DESIGN OF A LOGARITHMIC RESPONSE RECEIVER

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PREFACE

Many design requirements for electronic receivers in the past have demanded that the receiver be useful over an extremely wide range of input amplitudes. Accordingly, much work has been done in developing receiver gain characteristics that are approximately logarithmic in nature. The primary purpose of this type response, however, has been to avoid receiver overload and subsequent distortion with high level signal input. For this reason, responses that only approximate the logarithmic curve were developed.

The design requirements for this receiver required that an exact logarithmic response be developed in order that the output signal be a function of the input. Logarithmic recorder paper can be used in the recorder and the whole system calibrated to provide a direct readout of the input signal amplitude.

It is with the requirement for a purely logarithmic response that this investigation is primarily concerned.

This receiver was intended to be a subsystem of an ultrasonic system designed for bio-medical research. The entire system in specifically designed for Trans-Thoracic ultrasonic investigations and measurement of the Hemodynamic pulse.

The entire system was conceived and developed by Dr. Harry D. Crawford, Oklahoma State University. It was he who provided the interest and support in this undertaking. To him I am deeply indebted. I am grateful for the help and guidance provided by Dr. Harold T. Fristoe with the circuitry portion of this investigation as well as Bill Hugh Terrel, Larry Lee Lowcock, Carl Edward Hittle, Gary Lewis Johnson, and the rest of Dr. Crawford's research staff. I also acknowledge the United States Air Force, who made this investigation possible, and the National Institute of Health whose grant provided the financial support for this undertaking.

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CHAPTER I

INTRODUCTION

This receiver was designed as an ultrasonic receiver in the 2.25 to 2.5 mc. range for use in bio-medical research. Many unusual design requirements were imposed because of the associated electronic equipment and the nature of the bio-medical investigation. Accordingly, some of these considerations will not be discussed in this work.

The receiver is required to receive signals from an ultrasonic transducer after passing through a biological subject. The nature of the specimen determines to a great degree the amplitude of the incoming signal. For example, one of the primary areas of investigation is the Thoracic region (an ultrasonic signal is emitted from a transducer on one side of the region and received on the other). When the subject inhales, the ultrasonic beam travels through the lung area, which is predominately air, greatly attenuating the signal. When the subject exhales the ultrasonic beam travels through an area which is predominately tissue, providing very little attenuation. The difference between these two signals is approximately 80 db. Thus the receiver must accept signal strength variation over 80 db. and still provide a recorder output that is an accurate indication of the input signal. This requires the response to be as exactly logarithmic as possible in order for the output to be directly recorded on logarithmic ruled paper.

The ultrasonic system is essentially a closed system; therefore, immune to atmospheric noise. This added to the possibility of detecting signals of very low amplitude places a premium on low internal system noise and high sensitivity.

In addition to the amplitude modulation caused by breathing, other modulations of lesser intensity, such as FM, have been detected by previous equipment; therefore, the recorder output circuitry must have a frequency response from DC to at least 20 KC.

Parameters and Their Measurement

The frequency range of the receiver is 2.25 to 2.5 mc which is well below the frequency cut-off limits of all transistors used. All of the circuits in the receiver system are considered small signal with the exception of the audio power amplifier section and will be analyzed accordingly. The hybrid h parameters and their associated equivalent circuit seems most useful under these conditions and will be the predominant means of analysis.

The parameters were obtained by measuring the transistor under operating conditions with the circuits and procedures developed by Daniel M. Rukavina [1]. The circuit for measuring h_{11} and h_{21} is shown in Fig. 1-1. Figure 1-2 is the circuit for measuring h_{21} and h_{22}. Tables 1-1 and 1-2 provide measurement correction data and h,z,y parameter conversion formulas.

Method of Design

The design of the receiver was approached by determining the overall design requirements of system then dividing the system into smaller subsystems. The smaller subsystems were then designed independently with consideration of the input and output characteristics, available power, etc. only as design requirements of that subsystem. In this manner, each subsystem could be considered more or less independently yet still be an integral part of the whole receiver.

Figure 1-3 shows a block diagram of the receiver with each block representing a separate subsystem.

The remaining chapters will consider each subsystem individually.

Design Requirements

Very definite requirements were imposed upon the input circuitry by the external circuitry and conditions of operation. The receiver was designed for use with a lead metaniobate ultrasonic transducer especially manufactured for this application by Automation Industries of Boulder, Colorado. The transducer, with 25 feet of RG-59/u coaxial cable, presented a nominal generator or source impedance of 14-j90 ohms

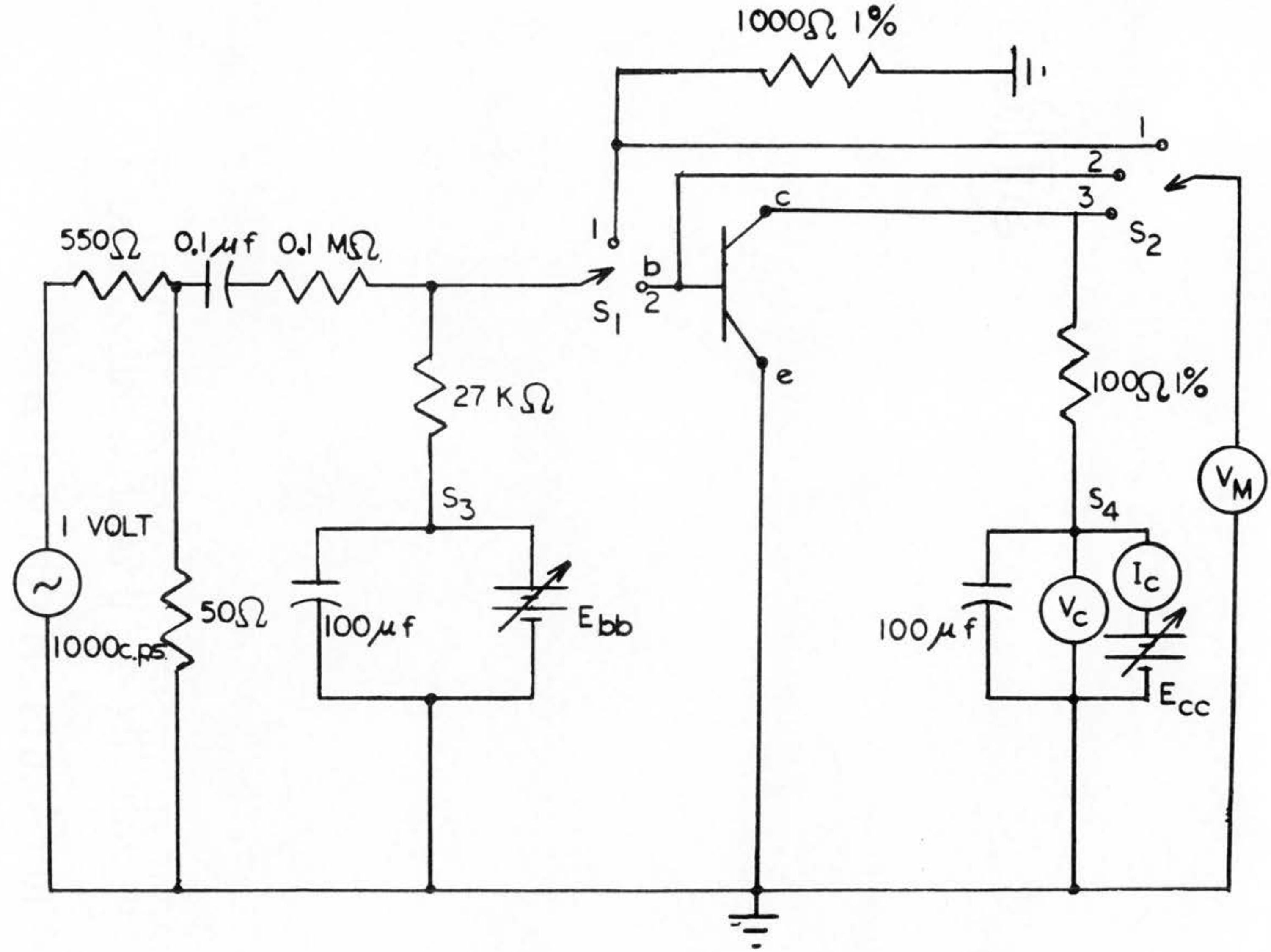


Figure 1-1. Circuit for Measuring h_{11} and h_{12}

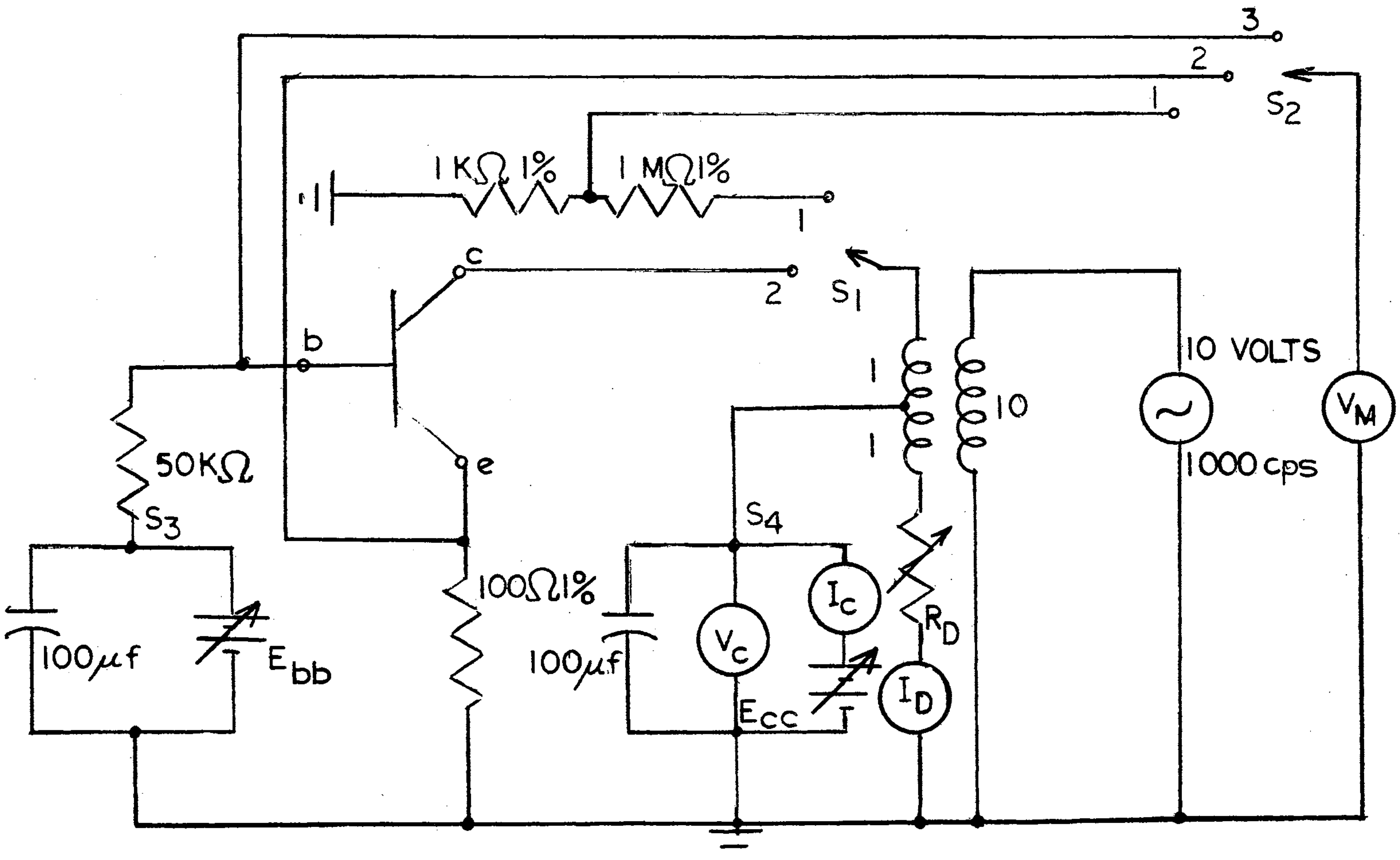


Figure 1-2. Circuit for Measuring h_{12} and h_{22}

TABLE 1-1

h PARAMETER MEASUREMENT CORRECTION DATA

Measured Value	Source Error Factor (S.E.F.)	Termination Error Factor (T.E.F.)	Parameter
$\dfrac{v_1}{i_{1\,cal}}$	$\dfrac{z_{s_1} + z_{1\,cal}}{z_{s_1} + h_{11}}$	$1 - \left(\dfrac{h_{12}\,h_{21}}{h_{11}\,h_{22}}\right)\left(\dfrac{1}{1+\dfrac{1}{h_{22}\,z_{T_1}}}\right)$	h_{11}
$\dfrac{v_1}{v_{2\,cal}}$	$\dfrac{\left(1 - \dfrac{z_{s_2}}{z_{s_2} + i/h_{22}}\right)}{\dfrac{z_{2\,cal}}{z_{s_2} + z_{2\,cal}}}$	$1 - \dfrac{1}{\left(1 + \dfrac{z_{T_2}}{h_{11}}\right)}$	h_{12}
$\dfrac{i_2}{i_{1\,cal}}$	$\dfrac{z_{s_1} + z_{1\,cal}}{z_{s_1} + h_{11}}$	$1 - \dfrac{1}{\left(1 + \dfrac{1}{h_{22}\,z_{T_1}}\right)}$	h_{21}
$\dfrac{i_2}{v_{2\,cal}}$	$\dfrac{\left(1 - \dfrac{z_{s_2}}{z_{s_2} + 1/h_{22}}\right)}{\dfrac{z_{2\,cal}}{z_{s_2} + z_{2\,cal}}}$	$1 - \left(\dfrac{h_{21}\,h_{12}}{h_{22}\,h_{11}}\right)\left(\dfrac{1}{1 + \dfrac{z_{T_2}}{h_{11}}}\right)$	h_{22}

TABLE 1-2

h, y, z, PARAMETER CONVERSION EQUATIONS

z_{11}	$1/y_{11}(1 - \tau)$	$h_{11}/(1 - \tau)$
z_{12}	$-y_{12}/y_{11}\ y_{22}(1 - \tau)$	h_{12}/h_{22}
z_{21}	$-y_{21}/y_{11}\ y_{22}(1 - \tau)$	$-h_{21}/h_{22}$
z_{22}	$1/y_{22}(1 - \tau)$	$1/h_{22}$
$1/z_{11}(1 - \tau)$	y_{11}	$1/h_{11}$
$-z_{12}/z_{11}\ z_{22}(1 - \tau)$	y_{12}	$-h_{12}/h_{11}$
$-z_{21}/z_{11}\ z_{22}\ (1 - \tau)$	y_{21}	h_{21}/h_{11}
$1/z_{22}(1 - \tau)$	y_{22}	$h_{22}/(1 - \tau)$
$z_{11}(1 - \tau)$	$1/y_{11}$	h_{11}
z_{12}/z_{22}	$-y_{12}/y_{11}$	h_{12}
$-z_{21}/z_{22}$	y_{21}/y_{11}	h_{21}
$1/z_{22}$	$y_{22}/(1 - \tau)$	h_{22}
$\tau = z_{21}\ z_{12}/z_{11}\ z_{22}$	$\tau = y_{21}\ y_{12}/y_{11}\ y_{22}$	$\tau = \dfrac{1}{1 - h_{11}\ h_{22}/h_{21}\ h_{12}}$

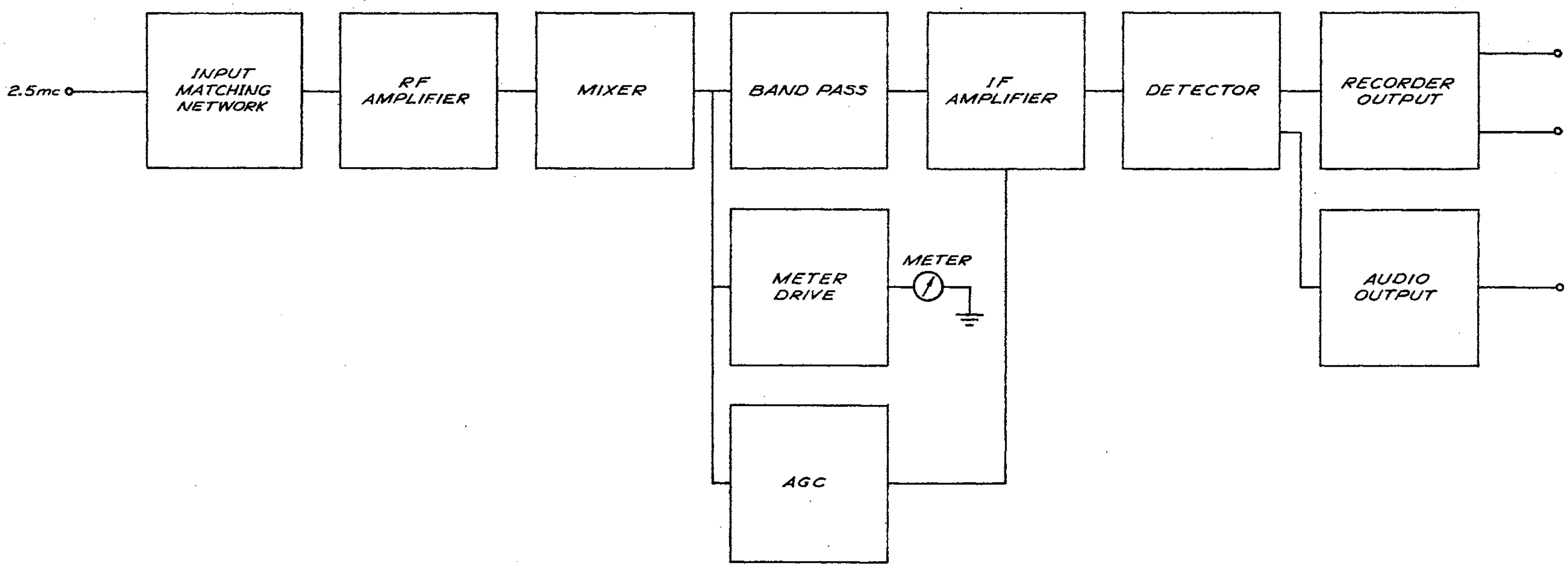


Figure 1-3 Receiver Block Diagram

at 2.5 mc. This impedence must be properly matched by the receiver input circuitry. The maximum voltage anticipated was 10 mv and the receiver must be sensitive over a range of at least 80 db. This required the RF circuitry to have a minimum sensitivity of at least 1 microvolt.

Since the incoming signal to the transducer is ultrasonic, the problem of external noise and antenna noise is not present as it would be in ordinary communications receivers; therefore, low noise circuitry presented a real advantage in this application, much the same as in receiver design for the UHF frequency range. For this reason, much thought was given to developing good low noise circuitry.

CHAPTER II

RF AMPLIFIER

Circuits Considered

Primary consideration in the design of the RF Amplifier portion of the receiver was to develop a low noise, stable circuit. Three basic circuits were considered; a cascade amplifier, a cascode amplifier, and a parametric type amplifier. The cascode amplifier consists of two or more common emitter amplifiers coupled together by transformer coupling or other means. This type of amplifier, though commonly used in the lower frequency ranges, was not considered further because of the practical noise figure attainable. The cascode amplifier is a direct coupled common emitter, common base amplifier which combines the best characteristics of both. The theoretical noise figure is the same as that of a cascade circuit but the practically attainable noise figure is much lower. It essentially provides a single stage noise figure with a two stage gain. The cascode amplifier is traditionally used in the VHF frequencies as RF amplifiers because of their low noise characteristics.

The parametric amplifier is usually employed in the micro-wave frequency region because of its low noise characteristics but a few attempts have been used at the lower[2] frequencies around 30 kilocycles[3] and at broadcast frequencies.[4] The parametric amplifier considered is an adaptation of an amplifier developed by Planinac.[4] This circuit was especially attractive because of its use of junction diodes. Fig. 2-1 is a junction diode parametric amplifier considered for this application. The lowest attainable noise figure for the parametric amplifier is 0.9 db.[5] Although the parametric amplifier has a lower noise figure it is also more complicated, more expensive, and more difficult to construct. With the availability of extremely low noise, low cost, transistors a cascode amplifier can be designed with approximately 1 db noise figure. For this reason, the cascode amplifier was selected for the RF amplifier portion of the receiver.

Design of the Input Matching Network

The cascode amplifier was selected because of its simplicity and the low noise figure attainable. The design input impedance was a nominal 10-j100 ohms to match the ultrasonic transducer input.

This required a matching network to match 10-j100 ohms required by the cascode amplifier. This was accomplished by the use of a series inductor and an L matching network. The series inductor neutralized the capacitive reactance and the L network matched the impedances. Figure 2-2(a)

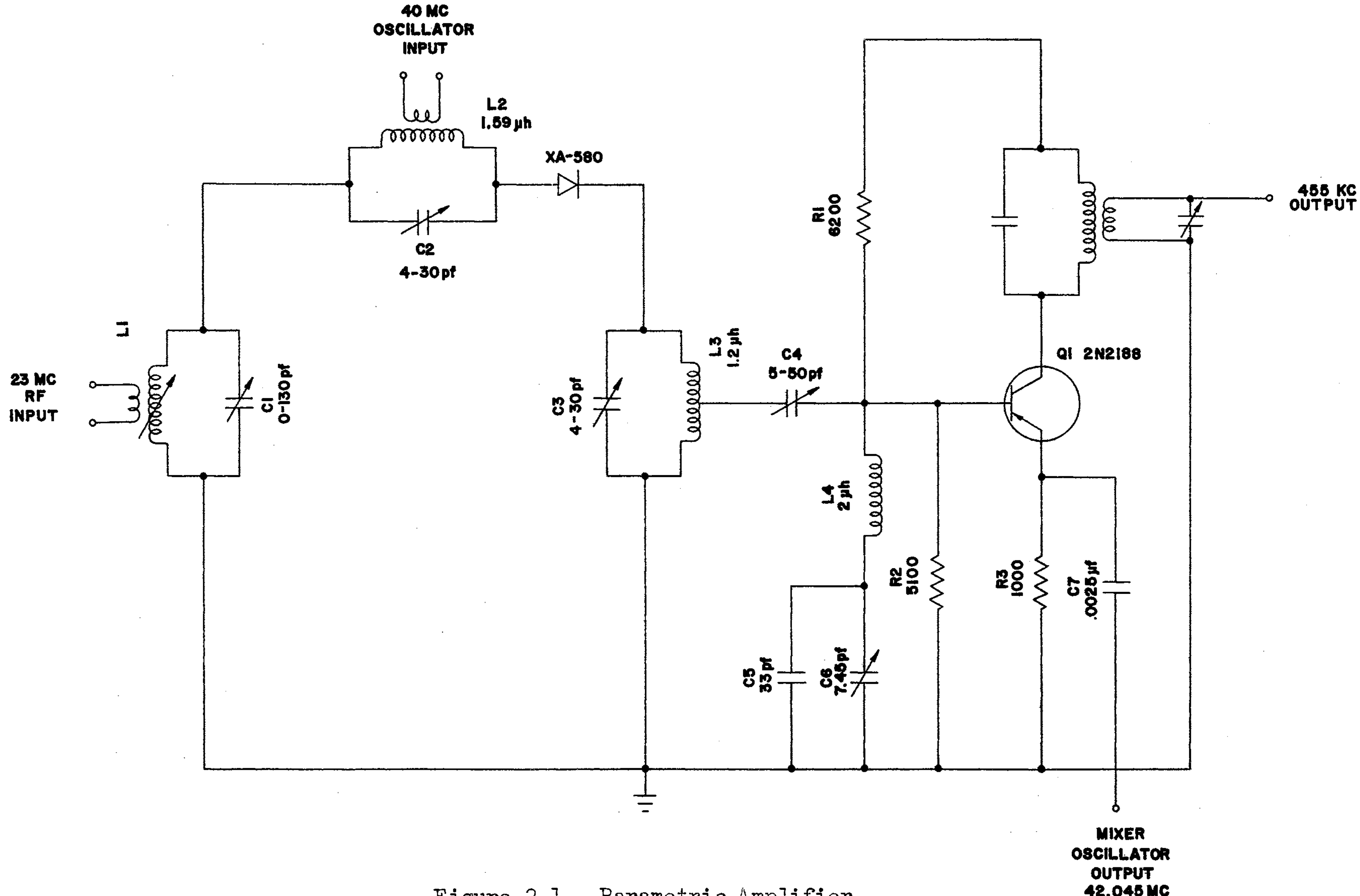


Figure 2-1. Parametric Amplifier

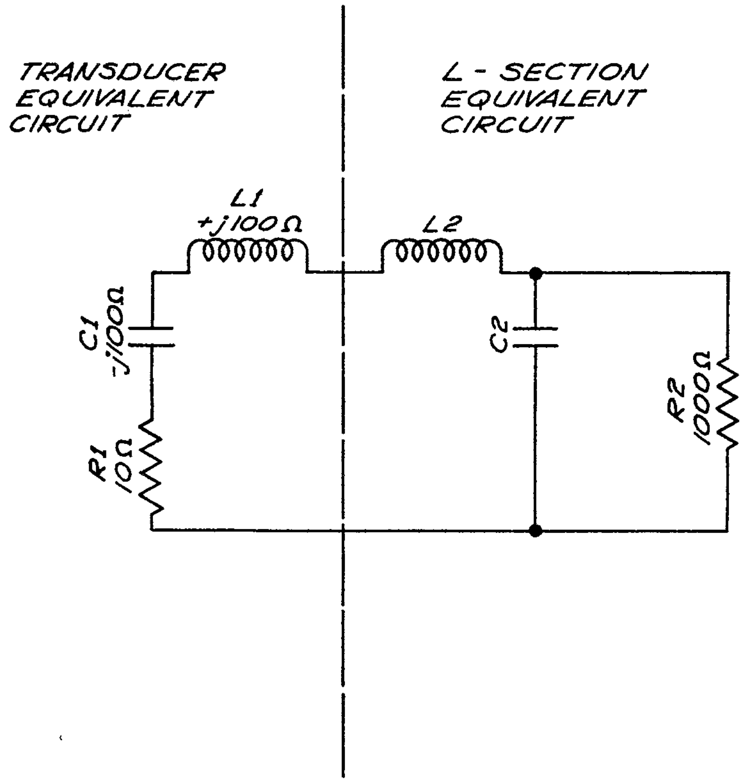


Figure 2-2a. Matching Network Equivalent Circuit

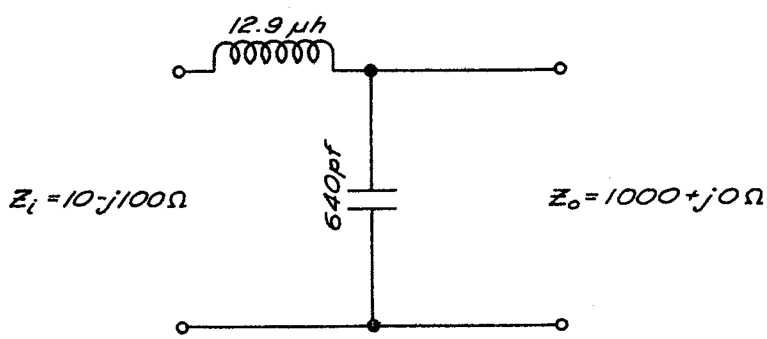


Figure 2-2b L-Matching Network

shows the lossless equivalent circuit.

The desired value of L_1 is such that it presents a reactance of 100 ohms at 2.5 mc in order to offset the capacitive reactance of the transducer.

$$Z = jwl = j100$$

$$L_1 = \frac{Z}{jwl} = \frac{j100}{j(2.5 \times 10^6)\ (2\pi)} = 6.5\mu h \qquad (1)$$

The L-matching network was designed with reference to H. T. Fristoe's "Q Equations"[6]

$$\frac{R_2}{R_1} = \frac{1000}{10} = 100$$

From chart II-I[6]

$$Q = 10$$

$$L_c = .016$$

$$C_c = .016 \qquad (2)$$

$$L_2 = \frac{L_c R_2}{F_r}\ \mu h$$

$$L_2 = \frac{(.016)\ (1000)}{2.5} = 6.4\ \mu h \qquad (3)$$

$$C_2 = \frac{C_2}{F_r R_1} \times 10^8 \text{ pf.}$$

$$C_2 = \frac{.016}{(2.5)\ (1000)} \times 10^8 = \frac{1.6 \times 10^{-2}}{2.5 \times 10^3} \times 10^8$$

$$C_2 = \frac{1.6 \times 10^3}{2.5} = .64 \times 10^3 = 640 \text{ pf.} \qquad (4)$$

Inductors L_1 and L_2 are combined in one resulting in the final matching network as shown in Fig. 2-2(b).

Referring to Figure 2-3, the first two transistors, Q_1 and Q_2, comprise the cascode RF amplifier. This figure will be referred to throughout the design analysis of the cascode amplifier.

Design of the Cascode Amplifier

The L-matching network provided an input impedance of 1000 ohms to the cascode amplifier. This was done in order to operate the amplifier under minimum noise conditions.

The transistor selected for the Cascode RF Amplifier was the TI-363, a low noise device manufactured by Texas Instruments, Inc. Under proper operating conditions, this transistor has a noise figure of 1 db at 2 mc and room temperature.

The design criteria for minimum noise operation of the cascode amplifier was,[7]

$$\begin{aligned} V_{ce} &= -6 \text{ volts} \\ I_E &= -.5 \text{ ma} \\ R_g &= 1000 \text{ ohms} \\ T_A &= 25^\circ \text{ C} \\ F &= 2.5 \text{ mc} \\ S &= 7 \end{aligned} \tag{5}$$

Where: V_{ce} = the collector to emitter voltage (volts)

I_E = Emitter current (ma)

R_g = generator resistance

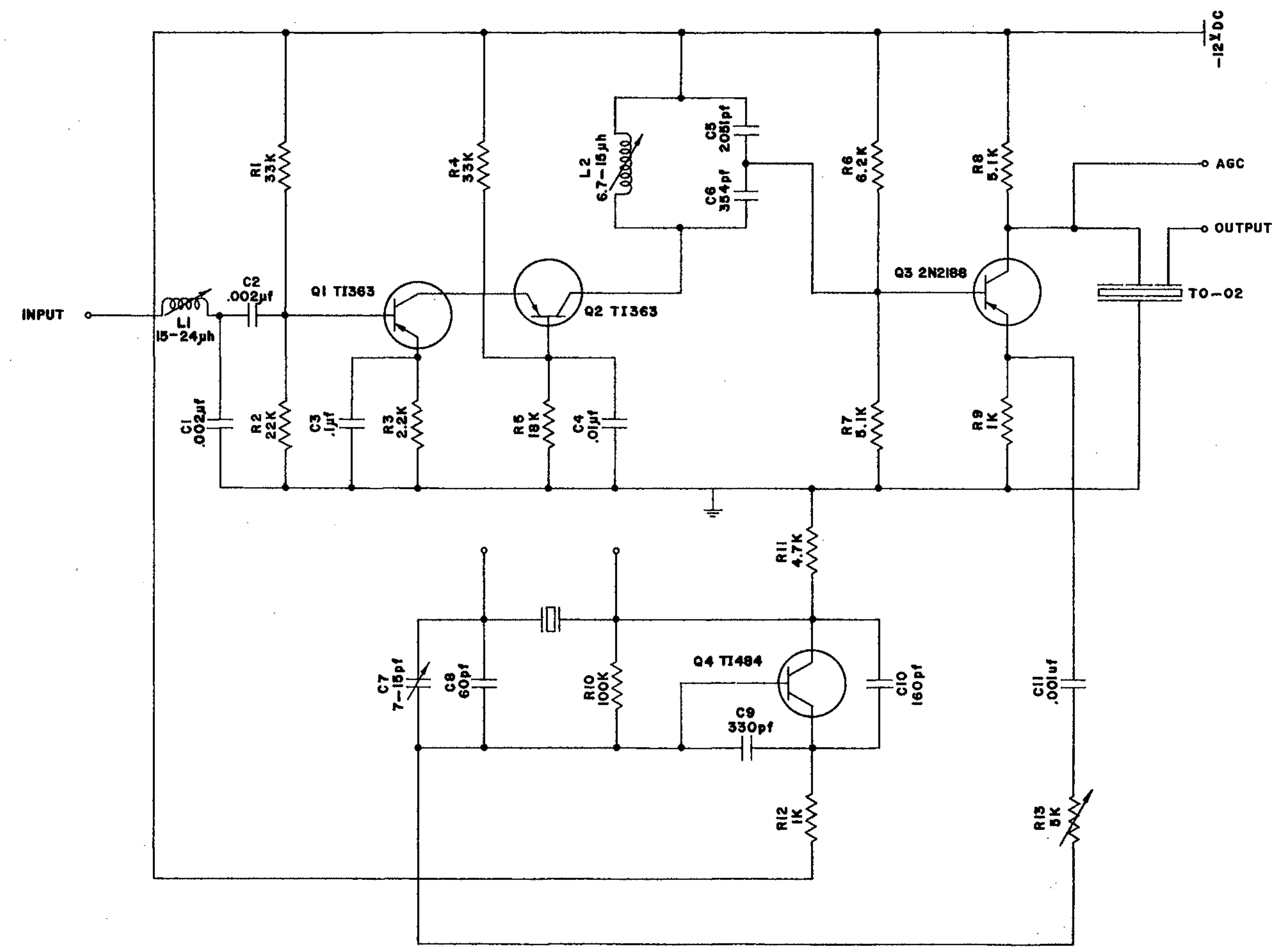


Figure 2-3. RF Amplifier, Mixer and Local Oscillator. Board 1

T_A = Test temperature (^oC) (6)

F = Frequency

S = Stability factor

The stability factor (S) of 7 was selected to provide a reasonable temperature stability without too much loss or gain.

$$S = \frac{I_c}{I_{co}} = \frac{R_3 + R_b}{R_3 + (1 - \propto R_b)} \quad (7)$$

R_3 is fixed by emitter current considerations at 2200 ohms. R_b consists of R_1 and R_2 in parallel. Since $\propto$ is .98 then $1-\propto$ is approximately zero and we can neglect the term $(1-\propto)R_b$. Solving for R_b,

$$S = \frac{R_3 + R_b}{R_3}$$

$$R_b = R_3 \ (S-1) \quad (8)$$

Substituting in values we get,

$$R_b = (2.2 \times 10^3) \ (7-1) = 6 \ (2.2) \times 10^3$$

$$R_b = 13.2 \times 10^3 \text{ ohms} \quad (9)$$

For proper bias, R_b consisted of the parallel combination of 33 K ohms for R_1, and 22 K ohms for R_2. R_4 and R_5 were similarly proportioned in order to properly bias Q_2. Capacitors C_3 and C_4 were chosen so as to properly bypass R_3 and R_5. Bypassing R_5 places the base of Q_2 at ground potential as far as the signal is concerned. A quiescent point for the cascode amplifier was determined by graphical analysis. Maximum current considerations set the value of

R_3 at 2200 ohms. This fixed the impedance of the parallel tuned circuit formed by L_2 and the series combination of C_5 and C_6.

In the actual circuit L_2 is a variable inductor tuneable over the range from 6.7 to 15 μh to resonate with a nominal 305 pf capacitor. The design Q is 10. Capacitors C_5 and C_6 in series present an equivalent capacity of 305 pf. The output is taken between the capacitors to provide an output impedance of 900 ohms in order to provide the required input impedance to the mixer.

The performance of the entire amplifier now was evaluated by means of a mathematical model. Nominal values for common emitter h parameters were found to be,

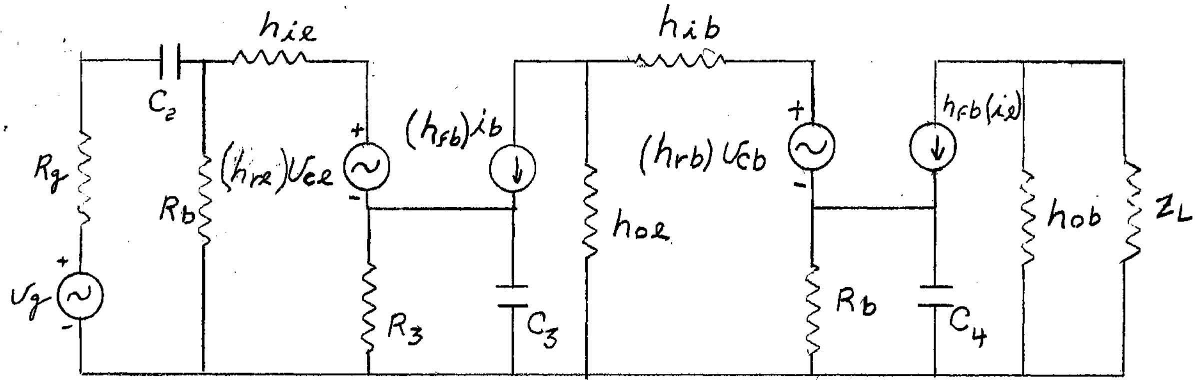
$$\begin{aligned} hie &= 3.46 \text{ K ohms} \\ hfe &= 60 \\ hre &= .0192 \\ hoe &= 329 \text{ mho} \end{aligned} \tag{10}$$

Utilizing the conversion factors of table 1 the common base h parameters are,

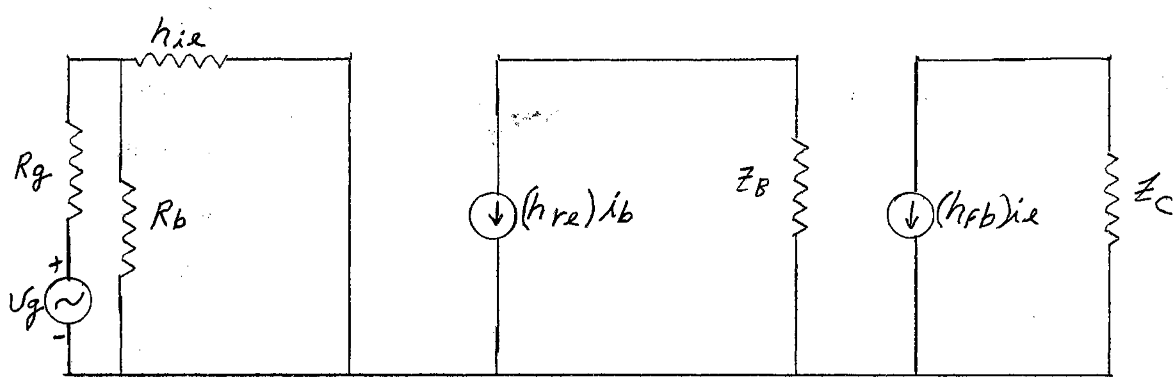
$$\begin{aligned} hib &= 56.8 \text{ ohms} \\ hfb &= .984 \\ hrb &= 5.6 \times 10^{-4} \\ hob &= 329 \times 10^{-6} \text{ mhos} \end{aligned} \tag{11}$$

For purposes of the mathematical analysis the input matching network was replaced with a voltage source in series with a generator resistance of 1000 ohms.

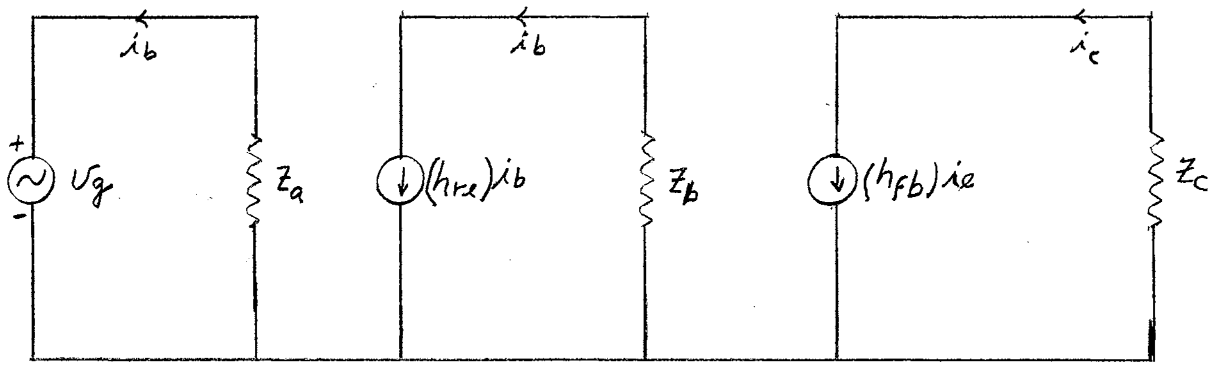
The h parameter[8] equivalent circuit for the cascade amplifier is,



The equivalent circuit can be simplified greatly by combining values and by making certain assumptions. h_{re} and h_{rb} are negligible compared to the rest of the quantities, therefore they were neglected. The parallel combinations of R_3 and C_3 and R_b and C_4 have approximately zero impedance and can be neglected. C_1 also presents a negligible impedance at signal frequencies. With these assumptions and combining elements the equivalent circuit reduced to,



Which can be further reduced to,



Where;

$$Z_a = R_g = \frac{hie\ R_b}{hie + R_b} \tag{12}$$

$$Z_b = \frac{hib\ hoe}{hib + hoe} \tag{13}$$

$$Z_c = \frac{hob\ Z_L}{hob + Z_L} \tag{14}$$

Substituting values into these equations we get,

$$Z_a = 3.74 \text{ K ohms} \tag{15}$$

$$Z_b = 55.4 \text{ ohms} \tag{16}$$

$$Z_c = 2.08 \text{ K ohms} \tag{17}$$

Now solving for values of ib, ie, and ic,

$$i_b = \frac{V_g}{Z_a} = \frac{V_g}{3.74 \times 10^3} \text{ amp.} \tag{18}$$

$$i_b = \text{hfe (ib)} = \frac{60\ V_g}{3.74 \times 10^3} \text{ amp.} \tag{19}$$

$$i_c = \text{hre } i = \frac{(.984)\ (60)\ V_g}{3.74 \times 10^3} \text{ amp.} \tag{20}$$

The output voltage (v_o) normally would be equal to (i_c) (Z_c) but, because of impedance matching,

$$v_o = \frac{1}{5}\, i_c\, Z_c \tag{21}$$

The voltage gain was then calculated to be,

$$\frac{V_o}{V_g} = \frac{(.984)\ (60)\ (2.08 \times 10^3}{3.74 \times 10^3} = 6.56 \qquad (22)$$

Similarly the current gain and power gain is found to be,

$$\frac{i_c}{i_b} = 294 \qquad (23)$$

$$\frac{P_o}{P_i} = \frac{V_o i_c}{V_g i_b} = (294)\ (6.56) = 1928 \text{ or } 33 \text{ db power gain.} \qquad (24)$$

These values were compared with a measured voltage gain of 5.1 and current gain of 247.

The observed power gain of 31 db compared very well with the theoretical value and represented a respectable gain considering the sacrifices made in the interests of low noise and stability.

CHAPTER III

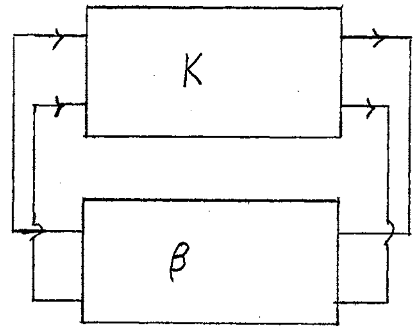
MIXER AND LOCAL OSCILLATOR

Design Considerations

The mixer circuit was designed to provide good mixing, a modest power gain, and good temperature stability while having relatively low noise. The local oscillator selected was crystal controlled and provides a stable, low impedance output to the emitter of the mixer transistor. The two portions of the circuit will be considered separately.

Local Oscillator

A necessary condition for oscillation is that the circuit provides a power gain at the desired frequency great enough to overcome circuit losses and to establish unity gain around the feedback loop.[8] A crystal controlled version of the Colpitts Oscillator[9] was selected for this application. The output was taken from the base of the oscillator in order to provide a low impedance output. An oscillator may be treated as an ordinary amplifier with a feedback or resonator loop. If K represents the gain of



the amplifier and β is the gain of the feedback loop, then oscillation occurs when,

$$\beta K = 1 \tag{25}$$

Arranging the Colpits oscillator in the same configuration the transistor is the amplifier with a gain of K while the capacitors and crystal form the feedback network with a gain of β. Referring again to Fig. 2-3, Q_4, R_{10}, R_{11}, and R_{12} is the amplifier portion of the oscillator while the crystal, C_7, C_8, C_9, and C_{10} is the feedback loop.

The transistor selected for this application is the TI-484, a double - diffused mesa silicon type. A silicon transistor was chosen to allow the use of degenerative feedback biasing without sacrifice of temperature stability. This type of biasing helps in maintaining a constant amplitude output. Resistors R_{10} and R_{12} provide the bias and temperature stabilization.

All components were selected for quality and temperature stability. All capacitors have mica dielectric and the fixed capacitors are epoxy encapsulated. The crystal is housed in a crystal oven with the temperature controlled by a heater - thermostat combination to 85° C.

Capacitors C_7 and C_8 are in series with the crystal in the feedback loop and their values were determined experimentally to provide a slight frequency adjustment and to adjust the feedback gain. The entire circuit serves as an extremely stable frequency source for the mixer. R_{13} and C_{11} couple the base of the oscillator to the emitter of the mixer and adjust the drive level.

Mixer

A transistor mixer circuit was selected for this application primarily because of ease of design and simplicity. A better choice would have been one of the sheet-beam switching tube series such as the 7360.[10] This would provide a greater gain and better suppression of the RF and local oscillator frequencies but would have required the use of a separate power supply for just one tube. The advantages did not justify the increased complexity. Accordingly, an emitter-fed transistor type mixer was selected.

A transistor mixer need not be able to amplify at the RF frequencies involved because it is equivalent to a diode mixer followed by a transistor amplifier.[8] The mixing in this case is accomplished at the base-emitter diode. The transistor must be able to amplify signals at the IF frequency. Transistors used as mixers must meet the following requirements:

1. Efficient base-emitter diode characteristics.

2. Low emitter input capacitance.

3. Good power gain at the IF frequency.

The mixing action of the transistor, as previously stated, takes place over the nonlinear impedance of the base-emitter diode producing a series of sum and difference frequencies. This series of sum and difference frequencies can be expressed in a Fourier Series.[11] Designating the RF signal as ω_1 and the oscillator as ω_2 the Fourier Series is,

$$F = A + B \sin \omega_1 t + C \sin \omega_2 t + D \sin (\omega_1 + \omega_2) t + E \sin (\omega_1 - \omega_2) t + F \sin 2 \omega_1 t + G \sin 2 \omega_2 t + \dots \quad (26)$$

One of these frequencies is selected as the IF frequency. The IF frequency selected for this application was the $E \sin (\omega_1 - \omega_2) t$ term. The RF signal is 2.5 mc and the desired IF frequency is 455 kc therefore, the local oscillator must be,

$$\frac{\omega_2}{2\pi} = 2.5 \text{ mc} - 455 \text{ kc} = 2.045 \text{ mc} \quad (27)$$

All other frequencies are suppressed by frequency selective output circuitry.

The transistor selected for this application was the 2N 2188, a germanium mesa type, which has a useful amplification up to 150 mc.[15] It had a spot noise figure of 4.5 db at 2.5 mc with $L_E = 1.5$ ma and $R_g = 1\text{K}$ ohms. The transistor was biased to operate under these conditions. A stability factor of 4 was selected for the mixer to provide an extremely stable circuit but only at the expense of considerable gain. Resistors R_6, R_7, and R_9 set the bias conditions and stability for the mixer circuit.

Proper operation of the mixer circuit required an oscillator signal of 630 mv to the emitter of Q_3.[13] The maximum design RF input to the mixer is 100 mv which gives an output of 880 mv for conversion voltage gain of 88. R_8 is the load resistor for Q_3. Its value is set by impedance and matching considerations and will be descussed in conjunction with the IF amplifier section. An unfiltered signal is taken from the collector of Q_3 for the automatic gain control (AGC) circuitry.

CHAPTER IV

AUTOMATIC GAIN CONTROL AND METER DRIVE

Introduction

Since one of the prime design objectives was to obtain a good logarithmic response, the automatic gain control (AGC) forms the heart of the receiver. In this area it is felt that the greatest contribution has been made. This design is rather unique for the following reasons:

1. The design is simple and straightforward as compared to more conventional approaches because no complicated feedback loops are employed.

2. A feedforward rather than a feedback loop was selected so that the attack and release time constants can be dictated by the desired operating characteristics of the receiver rather than the requirements of the circuitry.

3. The logarithmic voltage-current characteristics of a germanium diode were utilized to develop the AGC voltage.

Design Approaches Considered

Several good approaches to the problem of logarithmic response amplifiers have been made, particularly in television receivers. One approach is a series of pentode amplifier stages, each one biased at different levels by means of diodes.[14] Each stage begins amplifying at different input levels thereby approximating the logarithmic curve by means of a piece-wise linear approach. Another approach is to vary the bias of the IF strip by means of the conventional AGC feedback loop with the control voltage being developed by means of a non-linear amplifier.[15]

The major difficulty with these approaches is that they are either very critical or difficult to adjust in order to attain an accurately logarithmic response or they were originally intended to approximate the logarithmic curve. The primary interest in most television receiver circuits is to prevent over-driving the amplifier stages and not to develop an accurate logarithmic response. Accordingly several new approaches were tried.

The most interesting approach was to develop a logarithmic attenuator. If the center leg of a T-pad attenuator is replaced by an active device such as a tube, the attenuation could be varied while maintaining a relatively constant input and output impedance. This idea was developed and subsequently patented by Edwin C. Miller.[16] Figure 4-1 is an emitter follower-attenuator developed

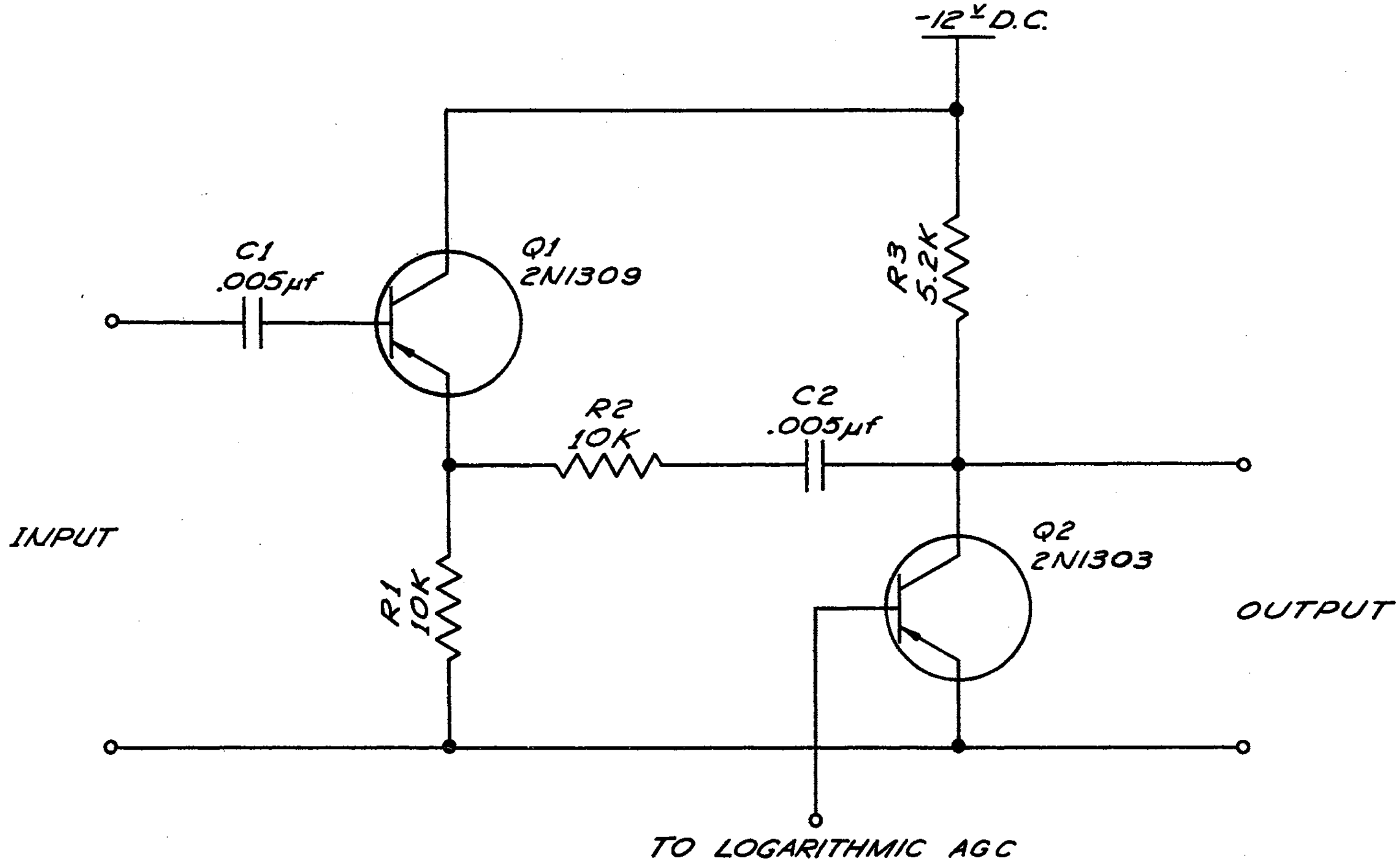


Figure 4-1 Logarithmic Attenuator

around this idea. This is a very good approach for attenuation up to 20 db. For greater attenuation several emitter follower-attenuator stages could be cascaded but the complexity becomes prohibitive in addition to the large insertion loss. To attain an attenuation of 80 db, 4 cascaded stages of attenuators would be required with an insertion loss of 35 db. This circuitry would be in addition to at least one extra IF stage to make up for the insertion loss.

AGC Circuitry

In an effort to simplify the circuitry as much as possible, an effort was made to utilize conventional AGC line to provide a control bias to the bases of the IF amplifiers. The logarithmic control voltage was developed by taking advantage of the characteristics of the "stabistor" produced by Texas Instruments, Inc. The stabistor is a germanium diode selected for its logarithmic characteristics. The characteristics of the stabistor approximates a logarithmic function according to the equation,[17]

$$V_D \approx \frac{AKT}{q} \ln \frac{I}{I_S}$$

Where:

K = Boltzmann's constant

T = ambient temperature in degrees Kelvin

q = charge on an electron

I = forward diode current

I_S = diode saturation current

$$I_S \approx 10^{-2} \text{ amps for } V > \frac{KT}{q} \qquad (28)$$

$\frac{KT}{q} \approx 25.8$ mv at room temperature

The stabistor diode selected was the G129.

The following analysis will be referred to Figure 4-2. The reference signal from which the AGC signal is derived is taken from the collector of the mixer transistor (Q_3 of Figure 2-3) to the AGC circuitry. This signal is unfiltered and is comprised of all of the components of the Fourier series, equation (26). If there is no incoming signal then the only component left in the series is the oscillator signal. This residual signal is used to provide a reference bias for the AGC circuitry and IF amplifiers. The incoming reference signal passes through a voltage doubler-rectifier formed by diodes D_1 and D_2 then is filtered by the filter formed by C_5, C_6, and L_2. The result is a d-c voltage which controls the base of Q_3.

The use of the feed-forward control loop eliminates the problem of adjusting the attack and release times in order to avoid "motor boating." The attack time of the AFC is determined in part by the input resistance and L_2. The input resistance is approximately the output impedance of the mixer or 5K ohms. The attack time τ_1, is,

$$\tau_1 = \frac{L}{R} = \frac{750 \times 10^{-6}}{5 \times 10^3} = .15\,\mu s. \qquad (29)$$

The release time τ_2 is determined by C_5, C_6, and L_2,

$$\tau_2 = L(C_1 + C_2) = 750 \times 10^{-6} \; (2000 \times 10^{-9})$$

$$\tau_2 = 1500 \text{ ps.} \qquad (30)$$

These time constants are added to the time constant formed by the AGC line resistors, capacitor C_9, and base-emitter capacitances of the IF amplifiers. τ_1 and τ_2 may be adjusted as the performance of the circuit requires. The values given in (29) and (30) were found to give the best overall performance.

The d-c voltage from the filter controls the base of Q_3, as previously mentioned. This directly controls the current through Q_3. Since Q_3 is operated in its linear range the voltage across D_3 is a logarithmic function of the incoming signal. The resistor R_4 provides an adjustment of the zero signal bias. The AGC voltage is taken as the voltage across the series combination R_4 and D_3. In practice the meter drive adjustment resistor R_3 and R_4 interact slightly, therefore the meter deflection and zero signal bias must be adjusted simultaneously.

The AGC voltage varies the biasing of the IF amplifiers. Using the figure on the following page, the AGC line actually places D_3 and R_4 in series with the individual bias adjustment potentiometers. The voltage across the D_3 is controlled by the external current source, which is Q_3 of Figure 4-2. As the incoming signal increases,

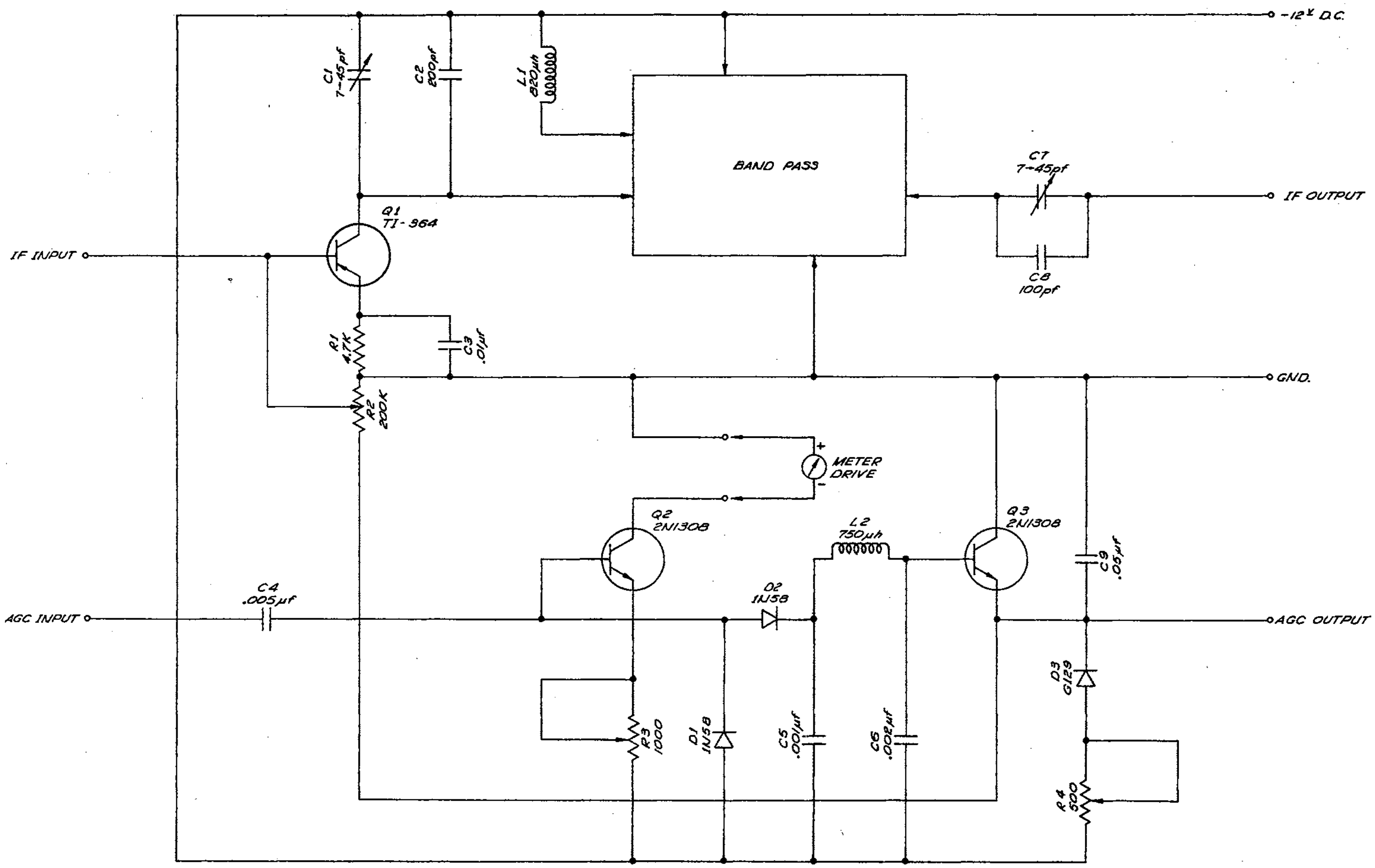


Figure 4-2. A.G.C. Band-Pass and Meter Drive Circuit. Board 2

the gain of the IF amplifiers are decreased. The individual bias adjustment resistors may be adjusted to provide different AGC response from each amplifier, thereby permitting the attainment of an accurate logarithmic response of the entire receiver.

Meter Drive

The information from the receiver is presented in three forms, a differential output for recorder purposes, a conventional audio output, and a meter output. The recorder and audio outputs will be treated later.

The meter face was calibrated to provide an accurate indication of the incoming voltage at very low modulation frequencies. As the modulation frequency increases the response time of the meter movement dampens the needle oscillation and the meter indication becomes the average value of the amplitude of the incoming signal. Referring again to Figure 4-2, transistor Q_2 provides the current drive for a 0-1 milliammeter located on the front panel of the receiver. Q_2 actually amplifies the positive portion of AGC reference signal. Since the meter movement will not respond to the high frequency components, the meter indication corresponds to the lower frequencies and the d-c component of the reference signal. R_3 provides a means for meter current adjustment but must be adjusted with R_4 because of the interaction previously mentioned.

CHAPTER V

IF AMPLIFIER

Design Considerations

The design of the IF amplifier section was dependent upon the method of developing the logarithmic response. The initial approach, involving the use of the logarithmic attenuator circuit of Figure 4-1, required an IF amplifier with linear characteristics since the receiver response was determined prior to the IF amplifier. The circuit of Figure 5-1 was developed.[9] This circuit takes advantage of the simplicity offered by complementary circuitry.

The amplifier utilizes a tuned-emitter direct-coupled configuration. Stability is achieved by mismatch, therefore neutralization is unnecessary. This arrangement results in several desirable features:

1. The amplifier can be designed so that the operating frequency is almost independent of transistor property variations and supply voltage changes.
2. Circuit complexity is reduced.
3. Alignment procedure is simplified.

The predominant disadvantage is that a lower overall gain results because of the use of mismatch for stabilization.

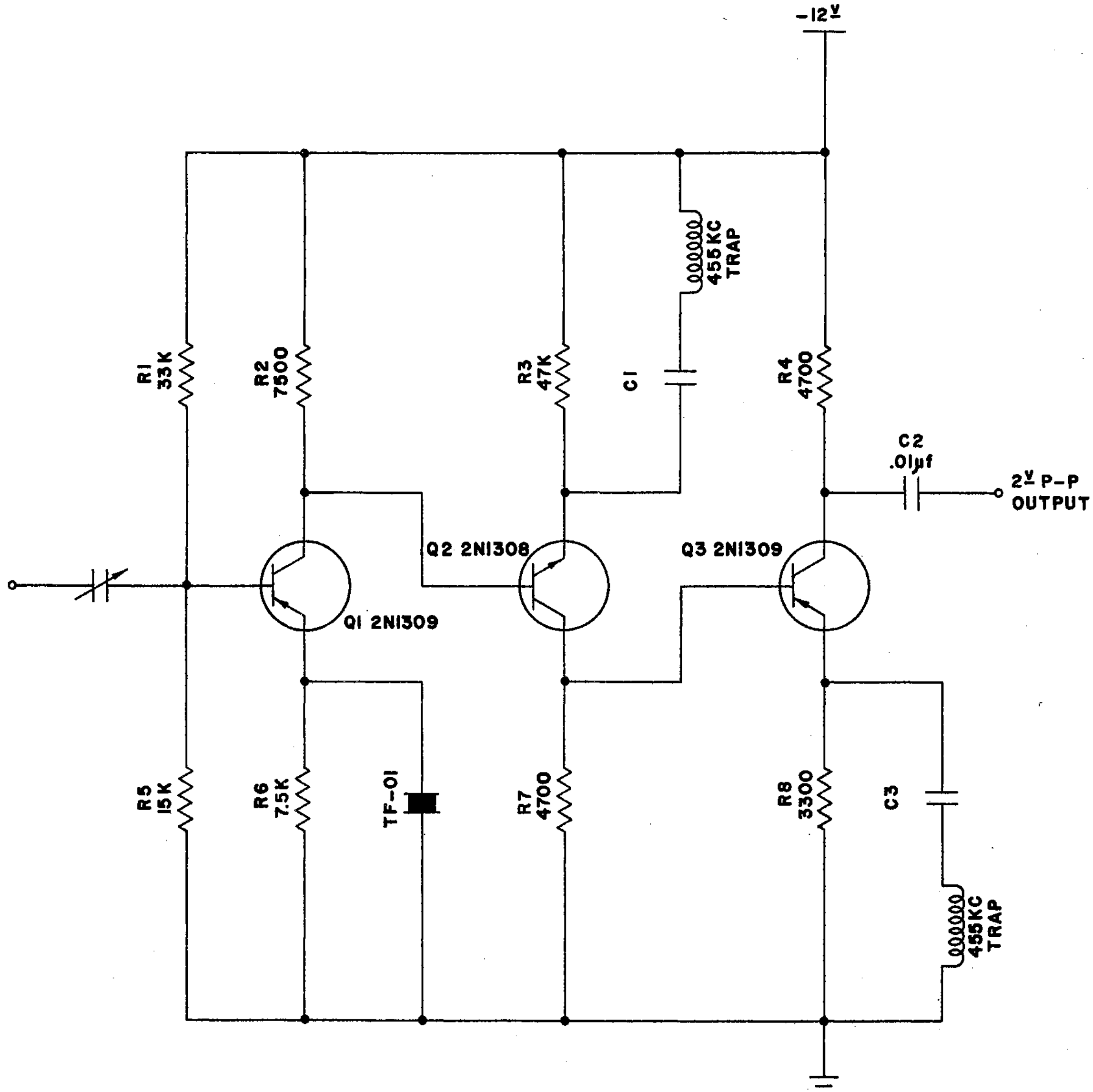


Figure 5-1. Direct-Coupled IF Amplifier

Additional selectivity was achieved by use of a TF-01, a ceramic by-pass device manufactured by the Clevite Corporation,[18] in place of the series tuned circuit of one emitter. This type of circuit readily accepts crystal, ceramic or mechanical filters in place of the series-tuned emitter by-pass networks for frequency selectivity. It is a very simple straightforward circuit for applications where an AGC loop is not required. The bias and temperature stability for the entire stage is provided by R_1 and R_2, since all of succeeding stages are direct coupled.

IF Amplifier Design

The decision to incorporate an AGC type of control in order to reduce overall complexity necessitated a new approach to the design of the IF amplifier. In order to provide simplicity of alignment and long term reliability a "Transfilter", manufactured by the Clevite Corporation[19] was chosen as the frequency selective and interstage coupling device. The transfilter is composed of two lead zirconate-lead titanate ceramic disks cut to resonate at 430 kc. These disks operate as a two-part piezoelectric device and are connected to an L-section filter. Since the resonate frequency is set at 430 kc the antiresonate frequency occurs at 455 kc, the desired IF frequency.

The transfilter selected was the TO-02 because of its impedance matching characteristics. It will match a source impedance range from 3.9K ohms to 15K ohms with an output impedance range of 680 to 3.0K ohms.

The transistor selected for the IF amplifier was the TI-364. This transistor is a germanium pnp device expressly design for low noise IF amplifier application. It exhibits its best noise characteristics with a generator impedance of 1K ohms.

The impedance transformation provided by the transfilter is determined by the load resistor of the preceding stage. The transfilter's load Q is determined by the equation:[21]

$$Q = \frac{Fo}{B_{3db}} \tag{31}$$

Where: Fo = The transfilter center frequency

B_{3db} = The desired band width between the 3db points.

For a 10kc bandpass,

$$Q = \frac{455 \times 10^3}{10^4} = 45.5 \tag{32}$$

The load resistor R_g now is determined by the desired input resistance to the following stage, in this case 1K ohms, according to the equation,

$$R_q = \frac{1}{R_L} \left[\frac{2}{2 \times 10^{-2}}\right]^2 = 5.2\text{K ohms} \tag{33}$$

This fixes the load resistor for all of the IF amplifier stages, and the mixer, with the exception of the first IF stage. This completes the impedance matching and frequency selectivity portion of the analysis.

The first stage of the IF amplifier has an output circuit which provides a selectable band-pass feature.

Referring again to Figure 4-2, the band-pass network consists of a parallel tuned circuit, two Collins mechanical filters, and switching provisions for selecting the desired output. This feature allows a choice of 2.1 or 6.0 kc band-pass. The natural band-pass of the IF stages is approximately 10kc. The mechanical filter is resonated on each end by the capacitors C_1, C_2, C_7, and C_8 according to design practices suggested by the manufacturer.[22] When the 10kc switch position is selected both mechanical filters are switched out of the circuit and replaced by an equivalent inductance L_1. Then L_1 and capacitors C_1 and C_2 form a parallel resonant circuit which acts as a load for Q_1. Other than the band-pass selection feature the first IF amplifier is identical to the following stages shown in Figure 5-2.

Referring now to Figure 5-2, resistors R_7, R_8, and R_9 are the bias adjustment resistors for the individual stages and are connected to the AGC line as previously described. With zero signal input conditions the AGC line potential is very nearly equal to -12 Volts. As the incoming signal increases the AGC line becomes more positive thereby controlling the gain of the stage. The temperature stability of each amplifier varies with the potential of the AGC line and is a minimum at zero signal conditions. The largest stability factor occurs under these conditions and is approximately 7.4.

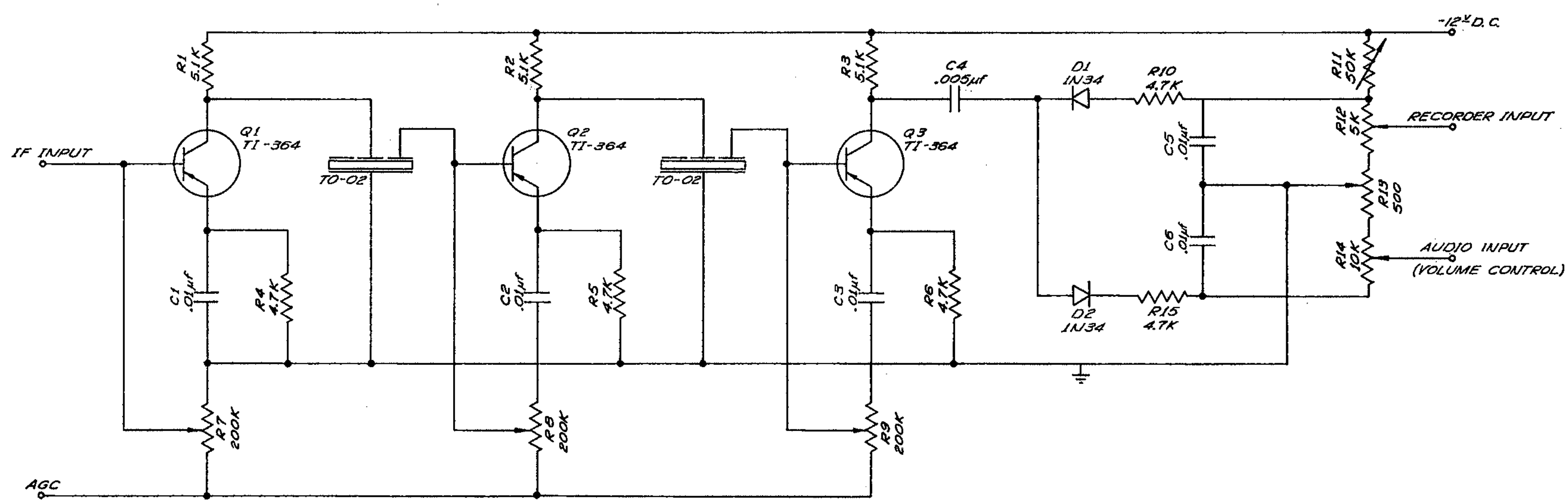


Figure 5-2. IF Amplifier and Detector. Board 3

Because of the non-linear amplification characteristics imposed by the AGC an equivalent circuit analysis is quite impractical. The amplifier was analyzed graphically under several bias conditions. The final step was to adjust each amplifier individually to obtain the desired overall response. In actuality each amplifier is adjusted to have slightly different amplification characteristics.

CHAPTER VI

DETECTOR

Design Considerations

The detector requirements were not much different from the normal, however the design finally arrived at is quite different than the usual approach. This was done in order to solve a problem which occurred in the output circuitry. An excellent analysis of the diode detector is contained in the Electronic Designer's Handbook.[23] Therefore it will not be covered here.

The diode detector extracts the AM modulation signal from the carrier. Since the modulation information is duplicated in either sideband the detector need detect only one sideband in order to obtain all available information. For this reason, the detector is biased slightly forward and detects only the positive or negative peaks, depending on the polarity of the diode. In practice the negative peaks are selected in the case of tubes and npn type transistor amplifiers and the positive peaks are selected for pnp transistor amplifiers in order to provide an AGC feedback.

In this application the AGC signal was not recovered from the detector, therefore, it was immaterial which polarity was selected. Advantage of this fact was taken in order to provide isolation between the audio and recorder output circuitry. The signal from the detector was required by both the recorder output and audio output circuitry, but at the same time the two outputs had to be completely independent. The recorder circuitry must be unaffected by drive level changes in the audio system. Moreover the recorder output system must have a d-c response and therefore must be directly connected to the detector output. This problem was solved by using two separate detectors.

Referring to Figure 5-2 the detector circuitry actually consists of two separate diode detectors, one responding to the positive peaks and the other to the negative peaks of the incoming signal. Diode D_1 and the filter circuitry formed by R_{10}, R_{12}, R_{13} and C_5 forms a detector which detects the negative peaks. On all negative peaks the rest of the circuitry is not conducting and is not a part of the acting circuit. The output of this detector is direct-coupled to the differential amplifier which provides the recorder output. In addition, resistors R_{11} and R_{12} provide a gain control and d-c biasing for one-half of the differential amplifier.

During the occurrence of positive peaks diode D_1 is cut off, isolating that portion of the detector circuitry. Diode D_2 and its filter network consisting of R_{13}, R_{14}, R_{15} and C_6 responds to the positive peaks of the incoming signal which is supplied to the audio output circuitry. In addition R_{14}, which is a front panel control, serves as the volume or gain control for the audio system. Since both halves of the detector circuit conduct alternately, the outputs are almost completely isolated from each other. Resistors R_{10} and R_{15} are part of the IF frequency filter network. In addition, they add to the network isolation characteristics of the detector although they decrease detection efficiency.[23] Resistor R_{13} serves as d-c balance adjustment for the detector.

CHAPTER VII

RECORDER AND AUDIO SYSTEM

Design Considerations

As previously mentioned, three types of information output were required. One of the circuits that was considered worked very well but was rather difficult to adjust. This circuit is shown in Figure 7-1. This approach was originally considered because there appeared no other way that a logarithmic meter output could be obtained. Later the meter drive was taken form the AGC reference line and this circuit was no longer needed. It is an interesting approach to the porblem, however, so it will be briefly described here. Transistors Q_{13}, Q_3 and their surrounding circuitry forms a differential amplifier designed to work with a push-pull detector. Transistor Q_3 is the meter driver which derives its voltage supply from the potential difference between ground and the tap setting of resistor R_2.

The current variation through the meter is adjusted by the positions the wiper arms of R_1 and R_2 resistors from the -12 volt ends and the d-c bias for Q_3 is adjusted by setting the arm of R_1 slightly more positive than that of

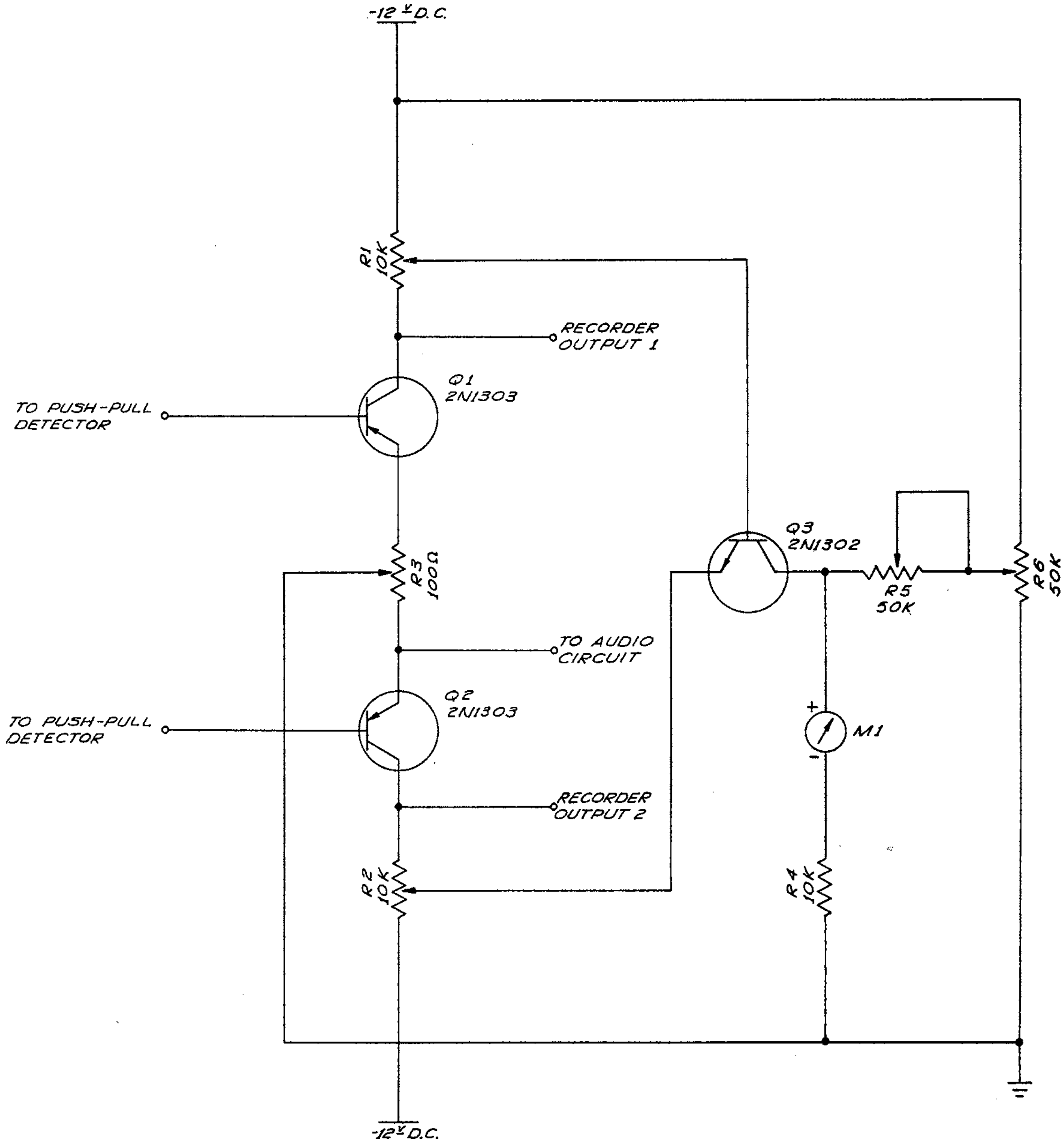


Figure 7-1. Recorder Output and Meter Drive Circuit

R_2. A signal was picked off from the emitter of Q_2 to go to the audio output circuitry. Resistors R_5 and R_6 provide a means for cancelling out the zero signal meter current. In addition to the obvious difficulty of adjustment a very slight interaction between the recorder and audio circuitry was detected. This was due to changes in input impedance of the first audio amplifier with change of the audio gain control setting.

The output circuitry was subsequently simplified and redesigned by taking the meter drive from the AGC reference voltage, previously discussed, and isolating the recorder and audio sources with the detector of Figure 5-2.

This simplification resulted in the circuitry of Figure 7-2.

Recorder Driver

The input requirements of the Sanborn Electrocardiagraph Recorder, which will be used with this equipment, dictated that a differential output be provided transistors Q_1 and Q_2 are germanium junction transistors. They were especially selected to have identical h_{fe} and temperature characteristics and were housed in a common heat sink. Because of these precautions the recorder output, taken as a difference between the collector potentials, is essentially independent of temperature. The resistors R_3 and R_4 provide a common emitter resistor and an

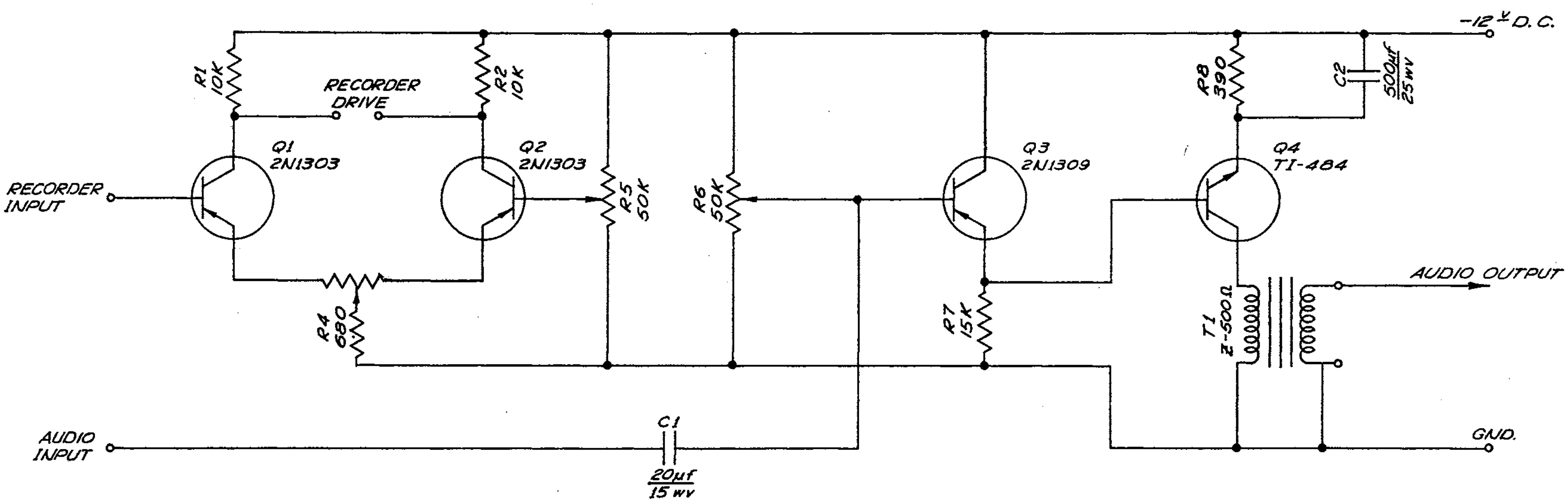


Figure 7-2. Recorder Output and Audio Circuits. Board 4

adjustment for current fluctuation reduction. Since matched transistors were employed a constant current source was not required.[24]

Resistor R_5 provides a fixed bias and temperature stabilization for Q_2 and eliminates the requirement for a push-pull input. The base of transistor Q_1 is directly connected to the detector circuit and receives both the signal drive and d-c bias as previously indicated. The maximum output voltage from the differential amplifier to a high input impedance recorder is 10 volts peak-to-peak.

Audio Amplifier

The audio amplifier utilizes an emitter-follower as a driver which is direct-coupled to the power amplifier.[25] Resistor R_6 provides an adjustable bias for the amplifier. The output transformer furnishes an output impedance of 4 ohms to a speaker.

CHAPTER VIII

ALIGNMENT AND ADJUSTMENT

Procedures

The adjustment and alignment procedures are quite simple to describe but rather tedious to perform since the entire system must be aligned together. The alignment procedure will be described step by step referring to Figures 2-2 (b), 2-3, 4-2, 5-2, and 7-2:

1. Figure 2-3 adjusts the RF Amplifier, mixer and local oscillator for maximum output to the first IF amplifier by adjustment of L_1 and L_2 and R_{13}. The signal input should be supplied from the specific transducer to be used with this receiver in order to insure proper input matching. There is a definite setting of R_{13} for best mixing action. In fact, each of three adjustments will give a definite amplitude peak when properly tuned.

2. Capacitor C_7 will vary the frequency slightly to align the transmitter and receiver frequencies precisely. The transmitter frequency should first be set by menas of a frequency counter if available.

3. Figure 4-2. With zero input signal adjust R_3 and

R_4 for minimum meter reading and for approximately -11.9 volts at the emitter of Q_3. This is a rough setting.

4. With about 40 microvolts RF signal input adjust R_2, C_1, and C_7 for maximum output from the bandpass network. The bandpass selector switch should be in the 6.0 setting. Capacitors C_1 and C_7 tune the input and output circuits of the mechanical filters and R_2 adjusts the bias of Q_1.

5. Figure 5-2. Adjust resistors R_7, R_8, and R_9 for proper operation.

6. Adjust gain control R_{14} to maximum volume.

7. Figure 7-2. Resistor R_6 can be adjusted now for maximum audio output. This is all the adjustment needed for the audio output. The audio gain control can now be adjusted as desired.

8. Figure 5-2. Adjust R_{11}, R_{12} and R_5 of Figure 7-2 for proper operation of the recorder output.

9. Figure 7-2. Resistor R_3 may be adjusted for maximum common-mode suppression.

10. Figures 4-2, 5-2, and 7-2. All adjustments in these figures should be reaccomplished to achieve an accurate logarithmic response for RF inputs up to 0.1 volt. With 0.1 volt to the receiver the meter reading should be 1 ma, input to the detector should be 2 volts, and recorder output should indicate 12 volts.

CHAPTER IX

CONCLUSIONS

The objective of this study was to develop an ultrasonic receiver with a logarithmic response over at least 80 db. This objective was successfully achieved. Two of these receivers and one complete unit of supporting equipment was delivered to the University of Minnesota Medical School, Minneapolis, Minnesota and placed in operation. In all, five receivers will be constructed for use in medical research.

The design resolved some of the problems encountered by the medical researchers but, of course, does not represent the ultimate in design. Several areas have arisen in which the receiver could be improved. One of these is that the AGC circuit, while presently adequate for this application, could be made more versatile by adding an amplifier in the AGC reference line prior to the meter drive. This would allow the maximum AGC level to be set over a larger range than now possible. This and other possible modifications, particularly in the mechanical construction area, have been suggested to Dr. Crawford's Research Staff.

BIBLIOGRAPHY

1. Rukavina, D. M. "Transistor h Parameter Techniques," (Unpublished M.S. thesis, Oklahoma State University, 1962).

2. American Radio Relay Handbook, American Radio Relay League, 1964, 404.

3. Davis, P. D. Jr. and Ezell, G. D. "Subaudio Parametric Amplifier For Ocean-Botton Seismometer." Electronics, XXXVI (March 1, 1963), 28.

4. Planinac, J. W. "An Investigation of Junction Diode Parametric Amplifiers at Medium Radio Frequencies." (Unpublished M.S. thesis, Oklahoma State University, 1960).

5. Uenohara, M. and Sharpluss, W. M. "An Extremely Low Noise 6-Kmc Parametric Amplifier using Gallium-Arsenide Point-contact Diodes." Proceedings of The IRE, XLVII (December, 1959), 2114.

6. Fristoe, H. T. "The use of Q Equations To Solve Complex Electrical Networks." Oklahoma State Engineering Research Bulletin No. 105, (June, 1961)

7. "pnp Alloy-Diffused Mesa Germanium Transistors, Types TI-363, TI-364." Texas Instruments Bulletin DL-S 622529, (April, 1962).

8. Texas Instrument Inc. Engineering Staff. "Transistor Circuit Design." New York: McGraw-Hill Book Company, 1963.

9. "Selected Semi-Conductor Circuits, Military Standardization Handbook MIL-H DBK-25, June 15, 1960.

10. Stoner, D. L. "New Product Detector Tube," CQ, May, 1960), 50.

11. Alley, C. L. and Atwood, K. W. Electronic Engineering. New York: John Wiley and Sons, Inc., 1962.

12. "pnp Alloy-Diffused Mesa Germanium Transistors, Types 2N2188, 2N2189, 2N2190, 2N2191." Texas Instruments Bulletin 623354, (June, 1963).

13. "Dalcom". Texas Instruments Bulletin 19562, (April, 1963).

14. Mathams, R. F. "A Voltage Operated Logarithmic Amplifier". Electronic Engineering, XXXI (August, 1959), 463-465

15. Crawford, H. D. "The Study of Transthoracic Propagation of Ultrasound" Final Report to National Institute of Health, Grant Number H.E.W. 06632-01.

16. Miller E. C. "Automatic Programming in Small A-M Stations". Electronics, XXX, (February 1, 1957), 146-148

17. "Silicon Forward Conductance Diodes, Types G129 and G130". Texas Instruments Bulletin DL-S 60217, (March, 1960).

18. "Ceramic Filter for Emitter By-pass Application". Clevite Electronic Components, Inc. Bulletin 9407, (February, 1962).

19. "Overtone Transfilter Matches Impedances As Interstage Coupler". Clevite Electronic Components, Inc. Bulletin 9406, (February, 1962).

20. Curran, D. R. and Gerber, W. J. "Piezoelectric Ceramic I.F. Filters", Clevite Electronic Components, Inc. Technical Paper TP-42.

21. "Circuit Applications for "Transfilter" Ceramic IF Filters TO-01 and TO-02". Clevite Electronic Components, Inc. Bulletin 94-35.

22. "Collins Mechanical Filters". Collins Radio Company Bulletin 1007.

23. Landee, R. W. Davis, D. C. and Albrecht, A. P. Electronic Designers' Hand Book, New York: McGraw-Hill Book Co., 1957.

24. Earle, W. E., "Designing Zero-Draft D-C Differential Amplifiers". Electronics, XXXVI, (November 8, 1963), 66-70

25. General Electric Transistor Manual (5th ed., 1960).

APPENDIX

SUPPORTING EQUIPMENT

The receiver was designed to operate as part of some ultrasonic medical research equipment. A complete unit consisted of:

1 - 2.5 mc receiver

1 - 2.25 mc receiver

1 - 2.5 mc transmitter

1 - 2.25 mc transmitter

1 - control and speaker panel

1 - Kepko power supply

The ultrasonic transmitter is of conventional tube design and is shown in Figure A-1. It is designed to deliver up to 15 watts to 50 ohm load. The drive level control provides a continuously variable output from 0.1 watt to the maximum output set by an internally adjustable maximum output control. The maximum output is presently set at 5 watts due to limitations imposed by the transducers.

Figure A-2 is the interconnecting diagram for the entire unit.

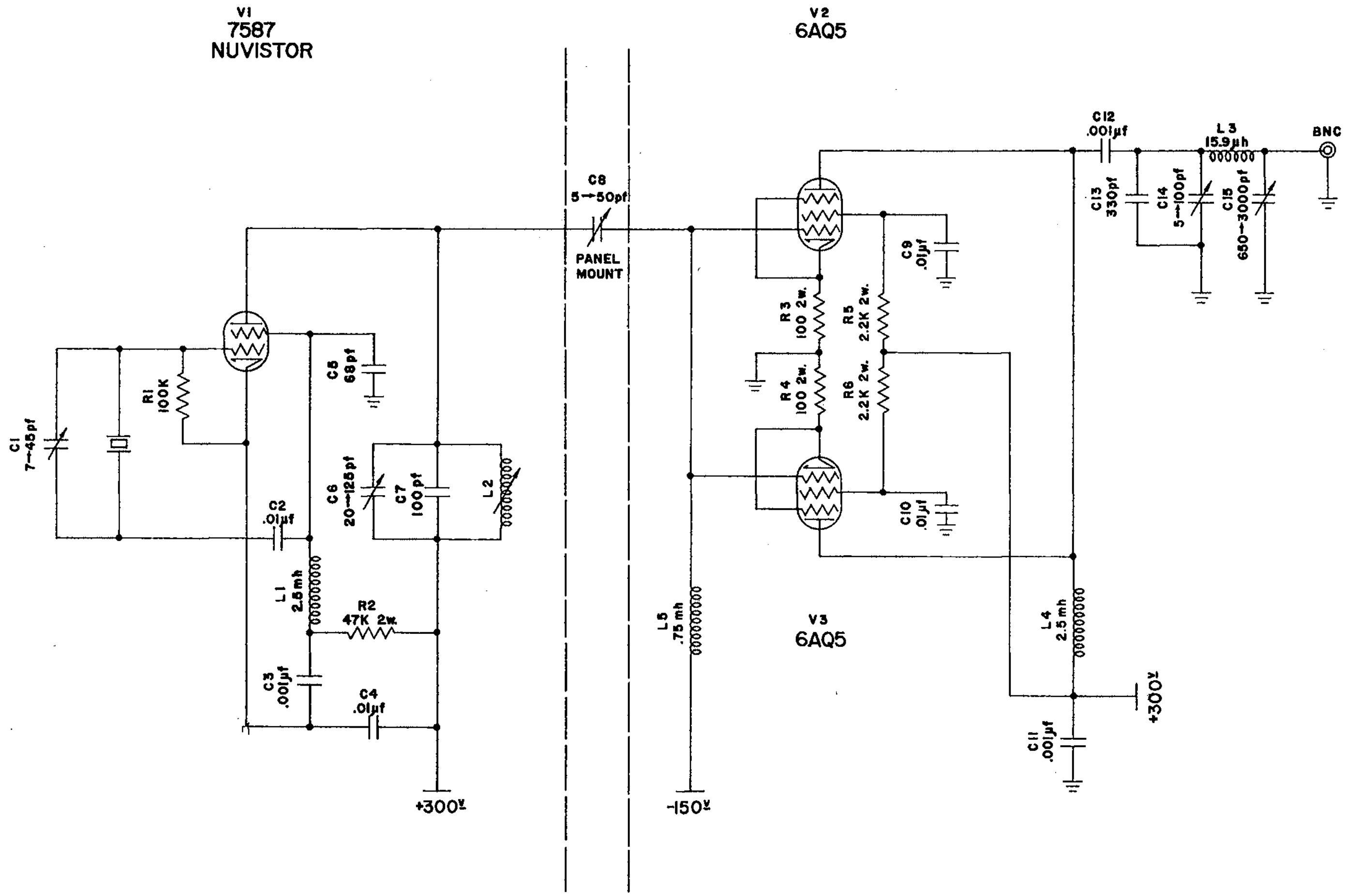


Figure A-1. Ultrasonic Transmitter

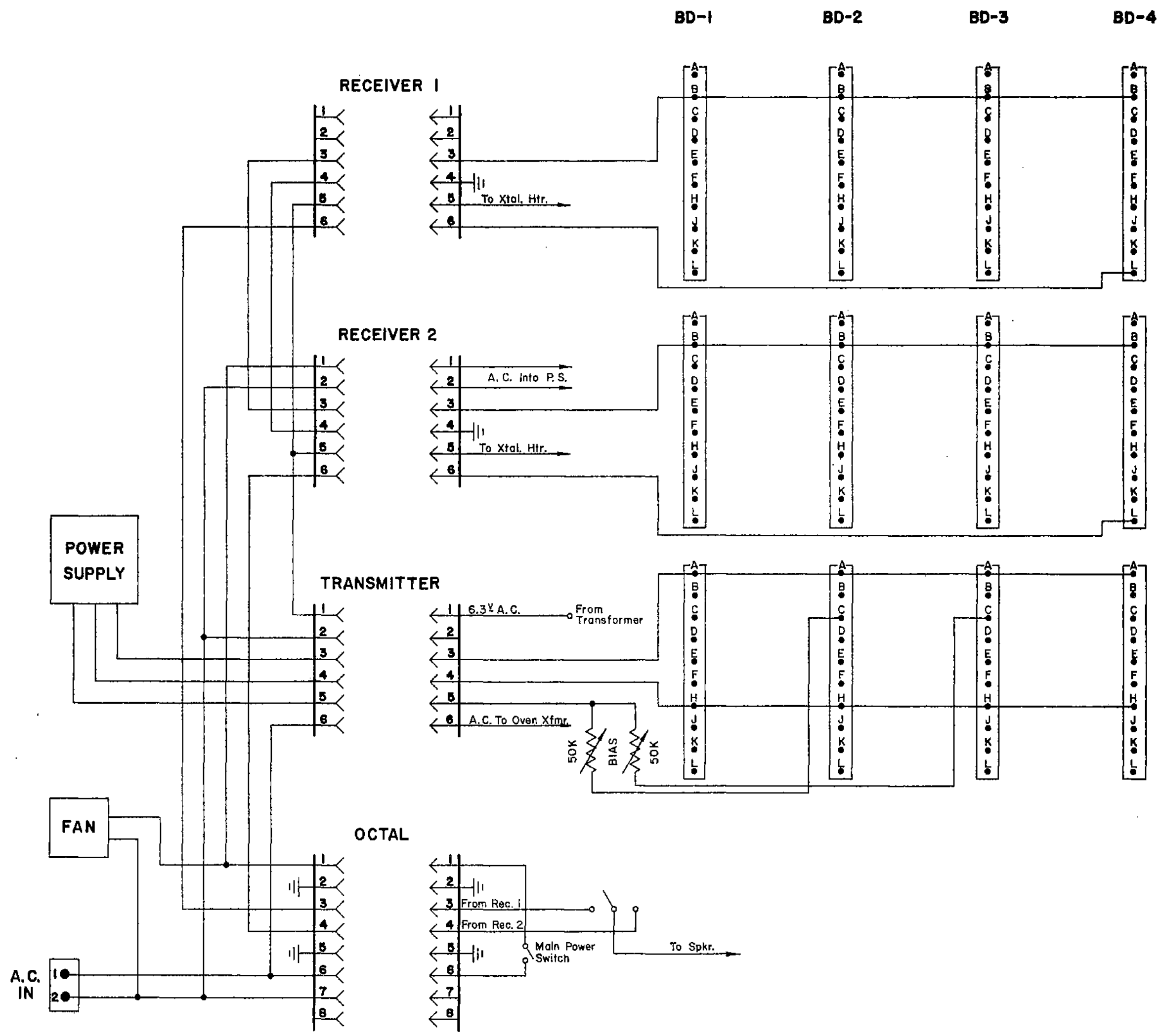


Figure A-2. Interconnecting Diagram

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