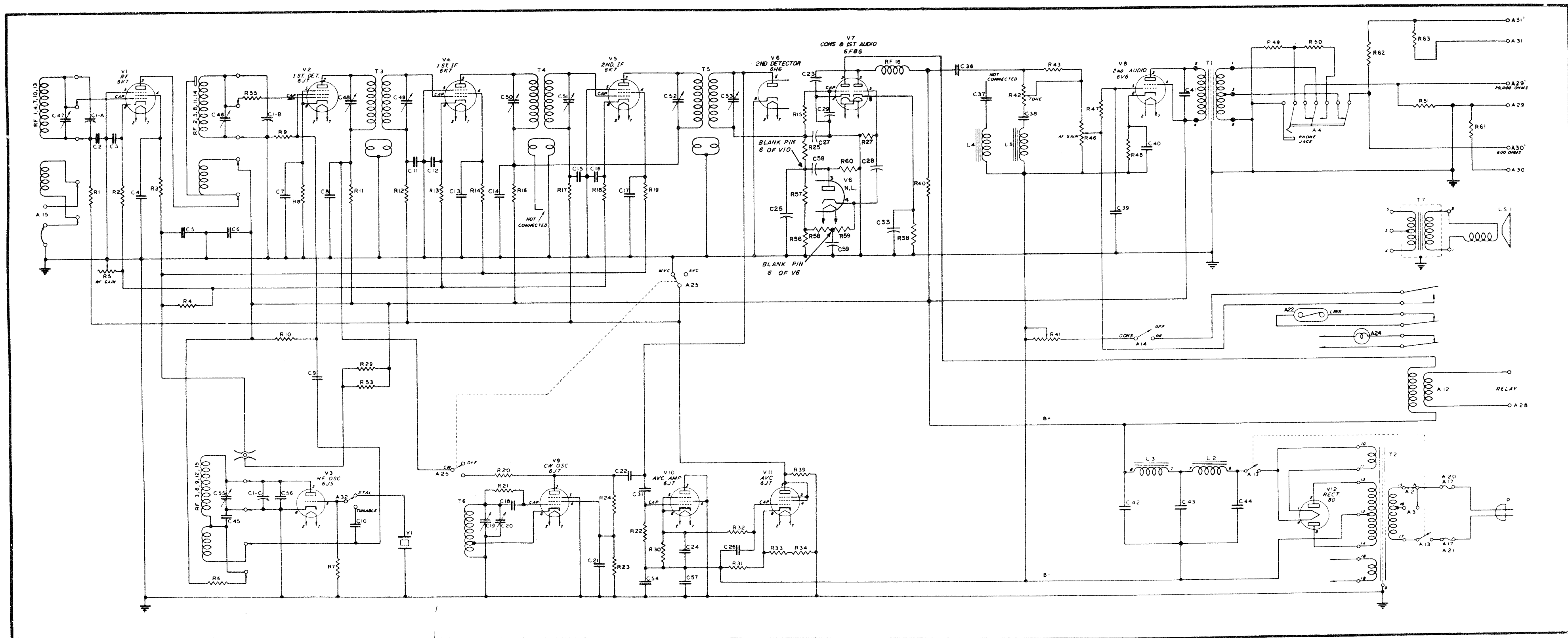
There are two unusual mechanical features which add to performance and reliability of this receiver. One of these is the tuning capacitor and dial drive mechanism which are built and mounted on the chassis as one unit. Inasmuch as the dial makes ten complete revolutions when turning the tuning capacitor through 180° , and the dial scale diameter is $4\frac{1}{2}$ inches, the effective scale length is about 12 feet. The dial scale has 500 divisions spaced $\frac{1}{4}$ inch apart. This arrangement makes tuning and logging both easy and precise, even on the highest frequencies, since the dial can be re-set with an accuracy better than one part in 2500.

Another mechanical feature is the system of automatically plugging in the RF amplifier and high frequency oscillator coils of the various ranges. All of these coils and their associated trimmer capacitors are mounted in a rigid

aluminum casting which slides the length of the chassis and is moved by the range changing knob on the front panel. The various coils are fitted with heavy contact pins which engage spring contactors mounted immediately under the variable tuning capacitor. The particular

coils in use are, therefore, in position affording the shortest leads to the tubes and variable capacitor, while at the same time they are thoroughly shielded from all other coils, and all unused coils are completely disconnected from the circuit.





NOTE: MODIFIED PER CED No. 41;
 EMBI No. 122;
 EEM No. 4 and No. 9.

Figure 12-4. RCP Receiver

CHAPTER 13

RHZ RECEIVER

13-1. GENERAL DESCRIPTION

The RHZ receiver is of the fixed-tuned, single-frequency type for ground station installation and designed to be used for the reception of radio-telephone signals from aircraft operating on a frequency to which the receiver has been pre-tuned in the range of 1600 to 10,000 kc. It employs ten tubes in a super-heterodyne circuit with a crystal-controlled High-Frequency conversion Oscillator (HFO) to provide a high degree of signal frequency stability. Automatic Volume Control (AVC), Carrier-Operated Noise Suppression (CONS), and Peak-Noise Limiter (PNL) are incorporated. Facilities are provided for operation from either a grounded or balanced antenna or a remote antenna connected to the receiver through a 70-ohm coaxial transmission line. The receiver is designed for mounting in a standard type FAA cabinet relay rack. See Figure 13-1 Front Panel. The component parts of the receiver are mounted on two vertical chassis behind the panel. See Figure 13-2 Receiver, Top View. The arrangement of components on the rear chassis is such that tube replacements and tuning adjustments may be accomplished from the rear of the receiver without necessity for removal from the rack. See Figure 13-2 Receiver—Top View.

To enable coverage at any frequency between 1600 and 10,000 kc, the proper set of coils (RF-1, RF-2, RF-3, RF-4 as shown on Figure 13-2) must be installed. Refer to Figure 13-2 and Figure 13-8 for replacement of coils and their connections to the various circuits. For the specified operating frequency, it is necessary to use a crystal of the proper frequency to mix with the operating frequency and produce the 455 kc intermediate frequency. After installation of a coil set for a new frequency, it is necessary to then perform

the alignment and adjustments as outlined in the instruction book.

The receiver is equipped with RF SENSITIVITY and AF GAIN controls. The sensitivity control varies the gain of all radio-frequency and intermediate-frequency stages simultaneously. The audio gain control permits attenuation of the 600-ohm output circuit to one-tenth milliwatt or less when the input RF voltage and the sensitivity control setting are such as to cause the AVC feature to function. Both gain controls, in the absence of AVC and CONS, produce an approximately exponential increase in audio output as the control knob is rotated clockwise. The sensitivity of the receiver, although somewhat dependent upon the operating frequency, will average approximately 150 milliwatts into a 20,000-ohm load circuit for an input signal of $2\frac{1}{2}$ microvolts modulated 30 percent at 400 cps. Under these conditions the ratio of the modulated to unmodulated carrier will be approximately 14 db. The choice of either a 600-ohm low-level line circuit or a 20,000-ohm high-level loud-speaker circuit is available upon operation of a panel switch marked SPKR-LINE. When in the line position, a 600-ohm headphone circuit is available at the panel jack marked PHONES.

13-2. Circuit Description

A. Antenna Circuit

The receiver input circuit is designed to operate from a grounded or balanced doublet antenna or a remote antenna connected to the receiver through a 70-ohm coaxial transmission line. When using either a grounded antenna or a remote antenna connected to the receiver with a 70-ohm coaxial line, the bottom terminal marked G should be directly connected to the chassis by means of the grounding strip pro-

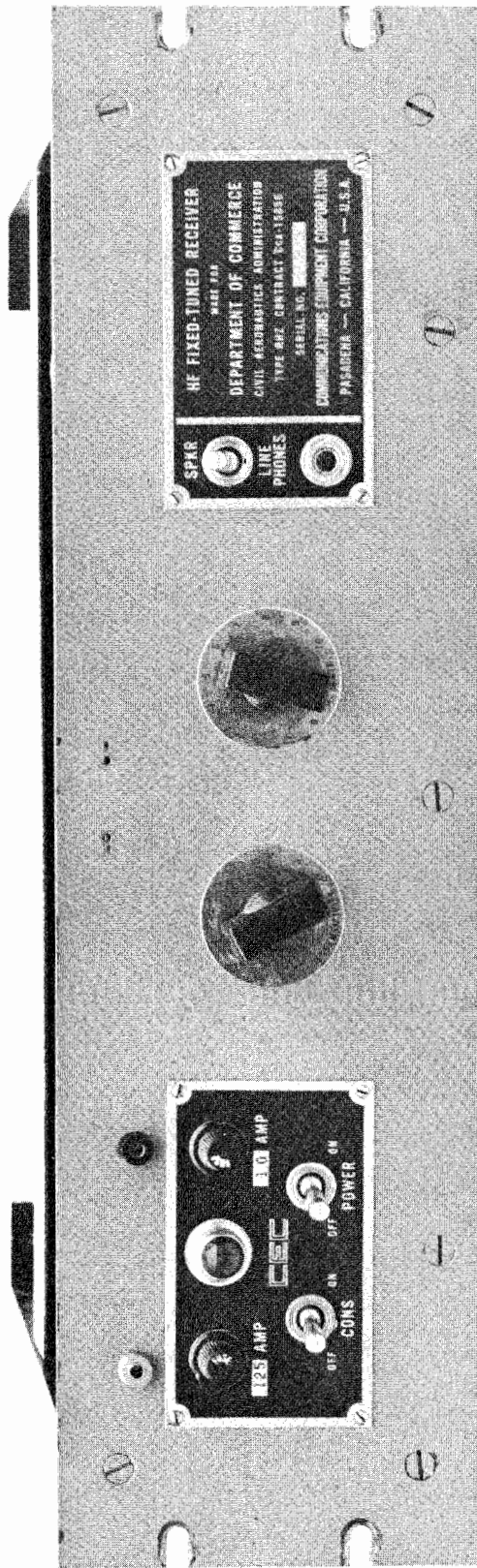


Figure 13-1. Front Panel of RHZ Receiver

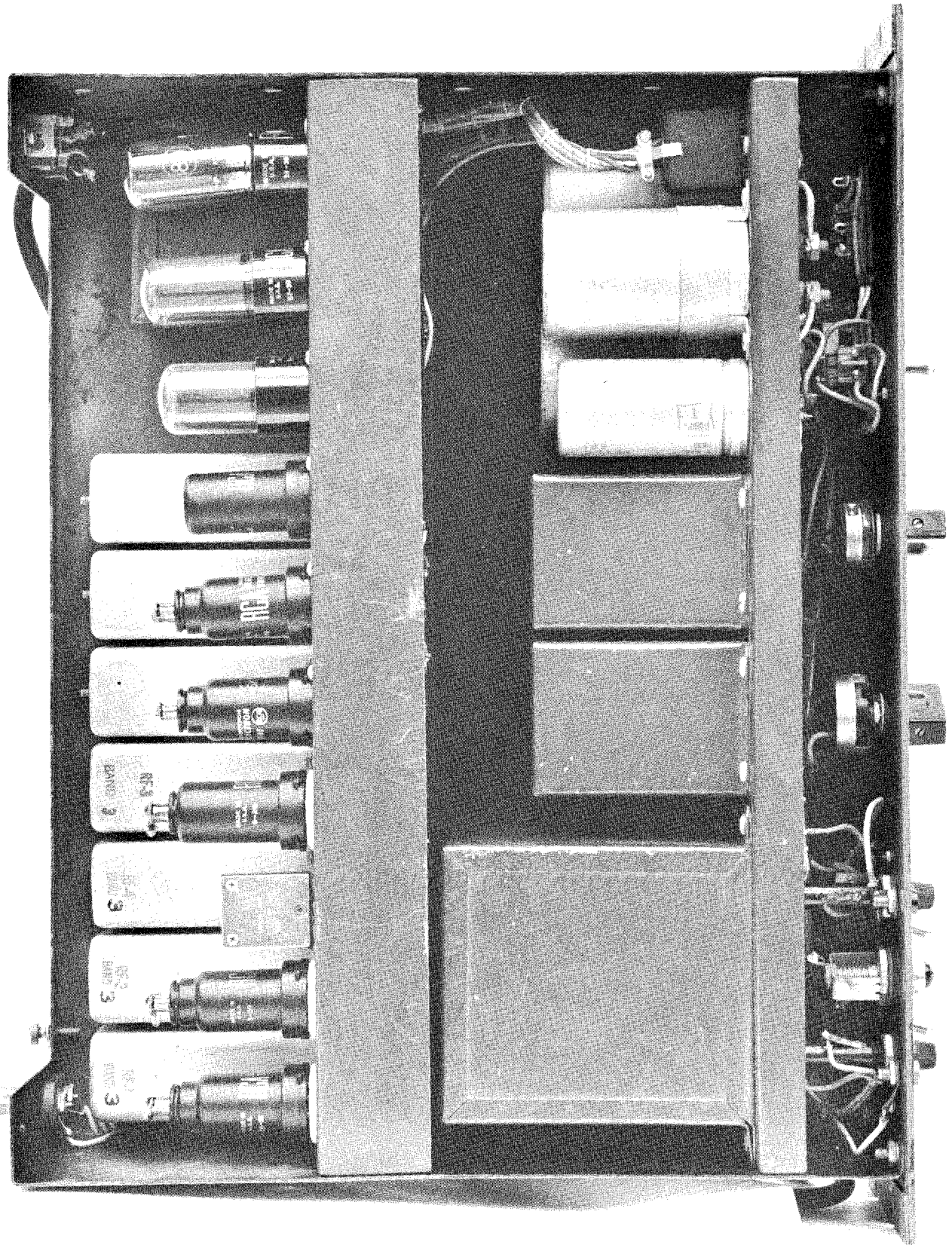


Figure 13-2. RHZ Receiver, Top View

vided, one end of which is permanently attached to the main chassis bracket.

B. Radio-Frequency Stages

Three tuned circuits, RF-1, RF-2 and RF-3, between the antenna line connections and the converter stage provide two stages of tuned radio frequency amplification for the receiver. This is a conventional preselector circuit utilizing 6K7 tubes providing a moderate amount of gain, a high signal-to-noise ratio, high image rejection ratio, and considerable attenuation of high signal levels upon the application of AVC voltages. Grid bias is provided by AVC action plus the voltage drop across the RF sensitivity control.

C. Frequency Converter Circuit

The oscillator and first detector circuits of the receiver incorporate a type 6K8 tube consisting of a triode and a hexode tube in a single envelope. This tube is designed to perform simultaneously the function of supplying a local oscillator frequency and mixing it with the signal input frequency to produce the resultant intermediate frequency.

The electrical arrangement of the 6K8 triode-hexode converter is shown in Figure 13-3.

It will be noted that the triode elements are on one side of the cathode, while the mixer

grid (an extension of the oscillator grid G_1) and other elements of the hexode are on the opposite side of the cathode. This design permits the complete isolation of the mixer plate and oscillator plate electron streams.

In normal operation the hexode section operates in much the same manner as the conventional pentagrid converter, with the voltage of the oscillator section developing on the mixer grid (G_1) with each positive pulse, and causing space current to form a "virtual cathode" on the cathode side of the signal grid (G_3). Voltage of signal frequency on the signal grid will control the number of electrons that the mixer plate (A_2) is able to draw from this "virtual cathode". Hence, the current actually arriving at the mixer plate is modulated by the oscillator and signal voltages, resulting in a difference frequency corresponding to the receiver intermediate frequency.

D. Intermediate-Frequency Amplifier

Two type 6K7 tubes are utilized in the two-stage IF system for amplification of the 455 kc component from the mixer. Coupling between the stages is accomplished by input transformer T-3, interstage transformer T-4, and output transformer T-5. Core type adjustment of transformer inductance is available for both the primary and secondary circuits of each

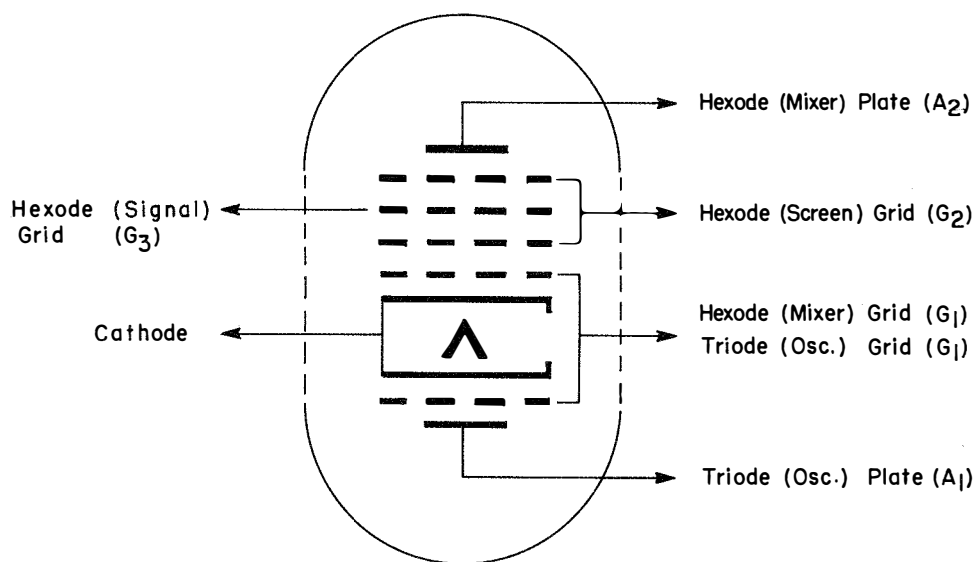


Figure 13-3. 6K8 Triode-Hexode Converter

transformer. Grid and plate connections are tapped down on the transformers to minimize shunting effects caused by capacity of the tubes, and to improve selectivity. The excitation voltage for the AVC circuit is obtained from the primary side of the last IF transformer.

The high degree of conversion stability and utilization of a crystal-controlled high-frequency oscillator has permitted the use of unusually high "Q" IF transformers for the purpose of obtaining signal frequency selectivity. The over-all response to frequencies off resonance is highly attenuated; however, to provide a pass band in which the response is down not more than two times (6 db) when two kc off resonance, the degree of coupling between the primary and secondary circuits has been chosen to effect this band width but still maintain the desirable attenuation to frequencies more than two kc off resonance. At six kc off resonance the response is down at least ten times (20 db) and at nine kc off resonance the response is down at least 100 times (40 db).

E. Automatic Volume Control Circuit

The AVC circuit makes use of a separate tube V-9, receiving its grid excitation from T-5 in series with capacitor C-39. Essentially the circuit functions as a linear dc amplifier for changes produced in its grid circuit by varying RF potentials. A filter C-40, R-35 and C-41 maintains the average dc plate potential developed independent of changes produced by varying percentages of modulation.

The components comprising the AVC circuit are shown in Figure 13-4 and Figure 13-8. A 6SJ7 tube is used as the "delayed" AVC amplifier to provide varying degrees of grid bias for the "controlled" tubes V-1, V-2, V-4 and V-5. This circuit is designed to regulate the bias of the "controlled" tubes in such a manner that the audio output of the receiver will remain fairly constant over a wide range of input signal intensities. With the sensitivity control R-2 at maximum, the audio output will not vary more than 4 db for any RF carrier input level between 10 microvolts and 0.1 volt.

With the sensitivity control, R-2, fully advanced (variable tap positioned at the junction

of R-2 and R-40), and no signal input to the receiver, the bias on the tubes V-1, V-2, V-4 and V-5, is the negative bias developed across R-40 from the B- supply. The receiver is at maximum signal amplification (minimum bias on V-1, V-2, V-4 and V-5).

A negative bias (B-) supply is developed by current flowing through R-41 in series with R-38 and R-39 to ground. (The relatively high resistance of the sensitivity control R-2 together with R-40 will have negligible shunting effect on the bias developed across R-38 and R-39). The portion of this negative bias developed across R-41, is applied to the control grid of V-9 through R-42, biasing the tube to cutoff. As a signal is applied to the input of the receiver, a 455 kc signal voltage will be applied through C-39 to the control grid of V-9, overcoming the fixed negative bias and causing the tube to conduct. The fixed negative bias supplied to the control grid of V-9 prevents V-9 from conducting until a signal strong enough (usually 2 to 10 μ v of RF carrier) to overcome this bias is applied; thus delayed AVC action is obtained. The cathode of V-9 is held at a fixed negative voltage, with respect to ground, by the fixed negative bias at the junction of R-41 and R-38; the plate of V-9, with no current flowing through the tube, is effectively at ground potential, which is positive with respect to the cathode. When the tube conducts, the current flowing from the plate through R-36 and R-40 to ground will develop a negative voltage at the junction of R-36 and R-35. This is added to the negative bias developed across R-40, and the resulting negative voltage is applied through R-35 to the AVC bus. The AVC bias voltage is applied to the control grid of V-1, V-2, V-4, and V-5 through decoupling networks R-3/C-6, R-5/C-10, R-15/C-20 and R-17/C-25, thus reducing the gain of these stages. A strong signal causes greater bias to be developed and proportionally reduces the gain of the controlled tubes.

With the AVC operative the maximum audio power output into the 20,000 ohm output circuit produced by a 0.1 volt RF input is 800 milliwatts when both the RF and AF gain controls are at maximum. When the sensitivity

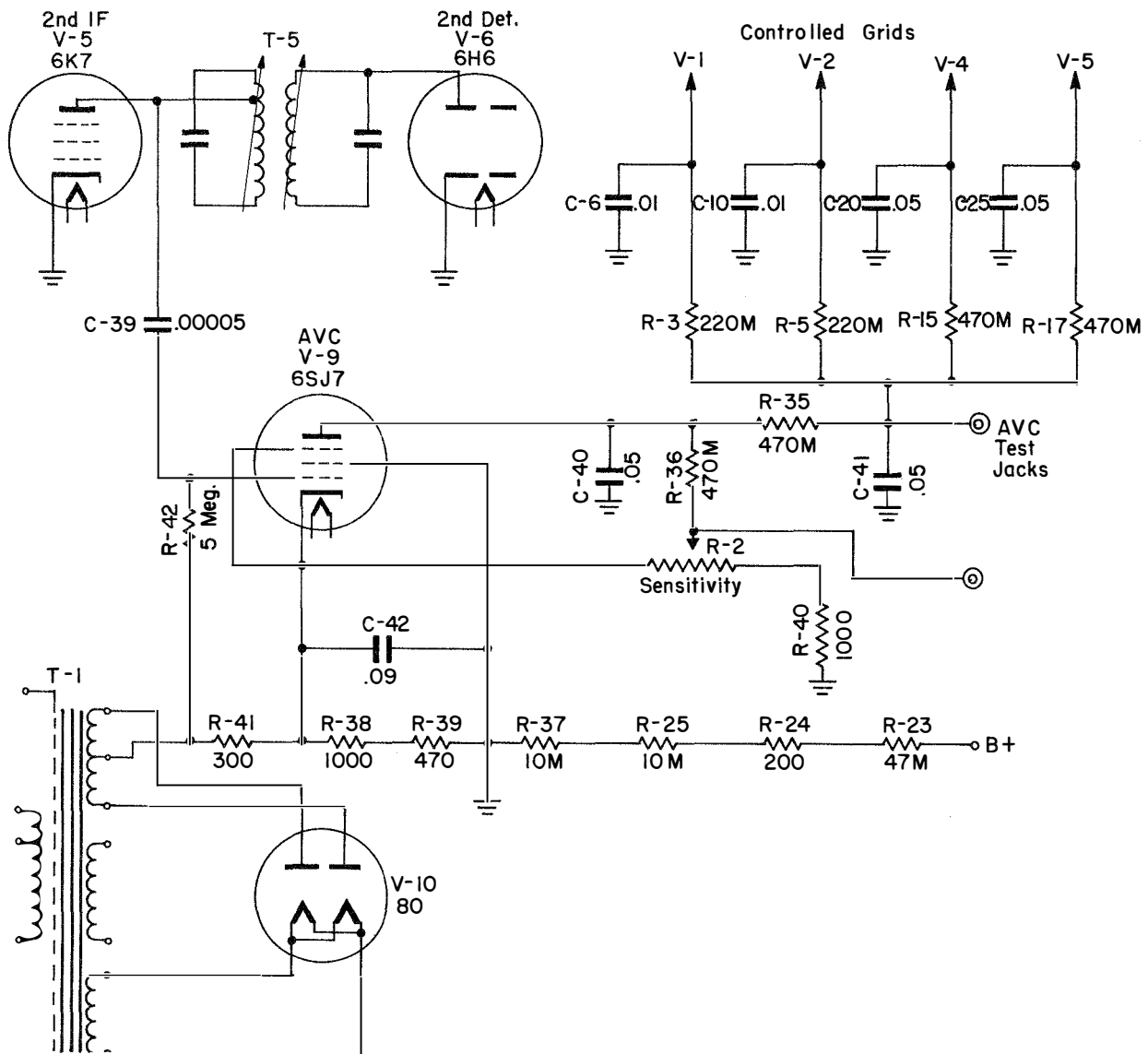


Figure 13-4. Automatic Volume Control (AVC)

control R-2 is set to other than maximum sensitivity position, increased fixed bias is applied to tubes V-1, V-2, V-4 and V-5, and higher input signal levels are then needed to produce AVC action.

A pair of test jacks have been provided on the front panel for the purpose of disabling the AVC bias voltage during certain performance measurements, and to provide a convenient test point for measuring the AVC bias voltage with a VTVM. This modification has been accomplished on all RHZ Receivers in accordance with EMBI No. 27 (Dated May 26, 1950.)

F. Second Detector

Figure 13-5 shows the elements of the diode detector circuit omitting all other associated circuits. The intermediate-frequency voltage induced in the secondary circuit of T-5 is rectified by the anode and cathode of V-6a (one half of the combined 2nd detector and noise limiter tube V-6). The diode load resistance consists of resistors R-21 and R-19 connected in series. Capacitor C-30 provides a low impedance path for the intermediate-frequency component of current flowing in the diode detector.

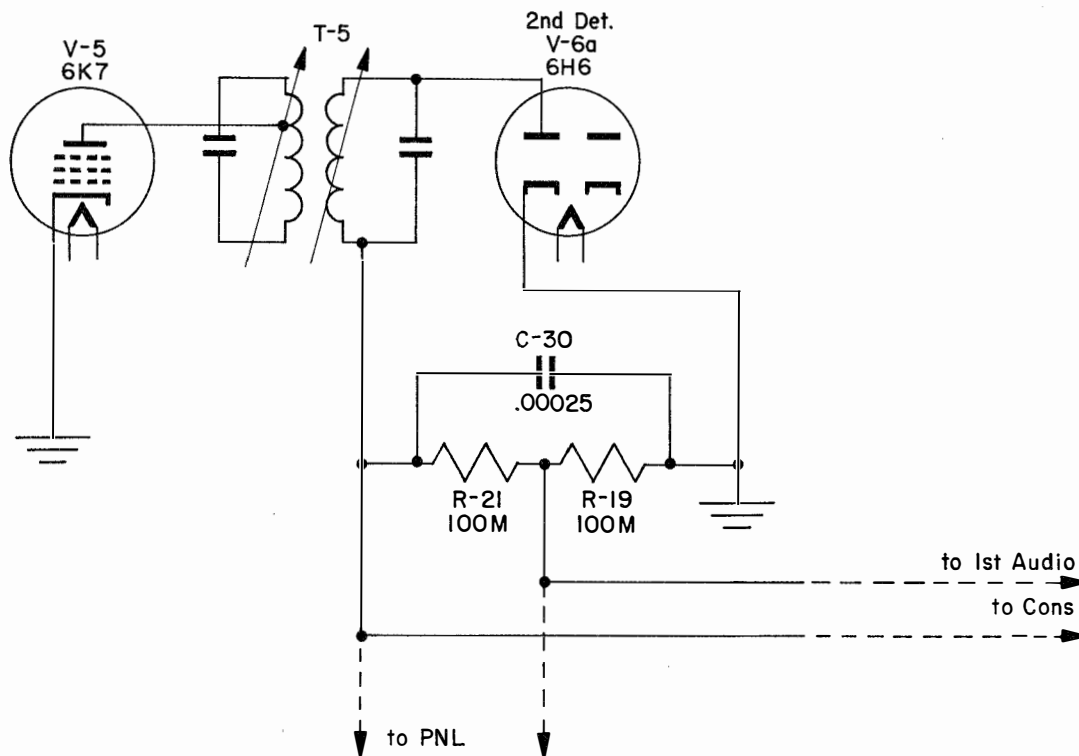


Figure 13-5. Second Detector

It will be noted that other circuits of the receiver are connected across the diode load resistor. In addition to the normal audio voltage that is developed for amplification through the audio system, rectified carrier potentials appearing across the diode load resistance are used for controlling the PNL and CONS.

The negative dc voltage developed across R-21 and R-19 will be approximately equal to the average signal voltage induced in the secondary of T-5.

G. Peak Noise Limiter (PNL)

Referring to the accompanying diagram Figure 13-6, positive peaks of RF voltage applied to the plate of V-6a cause electron current to flow. This causes plate of V-6b to be biased negatively with respect to the cathode of V-6b by rectification of steady-state noise signals of rectified carrier signal. Negative bias on the plate of V-6b is twice the negative bias on the cathode of V-6b by virtue of the

voltage divider circuit R-21 and R-19. R-22 and C-31 comprise an audio filter to remove audio ripple from the anode of V-6b. (Compute X_c at 400 cps . . . 443 ohms). R-22 and C-31 also provide a voltage-change delay circuit such that the plate voltage will not follow a quick voltage change at the junction of R-21 and R-22.

Any steady state signal appearing across the secondary of T-5 will produce a negative voltage at the junction of R-21 and R-22 that is twice the negative voltage at the junction of R-21 and R-19. If a sharp noise peak should be impressed across T-5 the rectified peak signal is then impressed across R-19, R-21 and R-22. Because the plate voltage of V-6b cannot rapidly follow the dc impressed across the R-21, R-19 combination, the cathode of V-6b will be momentarily driven negative with respect to its anode and current will flow in V-6b. Enough of the signal is shorted out that the remaining signal presented to the 1st audio

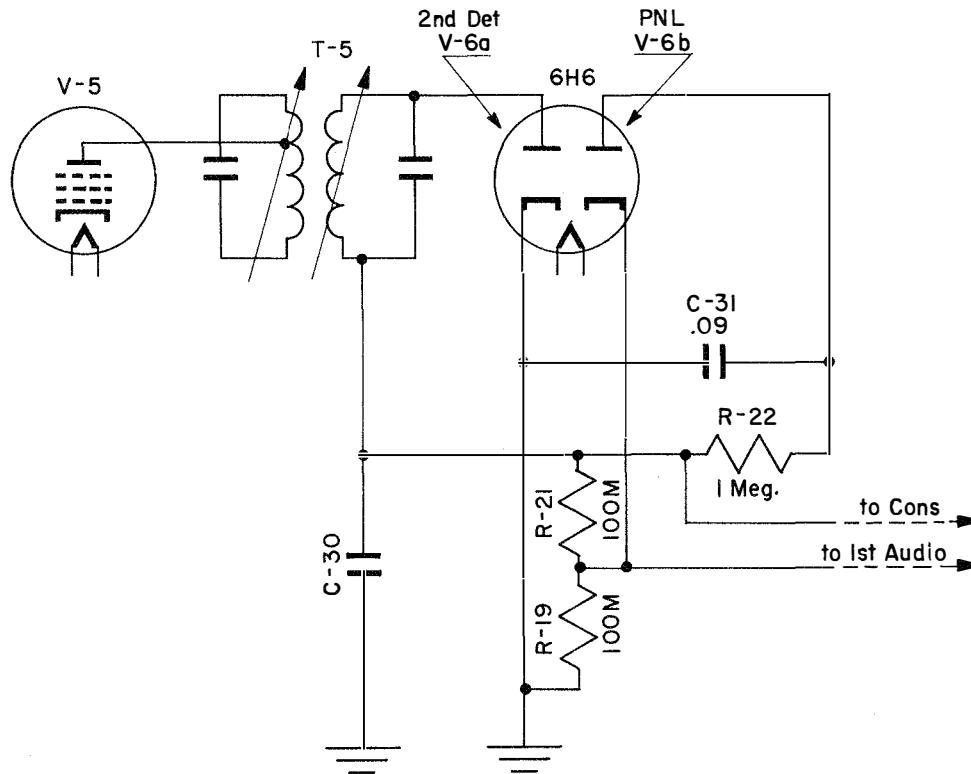


Figure 13-6. Peak Noise Limiter (PNL)

stage is so greatly reduced as to cause little disturbance in the audio output stage of the receiver.

If a 100% modulated signal is applied to the secondary of T-5, the resultant audio peaks appearing at the cathode of V-6b cause the cathode at this time to have approximately equal voltage to that applied to the plate of V-6b. Therefore, if any noise signal of voltage amplitude greater than the peak audio signal appears, it will drive the cathode temporarily negative with respect to V-6b plate and V-6b will conduct thus effectively shorting out the noise peak to ground through capacitor C-31.

H. Carrier-Operated Noise Suppressor (CONS)

Figure 13-7 shows the components of the Carrier-Operated Noise Suppressor (CONS). This circuit is designed to render the receiver inoperative and suppress noise during the absence of a carrier. A dual triode (6SL7-GT) performs the combined function of a noise suppressor and audio amplifier. For purposes of circuit description the individual triodes in the 6SL7-GT envelope will be considered as

separate tubes. Since the actual silencing operation occurs in the 1st audio triode under the control of the CONS triode, the two triodes will be referred to as the "controlled-tube" V-7b and the "controller-tube" V-7a respectively.

The ability of the receiver to "squench" noise is dependent upon the biasing and plate potentials of both these triodes. V-7a receives its biasing potential from the voltage developed across the diode load resistance R-19 and R-21. Its plate voltage is obtained by the voltage drop across R-25 and R-37. Cathode and grid returns for V-7b are connected to the receiver voltage divider in such a manner that, although the grid is negative with respect to its cathode by the amount of voltage drop across R-24, it is positive with respect to ground by the potential developed across R-25 and R-37. It will be noted that the plate return of V-7a and grid return of V-7b are common. Resistor R-26 acts as a plate load resistor for V-7a. R-20 and C-33 are filters to provide an average dc biasing potential on V-7a independent of varying percentages of modulation.

In normal operation with the controller-tube V-7a disconnected by switch A-3 the con-

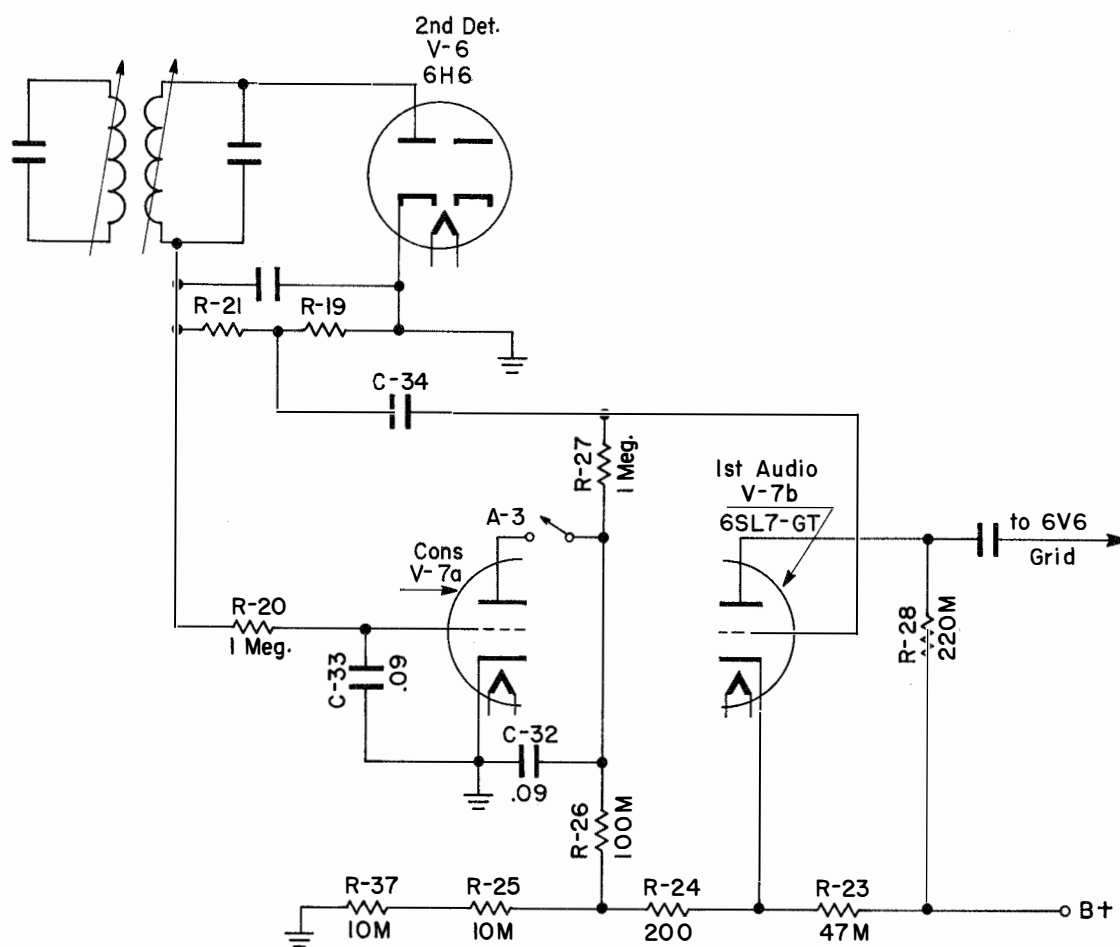


Figure 13-7. CONS and 1st AUDIO

trolled-tube V-7b acts independently as a straight audio amplifier; however, with the CONS switch on, *and in the absence of a carrier*, the tube passes current and offers a low resistance to ground for the grid return of V-7b. Since this grid return and V-7a plate are common, controlling bias will develop on V-7b grid depending upon the voltage drop across R-25 and R-37. In this manner V-7b is biased beyond plate current cut off, "squelching" the receiver completely.

With the appearance of a carrier, grid bias will be developed across R-19 and R-21 for V-7a causing it to block, and thereby release the controlling bias on the 1st audio tube, which will then function as an amplifier.

A variable threshold-adjustment has not been provided. Resistors R-24, R-25 and R-37 have each been predetermined so that, with the

receiver operating at maximum sensitivity, an input signal within a certain definite range will automatically develop across the detector load resistor a blocking bias for V-7a. At lower values of sensitivity a higher input signal will be required to develop the blocking bias for V-7a.

I. Audio Circuit

With exception of the 1st audio stage the audio system is conventional. A 6V6-GT is used in the output stage with its grid bias obtained partially from a tap on the receiver voltage divider and by the voltage drop across cathode resistor R-30. An audio volume control (R-1) is provided between the 1st audio and 6V6 grid. Output transformer T-2 provides the proper match between the 6V6 plate and the receiver output circuits.

J. Output Circuits

It is anticipated that this receiver will sometimes be located in the same room with the radio operator but, in other cases will be located at a remote point requiring that the receiver output be fed over a telephone line to the point where the operator is located. A relatively high output is provided for local operation of a loud-speaker, no further amplification being required for normal usage. For remote operation into a telephone line, a low output is provided because of the necessity of keeping the level on the line below the point that may cause cross talk. In addition to the above circuits, as shown on Figure 13-8 RHZ Schematic Diagram, a headphone output circuit is also provided. These three output circuits are described below:

K. High-Level Loud-Speaker Circuit

A 20,000 ohm high-level output circuit is provided for feeding a local loud-speaker. This circuit is fed directly from a transformer secondary to a double-pole double-throw panel switch which in the upper position connects the transformer secondary to a Jones connector socket A-9 (marked "SPKR") on the back of the chassis bracket. See Figure 13-2.

L. Low-Level Line Output

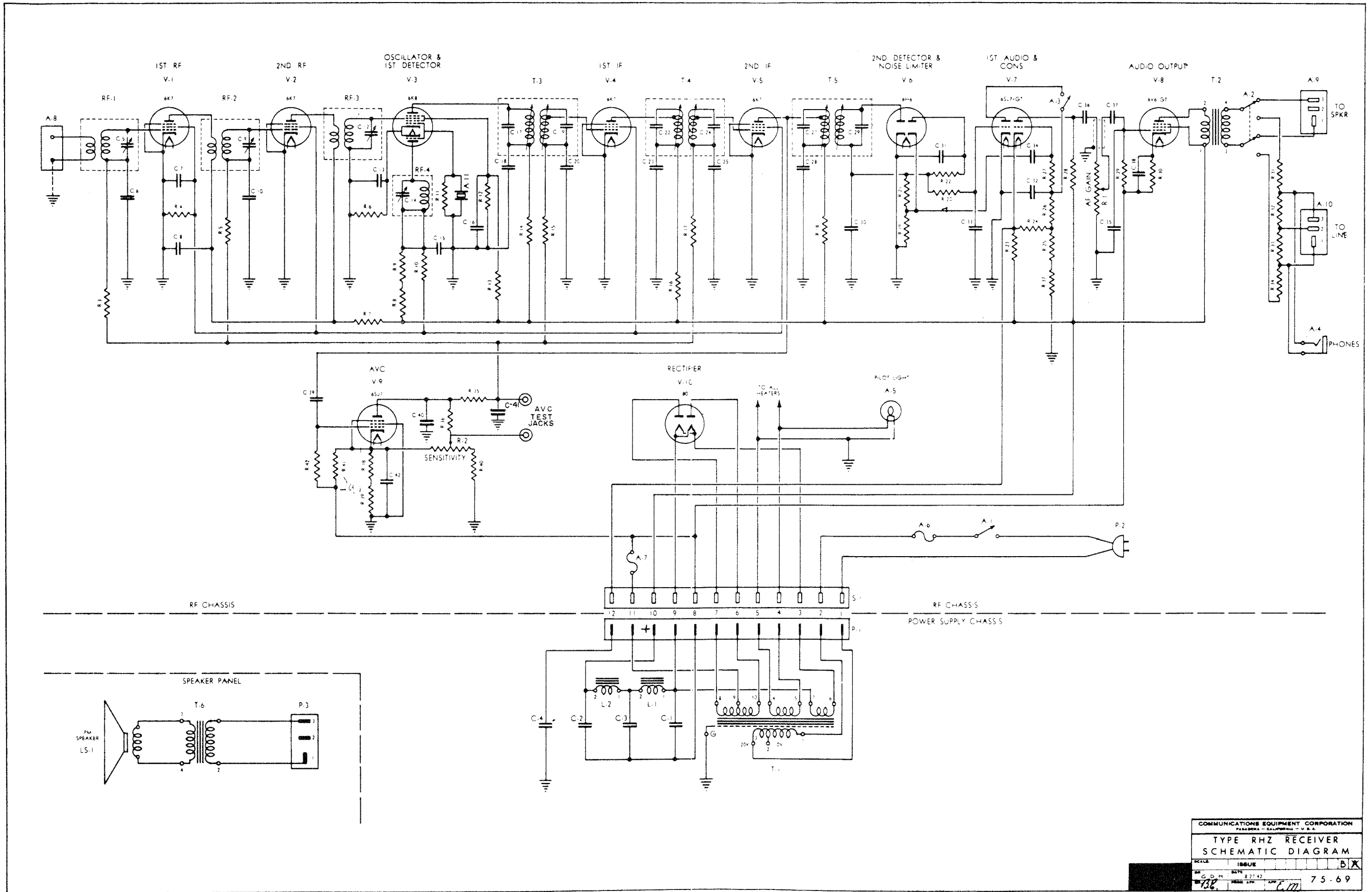
With the double-pole double-throw switch in the lower position a balanced "LL" pad is connected into the circuit between the 20,000 ohm transformer secondary and the Jones connector socket A-10 (marked "Line") located on the rear of the chassis bracket. See Figure 13-2. In this position both connections to the loud-speaker are opened. The "LL" pad is designed to feed a balanced telephone line and consists of two 10,000 ohm series resistors and two 300 ohm shunt resistors. This pad introduces an insertion loss of 21.2 db into the 600 ohm output circuit equivalent to a power ratio of 133/1.

M. Headphone Output

The 600 ohm audio output circuit is also connected to a headphone jack A-4 on the front panel. See Figure 13-1. This provides connection of the headphones directly in parallel with the line output. This jack is insulated so that no ground is placed on either side of the line output.

N. Technical Summary of Receiver Characteristics:

Frequency Range	1600 kc to 10,000 kc by substitution of coil sets
Coil Set Coverage	Band 1—1595 to 2175 kc Band 2—2175 to 2960 kc Band 3—2960 to 4040 kc Band 4—4040 to 5510 kc Band 5—5510 to 7500 kc Band 6—7500 to 10,000 kc
Intermediate Frequency	455 kc
Crystal Frequency	Signal-frequency plus IF frequency
RF Input Impedance	70 ohms
Output Impedances	20,000 ohms at Speaker Socket 600 ohms at Line Socket 600 ohms at Phone Jack
RF Sensitivity for 500 Milliwatt Output	4.5 microvolts
Maximum Undistorted Power Output	Into 20,000 ohms—1 watt Into 600 ohms—7.5 milliwatt
Operating Power	65 watts, 105 to 125 volts, 60 cps, single phase
Fuse Complement	1 ea type 3AG, 1 amp, slo-blo 1 ea type 3AG, .125 amp, slo-blo
Vacuum Tube Complement	4-6K7, 1-6K8, 1-6H6, 1-6SJ7, 1-6SL7-GT, 1-6V6-GT, 1-80



COMMUNICATIONS EQUIPMENT CORPORATION PALMDALE - CALIFORNIA - U.S.A.			
TYPE RHZ RECEIVER SCHEMATIC DIAGRAM			
SCALE	ISSUE	B X	
DATE	8-27-42		
738	75-69		

CHAPTER 14

RUP RECEIVER

14-1. GENERAL DESCRIPTION

A. *Equipment Purpose*

The RUP equipment was specifically designed for aeronautical ground station reception of amplitude modulated signals in the VHF band. It can be tuned continuously from 108 to 142 mc, either remotely or locally.

B. *Mechanical Construction*

The RUP receiver is designed to mount in a standard 19" relay rack, and requires 19 7/32" of vertical panel space. The remote control unit is designed to mount in a 10 1/16" rack or console, and requires 4 1/4" of vertical space.

The RUP receiver has a vertically mounted chassis with all the top chassis components projecting horizontally and to the rear of the unit. The fuse, local-remote switch, tuning control, dimmer, RF switch, power pilot lamp, carrier pilot lamp, tuning meter, and tuning dial are exposed for easy access on the front panel. The sub-chassis components and wiring are located behind the front panel door. A detachable sub-chassis, located in the upper left side of the main chassis, contains the RF amplifier, mixer, and the oscillator stages. The IF amplifiers, detector, AVC amplifier, CONS, and power supply circuits are located on the main chassis.

The remote control unit front panel has: RF gain control, muting on-off switch, tuning control, tuning dial, and tuning meter exposed for easy access on the front panel. The front panel also serves as the chassis on which the components are mounted.

14-2. THEORY OF OPERATION

A. *Input Circuit* (Figure 14-1)

The RF input circuit of this equipment is composed of two tuned circuits loosely cou-

pled to obtain maximum Q for RF selectivity. The input circuit is tuned by the main tuning capacitor (C-101A & B). The input is unbalanced and the impedance is 52 ohms. The input jack (J-105) is a coaxial receptacle, type UG 58/U.

B. *Radio Frequency Amplifier* (Figure 14-1)

The RF amplifier stage uses a type 6AK5 tube (V-101) connected in a conventional pentode circuit. The input circuit is tapped down to allow the tube to be operated at optimum voltages and still control the over-all gain. The tank circuits are tuned by C-101B, C, and D to a frequency in the range of 108 to 142 mc. The filament, plate, and screen circuits include RF decoupling networks. The gain of this stage is controlled by AVC.

C. *Mixer* (Figure 14-1)

The mixer employs a type 6AB4 tube (V-102), and operates as a biased detector; and the RF signal from V-101 and the injection frequency from V-112 are applied to the control grid of V-102 in parallel. The resulting signal in the plate circuit is the 12 mc IF, due to the filter action of C-113 and L-110, designed to prevent RF other than the IF from reaching Z-101. The injection frequency is higher than the input frequency from the RF amplifier. C-114 and R-106 act as a decoupling network to keep RF out of the power supply. The filament circuit also has an RF filter network.

D. *Oscillator Section* (Figure 14-2)

The oscillator section employs three stages, using V-110 (6BH6) as an oscillator-doubler, V-111 (6BH6) as a quadrupler, and V-112 (6AK5) as a doubler.

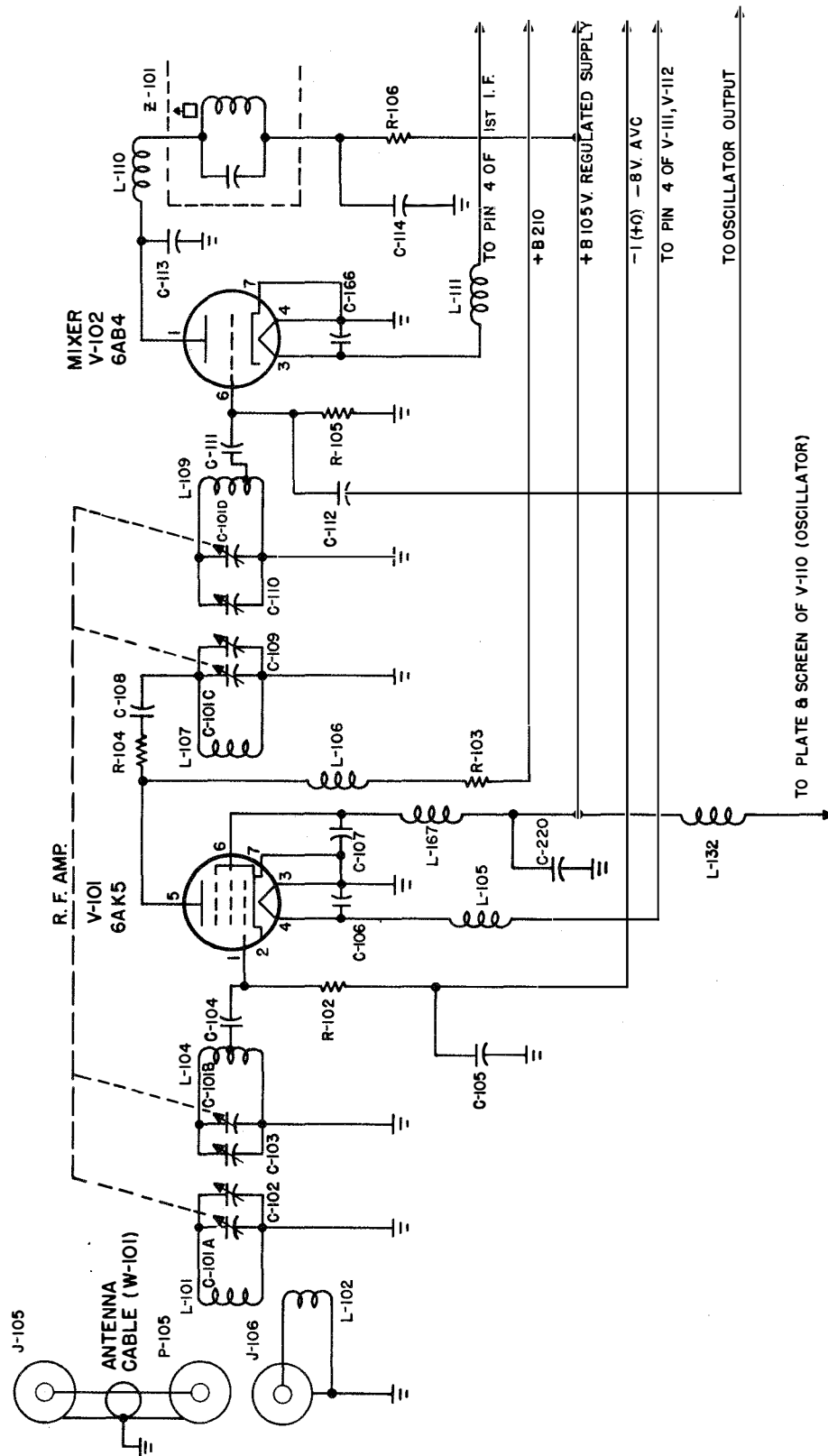


Figure 14-1. RF Amplifier and Mixer

The oscillator stage consists of an electron-coupled oscillator (modified Colpitts). C-198 and C-119 are temperature compensated capacitors and act as a voltage divider network. Since their value is large, as compared with the input capacitance of the tube, the effect of the tube capacitance on the resonant circuit is very small, resulting in higher stability. The resonant circuit composed of L-130, C-194, C-195, C-165, and C-193 has a frequency range of 7.4375 to 9.6875 mc. This circuit is coupled to the grid of V-110 through capacitor C-197 which is temperature compensated. C-194, main tuning, is coupled mechanically to C-101 by means of a gear drive. C-165 and C-193 are temperature compensated padders, used to provide proper tracking, because one end of the tank coil is at ground potential. C-197, C-198, and C-199 all affect the LC ratio of the tank and, therefore, the frequency. The resistor R-162 keeps the cathode at the proper RF potential while at the same time provides a dc path to ground. The plate and screen voltages for this stage are taken from the regulated supply. The oscillator energy is electron coupled to the plate of the tube and thence through C-201 to the resonant circuit of L-133, C-101E, and C-202 which has a frequency range of 14.875 to 19.375 mc.

The QUADRUPLER stage employs a type 6BH6 (V-111) and has one tuned circuit. The input signal is coupled through C-203 to the grid of V-111 and the output circuit is tuned to the fourth harmonic of the input frequency

in the range of 59.5 to 77.5 mc. The rotor C-101F, the main tuning capacitor, is grounded which makes it necessary to use the blocking capacitor C-206. Energy is coupled to the next stage due to the proximity of L-136 and L-137. The plate and screen circuits have decoupling networks.

The final multiplier stage of the oscillator section employs a type 6AK5 tube (V-112) and has two tuned circuits. The input circuit tunes to the same frequency as the preceding stage and the output circuit is tuned to the second harmonic of the input frequency, in the range of 119 to 155 mc. The rotors of the two tuning capacitors are grounded which makes it necessary to use the two blocking capacitors C-209 and C-211. Both plate and screen circuits have decoupling networks.

E. IF Amplifiers (Figure 14-3)

The first four IF amplifier stages are exactly alike and use the type 6BJ6 tube, V-103, V-104, V-105, and V-106 respectively. AVC is applied to each stage through a decoupling resistor, R-107, R-110, R-112, and R-114 respectively. The first IF amplifier (V-103) is capacitively coupled to Z-101 (tuned to 12 Mc) through C-115. A coaxial cable (W-102) is used to link Z-101, on the RF deck, to V-103 which is on the main chassis. The plates of the first four IF amplifiers are coupled to the next succeeding stages through Z-102, Z-103, Z-104, and Z-105 respectively. The IF trans-

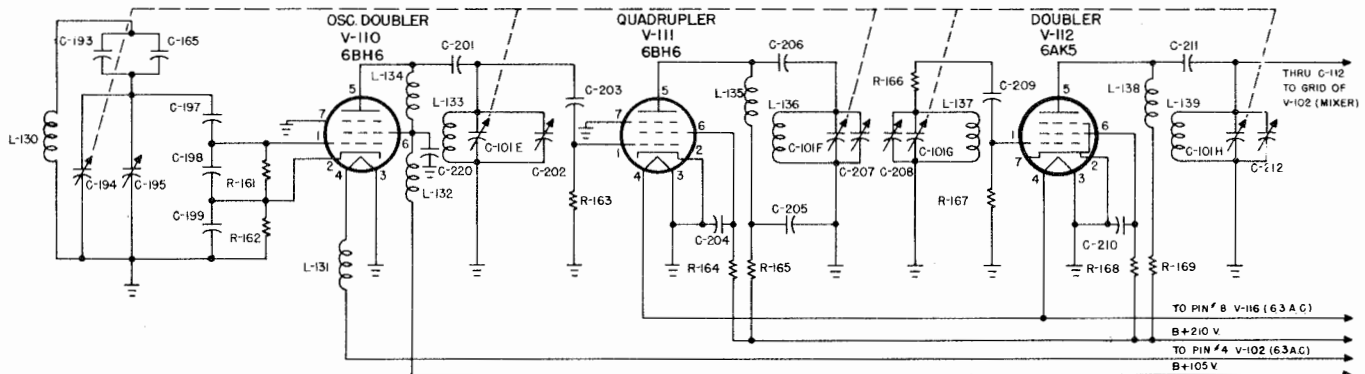


Figure 14-2. Oscillator and Frequency Multiplier Circuits

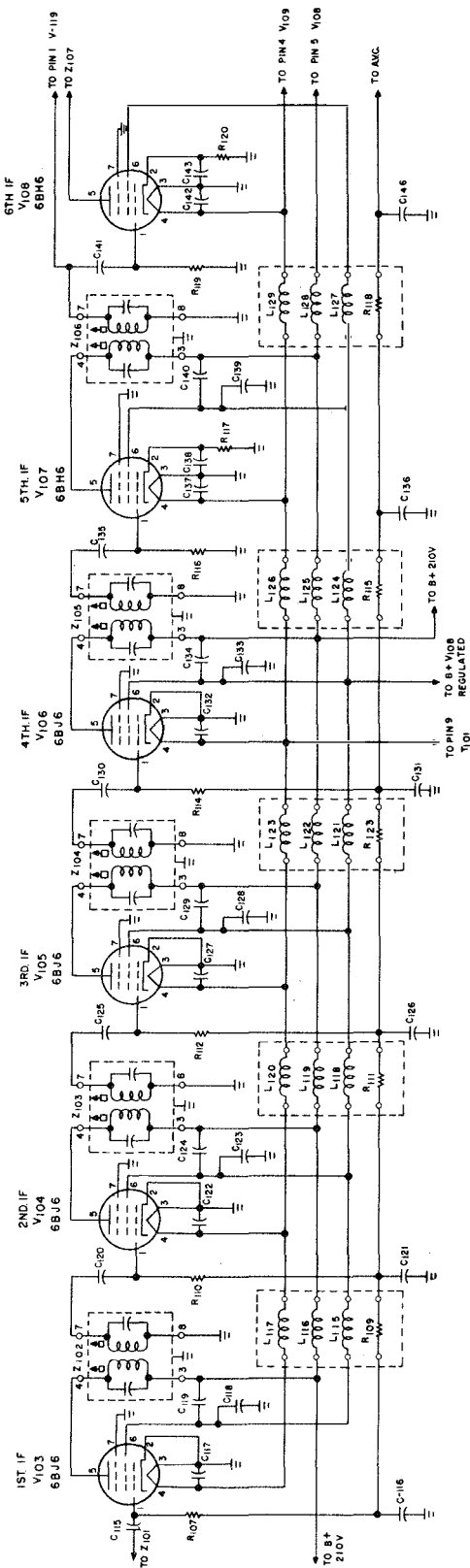


Figure 14-3. IF Amplifier Circuit

formers are inductively tuned by means of movable iron cores. Plate, screen, and filament circuits all have decoupling networks.

F. The Fifth IF Amplifier (Figure 14-3)

This amplifier employs a type 6BH6 tube (V-107). This stage is not controlled by AVC, but is biased by the cathode bias resistor, R-117. In this stage as in the other IF amplifiers, the input grid is capacitively coupled to the IF transformer. The plate is connected to the IF transformer, Z-106, which is inductively tuned by means of movable iron cores. Plate, screen, and filament circuits have decoupling networks.

G. Sixth IF Amplifier (Figure 14-3)

The sixth IF amplifier employs a 6BH6 (V-108) and is capacitively coupled to IF transformer Z-106, through C-141. The output of Z-106 is also coupled to the input of the AVC amplifier, V-119. V-108 is not controlled by AVC, but receives its bias from R-120, cathode bias resistor. The plate of V-108 is connected to Z-107 which is tuned by means of movable iron cores. Plate, screen, and filament circuits have decoupling networks.

H. Detector (Figure 14-4)

The signal voltage developed by the secondary of Z-107 is rectified by diode action in tube type 6BF6, V-109. The plates (pins 5 and 6) are connected to one terminal of the secondary and a detector filter network is connected to the other terminal. The cathode (pin 2) of V-109 is returned to ground through the meters M-101, M-201 and the LOCAL-REMOTE switch S-102. With the switch set at LOCAL only M-101 is in the circuit. When S-102 is set at REMOTE both M-101 and M-201 are in series to ground. The amount of current through the cathode will vary in accordance with the input signal; the meters will therefore serve as tuning indicators. R-177 is included in the grid circuit of V-109 to permit the signal voltage developed across R-122 to adjust the effective bias level of V-109, thus insuring operation of the tuning indicator meter at low levels of input signal. The detector filter network consists of R-101, C-147, and

C-148. The series combination of R-122 and R-176 serves as the detector load. The portion of the detected signal that is developed across R-122 is coupled to the grid (pin 1) of V-109 through C-149, and amplified. This amplified voltage is coupled to the cathode of V-120, series noise limiter, through C-150.

I. ANL (Series Noise Limiter) (Figure 14-4)

This stage employs a type 6AL5 tube, V-120. The cathode of V-120 is connected to ground through R-126 and R-125 in series with the detector load, R-122 and R-176. The resistance value of the detector load is much greater than the combined resistance of R-125 and R-126. The plate (pin 2) of V-120 is connected to ground through R-128 and R-129. The current caused by a detected signal will flow through the detector load to ground, which places the plate of V-120 positive with respect to the cathode, permitting signal current to flow through the tube and thus, develop an excitation voltage for V-114 across R-128, coupled to R-160 by C-125. The amplified signal from V-109 is also coupled to the cathode of V-120. This coupled voltage is opposite in phase to that supplied by the signal voltage across the detector load; therefore the total cathode current is determined by the difference between the two voltages. One result of this is to cause R-125 to act as an inverse feedback resistor, which corrects the phase shift and frequency response of the audio amplifier section of V-109. The values and voltages in this circuit are so chosen that a signal will pass through the diode as long as the input signal is held at a certain level. This level is determined by the bias on V-120 as set by R-124, R-132, R-136, R-158 (RF GAIN), R-159, and the bias supply. Refer to Fig. 14-4. The region of contact potential is avoided because this diode is operated at a higher level than usual. If a noise transient is received that is higher than the acceptable carrier signal, the resultant amplification of this signal causes the voltage drop across R-125 and R-126 to equal or exceed the potential across the detector load. This causes the diode, V-120, to become momentarily non-conducting. The clipping effect on the noise pulse is approximately 70%, due to the effect of the time constant of the circuit.

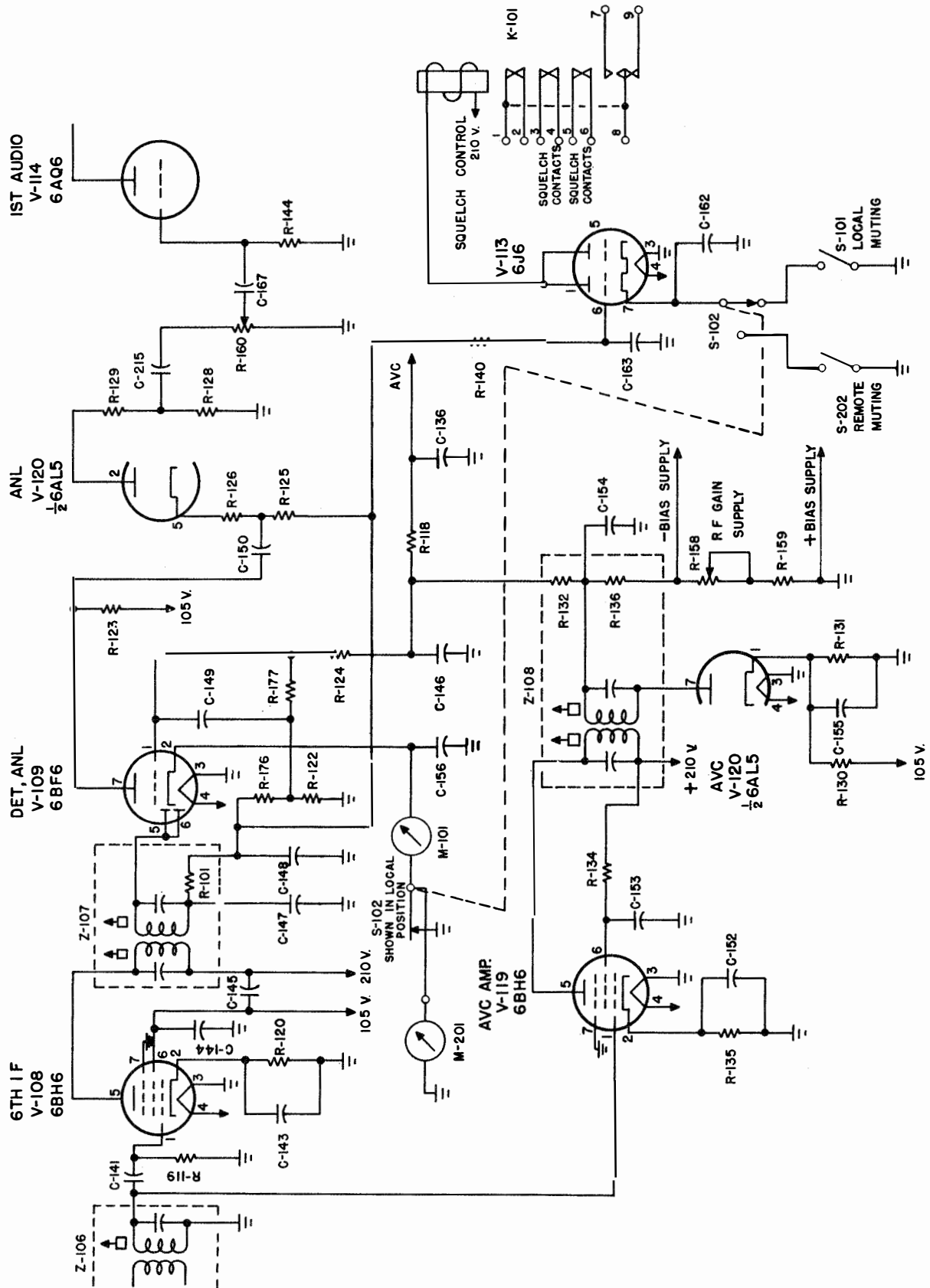


Figure 14-4. Detector, ANL and AVC Circuit

J. Squelch (Figure 14-4)

The squelch control circuit employs a type 6J6 tube, V-113, and is coupled through R-140 to R-125 and R-126 in the cathode circuit of V-120. The duo-triode, V-113, is connected as a single unit, using all elements. The grid (pins 5 and 6) are bypassed to ground by C-163. The squelch relay (K-101) coil is in series with the plate lead to the B plus supply. When the signal strength drops below a certain value, dependent on the setting of the RF GAIN control, the bias voltage applied to V-113 changes permitting 8 ma of plate current to flow, thus activating the relay. When this circuit is activated, contacts 7 and 8 of relay K-101 (Fig. 14-5) route the audio signal from the 1st audio amplifier V-114 through R-148 to the next audio amplifier V-116A; R-148 can be set to vary the muting to any level between 6 and 30 db below the unmuted level. With a sufficiently strong input signal, the bias applied to the squelch relay tube, V-113, is sufficient to hold the plate current of the tube to a level that prevents activation of the squelch relay, which opens contacts 7 and 8 of relay K-101 thereby removing R-148 from the audio path, and closes contacts 8 and 9 to provide an alternate path for the audio signal; R-148 now has no effect on the amount of audio voltage applied to the grid of V-116A.

K. Automatic Volume Control (Figure 14-4)

The AVC amplifier employs a 6BH6, V-119, and is coupled directly to the output of Z-106. The plate is connected to the primary of Z-108, a 12 mc IF transformer, tuned by means of a movable iron core. The secondary of Z-108 is coupled to one-half of V-120, duo-diode, the rectified output of which is used for AVC. The lower terminal of the secondary of Z-108 is connected to the plate (pin 7) of V-120. The upper terminal connects to a voltage divider network comprised of R-132, R-136, R-158, and R-159 in series between the AVC line and ground. The bias voltage from V-118 is dropped across R-158 and R-159 with the positive side grounded; thus a negative bias voltage is placed on the plate (pin 7) of V-120. R-158

controls the bias; not only on V-120, but also on the RF amplifier, the first four IF amplifiers, and the detector; thus R-158 functions as an RF GAIN control. The cathode (pin 1) of V-120 is returned to ground through R-131. A positive 105 volts is coupled to the cathode through R-130. The result of this network is to place a positive bias of approximately 30 volts on the cathode of V-120. The AVC action begins to function when the input signal is sufficient to overcome this bias. Thus it can be seen that the setting of the RF GAIN control also determines the operating point of the AVC circuit.

L. Audio Amplifier (Figure 14-5)

The audio amplifier section has two output stages, and facilities for three outputs. The main audio output will furnish 1 watt into a 20,000-ohm load, while the low level output will furnish 10 milliwatts into a 600-ohm load. The main audio may be monitored through terminals A and B of J-102; and the low level audio may be monitored through terminals D and E of J-102, or through the phone jack, J-101. The frequency response of the audio amplifier section is from 200 to 4000 cps with the high end attenuated 20 db at 8,000 cps.

The signal from V-120 is developed across R-160, the AUDIO GAIN control. Thus, the audio voltage coupled to the grid of V-114 (6AQ6) can be varied manually from the front panel. The output of V-114 is coupled through C-213 to a voltage divider network consisting of R-143, R-147, and R-148 in series to ground. This network is connected directly to terminals 7 and 9 of the squelch relay, K-101. (Refer to Fig. 14-4. The squelch relay provides for attenuation of the objectionable background noise when the receiver is not tuned to a station. The noise level is reduced by switching the input for V-116A ($\frac{1}{2}$ 6SN7) to a lower point on the voltage divider network. The amount of noise attenuation is determined by the potentiometer, R-148, MUTING control. Fixed resistances, R-143 and R-147, fix the upper and lower limits for the operation of this control. In the open position, the squelch relay couples the high end of the voltage divider network to the input of V-116A for full audio output. The

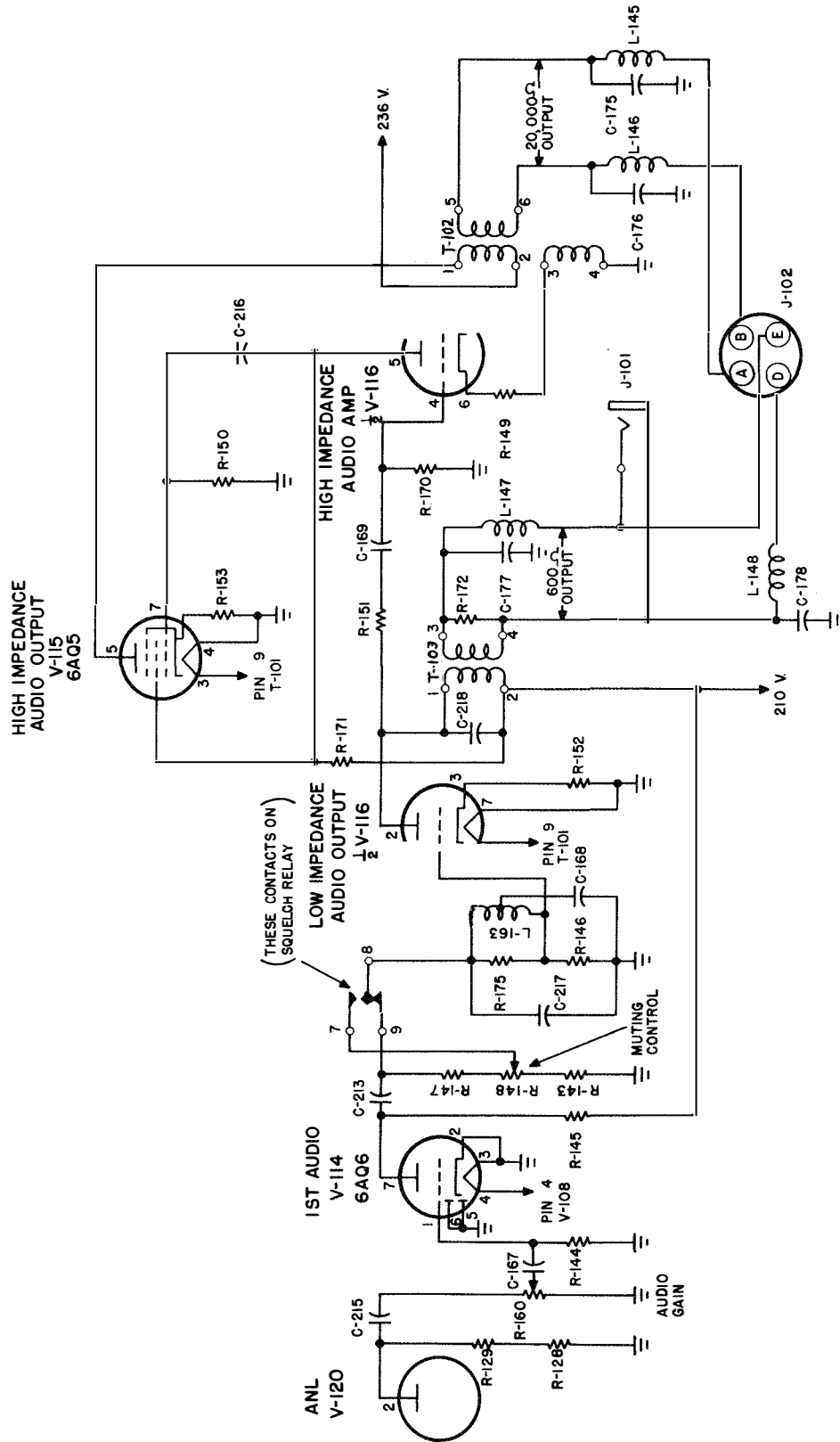


Figure 14-5. Audio Amplifier Circuit

audio that is coupled through the squelch relay is attenuated below 200 cps and above 4,000 cps by the AF filter network, L-163, C-168, C-217, and R-175; which determines the band of audio frequencies that will be amplified by V-116A.

The output of V-116A is coupled to the low level output, terminals D and E of J-102; also to J-101 in parallel. The impedance of this output is 600 ohms. There are RF filters on this output to reduce any strong RF pickup (i.e. C-178 and L-148, C-177 and L-147). The output of V-116A is coupled through R-151 and C-169, a gain and frequency compensating circuit, to V-116B ($\frac{1}{2}$ 6SN7). Cathode bias is used on V-115 and both sections of V-116. The output of V-116B is further amplified by V-115 (6AQ5) and coupled through T-102 and an RF filter network (i.e. C-175 and L-145, C-176 and L-146), similar to that used in the low level output, to terminals A and B of J-102. Possible damage to T-102, when operated with no load, is reduced by using the separate winding to provide inverse feedback to the cathode of V-116B; which also improves output load regulation.

M. Power Supply (Figure 14-6)

The receiver is connected to the power line through P-102 and P-104. Power is turned on and off by S-104, a STDP switch. Inductor-capacitor networks have been inserted in each power line to filter out any strong RF pickup (i.e. C-180 with L-150 and C-181 with L-151). The output of T-101 supplies the following voltages:

1. 6.3 volts to the receiver filaments, except for V-117.
2. 5 volts to the filament of V-117, type 5U4G.
3. 300 volts to the High Voltage supply.
4. 150 volts to the Low Voltage supply.

The high voltage supply employs a full-wave choke-input rectifier circuit. The two plates of V-117, type 5U4G, alternately conduct as they are driven positive with respect to the cathode by the applied transformer voltage. The two filter networks (L-164, C-160, and C-161; and L-165, C-158, and C-159) smooth out the ripple present in the output of V-117. The high volt-

age for the audio output stage is taken off the filter line between L-164 and L-165. Voltage regulator V-121, type VR-105, provides a stabilized voltage for:

1. The screens of the IF and RF amplifier tubes.
2. The cathode bias on the ANL diode.
3. The plate and screen of the local oscillator tube.

Resistor R-133 serves as a current limiting resistor for the VR tube. Its value lies between that which just permits minimum tube current to flow when the supply is loaded; and, that which just passes the maximum allowable tube current when the supply is unloaded. The positive voltage appearing at the high side of R-133 is supplied to:

1. The plates of the RF and IF amplifier tubes.
2. The screens and plates of the oscillator multiplier tubes.
3. The screen and plate of the AVC amplifier tube.
4. The plate of the audio input tube.
5. The plate of the squelch relay tube.

The low voltage supply employs a full-wave resistor-capacitor rectifier circuit. The two plates of V-118, type 6H6, alternately conduct as their respective cathodes are driven negative by the applied voltage from the power transformer. The resistor input filter network made up of R-139, C-164A, R-138, C-164B, R-137, and C-157 smoothes the ripple present in the output of V-118.

14.3. TUNING MECHANISM

A. Electrical Operation (Figure 14-7)

At any time that the ON-OFF switch, S-104, is in the ON position and power is being applied to the unit, the rotors of the tuning motor, B-101, the selsyn transmitter, B-102, and the selsyn receiver, B-201, (if the remote unit is connected) are all connected in parallel across the 120 volt ac input. The tuning motor, B-101, is a two-phase motor. Since the power source required for the receiver is single phase, a capacitor, C-214, is used to provide phase splitting. Resistor R-174 is used to limit arcing at the motor contacts.

When the TUNING switch, S-103, is in the neutral position, voltages applied to the two

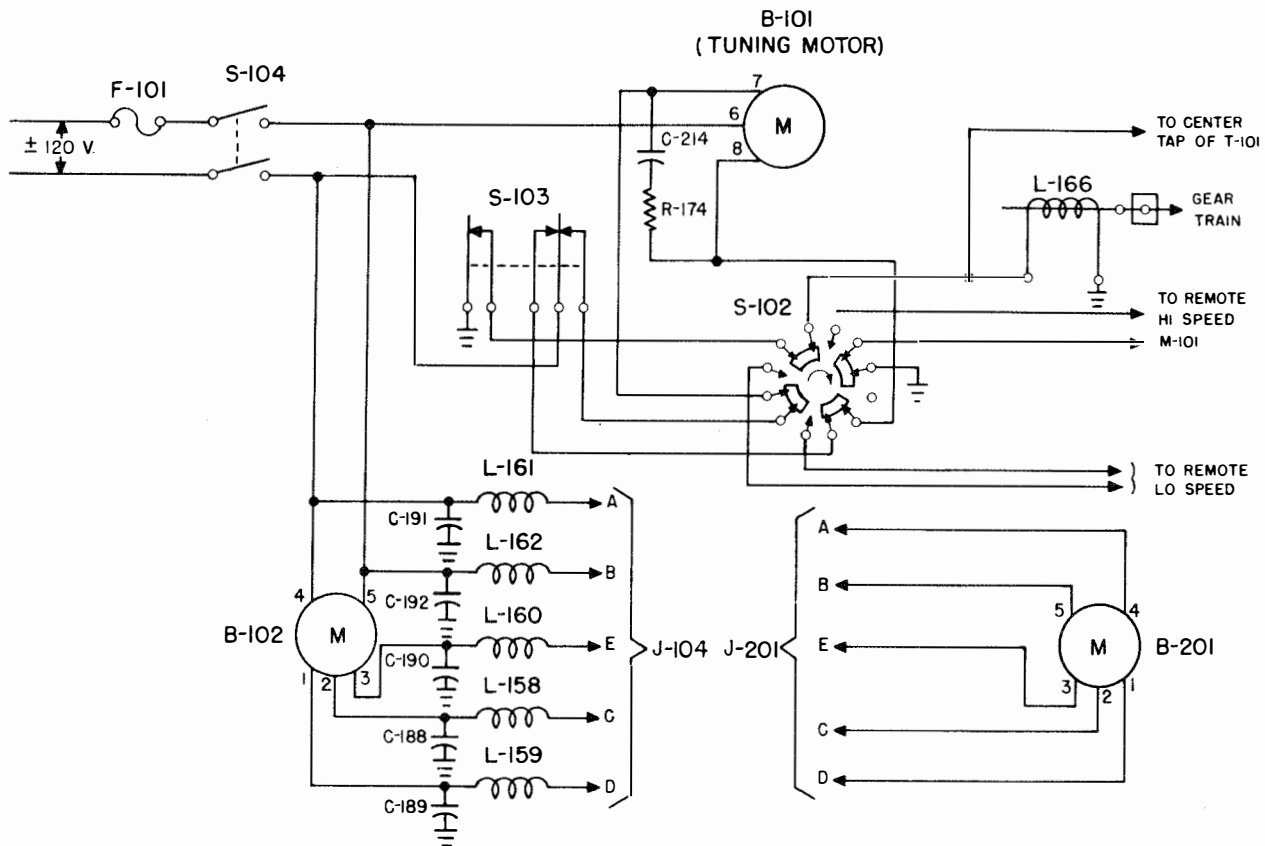


Figure 14-7. Electrical Circuit of the Tuning Mechanism

windings of B-101 are in phase, therefore the motor will not operate.

When the TUNING switch is turned counter-clockwise just enough to break contacts 4 and 5 but not enough to break contacts 1 and 2 (the slow tuning position) of S-103, current flows through B-101 in such a direction as to cause the rotor to rotate in a clockwise direction. The drum dial and the dial pointer are connected through gear train to B-101 in such a manner that the frequency indicator moves from the high frequency to the low frequency end of the drum. In the slow tuning position, the drum dial rotates at a speed of 1.37 rpm. With the switch, S-103, in the slow tuning position, the tuning speed solenoid, L-166, is shorted out by contacts 1 and 2 of S-103.

When S-103 is turned fully counterclockwise, contacts 1 and 2; 4 and 5, are broken (the fast tuning position), and current flows

through B-101 in the same direction as before and the dial pointer still moves from the high frequency to the low frequency end of the band. The drum dial now rotates at a speed of $62\frac{1}{2}$ rpm. This is due to the short circuit being removed from across the solenoid, L-166, by the breaking of the contacts 1 and 2 of S-103.

When S-103 is turned clockwise just enough to break contacts 3 and 4, but not enough to break contacts 1 and 2, the tuning speed solenoid is again shorted out and the drum dial will rotate at slow speed. Under these conditions the tuning motor will turn in a counter-clockwise direction and the dial pointer will move from the low frequency to the high frequency end of the drum dial. As S-103 is turned fully clockwise, contacts 1 and 2; 3 and 4 are broken and the tuning-speed solenoid

is again activated; the drum dial will rotate at fast speed and the dial pointer will move from the low frequency to the high frequency end of the drum dial.

The action just described for the electrical operation of the tuning mechanism will be the same whether the receiver is operated locally

or by a remote unit. Tuning switches S-103, local and S-201, remote, are identical in operation. The LOCAL-REMOTE switch, S-102, throws the TUNING switch, S-103, into the circuit when turned to LOCAL. TUNING switch, S-201, is thrown into the circuit when S-102 is turned to REMOTE.

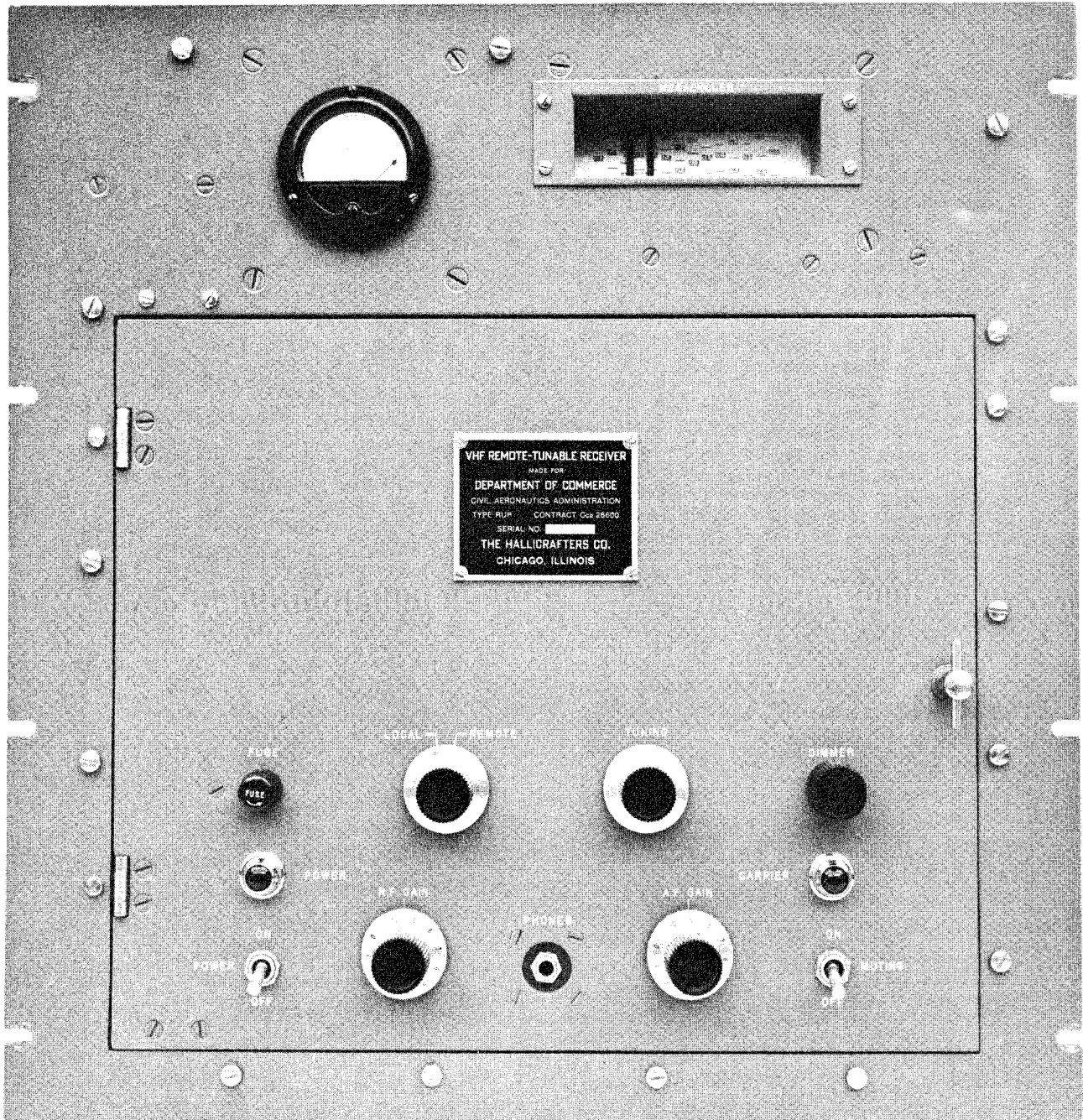


Figure 14-8. Front View of Receiver with Panel Closed

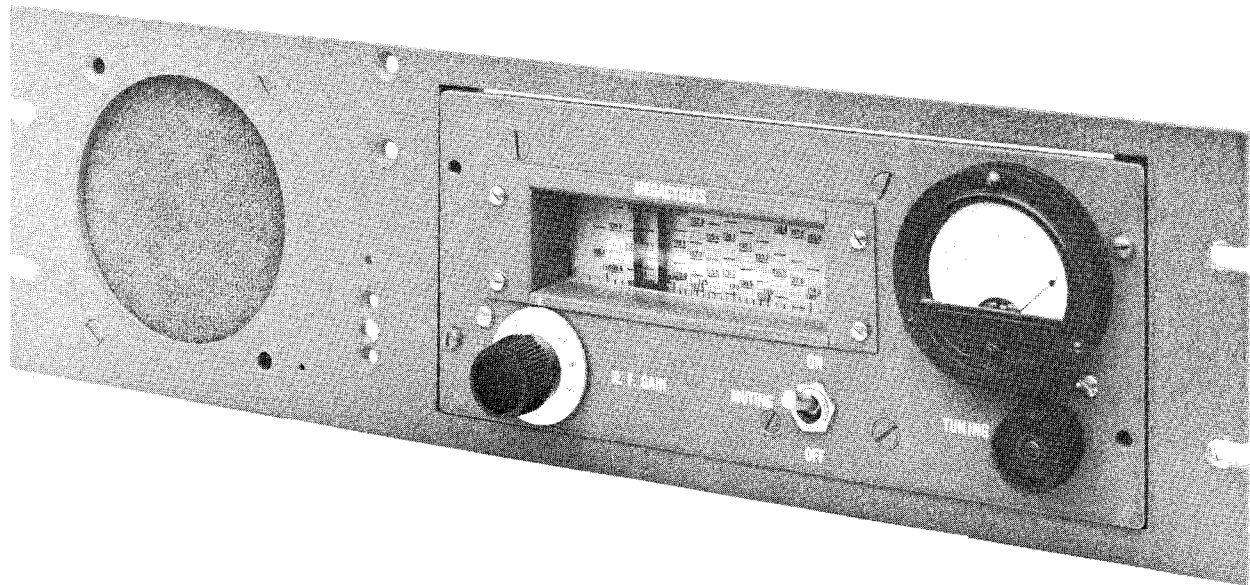


Figure 14-9. Remote Control Unit—Front View

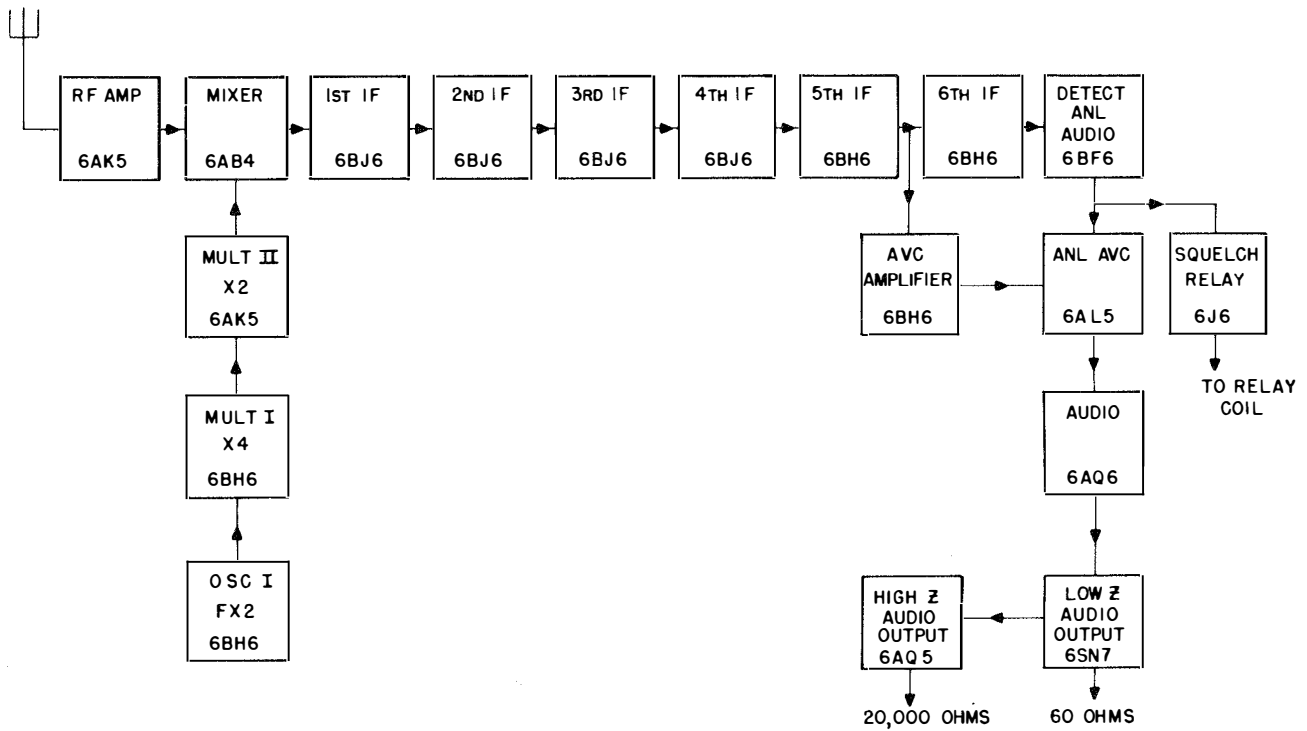


Figure 14-10. Block Diagram

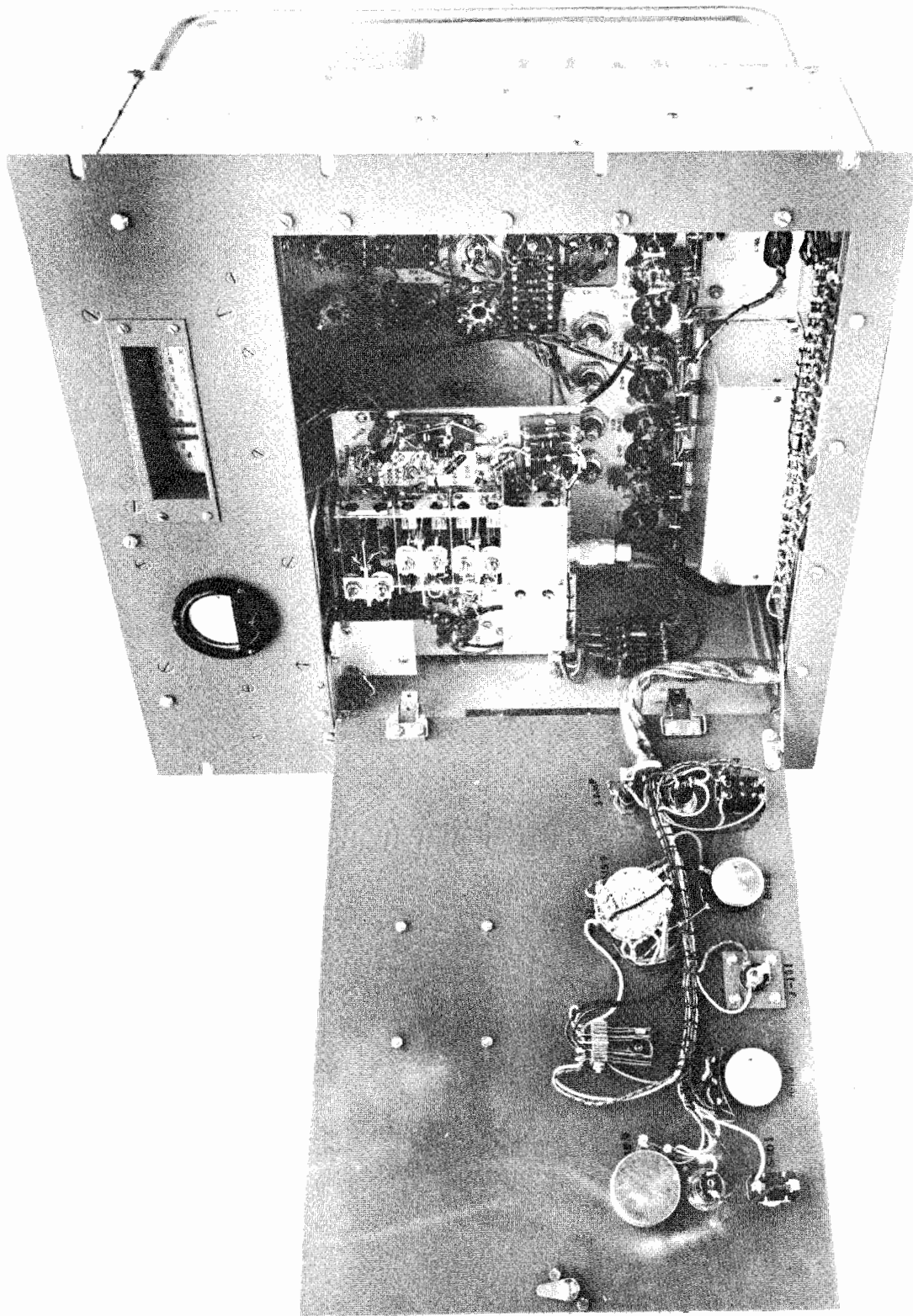


Figure 14-11. Front View of Receiver with Panel Open

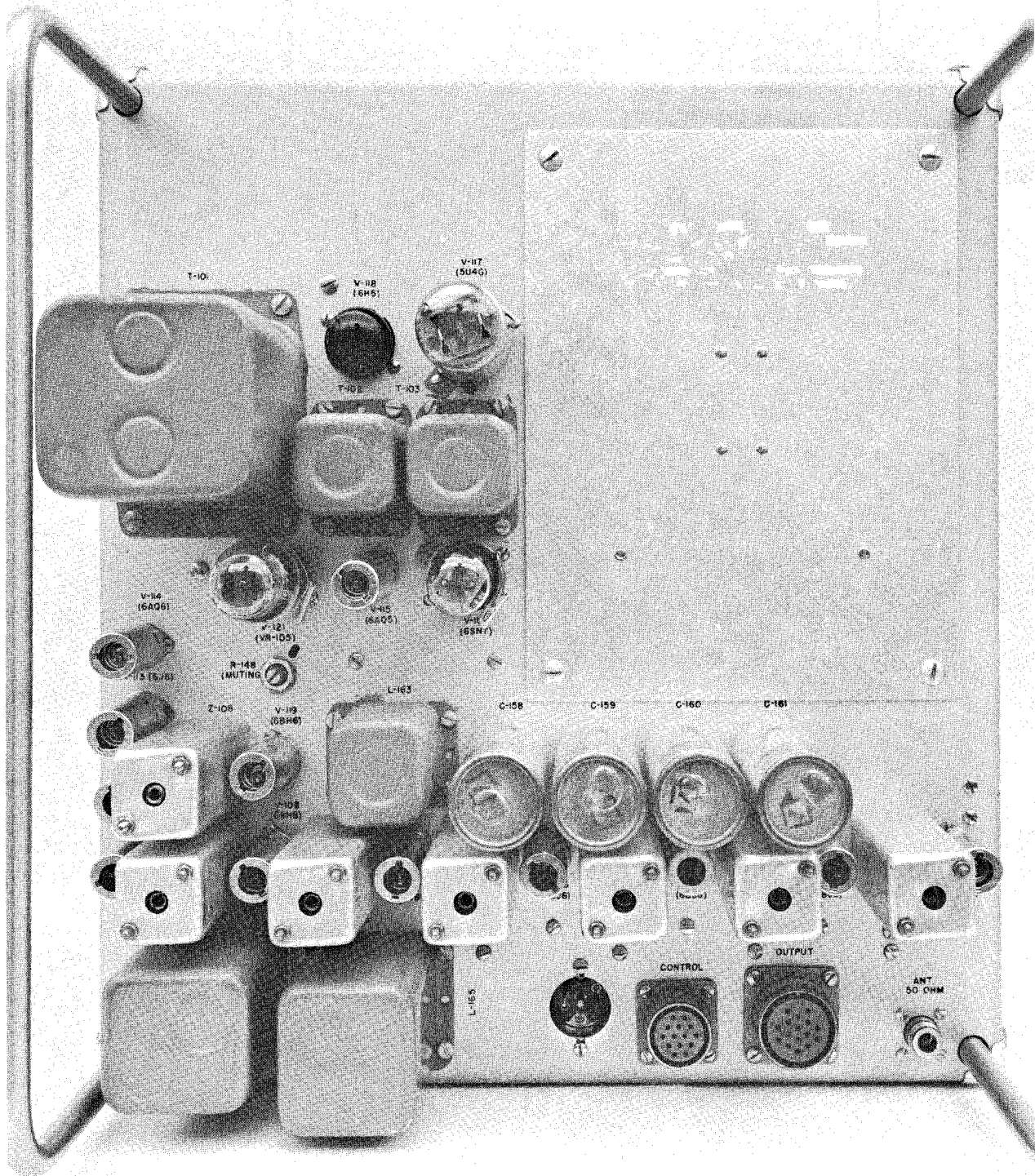
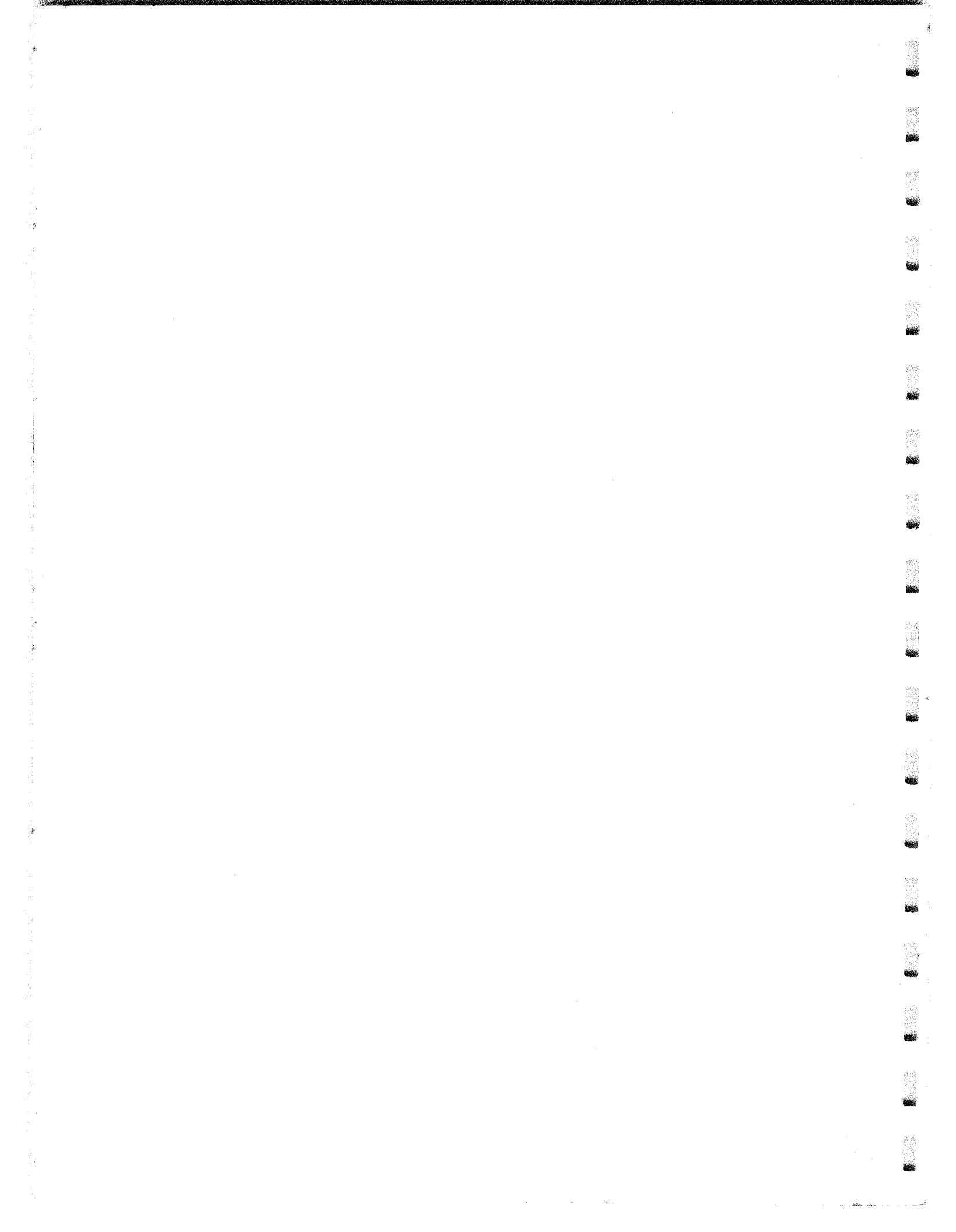
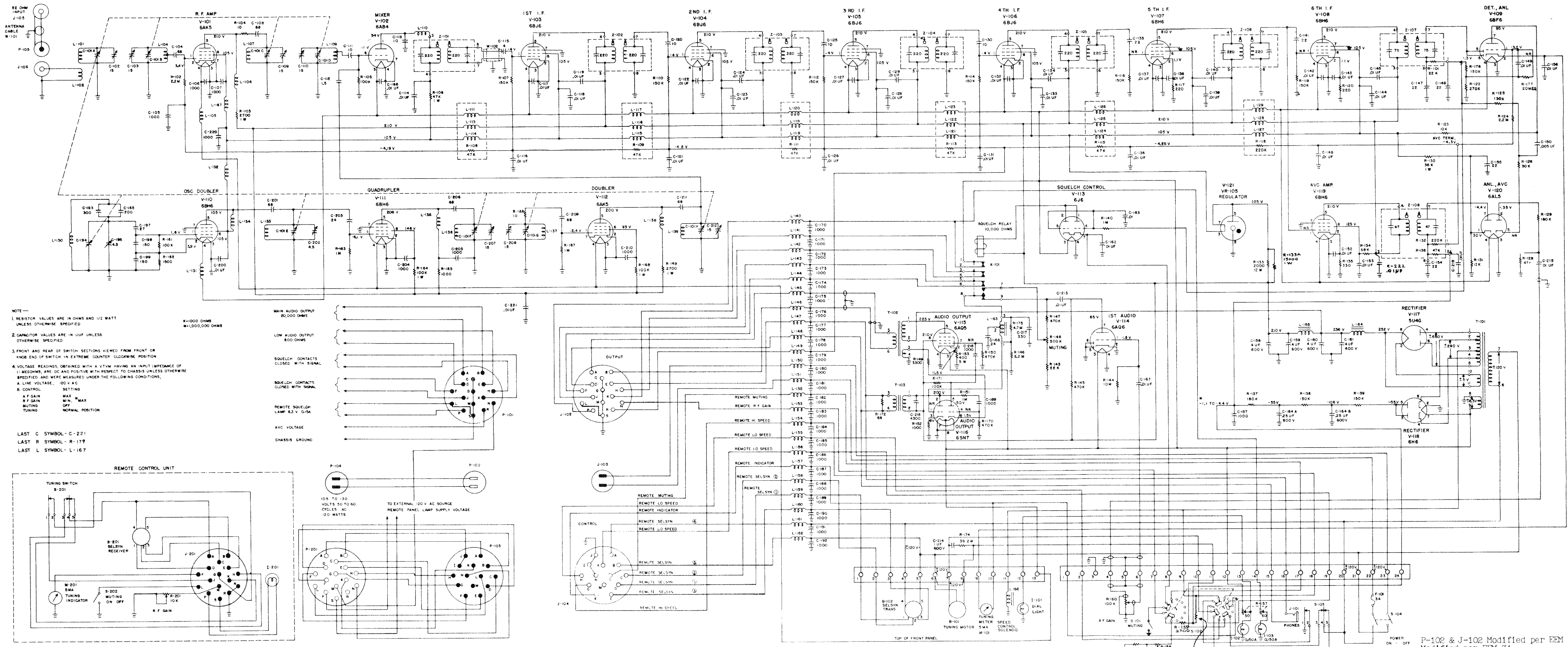


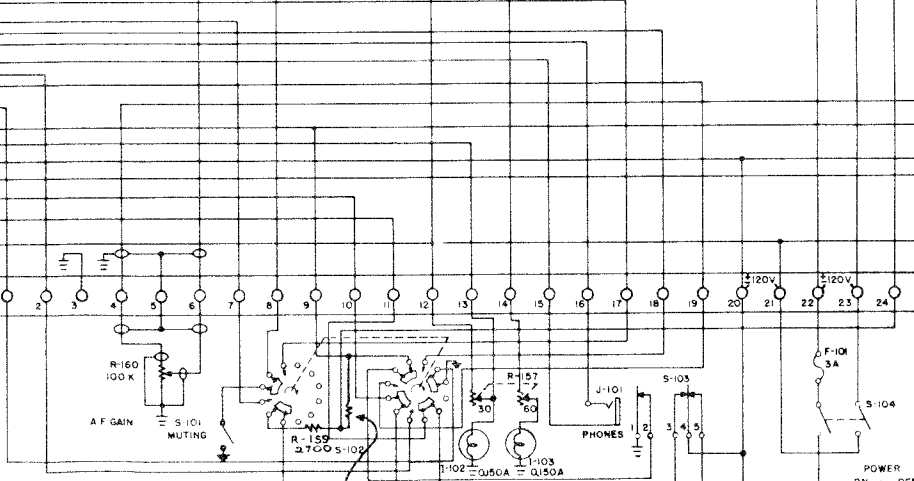
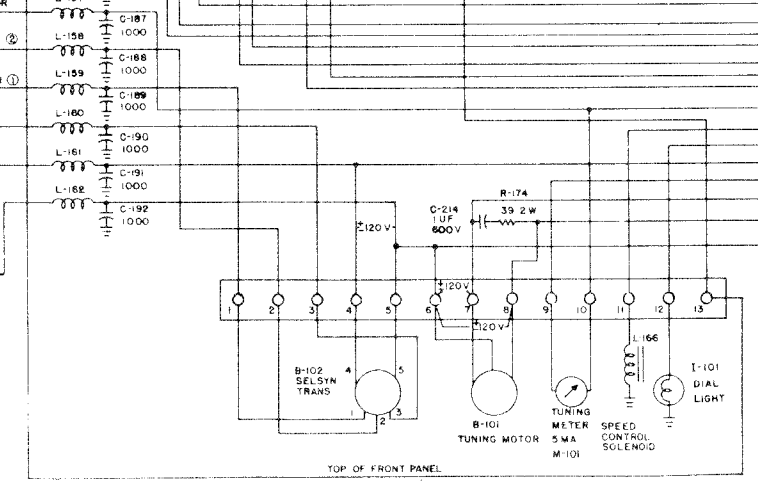
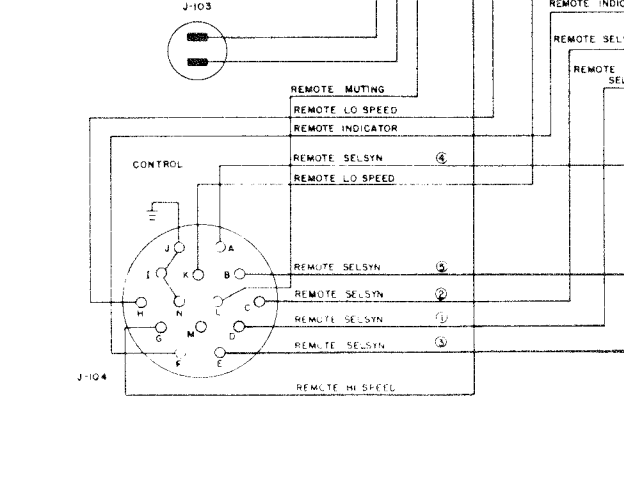
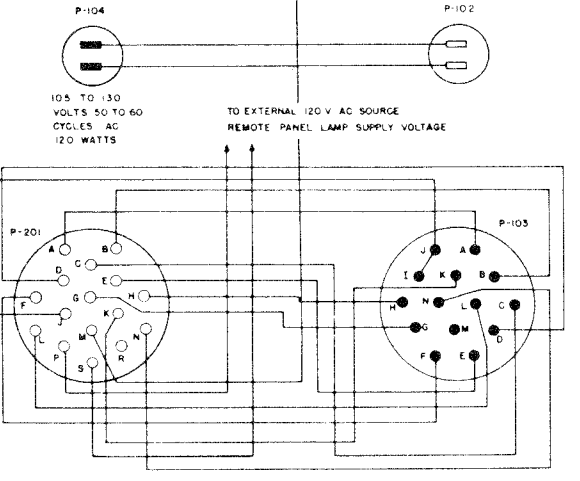
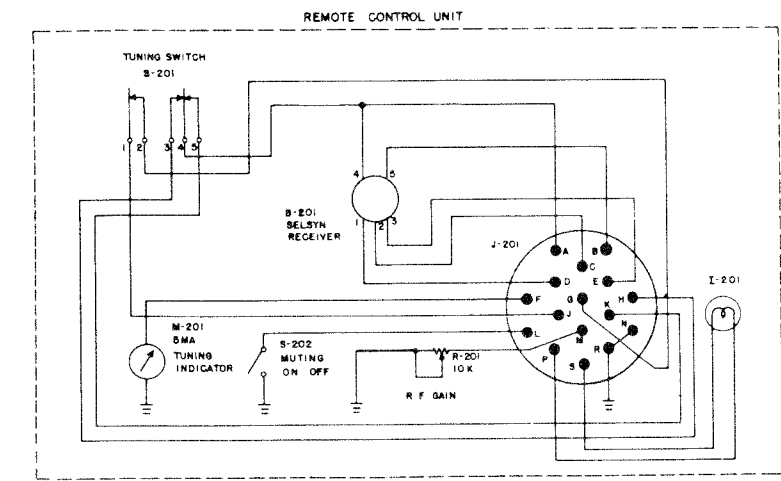
Figure 14-12. Rear View of Receiver





NOTE —
 1. RESISTOR VALUES ARE IN OHMS AND 1/2 WATT UNLESS OTHERWISE SPECIFIED
 2. CAPACITOR VALUES ARE IN UF UNLESS OTHERWISE SPECIFIED
 3. FRONT AND REAR OF SWITCH SECTIONS VIEWED FROM FRONT OR KNOB END OF SWITCH IN EXTREME COUNTER CLOCKWISE POSITION
 4. VOLTAGE READINGS OBTAINED WITH A VTVM HAVING AN INPUT IMPEDANCE OF 11 MEGOHMS, ARE DC AND POSITIVE WITH RESPECT TO CHASSIS UNLESS OTHERWISE SPECIFIED AND WERE MEASURED UNDER THE FOLLOWING CONDITIONS:
 A. LINE VOLTAGE, 120 V AC
 B. CONTROL SETTING
 R.F. GAIN MAX
 R.F. GAIN MIN
 MUTING OFF
 TUNING NORMAL POSITION

LAST C SYMBOL - C-221
 LAST R SYMBOL - R-179
 LAST L SYMBOL - L-167



P-102 & J-102 Modified per EEM 39
 Modified per EEM 34
 Modified per EEM 63

CHAPTER 15

L/MF RADIO RANGE MONITOR RECEIVERS - RIS, RIU, RIV AND RIW

15-1. INTRODUCTION

The receivers in this group operate in the 200-400 kc frequency range and are intended for continuous duty monitoring of *Compass Locator* and *Low Frequency Radio Range* transmitters.

15-2. GENERAL DESCRIPTION

All of these receivers are equipped with both visual and aural alarm indicating devices. There are two indicating lights, one green to indicate normal signal from the equipment being monitored and the other red to show abnormal conditions of the received signal or possible trouble in the monitor itself. There is also a buzzer which sounds when the received signal is not up to certain requirements, or when there is a malfunction in the monitor itself.

Automatic Volume Control is not used; consequently the receivers may easily be overloaded. As a result the RF gain control should never be advanced beyond the setting that is required for proper alarm operation.

The RIU, RIV and RIS receivers are of the *Tuned Radio Frequency* type. The RIV has an improved alarm circuit. The RIW uses the same alarm system as the RIV but its receiver section is a dual conversion superheterodyne which gives superior results as compared with the other three sets. Both RIV and RIW have voltage regulation of their high voltage supply to give more stable operation.

15-3. CIRCUIT ANALYSIS AND OPERATION OF THE RIU RECEIVER

A. *General Description*

This receiver employs three stages of tuned

RF amplification. The detector consists of both diode sections of a duo-diode-triode. The rectified signal is applied to the grid of a power output tube for operation of the loudspeaker and the auxiliary output. The rectified signal is also applied to the grid of the triode section of the duo-diode-triode to control the operation of the alarm actuating relay, RL-1. See RIU schematic, Fig. 15-4.

B. *RF Stage Circuit Description*

Type 6SK7 tubes are used in the three RF stages. Capacitive interstage coupling is used in the output of these three stages to reduce the tendency toward oscillation at the higher frequencies. The use of capacitive interstage coupling also simplifies the design of the RF amplifier, giving relatively sharp selectivity, while at the same time insuring the desired band pass characteristics.

C. *AF Stage Circuit Description*

The output of the 3rd RF stage is rectified in the diode section of V-4. The audio frequency component of the rectified signal drives the grid of V-5 in the usual manner. The direct current component of the rectified signal is utilized to operate the alarm system in a manner which will be explained in detail later.

D. *Types Of Alarm Operation*

A switch SW-2 on the back of the receiver near the relay shield housing permits the selection of either of two forms of alarm operation. The switch positions are marked INSTANT-ANEOUS and TIME DELAY. These two conditions of operation are briefly described as follows:

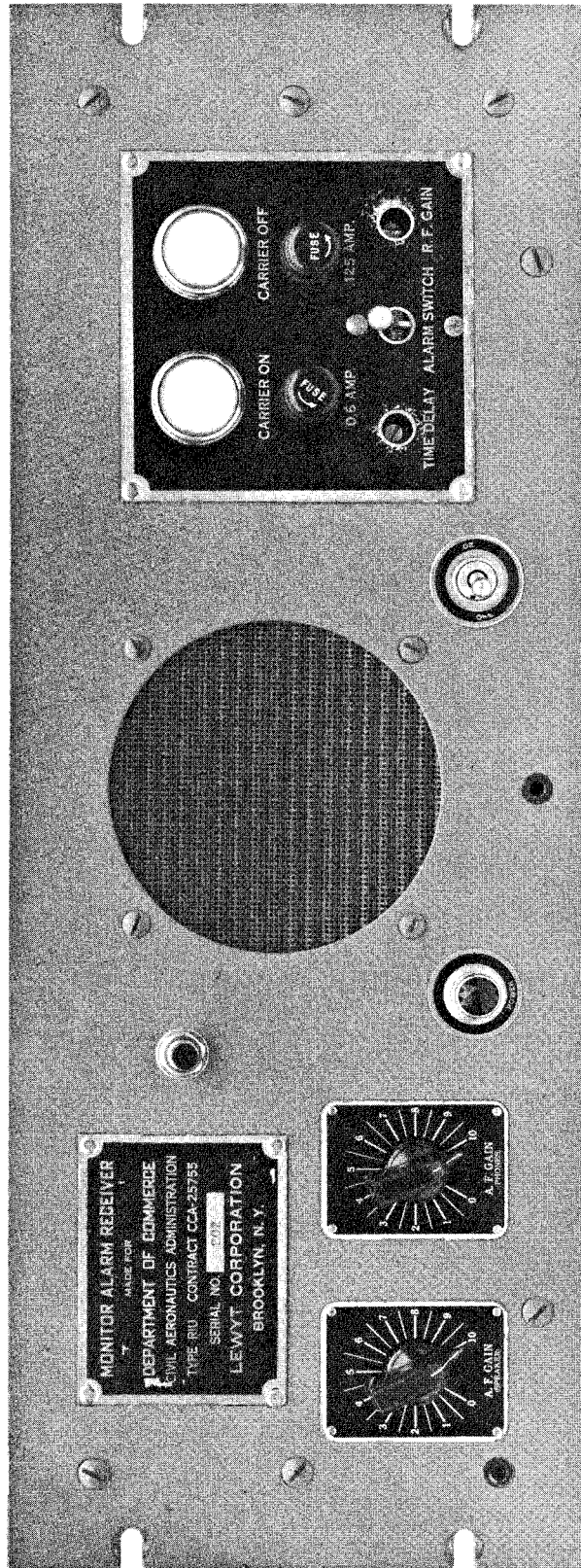


Figure 15-1. RIU Receiver

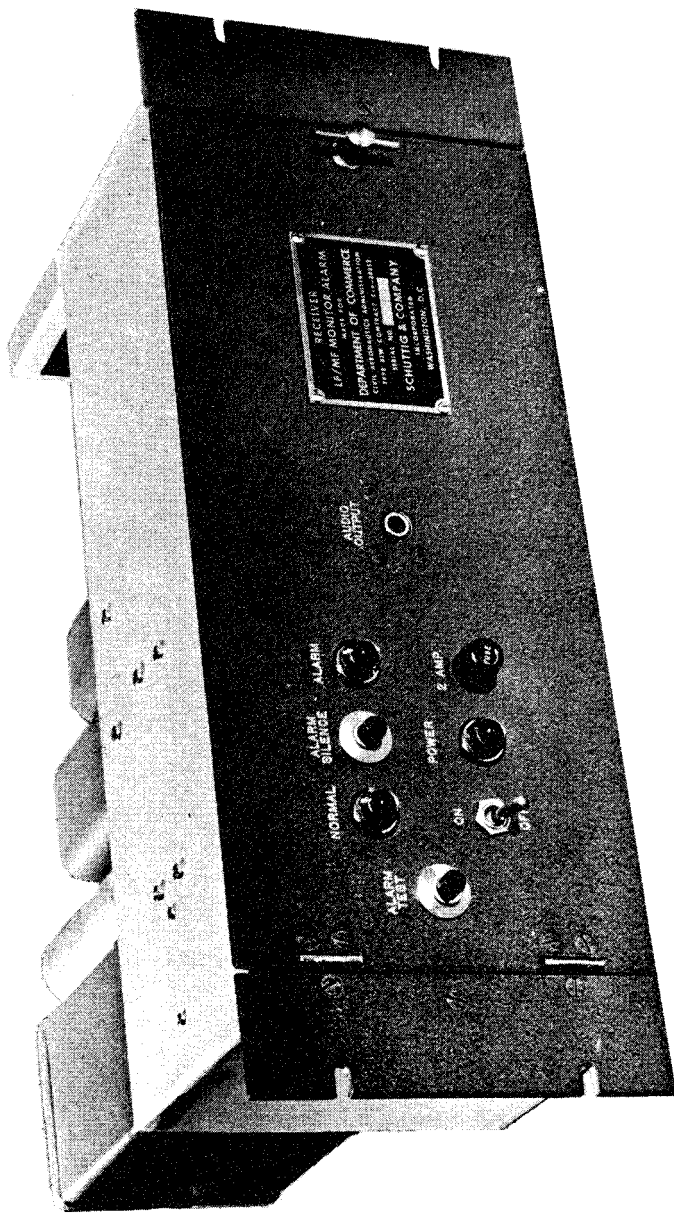


Figure 15-2. RIW Receiver

1. **INSTANTANEOUS ALARM:** With the switch in this position, the aural alarm in the receiver will be operated by short interruptions in the received carrier. An interruption of approximately 1/10 second is adequate to operate the alarm. A capacitor connected across the coil of the alarm relay in the receiver causes the alarm to sound for a second or two, even for such a short carrier-interruption. The *Instantaneous* condition is always to be used for monitoring facilities which radiate a carrier continuously such as a simultaneous radio range.

2. **TIME DELAY ALARM:** In this condition the alarm will not be operated by short carrier interruptions that will occur when the identifications of a facility such as a non-simultaneous range are keyed, but it is possible to so adjust the delay circuits in the receiver that interruptions appreciably longer than those caused by identification keying will set off the alarm. This *Time Delay* condition is always to be used on radio ranges or other facilities which key the carrier during the identification period or which are subject to similar brief carrier interruptions.

3. **MANUAL CONTROL OF AURAL ALARM:** For both *Instantaneous* and *Time Delay*, the alarm in the receiver continues to sound if the carrier interruption persists, unless switch SW-3 on the front panel is momentarily depressed thus *locking out* the alarm, inasmuch as relay RL-2 will remain closed as long as Relay RL-1 is in the closed position. Should the carrier come on again, the lockout circuit opens, so that any subsequent failure of the carrier will again operate the alarm.

E. Operation of Alarm Circuit

1. **INSTANTANEOUS OPERATION:** The operation of the alarm circuit can be followed by referring to the receiver diagram. Relay RL-1 has flowing through its coil the plate current of the triode section of the 6SR7. The diodes of this same tube serve as the detector, and the voltage developed by the detector is applied to the 6SR7 grid. Thus, as the RF input to the receiver increases, the bias on the 6SR7 grid increases, its plate current decreases, and the relay de-energizes, opening the relay contacts. Should the signal being received fail completely or drop to such an extent that the current

through the relay rises sufficiently to again pull the relay armature closed, both relay contacts close. If switch SW-2 is closed (instantaneous position), the relay RL-3 will immediately be energized and the alarm will be sounded.

2. **TIME DELAY OPERATION:** With switch SW-2 in the open, TIME DELAY, position, operation is slightly different from that described in section 15-3-D-2.

When the current through the coil of relay RL-1 increases to the point where the armature closes, one of the contacts on this relay closes the heater circuit of V-7, a type 84 rectifier tube. After the time required for the cathode of this tube to reach operating temperature, current flows through the tube, through contacts on relay RL-2 and energizes relay RL-3 which causes the alarm to sound. It will be noted that switch SW-2 provides *Instantaneous* operation merely by shorting the cathode and anodes of V-7 so that the time delay provided by the heating of this tube is eliminated. The function of the second pole of SW-2 will be discussed in a later paragraph.

3. **ADJUSTMENT OF TIME DELAY PERIOD:** A variable resistor, R-21 has been provided for lowering the heater voltage of V-7 so that the time required for this tube to reach operating temperature may be adjusted. It has been found however, that with R-21 completely out of the circuit so that the heater voltage is 6.3 volts, a time delay of from 10 to 12 seconds occurs between the application of heater voltage and the passing of current from cathode to the anodes. This should, of course, be all that is needed for facilities at which the carrier is interrupted for identification keying, inasmuch as the total required for identification is only 7½ seconds. R-21 should be turned counter-clockwise until the proper delay is obtained.

4. **OPERATING LEVEL OF ALARM CIRCUIT:** Relay RL-1 does not close and open at the same current values. A current of approximately 2.3 ma, is required to close it, and it will remain closed thereafter until the current drops below 1.4 ma. These values of current correspond to the grid bias voltages of approximately -9 and -13 volts respectively on the 6SR7 grid if an unmodulated signal is

being received. For a modulated signal the voltages are somewhat higher, particularly the voltage corresponding to the 1.4 ma current. For the unmodulated signal the ratio of the two voltages is 9/13 or 0.7. For a 30 percent modulated signal the ratio has been found to be about 0.58. Thus, if the gain in the receiver is adjusted so as to just open relay RL-1 when receiving a signal such as from a simultaneous radio range station, the relay will not close again and set off the alarm until there has been a 42 percent drop in field strength. Although the ratio will be even less than 0.58 for 100 percent modulated facilities such as loop type radio ranges, this will present no problem inasmuch as the carrier is interrupted between identification characters; therefore, we are concerned only with the voltage required to release the relay as transmissions start.

5. **EFFECT OF PERCENTAGE MODULATION ON RELAY OPERATION:** As the reason may not be apparent for different degrees of modulation affecting the receiver's performance in the manner outlined above, this condition will be discussed further. With no bias on the 6SR7 grid, the plate current which flows through the coil of relay RL-1 is 5.5 ma. As indicated in the preceding paragraph, a bias of -13 volts produced by an unmodulated signal reduces the current through the relay to 1.4 ma so that it opens. If this unmodulated carrier is now 100 percent modulated, however, so that the instantaneous grid voltage of the 6SR7 varies between 0 and -26 volts (2×13), the plate current will vary between 5.5 and 0 ma, and will no longer have an average value of 1.4 ma. To again get a 1.4 ma current, it will be necessary to increase the average (or carrier) bias to a value well above -13 volts. The current through the coil of relay RL-1 is, of course, substantially free of ripple even when a modulated signal is being received due to the filtering action of the capacitor from the 6SR7 plate to ground and the inductance of the relay itself.

6. **FILTERING 6SR7 BIAS VOLTAGE:** If it is ever found difficult to obtain a sufficiently high signal level on the 6SR7 grid to open the relay with a 100 percent modulation input, the condition can be corrected by filtering the audio modulation from the voltage on the grid so that

only the average (or carrier) voltage is applied. This can be done by connecting a capacitor about 0.001 μ f from the 6SR7 grid to ground, and inserting a 1-megohm resistor directly in the lead from the diode detector circuit to the grid; these components have been shown by dotted lines on the receiver circuit diagram.

F. Audio Gain Control

The two gain controls on the front panel of the receiver are for loudspeaker and headphone levels. The RF gain control is adjusted with a screwdriver through a hole in the front panel, and it should be set to give just a little greater output than is required to keep the alarm from operating.

1. **CONTROL OF AUDIO GAIN BY SWITCH SW-2:** A study of the schematic wiring diagram (Fig. 15-4) will reveal that the same voltage applied to the grid of the 6SR7 is applied to the grid of the 6V6-GT except that dc voltage is blocked off the latter grid by a capacitor. Obviously the 6V6-GT will be overloaded with a signal of about 20 volts, 100 percent modulated, such as that applied to the 6SR7 grid unless some precautions are taken. Inasmuch as switch SW-2 will always be in its open or TIME DELAY position for 100 percent modulated range signals, this switch has been so connected that it opens the cathode by-pass capacitor of the 6V6-GT tube. This greatly reduces the sensitivity of this tube and permits the grid voltage obtained with 100 percent modulated signals to be handled without grid current being drawn.

G. Auxiliary Alarm Light Provisions

An alarm light, utilizing a type S-6 lamp for 115 volt operation (red) has been incorporated and connections to this light have been brought out to terminals on the rear of the chassis so that, if desired, an external alarm light or alarm operating relay can be energized simultaneously. Another indicating light denoting proper operation of the monitored facility has been provided. This light also utilizes a type S-6, 115 volt, lamp (green) and lights when the monitored signal is normal. Connections to this light have also been brought out to terminals on the rear of the chassis.

H. Auxiliary Circuit Provisions

An auxiliary 600-ohm output, brought out to terminals on the rear of the chassis, has been provided, which is suitable for use with a pair of 600-ohm headphones. A 600-ohm pad R-16, marked AF GAIN PHONES on the front panel, has been provided across this circuit. When properly loaded, the operation of this pad will not alter the impedance of the auxiliary output circuit.

I. Addition of Alarm-Test Switch

The circuit of the RIU and RIS receivers has been modified (See EMBI No. 7 and No. 88) to make it possible by use of a switch to simulate a loss of signal from the transmitter being monitored and thus test the operation of the alarm circuit.

15-4. MODIFICATION OF RIU AND RIS RECEIVERS — EEM NO. 238

A. Object

To provide *fail-safe* operation and improve the alarm circuitry of the RIS and RIU receivers.

B. Reason for Modification

The RIS and RIU receivers as originally designed did not provide fail-safe operation.

C. Results of Modification

The 6SR7 has been replaced by a 6SN7GT and the circuitry changed to provide fail-safe operation of the monitor receiver. After modification, operation of the receivers will be as follows:

As the RF input to the receiver increases, bias on the 6SN7GT (V-4 triode section) decreases, increasing the plate current and causing RL-1 to operate. This requires about 4 microvolts RF input to the receiver depending on the setting of the new 50,000 ohm potentiometer (replaced TIME DELAY potentiometer). Correspondingly, a decrease in RF input will increase the 6SN7GT (V-4 triode section) grid bias and cause less plate current to flow, releasing RL-1 and producing an

alarm. This value of input to the receiver will be about $2.7 \mu\text{v}$ (full RF gain). A somewhat lowered sensitivity may be encountered at the extreme ends of the tuning range of the receiver. With SW-2 in the instantaneous position, the alarm will take place in about 1 second. With SW-2 in the time delay position, the discharge of the $5 \mu\text{f}$ capacitor through the 1 megohm resistor will take approximately 12 seconds to lower the grid voltage of the V-4 triode section sufficiently to cause RL-1 to drop out. Also, a fuse, tube or component failure will cause RL-1 to release and give an alarm indication.

The new potentiometer R-26, installed in place of old R-21, serves as a means for adjusting the alarm to the desired point, based on reduced signal input. The adjustment of RL-1 also effects the point of alarm. The over-all sensitivity adjustment should be such that, to the extent practicable, a reduced carrier will cause an alarm regardless of background noises (QRM, QRN).

In lieu of the remote monitoring feature originally provided for the RIS and the auxiliary alarm light provisions for the RIU, both receivers will now have the form C contacts of relay RL-1 brought out to a terminal for external connections, as may be required.

The operation and utilization of switches SW-1, SW-2 and SW-3 have not been affected by the modification. The purpose of the new switch SW-4 is to provide a means of disabling the local alarm (buzzer) without disturbing the operation of SW-3.

D. Tests After Modification

Connect a signal generator to the antenna terminals and align the receiver in the manner prescribed in the maintenance section of the instruction book. With carrier 30% modulated by a 400 cps signal, full RF gain and SW-2 in the instantaneous position,

- (1) A reduction of level to approximately $2.7 \mu\text{v}$ should cause the receiver to alarm, and,
- (2) The signal-to-noise ratio should be approximately 19 db for the RIU and 15 db for the RIS.

These values may be slightly higher on the extreme ends of the tuning range.

15-5 THE RIW RECEIVER

A. Radio Frequency Section

1. **INPUT:** Signals are fed into the receiver input through either of two input connections on terminal board TB-1 (Fig. 15-3). Terminals 2 and 3 are for connecting an unbalanced 70 ohm transmission line to the input; terminals 1 and 2 are for connecting a whip or wire antenna. Terminal 2 is the ground connection for both antennas. The antenna is coupled to the RF amplifier through the permeability-tuned antenna coil contained in Z-1. The signal is applied to the RF amplifier through the RF gain control, R-2 a potentiometer controlling the amount of signal reaching the grid of the RF amplifier. The RF amplifier, V-1 is a type 6BA6 pentode, resistance coupled to the following stage.

2. **FIRST MIXER STAGE:** The first mixer V-2 is a type 6BE6 pentagrid converter. The RF signal from V-1 passes through a low-pass filter, Z-2 before reaching the third grid of V-2. Z-2 attenuates all frequencies above 420 kc and thereby helps eliminate spurious response above this frequency. This filter also helps attenuate oscillator signals that might appear on the antenna. V-2 is connected in such a way that it combines the functions of both local oscillator and mixer. L-7, C-8 and C-35 form the tank circuit of the local oscillator. C-35 is a temperature compensating capacitor used to stabilize the oscillator frequency. C-7 is the oscillator coupling capacitor and R-6 is the grid leak resistor. R-5 is the cathode resistor and C-14A is the cathode bypass capacitor. The local oscillator operates in the approximate range 1300 to 1600 kc, producing an intermediate frequency of about 1145 kc.

3. **SECOND MIXER STAGE:** Z-4, the 1145 kc IF transformer, couples the output of V-2 to V-3, a type 6AS6 pentode. Like V-2, this tube is a combination oscillator-mixer. A special space-charge oscillator circuit provides high conversion efficiency. The oscillator tank circuit consists of L-11 and capacitors C-22 and C-23. C-23 is a temperature compensating capacitor which minimizes frequency drift due to temperature change. L-10 is a tickler coil

in the plate circuit of V-3, and provides oscillator feed-back by inductive coupling to L-11. The nominal frequency of the oscillator circuit in this stage is 995 kc. When mixed with the 1145 kc first IF, the 995 kc oscillator output converts it to the second IF, 150 kc. The output of V-3 is coupled to V-4 through a 150 kc IF transformer, Z-6, consisting of a tuned primary and secondary. This IF signal is amplified by V-4, a type 6BA6 pentode used in a conventional IF amplifier circuit.

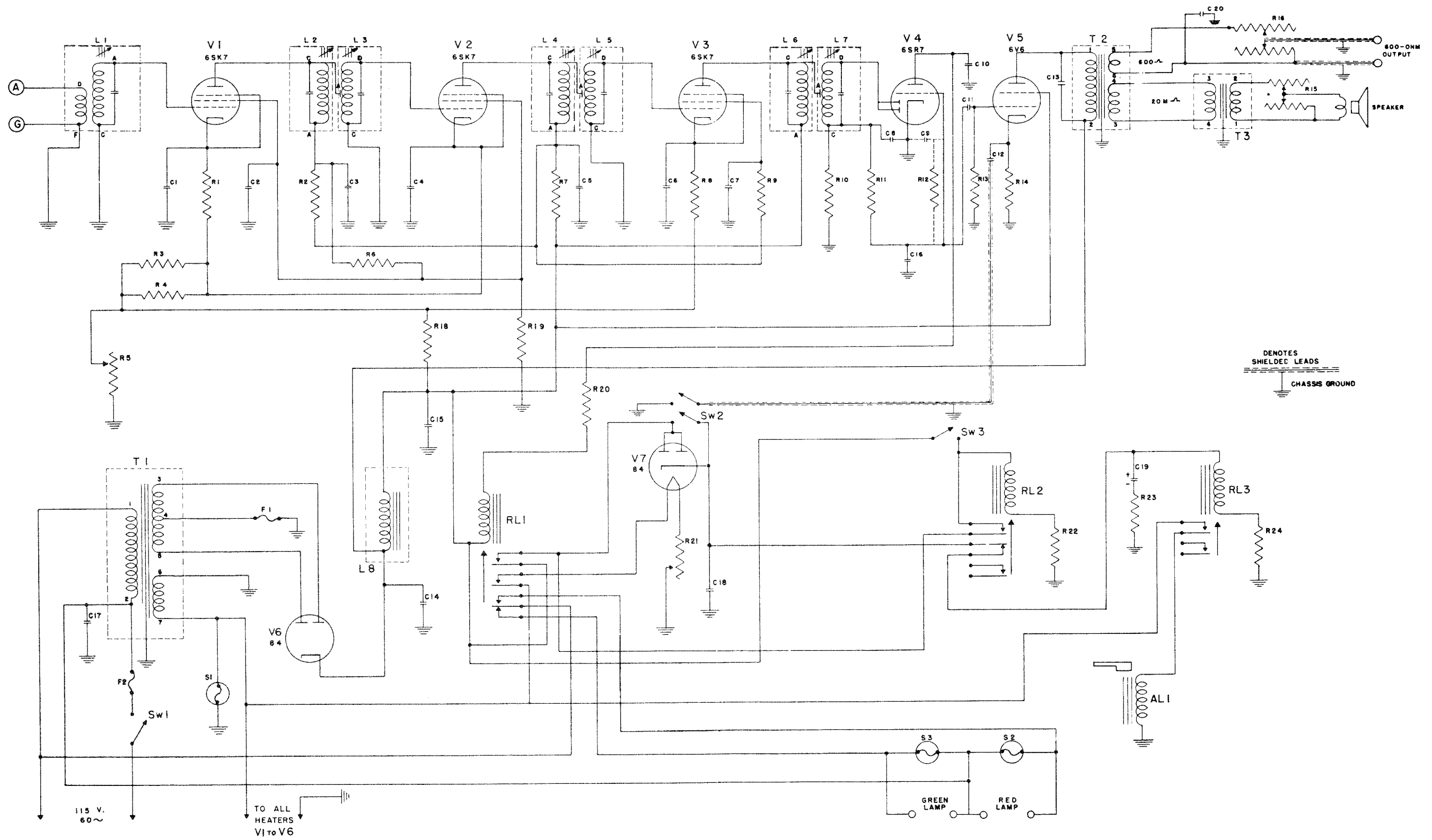
4. **DETECTOR:** The output of V-4 is coupled to the detector, V-5 by means of IF transformer Z-7. V-5 is a type 6AL5 dual-diode. This tube is connected to provide two separate outputs. One section of V-5, pins 2 and 5, develops a dc voltage across load resistor R-41, ALARM SENSITIVITY control. This dc voltage is filtered by C-24, R-15 and C-13C to remove audio modulation and then applied to the grid of V-8.

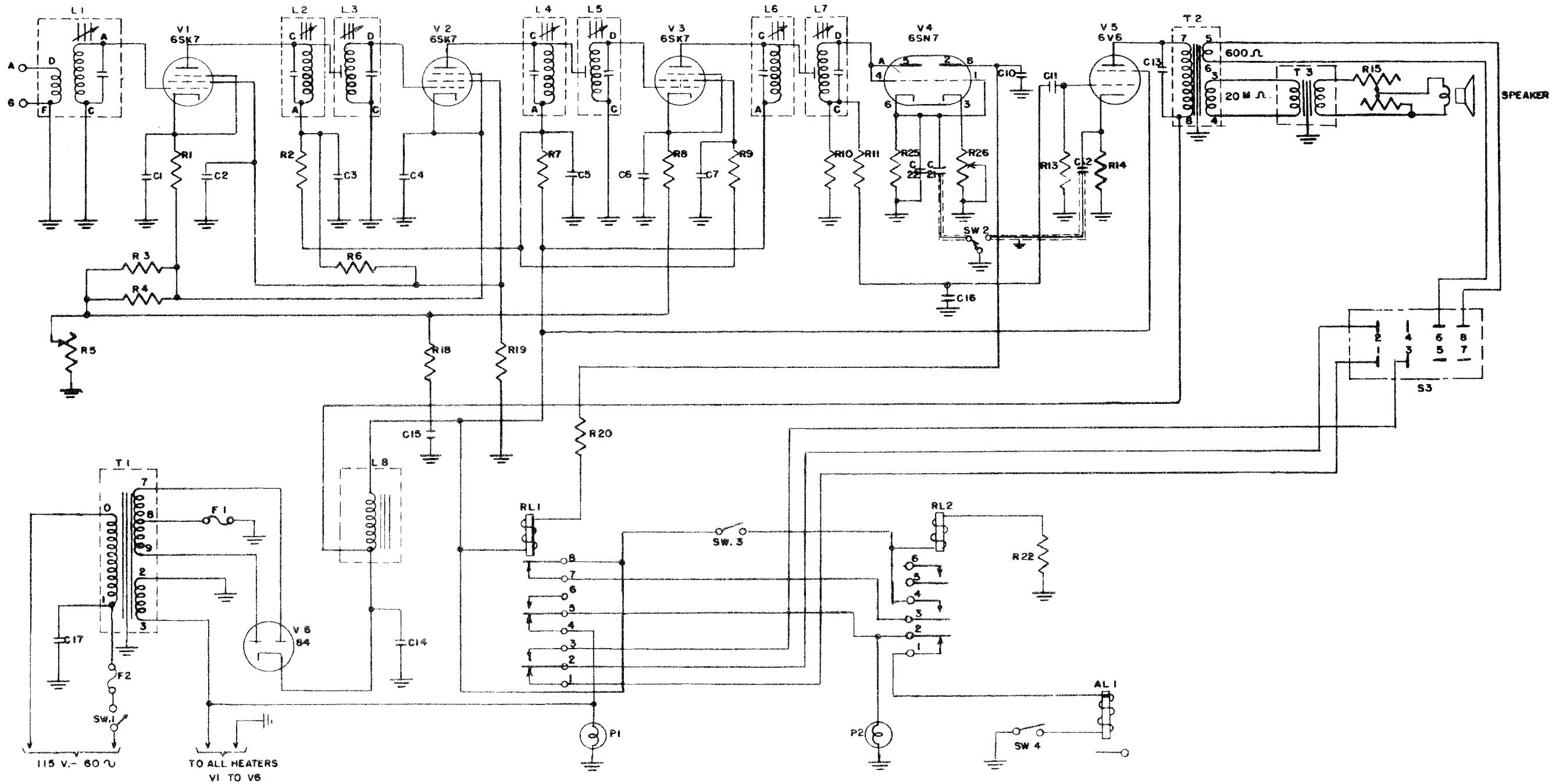
The alarm sensitivity control is a potentiometer used as a voltage divider to apply a portion of the detector voltage to the alarm section. The voltage appearing at the grid of V-8 determines whether the receiver indicates ALARM or NORMAL. Therefore, the RF level at which the receiver goes into alarm is determined by the adjustment of the RF gain control, R-2 and the alarm sensitivity control R-41.

B. Alarm Section

DC voltage from the detector load is applied to the grid of the alarm tube, V-8 via R-15. V-8 is a type 6BN6 gated beam tube. In this application, V-8 is biased such that it is non-conducting when the output voltage from V-5 drops below approximately $4\frac{1}{2}$ volts, and conducting when this voltage exceeds about 5 volts. The exact signal level at which V-8 is cut off is determined by the adjustment of the alarm sensitivity control, R-41. Cathode bias for V-8 is provided by R-14 and R-16. R-14 has a negative temperature coefficient and stabilizes the alarm response over large variations of ambient temperature. R-39, R-24,

R-23 and R-31 form a voltage divider from which is obtained voltages for cathode, screen and quadrature grid of V-8. Whenever the line voltage shifts up or down, these voltages maintain the same relationship to each other and help prevent changing the receiver alarm characteristics.





SCHEMATIC DIAGRAM
RIS
MODIFIED PER EEM#238

10-4-57

FC-141-1

Figure 15-5. RIS Receiver

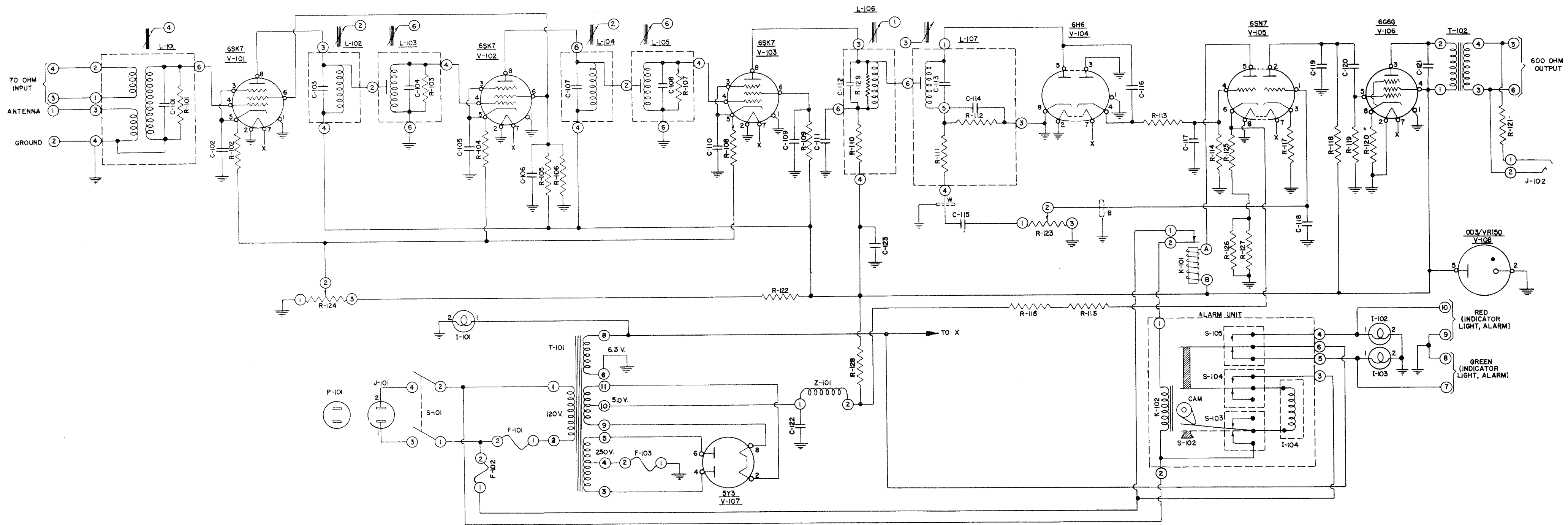


Figure 15-6. RIV Receiver