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AERONAUTICAL SYSTEMS DIVISION
AIR FORCE SYSTEMS COMMAND
WRIGHT-PATTERSON AIR FORCE BASE, OHIO

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(Prepared under Contract No. AF 33(600)-40955 by
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Hicksville, Long Island, New York
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FOREWORD

This report was prepared by Radio Receptor Co., Inc., a Division of General Instrument Corporation, Hicksville, L.I., on work performed on U.S. Air Force Contract AF 33(600)40955 under Task No. 410709 of Project No. 4107, Transmitter Group AN/AKA( ).

The work was administered under direction of the Electromagnetic Warfare and Communication Laboratory, Aeronautical Systems Division, Air Force Systems Command, Wright-Patterson Air Force Base, Ohio. Mr. Stanley E. Weber was the Project Engineer for the Laboratory.

The work presented in this report began on 29 March 1960 and was concluded 29 March 1963.

The work presented in this report represents a group effort by the staff of Radio Receptor. However, specific acknowledgement is given to the following group for their patience and devotion to the project.

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ABSTRACT

This report covers the development of the AN/AKA( ) microwave telemetry transmitter. It describes the study and development phase related to the design of the equipment. In addition, a complete description of the final configuration is included along with typical test data.

A method for multiplying the voltage vs. capacity characteristics of a varactor is described along with a derivation of the necessary design equations. In addition, a "Figure of Merit" for a varactor is defined for frequency modulation application.

The presence of a grid structure resonance in present day UHF planar triodes is noted. The effects of these characteristics on transmitter performance are described in detail.

PUBLICATION REVIEW

This technical documentary report has been reviewed and approved.

FOR THE COMMANDER:

Ronald G. Stimmel

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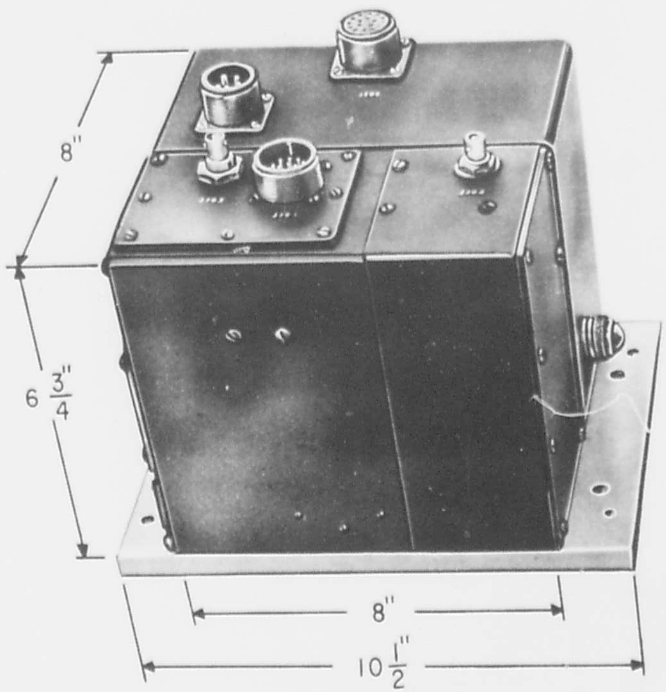
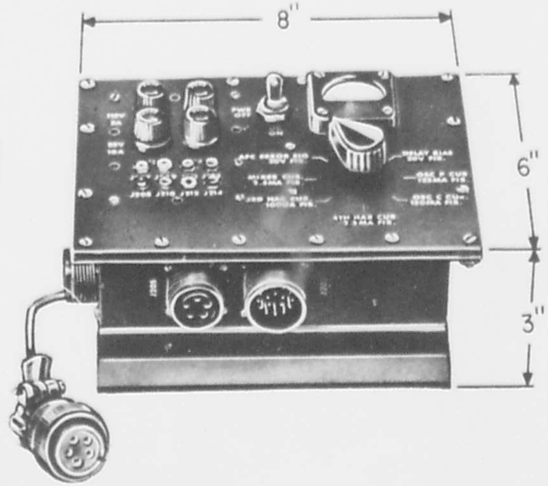


TRANSMITTER-POWER SUPPLY



CONTROL UNIT

1.0 INTRODUCTION

This report is a summary of the study and development work performed by the Radio Receptor Division of the General Instrument Corp., on the AN/AKA( ) microwave telemetry transmitter, to exhibit No. WCLCE-56-1A under contract AF 33 (600)-40955 for the Aeronautical Systems Division of the Air Force Systems Command. Both study and development phases are included as well as typical results obtained with the unit. Conclusions and recommendations for future work are also made.

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2.0 TRANSMITTER STUDY PHASE

The purpose of the study phase was to formulate a FM microwave transmitter configuration which would meet the objectives of the specification. There were three general electrical problems related to the design of the transmitter. They included:

A. Power Generator

B. Frequency Modulation

C. Frequency Stability

Although these factors represented distinct parameters of the system, it was evident that above consideration were interrelated. As such, the final configuration chosen would have to be a compromise between these parameters as well as size, weight, efficiency and cooling.

2.1 Power Generator

The objective for the Power Generator was to develop 5 watts at "S" Band which was capable of being frequency modulated and stable over a wide temperature range. The signal developed would then be fed to a Power Amplifier which would bring the available power to the antenna to a fifteen watt level. Two general methods were open to consideration to meet these objectives; they were:

a. A low frequency oscillator in conjunction with an amplifier multiplier chain.

b. A power oscillator operation at the desired frequency.

2.1.1 Low Frequency Oscillator

A block diagram of the system is shown in Figure (1). This configuration consists of a low frequency oscillator followed by a series of amplifier and multiplier stages. Frequency modulation is accomplished at low levels and suitable circuitry is provided for automatic frequency control when necessary.

The ideal low level oscillator would, of course, be crystal controlled, thus eliminating the need for an AFC circuit. Using a reliable crystal frequency in the order of 60 MC the necessary multiplier following the oscillator would be in the order of 36 times. In the multiplier process the modulation index would also be multiplied so that for a ± 500 KC carrier deviation at "S" Band the required deviation at the oscillator would be ± 13.9 KC. No known crystal was then available that would allow the crystal to be pulled by the amount and still maintain the required stability. This, therefore, indicated that the oscillator would have to be of the non-crystal controlled type with an AFC unit incorporated for stability requirements. Various modulation and stability techniques will be described in other sections of this report. With these requirements in mind, an analysis of the required multiplier amplifier chain was made. It was immediately evident that the multiplier amplifier chain would have to be done with tubes since varactor and transistors available in 1960 were extremely limited for high power, high frequency application.

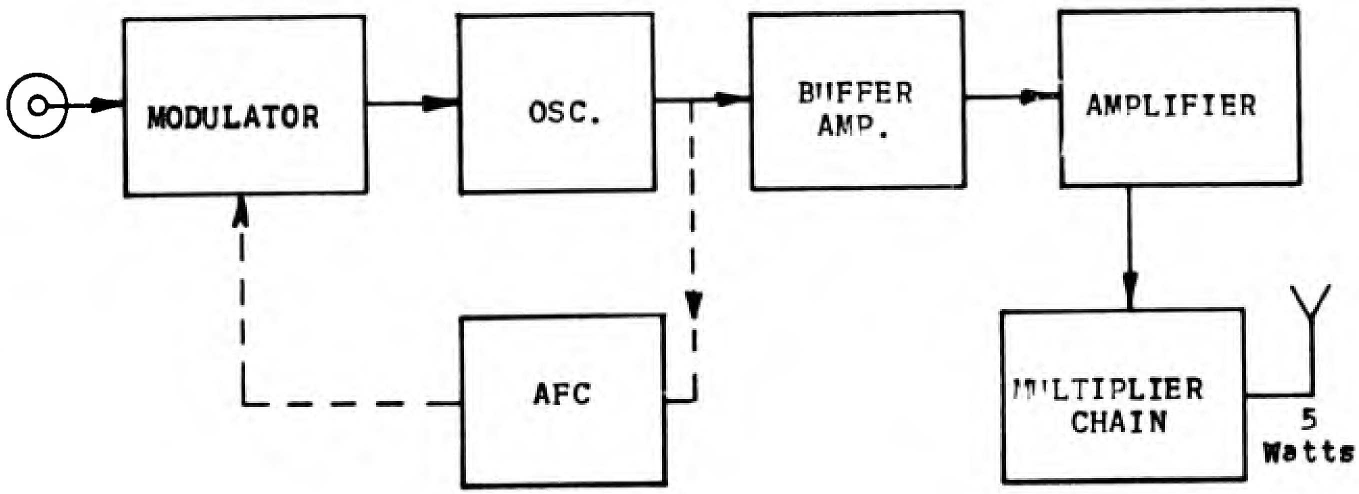


Fig. (1) Low Frequency Oscillator

A 36 X multiplier chain was used as a sample analysis to determine the feasibility of the approach. The circuit is shown in block diagram form in Figure (2).

A survey was made of the available tubes that could deliver 5 watts at 2250 MC. The 6442 was classified as a marginal choice since it would require using the tube at its extreme limits to get the necessary output power. A tube in the 2C39 - 3CX100A5 family was chosen. The efficiency of an amplifier tube is given[1] as

$$N = 0.7 \left(1 - \frac{Nfd}{\sqrt{E_p}}\right) \%$$

Where f = frequency of operation in megacycle

d = Cathode grid spacing in inches

E_p = Plate voltage

N = 1.2 for amplifier

N = 1.75 for oscillator

For the tubes in the 2C39 family, d is in the order of 0.005 at 750 MC and a 700 volt plate voltage

$$N = 0.7 \left(1 - \frac{(1.2)(.750)\ (5)}{\sqrt{700}}\right)$$

N = 55%

The expected efficiency of a tripler is given as 40% of that of a straight through amplifier[2] so that expected efficiency for the stage would then be 22%. For 5 watts output this would mean an input power of 22.7 watts plus an additional 6 watts for

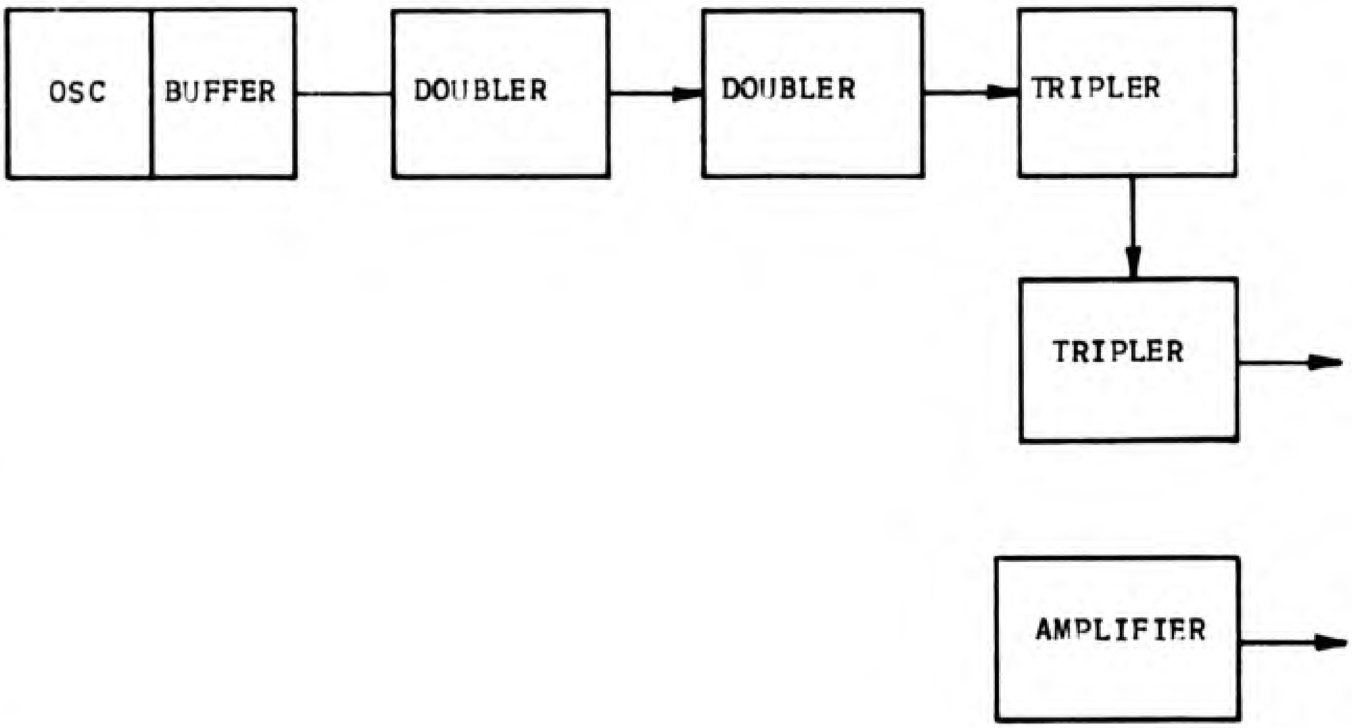


Fig. (2) Block Diagram
36 X Multiplier

filament power. The reliability of the above calculation has been demonstrated with the practical results[3] obtained with these tubes at S Band frequencies. Manufacturer's data also stated that a gain of 4 db was to be expected for this stage. The driver tripler would have to deliver 2 watts of drive power. Using a 6442 tube an efficiency of 27% is expected at 250 MC so that an input power of 7.3 watts is reached; in addition 5.4 watts are needed for filament power. Thus the total power for the two stages above would be 41.4 watts. In addition, the power amplifier would require 45 watts of plate power, plus 6 watts of filament power for a total consumption of 91.4 watts. It was evident that this system would far exceed the required input power considering the additional required circuitry. Besides the power requirements, the system could not be packaged in the desired volume and would be extremely difficult to tune. Considering all factors it was concluded that this design approach would not be feasible.

2.1.2 Power Oscillator

In this configuration a power oscillator is constructed at S Band to deliver the required 5 watts of output power. Since the inherent stability of such a device far exceeds the specification requirements, an automatic frequency control system is incorporated. The addition of a modulator and power supply completes the system. A Block Diagram of the configuration is shown in Figure (3).

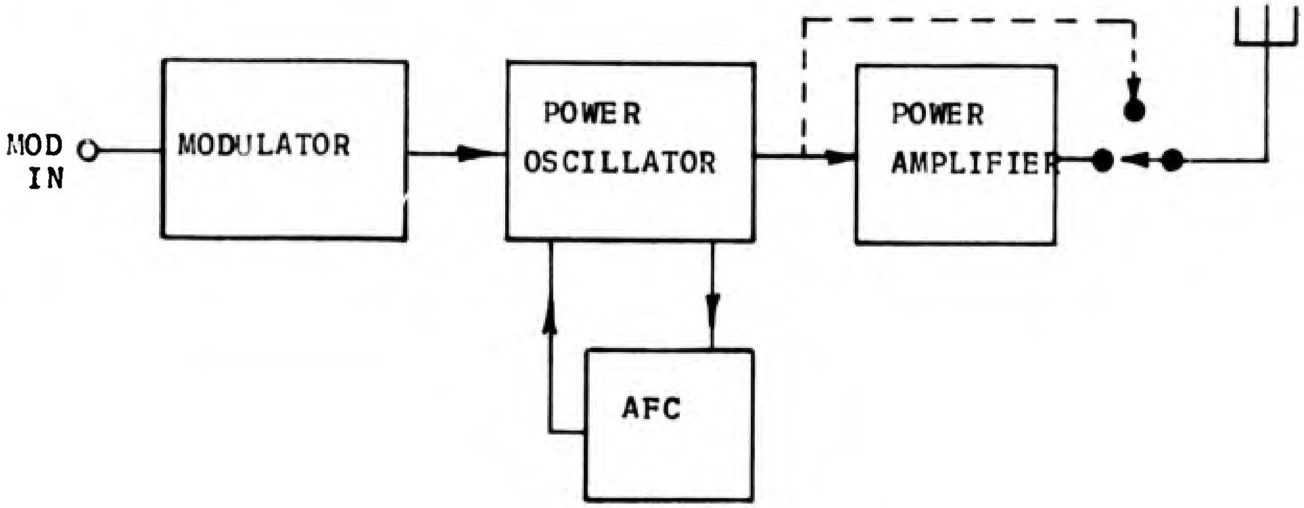


Fig. (3) Power Oscillator

An investigation of available tubes for the power oscillator was instituted. The selection was limited to those tubes which could operate with radiant cooling as forced air or water cooling was not practical. Among those considered were the GL6442 series as well as the 2C39-3CX100A5 family of planar triodes. Information supplied by the General Electric Company indicated that 5 watts output power could be achieved at 2350 MC using the GE6442. However, this would require operation of the tube beyond its maximum ratings and performance would not be guaranteed. In addition, a substantial reduction in life expectancy of the tube could be anticipated. Further data indicated that the tubes in the 2C39-3CX100A5 could meet the required output power specification with operation well within the maximum tube ratings. Since the mechanical configuration of tubes in this class were extremely similar, a cavity could be designed that could be applicable to all tubes of this series. Using the previous mentioned equation for tube efficiency (sec. 2.1.1), the expected efficiency at 2350 MC is 15%. Thus the plate power required for 5 watts output is 33.4 watts. With an additional 6 watts for filament, this brings the total input power for this stage to 40 watts. This power, combined with the required power for the power amplifier, yields an expected total of 91 watts which made the 125 volt-ampere equipment requirement feasible. In addition to this, it was estimated that the power oscillator volume would be 50% of that needed for the amplifier multiplier configuration. It should also be pointed out that the ease of tuning, maintenance and life made this configuration far more desirable.

2.1.3 Conclusions

In conclusion, the following advantages were attributed to the power oscillator configuration compared to the amplifier chain:

1. Power input would be lower and the objective equipment input power was feasible.
2. A considerable reduction in volume is realized.
3. Tuning and maintenance would be simplified.
4. Increased equipment life could be expected due to the reduction of stages.

The following disadvantages were noted:

1. Modulation would have to be performed at a higher level.
2. Care would have to be taken in the cavity design to insure minimum residual FM under shock and vibration.

It was concluded that the advantages of the Power Oscillator configuration outweigh the disadvantages and that these disadvantages could be minimized with good engineering design.

2.2 Frequency Modulation

Having decided upon the Power Oscillator configuration, a high level modulator had to be formulated. The modulator was broken down into two sections, the modulator amplifier and the actual transducer elements needed to convert time varying signals to time varying reactance. Since the gain of the amplifier would be determined by the transducer element, the latter had to be selected first. Two practical possibilities existed for the transducer element; the voltage variable capacitor or the microwave equivalent of the reactance tube amplifier.

2.2.1 Transducer Element

The reactance tube amplifier required that an auxilliary cavity, incorporating a miniature planar triode, be coupled into the oscillator cavity through a 90° phase shifting network. An obvious disadvantage to the system was the additional power and volume required. If a GL6299 was incorporated in the auxilliary cavity, it would require an additional 2.5 watt of power. The cavity and phase shifting network would utilize 10 in^3. Since the unit was required to operate over a 200 MC band, the auxilliary cavity and the phase shifting network would require adjustment. However, the configuration did offer the possibility of achieving excellent modulation linearity.

With these factors in mind, a study was instituted on the use of the voltage varying capacitor or "varactor" as it is better known. The capacity vs voltage characteristics of this device is given as

$$(1) \qquad C = \frac{Co}{\left(1 - \frac{V}{\phi}\right)^{1/N}}$$

where N = an integer dependent upon composition and processing of a junction

N = 2, for an Abrupt Junction

$N = 3$, for a Linear Junction

ϕ = Contact potential in volts

ϕ = .32 for silicon

It is obvious from the above equation that if a linear change in capacity was desired, only a small section of the characteristics could be used. Of course, a small signal swing would also mean a small change in capacity and a resultant small frequency deviation. However, it is possible to amplify these characteristics with the aide of a section of transmission line. The variable capacity is used to terminate a section of transmission line, which is connected across a tank circuit. A diagram of the system is shown in Figure (4). The impedance seen at the end of a lossless section of transmission line is given by:

$$Zin = Ro \left[\frac{Z_L + j\,Ro \tan Bl}{Ro + j\,Z_L \tan Bl}\right] \qquad (2)$$

Where Ro = Characteristics impedance of line

Z_L = Load impedance

$Bl = \dfrac{2\pi l}{\lambda}$

l = Length of transmission line

λ = Wavelength of excitation source

If the line is terminated in a lossless variable capacitor then

$$Zin = j\,Ro \left[\frac{\dfrac{-1}{\omega c_1} \cdot \dfrac{1}{\Delta} + Ro \tan Bl}{Ro + \dfrac{1}{\omega c_1}\,\dfrac{1}{\Delta} \tan\ Bl}\right] \qquad (3)$$

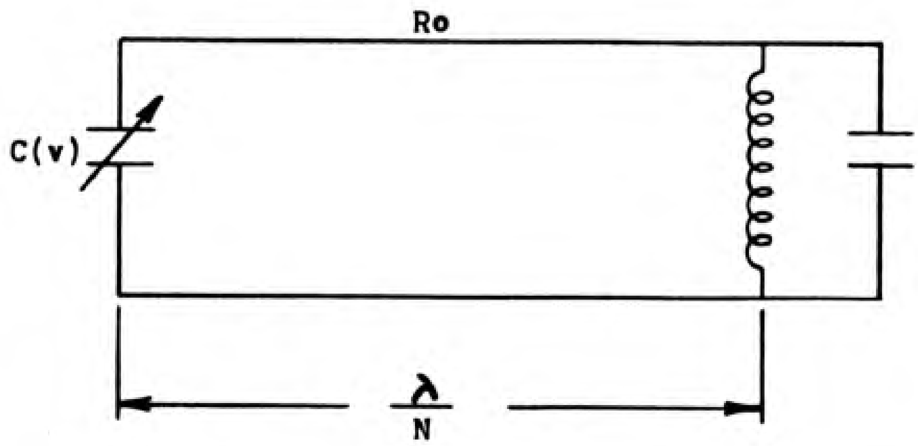


Fig. (4) Capacity Multiplier

from equation (1)

$$\frac{C}{C_1} = \sqrt{\frac{1 - \frac{V_1}{\phi}}{1 - \frac{V}{\phi}}} \equiv \Delta$$

where c = capacity at a voltage V

c_1= capacity at a voltage V_1

Delta is then defined as the fractional change in capacity and equation (3) defines the input impedance in terms of this fractional change in line length. In most applications V_1 is a bias point and is much greater than the contact potential. In addition, the signal swing is considered small so that V is also much greater than the contact potential.

$$\therefore \quad \Delta \equiv \frac{c}{c_1} \approx \sqrt{\frac{V_1}{V}} \qquad (4)$$

An enumeration of equation (3) indicates that the input impedance of the line can be made to look infinite. For V equal to V_1, Delta equal to one.

Let $\tan Bl = -Ro\,\omega c_1$

then from (3)

$Zin = \infty$ or $C = o$

By choosing a longer length of line, the input impedance will be capacitive and a shorter length will appear inductive. We therefore let $\tan Bl = -Ro\,\omega c_1 K$ where K is constant, which determines line length and simplify the equation by letting

$Ro = \frac{1}{\omega c_1}$

then from (3) $Zin = -j\,Ro \dfrac{\left[\dfrac{1}{\Delta} + K\right]}{\left[1 - \dfrac{K}{\Delta}\right]}$ (5)

For a length of line between $\frac{3}{8}\lambda$ and $\frac{\lambda}{2}$ the circuit will appear capacitive.

$$Zin = -j\, X_{c(v)}$$

$$C = \frac{1}{\omega Ro}\left[\frac{\sqrt[N]{\frac{V_1}{V}} - K}{1 + K\sqrt[N]{\frac{V_1}{V}}}\right] \qquad (6)$$

The capacity seen at the end of the transmission line is then defined in terms of the signal voltage across the varactor with the line length as a parameter.

at $V = V_1$

$C = C_1$

and

$$C_1 = \frac{1}{\omega Ro}\left[\frac{1 - K}{1 + K}\right] \qquad (7)$$

dividing (L) by (7)

$$\frac{C}{C_1} = \frac{\sqrt[N]{\frac{V_1}{V}} - K}{1 + K\sqrt[N]{\frac{V_1}{V}}} \cdot \left[\frac{1 + K}{1 - K}\right] \qquad (8)$$

if $K = o$ or $l = \frac{N\lambda}{2}$

$$\frac{c}{c_1} = \sqrt[3]{\frac{V_1}{V}}$$

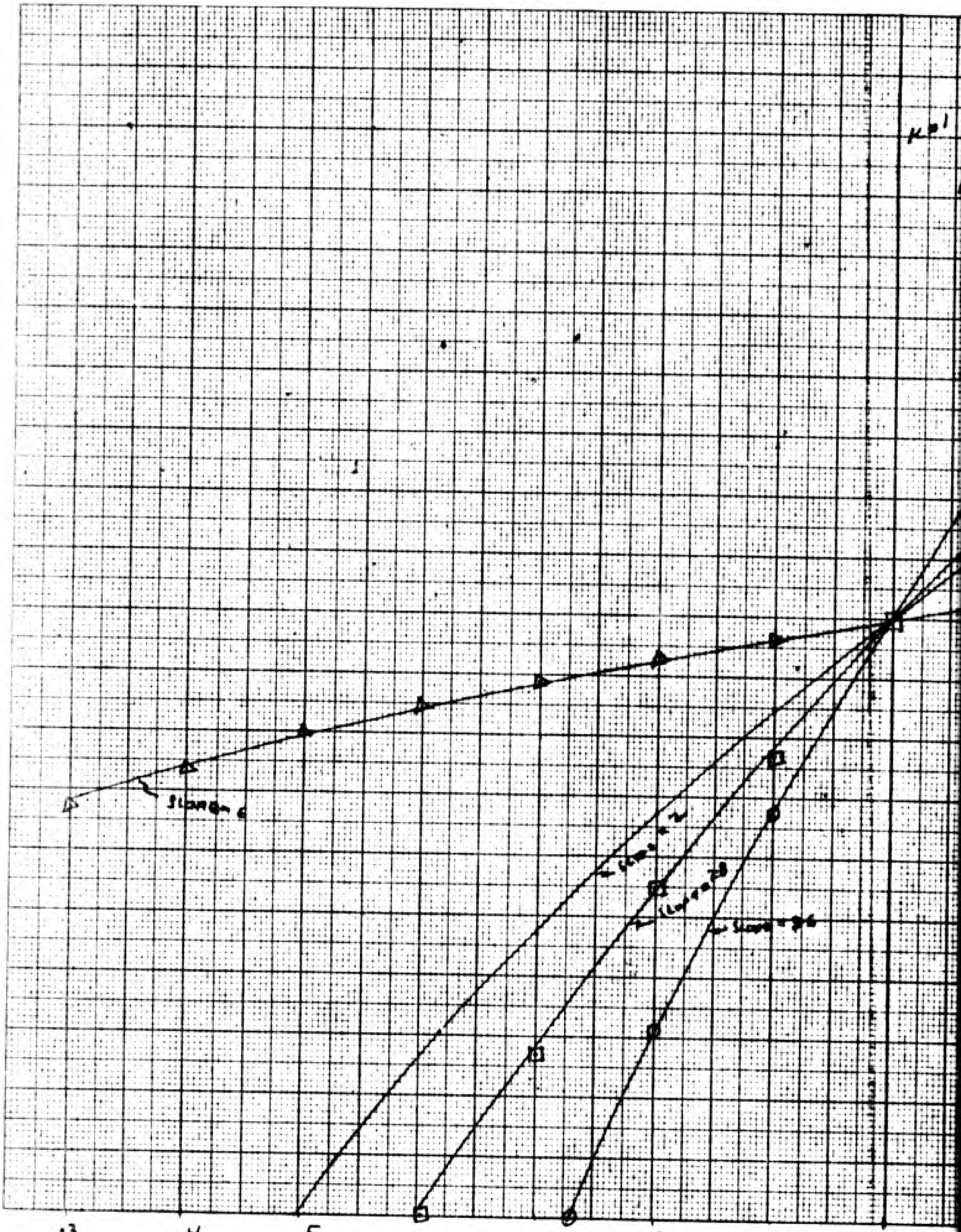
Which is recognized as equation (4) and is the capacitor law of the varactor by itself. A plot of equation (8) is shown in Figure (5) for five values of line length and n equal to 3. The varactor characteristics by themselves are represented by K = 0. For K equal to 0.8, 0.85 and 0.90 the increases in slope are 3.35, 4.65 and 6 respectively. That is, the capacitor voltage characteristics of the varactor have been expanded so that for the same voltage change a larger capacity change is realized. A more graphic form of the process may be shown with the aid of a "Smith Chart". We enter the Smith chart at the point 0 - j 1; that is for the case where the characteristics impedance of the line is equal to the quiescient varactor reactance. By varying the value of K, this point is transformed along a line of constant VSWR. For K = 1, the impedance is transformed to an open circuit or zero capacity point. For value of K greater than 1, the impedance is inductive and to K less than 1 it is capacitive. Examination of the reactance around K = 1 would show rapid changes in reactance for small changes in capacity. It should be pointed out the signal swing is limited to:

$$\frac{V_1}{V} > K^3 \quad \text{and } K < 1$$

so that as K approaches one, the increase in slope approaches infinity but the allowable signal swing goes to zero. The derivation can be extended for value of K>1 by setting equation (5) equal to + $j\omega L$.

In this case

$$\frac{L}{L_1} = \frac{K - 1}{K + 1} \cdot \frac{1 + \left(\frac{V_1}{V}\right)^{1/N} K}{K - \left(\frac{V_1}{V}\right)^{1/N}}$$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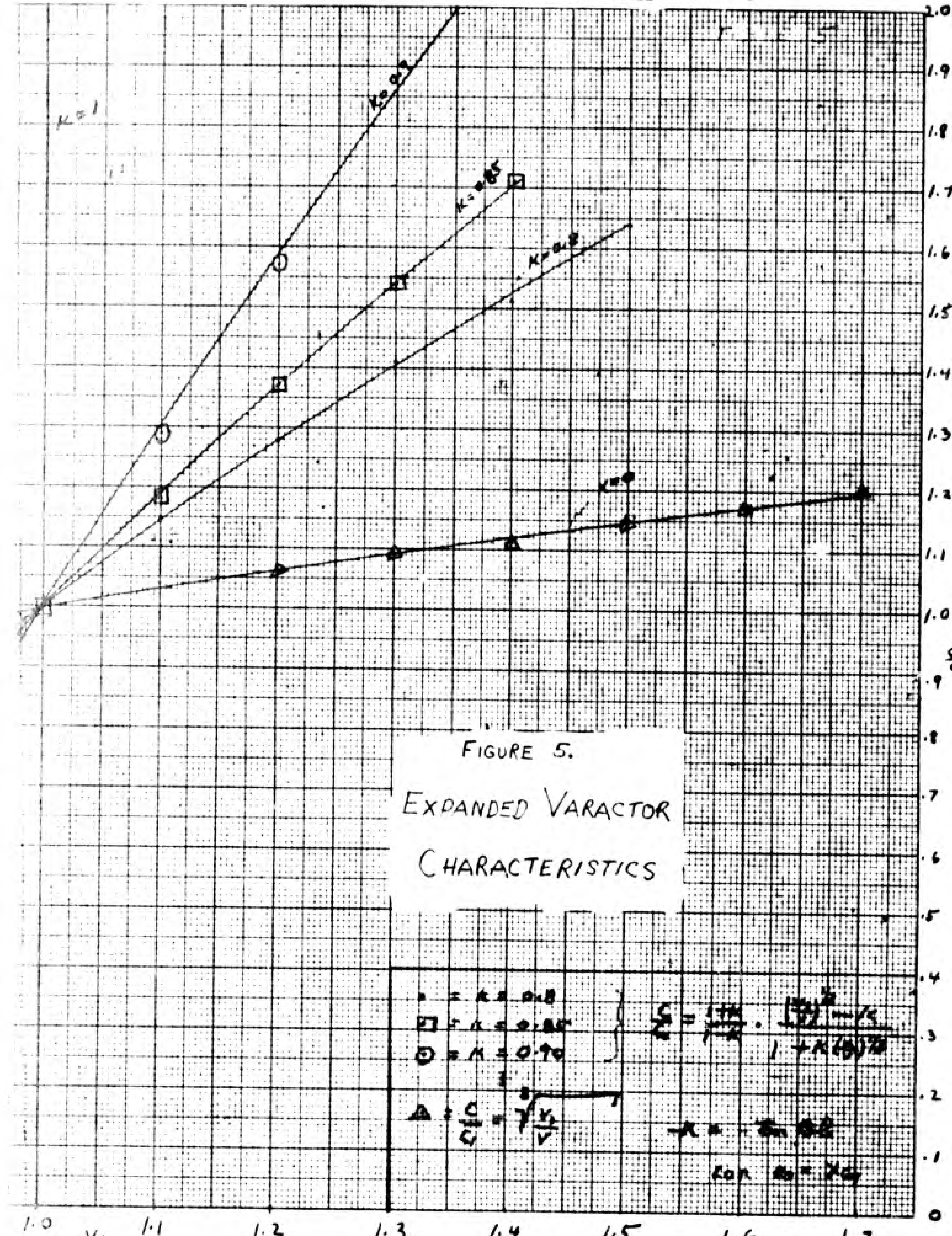


FIGURE 5.

EXPANDED VARACTOR CHARACTERISTICS

The same gain in slope and signal swing limitation is realized for this case. If the variable reactance is to be coupled in parallel with a tank circuit, then K should be less than 1 since the total capacity is the addition of all capacity. If the variable reactance is to be coupled to a series resonant tank circuit, then K should be greater than one since the total inductance is the sum of all inductances. It should also be pointed out that in the practical case the varactor is not a pure variable capacitor, but included a small amount of series resistance. This means that instead of transversing the Smith chart on an infinite VSWR circle the variable impedance is transformed along a finite one. For small values of series resistance, the loss in slope magnification is small which indicates that the varactor should have the lowest possible value of series resistance. Based upon the concept of capacity amplification described above, it was decided that the varactor configuration would be employed. The system was passive and thus required no power and could be fabricated in a small volume.

2.2.1 Modulation Amplifier

The Modulation Amplifier was required to have a 470 K ohm input impedance with a bandwidth capability of 500 K cps. Preliminary design conditions indicated that a circuit gain of from 5 to 10 was needed for a 1 volt input signal. The gain bandwidth considerations were felt not to be difficult since several relatively simple methods could be used to achieve the results.

However, the high input impedance over the large bandwidth did present a severe problem. The emitter follower configuration offers a high input impedance which is essentially Beta times the load resistor shunted by the parallel combination of collector resistance and capacitance. By using several of these circuits in cascade, the input impedance can be made to be in the megohm region. The disadvantage of this configuration is in the relatively poor bias stability, shunting effort of the collector capacity and limiter frequency response. Other techniques for providing high input impedance make use of negative feedback; that is the voltage gain and output impedance is lowered in exchange for a high input impedance. A circuit using this principle was described by Evans[(4)] in which impedance in excess of 1 megohm were obtained on a wide bandwidth.

In addition, Middlebrook and Mead[(5)] described a high input impedance amplifier which utilized both positive and negative feedback. A diagram of this circuit is shown in Figure (6). Here negative feedback is accomplished through the common load impedance R_1. It can be shown that the input impedance is in the order of $Beta_1$, $Beta_2$ times R_1. However, this impedance is now shunted by the effective Bias impedance of the input transistor, as well as its collection capacity and resistance. In addition, stray capacity introduced by cable and connector effectively shunt this impedance. A positive feedback loop consisting

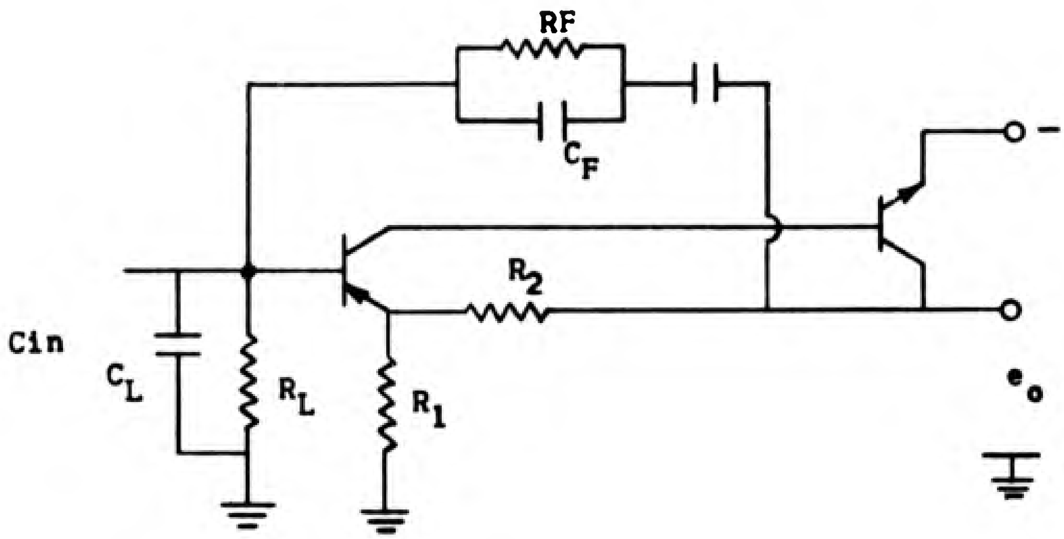


Fig. (6) High Input Impedance Amplifier

of RF and CF are utilized to minimize this effect. This cancellation can best be explained with the aid of Figure (7) where A_1 and A_2 are the amplifier and B_1 is the negative feedback path. The effective R_f, C_f combination form an all pass filter in conjunction with $R_L C_L$; that is, $R_f C_f$ (effective) is equal to $R_L C_1$. Thus stray capacity seen at the input terminals is compensated for. Amplifiers of this type have been built with several hundred megohm input impedances shunted by no more than a couple of micro-micro farads.

It was decided that two prototype configurations would be constructed. The first would consist of the negative feedback amplifier similar to that described by Evans. In addition, the advantages of the combination type amplifier would be investigated so that the best possible configuration would result.

2.3 Frequency Stability

Several techniques were considered for use in the automatic frequency control circuit. These included the "phase Lock Loop", the crystal discriminator, and a frequency lock loop. A brief description of these techniques follows.

2.3.1 Phase Lock Loop

A diagram of a typical phase lock loop is shown in Figure (8A). Here the output of the transmitter is converted to a lower frequency and is then compared with a reference oscillator is a phase detector.

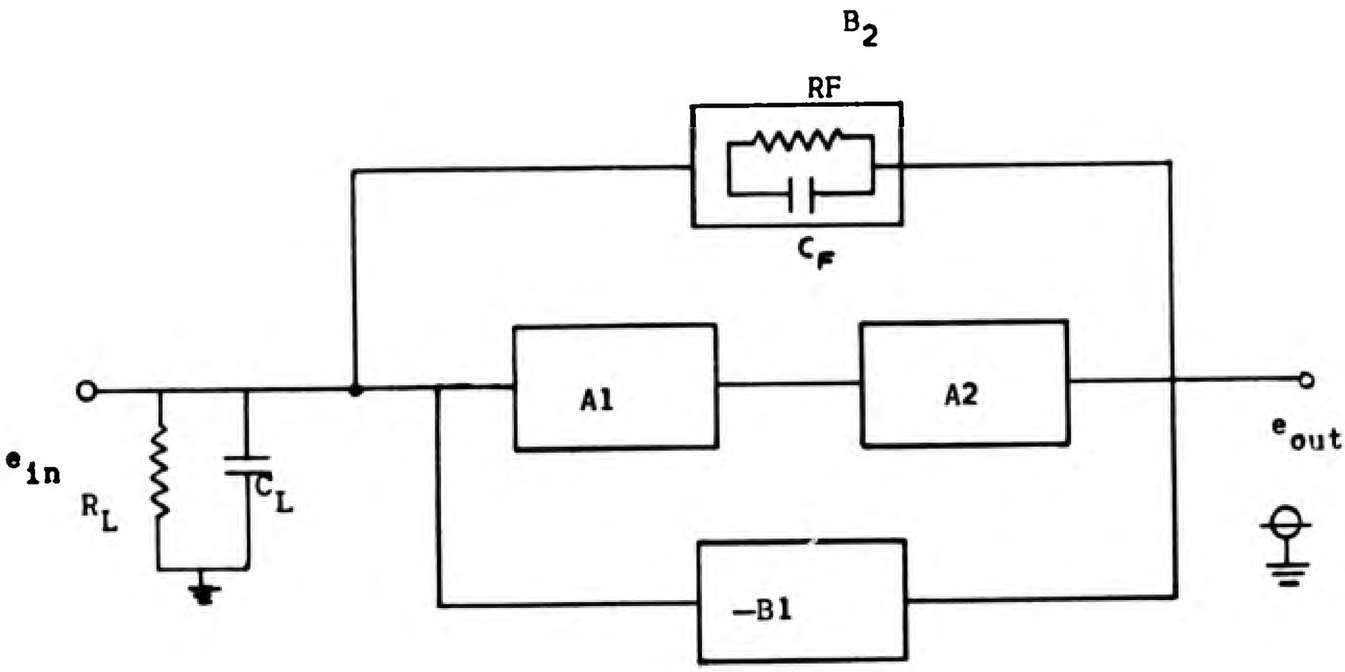
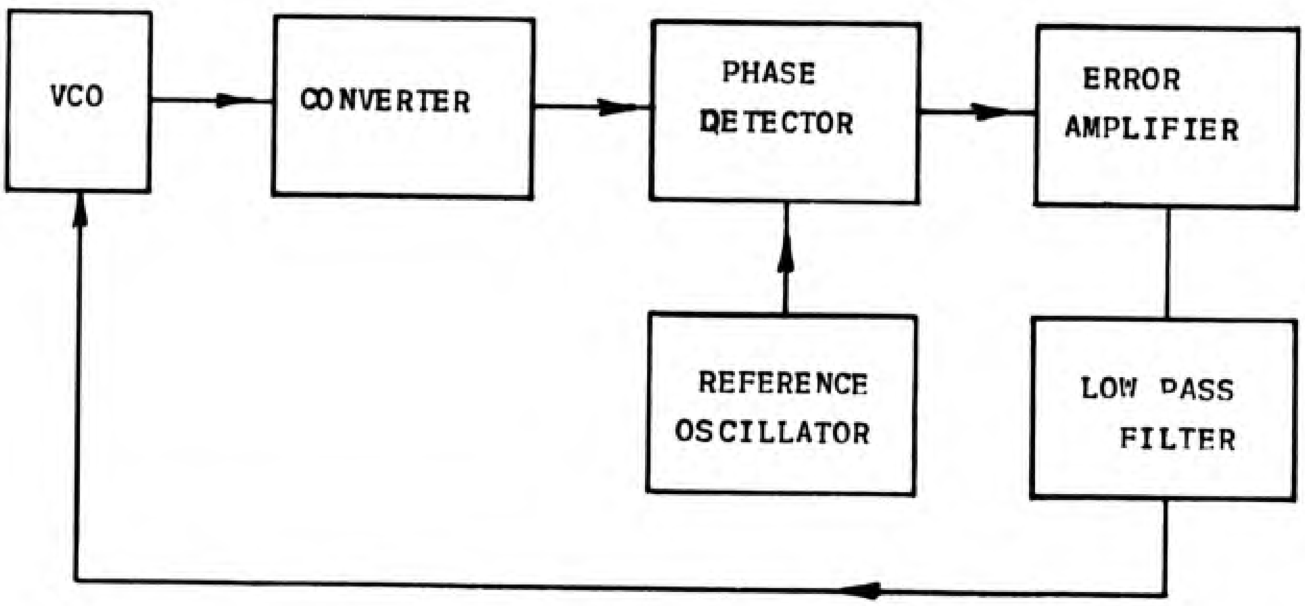
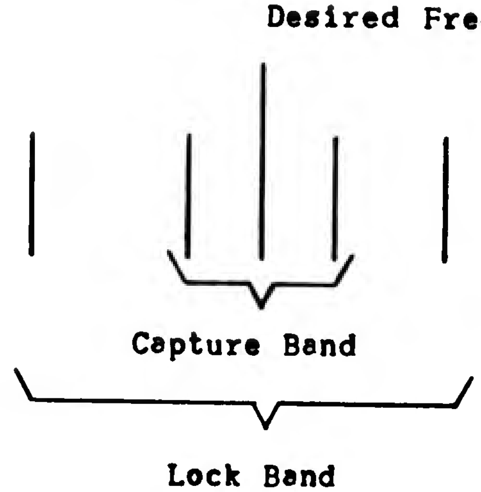


Fig. (7) Capacity Cancellation Circuit



(a)



(b)

Fig. (8) Phase Lock Loop

An error output, proportional to the difference is phase between the two signals, is amplified and fed through a low pass filter to the frequency control device. The system has a theoretical zero frequency error for a finite phase error. The lock range of the system, which is defined as that frequency band for which the circuit will hold the transmitter on frequency, is dependent upon the loop gain. The capture range of the system, which is defined as that frequency band which the transmitter must be adjusted within in order for frequency locking to take place, is dependent upon the characteristics of the low pass filter. These bands are shown in relationship to the desired frequency in figure (8B). In operation the transmitter frequency is adjusted to within the capture range of the system and is then controlled by the loop. If the transmitter is brought to a region between the lock and capture band, the loop is unstable and the transmitter will hunt between the two bands. Momentary drifts out of the lock band may result in an open loop condition. Now, since the transmitter was to be frequency modulated, the loop had to be designed so that for all rates of change of frequency within the modulation band, no error signal would be fed back to the transmitter. If this was not done, the loop would tend to decrease the amount of frequency deviation on the carrier. Therefore the low pass filter had to be designed so that the lowest sub-carrier frequency (400 cps) would not effect the loop. In doing this, the capture range was greatly reduced adding to the complexity of tuning the transmitter. It was also necessary that the lock range be as wide as possible to prevent an open loop condition due to frequency pertibation caused by shock and vibration environments. The required gain for a 2 or 3 megacycle lock range was considered excessive when compared to other systems.

2.3.2 Crystal Discriminator

Several systems which utilize a crystal discriminator were investigated. A block diagram of a typical system is shown in Figure (9). Here the transmitter frequency is converted down to a lower frequency and then fed to a crystal discriminator. The center frequency of the discriminator is chosen to be the desired transmitter frequency divided by the conversion factor so that a DC voltage proportional to the error frequency is obtained. This signal is fed through a low pass filter and used to correct the transmitter frequency. The system performance is directly related to the accuracy of the crystal discriminator and the loop gain of the system. In addition, if a local oscillator is incorporated in the converter, its stability will also be a determining factor in the system. Since the loop is to be deactivated for rates of change of frequency in the modulation band, a low pass filter is required. The effective lock in frequency of the system is determined by the bandwidth of the crystal discriminator and typical units are in the order of ± 100 KC. In order to accomodate the large tuning range, either the local oscillator or the crystal discriminator frequency had to be returned. The many disadvantages of this system showed it to be unaccpetable for a wide band FM transmitter.

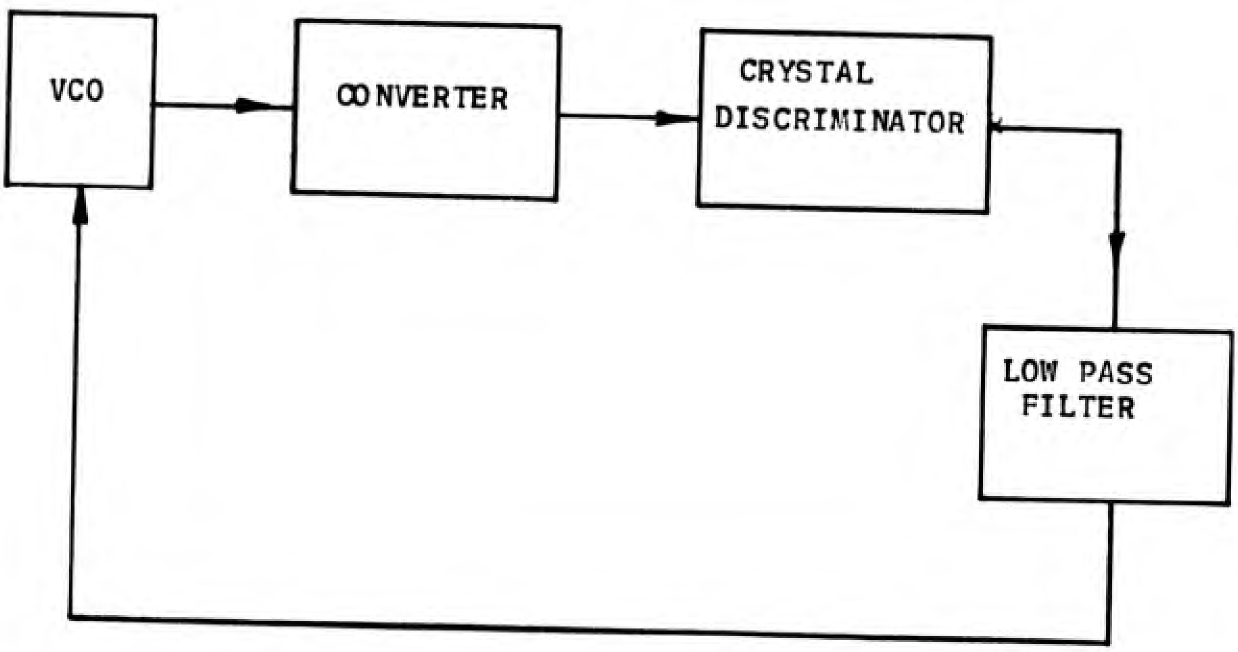


Fig. (9) Crystal Discriminator

2.3.3 Frequency Lock Loop

A diagram of the basic configuration is shown in Figure (10). A reference oscillator is used in conjunction with a converter to bring the transmitter frequency down to a lower frequency which is then fed to a gated amplifier. At the same time, an additional gated amplifier is fed with a sample of the reference oscillator frequency. The amplifiers are alternately gated on and off by an oscillator or switch so that the input to the discriminator is alternately the reference oscillator and then the transmitter sample. The output of the discriminator is a signal at the gate frequency whose amplitude is proportional to the frequency differences between the reference and transmitter frequencies. Again a low pass filter is utilized to deactivate the loop to the lowest sub-carrier frequency. After passing through the filter, the signal is given phase sense by comparing it with the gate oscillator in a phase detector. The resultant DC is then used to control the transmitter frequency. The system has several advantages in terms of stability and lock in range. The absolute limit on the transmitter stability is determined by the stability of the reference oscillator and the loop gain of the system. The center frequency and linearity of the discriminator are relatively unimportant. In addition, the lock in range is determined by the bandwidth of the discriminator amplifier combination and since it is not crystal controlled can be made extremely wide. The necessary loop gain for a given stability is analyzed with the help of Figure (11).

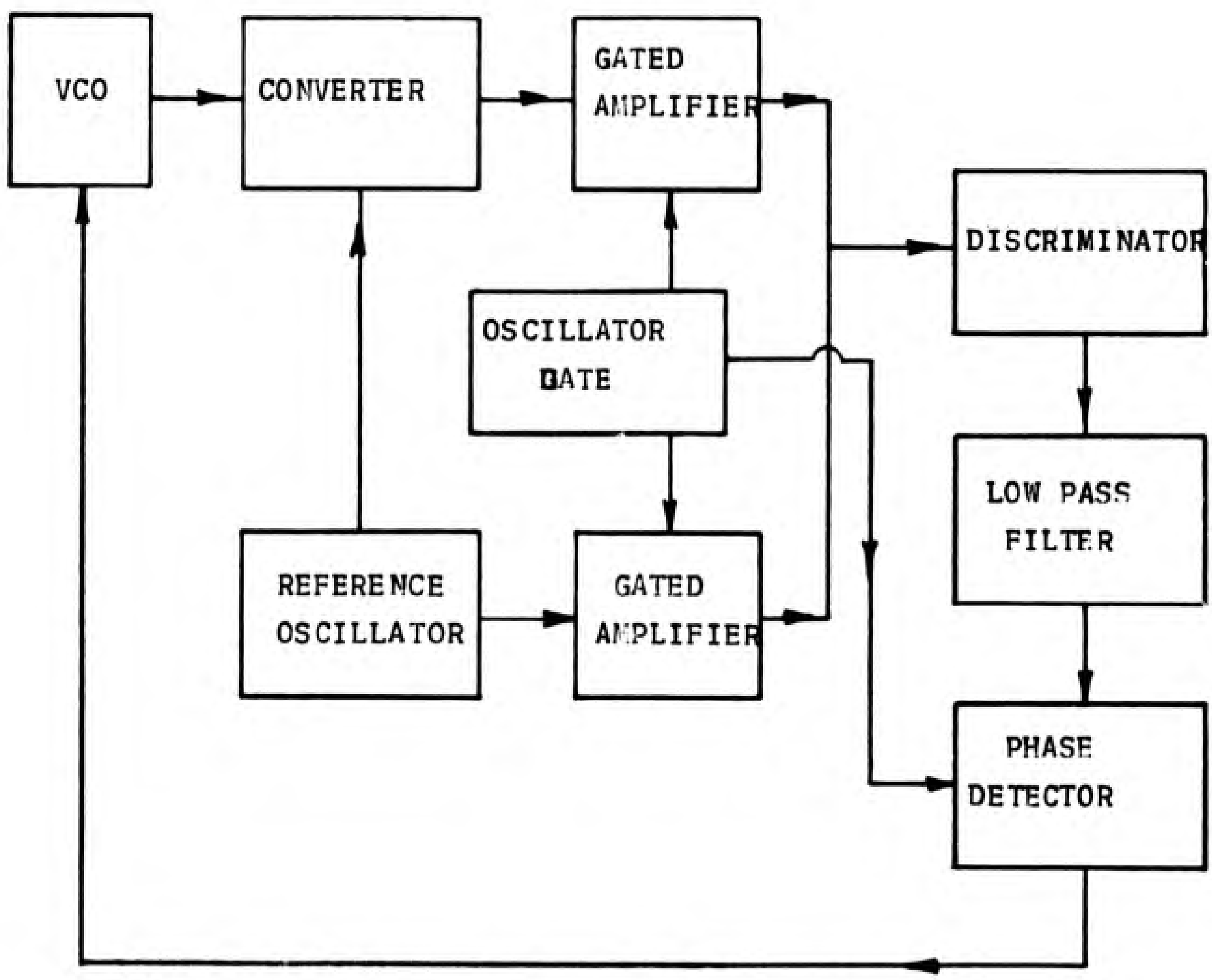
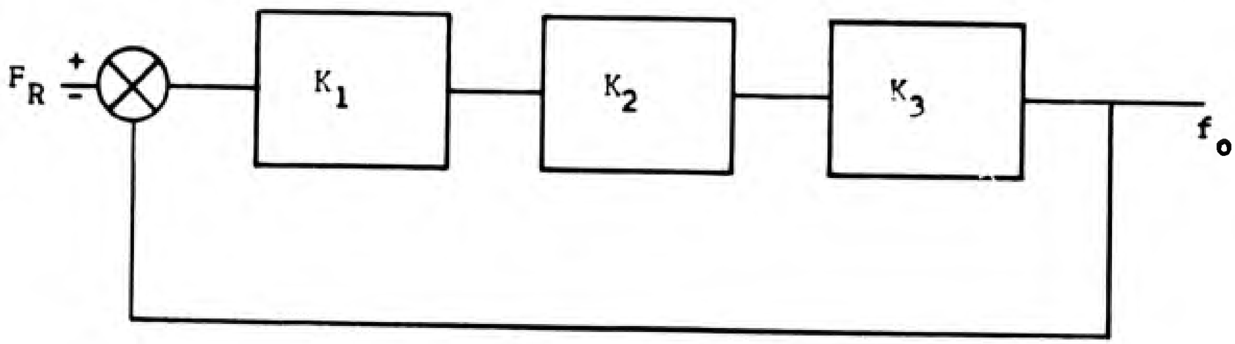


Fig. (10) Frequency Lock Loop



$6\ (s) = K_1K_2 = K_3$

Fo = Output Frequency

Fr = Reference Frequency

Fig. (11) Loop Gain Diagram

The system is recognized as a type zero servo loop with maximum steady state error equal to:

$$\Delta f = \frac{\Delta F}{1 + 6\ (o)} = \frac{\Delta F}{1 + K_1 K_2 K_3}$$

where: Δf = maximum steady state frequency error

ΔF = expected frequency drift due to thermal, vibration, and shock environment

K_1 = Discriminator constant with dimension of $\frac{\text{volts}}{\text{meg cps}}$

K_2 = Amplifier gain with dimension of $\frac{\text{Volts}}{\text{Volts}}$

K_3 = VFO transfer function with dimension of $\frac{\text{MCps}}{\text{Volts}}$

If we define the stability factor as $S \equiv \frac{\Delta f}{\Delta F} = \frac{1}{1 + K_1 K_2 K_3}$

then we see that the larger the loop gain the smaller will be the frequency error. The required stability was $\pm$ .005% or 50 parts per million. If we allow half this figure for the reference oscillator then the maximum drift of the transmitter is 25 parts per million. The expected uncontrolled drift of the oscillator is about 1000 parts per million. Hence a stability factor

$$S = \frac{25}{1000} = \frac{1}{1 + K_1 K_2 K_3}$$

or the loop gain

$K_1 K_2 K_3 = 40 - 1 = 39$

Based upon the previous discussion the above system offered the following advantages:

1. No difference between lock and capture range.

2. Lock range of system could be made relatively large, then simplifying tuning and reduce the risk of an unlocked loop.

3. Overall stability is determined by loop gain and stability of the reference.

4. Characteristics of discriminator relatively unimportant.

5. Lock range independent of low pass filter time constant.

It was concluded that this system offered the best possible solution to the problem of frequency stability for an unmanned FM transmitter.

3.0 TRANSMITTER - DEVELOPMENT PHASE

The development of the transmitter was broken down into six categories. They were:

1. Power Oscillator
2. Power Amplifier
3. AFC Circuit
4. Modulator
5. Power Supplies
6. Control Box

Although these boxes were interrelated, work was begun on the first four sub-sections with the power supplies and control box initiated as more information became available.

3.1 Transmitter-Block Diagram

A block diagram of the initial design is shown in Figure (12). A power oscillator designed to deliver 5 watts was to be fabricated which would drive a power amplifier for the necessary 15 watts output. A frequency lock loop would be used to hold the power oscillator on frequency. Due to the critical operating characteristics of the varactor, seperate varactors would be used for the AFC and modulation circuits. The varactor would be coupled to the power oscillator to convert the error voltage to a change in capacity and thus control the frequency of the power oscillator. A modulation system incorporating the capacity multiplier varactor circuit completed the system.

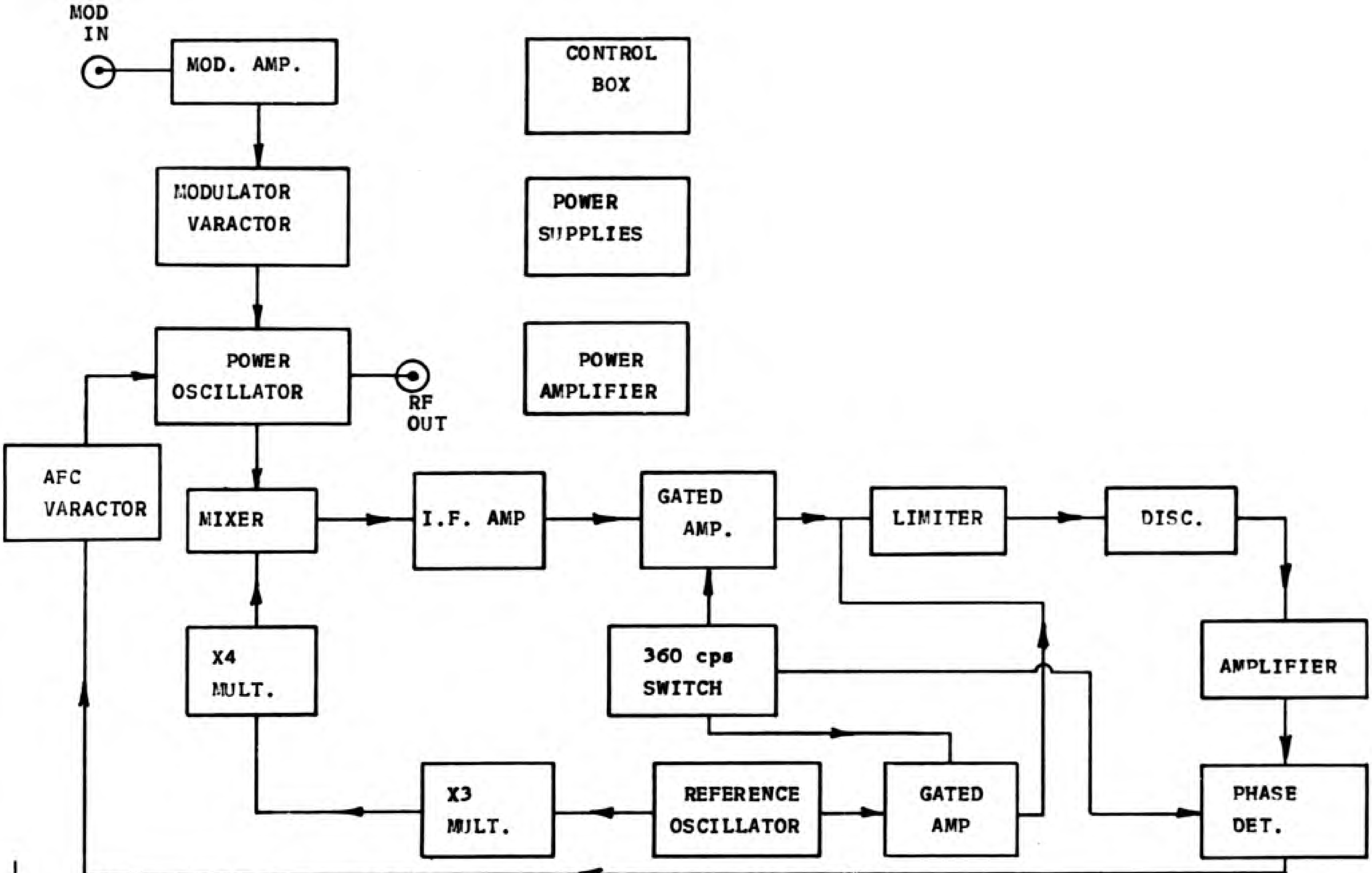


Fig. (12) Initial Transmitter Block Diagram

3.2 Power Oscillator

As initially conceived, the AN/AKA-( ) transmitter consisted of a five watt oscillator and a fifteen watt power amplifier. The initial investigation indicated that the oscillator would utilize a tube in the 2C39 family. In order to achieve maximum heat transfer it was decided that the tube would be operated in the grounded plate configuration. The cavity was of the grid separation type with the anode folded back over the cooling fins of the pnalar triode. Tuning of the anode line was accomplished with a capacitive probe while the cathode cavity was tuned with an adjustable spring fingered short. The grid was insulated from ground with a mica bypass capacitor and provision was made for bias adjustment. Initial tests on the cavity were not encouraging. Considerable difficulty was encountered with voltage breakdown in the grid bypass capacitor and the unit exhibited a tendency to oscillate at "L" band. Even when these problems were solved, it was noted that the oscillator had a tendency to "squegg". In an effort to facilitate the design of the unit, a commercial cavity for the 2C39 was purchased. This cavity exhibited squegging characteristics which were more severe than those of the developed cavity. Further investigation indicated that this cavitv, as well as most of the cavities available for the 2C39 had evolved from a test cavity utilized by the General Electric Company in the development of their planar triodes. Drawings of this cavity were acquired and a model was built. Initial tests on this cavity showed a squegging condition as well as considerable "L" band output power. After considerable consultation with General Electric's application department, modifications were made on the cavity and the "L" band output power was reduced. However, no success was obtained in eliminating the "squegg" problem. Further tests were run by the General Electric Company on their own cavity and their data confirmed the presence of the "squegg" condition. It was found that this squegg could only be eliminated with operation of high anode currents. The data was confirmed by both the General Electric Company and Machelett Laboratories. Successful operation of the cavity was eventually obtained over the band (2150 to 2350) with a minimum 15 watts of output power.

It was then noted that the successful operation of the cavity at this level of power would eliminate the need for a power amplifier stage. Further tests and modifications were made on the General Electric Test cavity which indicated satisfactory results for the desired 15 watts power level. Based upon these tests, agreement was reached with the government to eliminate the power amplifier from the transmitter configuration. This was advantageous in terms of producing a smaller, more efficient and lighter transmitter package.

In final breadboard form, the power oscillator consisted of a coaxial type tuned plate, tuned grid cavity. The plate grid cavity was folded over the cathode cavity to conserve space. Both of these cavities were tuned with capacitive plungers. Tests were conducted on various planar triodes, which included: 2C39B, GL6897 and 3CX100A5. The 2C39B/GL6897 was found to produce the most reliable and repeatable performance and was then incorporated in the final configuration. Temperature runs were made on several of these cavities made of Invar, Brass, Stainless Steel and Aluminum or combinations of these. It was found that an all Invar cavity showed a minimum frequency drift for temperature changes. This cavity showed a maximum 300 KC drift over a 150°C temperature excursion. Since it was desirable to keep this drift to an absolute minimum, the Invar configuration was utilized in the final cavity.

The development of the cavity then proceeded with the design of a set of Invar cams which would drive the tuning plungers. Temperature tests again confirmed the suitability of this modification and a program incorporating varactor modulation proivisions was initiated. It was found that power limitations in available varactor diodes prohibited modulation of the anode cavity. However, attempts to modulate via the cathode cavity met with some success. This success was eventually optimized by redesigning the feedback loop and the tuning plunger cams.

The present cavity is the outcome of work on three distinctly different oscillator designs (two coaxial mode and one radial mode) and emobódies a large number of "firsts", the most notable being the use of sintered beryllium oxide disc to heat sink the anode of the planar triode to the transmitter case and the use of sintered aluminum oxide rods to transmit motion through the anode cavity to the cathode tuning plunger.

In retrospect, the most formidable cavity design problem was the overcoming of "squegg". This phenomenon is barely mentioned in the literature and when it is mentioned, it is only to acknowledge that it had been observed. The most enlightening reference to be found consisted of a statement in the Radio Amateur's Handbook which defines "squegg" as a phenomon which can occur when the electron transit time within a tube becomes an appreciable portion of a RF cycle. The revelation of the knowledge of "squegg" gained in this program is rightfully the subject of another paper so suffice it to say that the phenomenon is directly related to cathode bypass capacitance, feedback magnitude, feedback phase, anode bypass capacitance, grid current, anode current, power supply impedance, and the physical structure of the tube employed.

3.3 Power Amplifier

The initial plan was to develop a 5 watt power oscillator and using this basic configuration the feedback loop would be removed and a connector added to the cathode line. However, when time consuming difficulties were encountered with the oscillator, it was decided not to schedual the amplifier development as a serial program but to investigate the oscillator breadboard that had been previously discarded as unsatisfactory. This parallel program investigated two radial mode and one coaxial mode amplifiers only to find them unsatisfactory. The configuration that finally proved satisfactory was a coaxial mode cavity which never went beyond the breadboard stage in view of the ensuing success with the high power oscillator.

3.4 AFC Circuit

The initial AFC Circuit is shown in Figure (12). It consisted of a transistorized crystal controlled oscillator, which served as the reference for the system, operating in the 60 MC region. Two outputs were taken from this circuit; the first fed a Harmonic Multiplier Chain and the second was used as an input to a gated amplifier. The Harmonic Generator consisted of a 4th harmonic multiplier in cascade with a 9th harmonic unit. The 9th harmonic generator was built into a coaxial cavity which also included a Mixer. The multiplier output was then heterodyned with the sampling from the transmitter in the mixer and the difference frequency selected in an IF Amplifier. This amplifier fed a gated amplifier which was turned on and off at a low frequency rate. The two gated amplifiers had a common collector load so that the input to the limiter was alternately the sampled transmitter frequency and reference frequency. The output from the discriminator was then a square wave at the switch frequency whose amplitude was proportional to the error of the carrier frequency. This error was then amplified and fed to a phase detector whose reference was taken from the same switch source which fed the gated amplifier. The output of this detector was a DC Voltage proportional to error and either of positive or negative polarity depending on whether the transmitter was higher or lower than the desired frequency. The DC voltage was coupled through an emitter follower and severed as the bias source for the AFC Varactor. A detailed description of the development follows.

3.4.1 Design Consideration

The frequency of operation for the oscillator is determined in the following way.

Let: ft equal to the desired transmitter frequency

fr equal to the reference oscillator frequency

N equal to the order of frequency multiplier

Then

(3.1) $(N + 1)\ fr = ft$

So that the IF frequency is given by

(3.2) $F_{IF} = F_T - Nfr$

Comparison of the equation 3.1 and 3.2 shows that the IF frequency is equal to the reference frequency. Thus as the transmitter frequency is changed, the reference oscillator frequency is changed. In order not to retune, the bandwidth of the system must encompass this frequency change. So that

$$\Delta F_{IF} > \frac{F_T}{N+1}$$

Where ΔF_T = Transmitter Tuning Range

ΔF_{IF} = System Bandwidth

It can therefore be seen that the choice of nominal reference oscillator frequency is a compromise between practical crystal frequency, order of multiplication for the harmonic generator and IF frequencies. A survey of available crystals indicated that a nominal 60 MC unit was the best compromise between drive power, stability and bandwidth consideration. This dictated a multiplication factor of 36 for the harmonic generator and a practical 60 MC frequency for the transistorized IF strip.

Having decided upon frequency and bandwidth considerations, the necessary power levels were investigated. Since the available power from the multiplier chain was limited, it was concluded that the transmitter would supply the necessary pump power for the mixer circuit. The circuit would behave as a superhetrodyne receiver with the transmitter sample acting as a local oscillator and the output of the multiplier chain as the received signal. Investigation indicated that the available power from the reference oscillator would be in the order of plus 17 DBM. Page[6] had shown that the efficiency of a harmonic generator employing ideal diodes is limited to $1/N^2$, where N is the order of multiplication. Although the use of varactors as harmonic generators was limited at this time, indications were that much greater efficience could be accomplished. Leeson and Weinrab[7] reported a conversion loss of 1 DB for a doubler with a predicted loss of 5 DB and 2 DB for a 4th harmonic and 3rd harmonic multiplier respectively. Additional literature[8] indicated that the expected loss in DB for a multiplier was 1.5 N for N greater than two. Little or no work had been done with higher order multiplier but it was the general conclusion of most authors that the efficiency of these circuits would approach $1/N^2$. In order to reduce problem of cascading stages and simplify tuning, a maximum of two stages would be incorporated in the multiplier chain. Two schemes would be evaluated. The first would consist of a 3rd harmonic generator feeding a 4th harmonic generator and utilized 3rd harmonic mixing in the mixer stage. The second configuration would incorporate a 4th harmonic generator in cascade with a 9th harmonic generator and use fundamental mixing. The expected power available at the IF amplifier for the first configuration was approximated to be -23 DBM considering a 30 DB loss for 3rd harmonic conversion efficiency. The estimated power available from the 2nd configuration was in the order of -16 DBM assuming a 6 db loss for the mixer. Setting zero dbm for the limiting level, a gain of 40 DB was concluded for the IF amplifier and switch circuit. This would insure adequate limiting tolerance.

3.4.2 Reference Oscillator

A schematic of the basic oscillator is shown in Figure (13). It consists of a modified Hartley oscillator with a series resonant crystal in the feedback loop. At frequencies other than the crystal frequency the feedback is reduced considerably and the circuit will not oscillate. The crystal employed was a 3rd overtone unit with a non-oven stability of 25 parts/million over a temperature range of -55 to 105°C. Initial test showed that external loading could produce frequency pulling of the crystal frequency and indeed, in extreme cases, stop operation of the circuit. It was concluded that this was the result of the finite Q of the crystal which determines the differences between the series and parallel resonance frequencies. To keep this

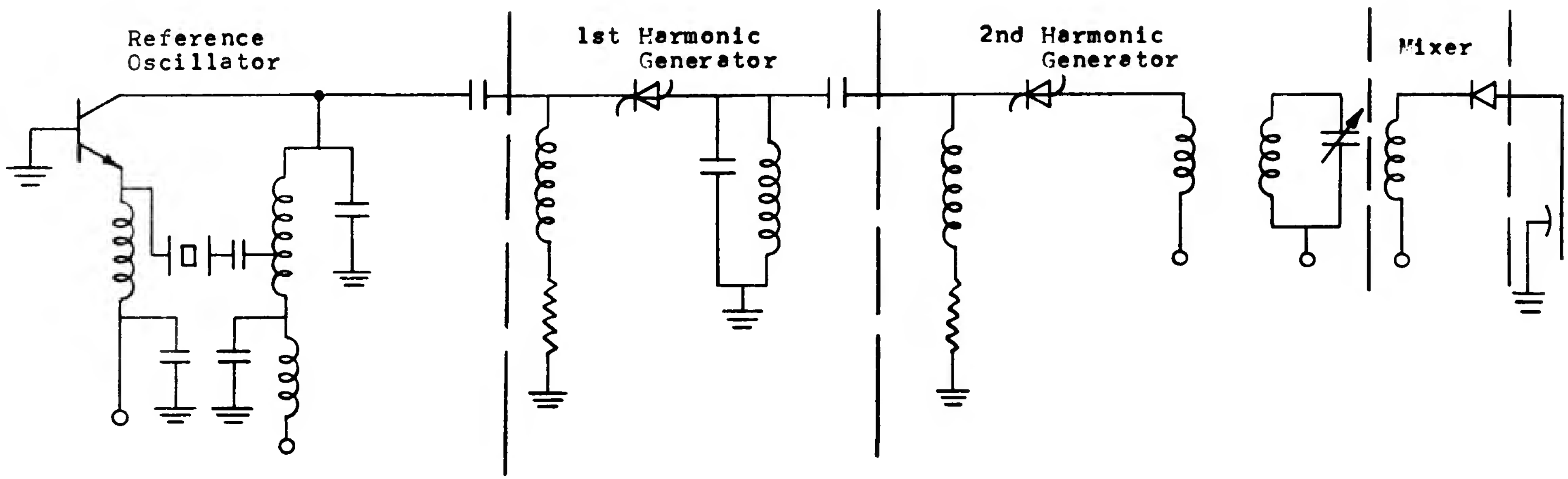


Fig. (13) Basic Oscillator Multiplier Chain pulling at a minimum the collector tuning and coupling components were chosen for low temperature coefficiencts. Final tests on the breadboarded unit are shown in Figure (14). With a temperature excursion of -65 to +115°C the output power changed by only 3 db while the frequency drift was in the order of 2 KC. Similar results were obtained at frequencies at the extremes of the band.

3.4.3 Multiplier Chain

Relatively little was known about varactor multipliers when the chain was initially designed. Indeed, the work bordered on the "State of the Art". However, their use was essential to meet overall power and packaging requirements. The varactors chosen for the chain were the Microwave Associates type MA460A, which had a cutoff frequency of 20 KMC. The basic configuration employed is shown in Figure (13) and is defined in the literature as the series mode of operation. No idler tanks were used in order to simplify tuning and conserve space. Initially, both varactors were back biased such that the RF swing would not drive them into the forward conduction region. It was the general opinion of those in the field that if forward conduction took place in the varactor, the efficiency of the stage would be drastically lowered. However, the reverse of the theory was found to be true; maximum efficiency was obtained when the units were operated with a small self bias induced by forward current flow. It is interesting to note that a recent publication by Microwave Associates[9] has indicated that self bias is now considered the optimum operating mode.

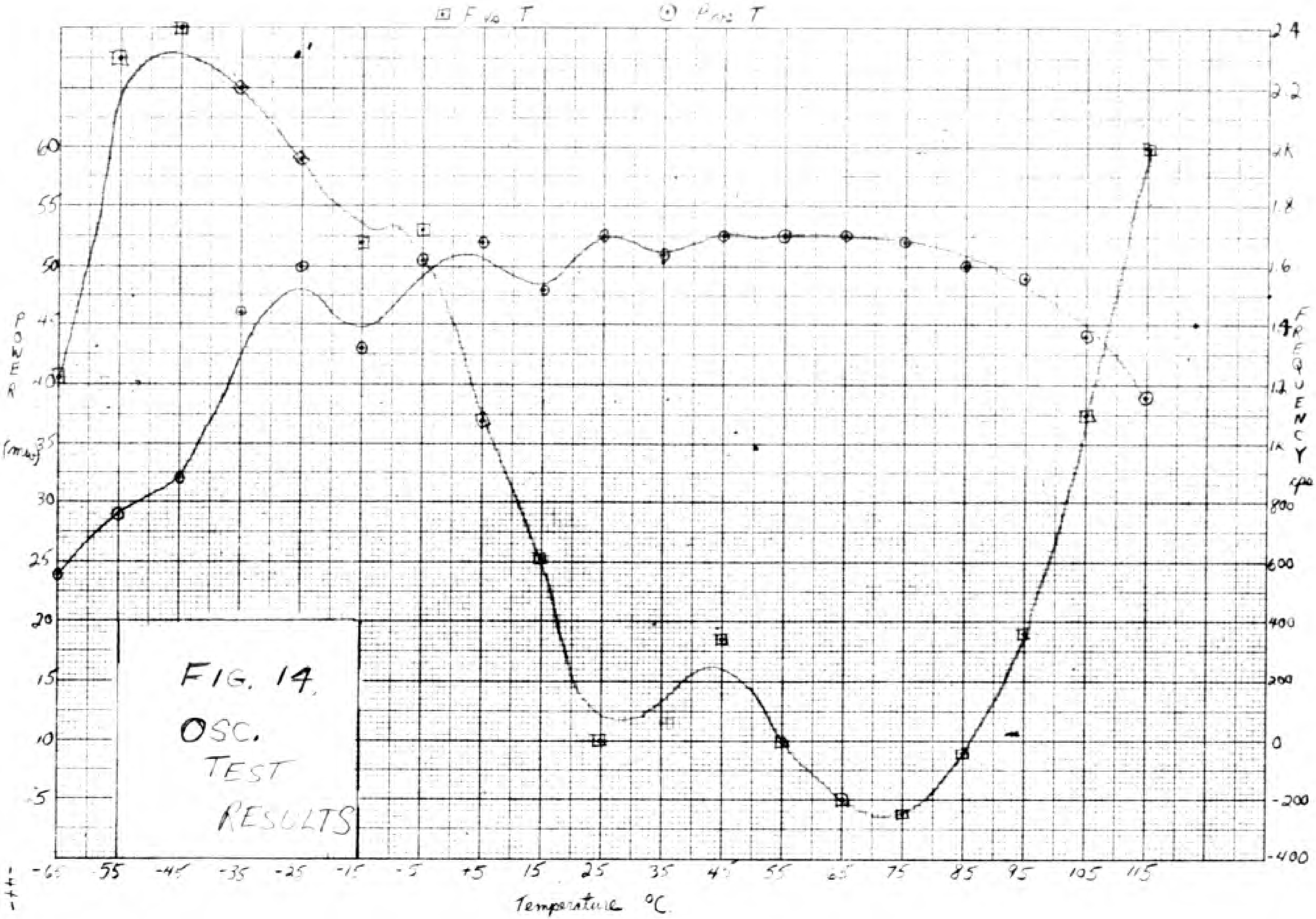


FIG. 14. OSC. TEST RESULTS

An initial 4 times multiplier was designed and breadboarded. Estimated conversion loss of this circuit was 7 db at 240 MC. However, when attempts were made at operating the circuit over the band without returning, the conversion loss increased rapidly. When two of these stages were cascaded together, the circuit became extremely unstable with parasitic oscillation occurring at various harmonic frequencies. In addition, the loading to the reference oscillator was severe and light coupling was necessary for stable operation. It was concluded that a fixed tuned system was unfeasible for the required bandwidth needed. However, a logical tuning procedure was formulated which made adjustment of the circuits practical. It was found that by sampling the DC current flowing through the varactor and adjusting it for a maximum by tuning the previous stage, optimum output was obtained. Although the magnitude of the current changed for a given drive level when the varactor was replaced with another unit, optimum tuning occurred when the current was maximized. With the completion of the 4th harmonic generator a 3rd harmonic unit was constructed. Since the output circuit would be tuned to a nominal 720 MC lumped parameter circuits was considered unpractical. Therefore the circuit was designed into a coaxial cavity together with the mixer diode. The cavity was fabricated from a section of waveguide whose dimensions were chosen so that the frequency of operation was well below its cutoff frequency. The center section of the coaxial line is formed with a circular post with the walls of the guide acting as the outer conductor.

The end capacity of the line is varied by a probe placed in proximity to the center conducter thus changing its resonant frequency. A drawing of this unit showing the mixer and varactor diode mounting is shown in Figure (15). Tests results with this configuration showed a level of -30 dbm available at the IF amplifier. Similar tests were run with the 2nd Harmonic generator stage modified to tune to the 9th Harmonic. A 6 db improvement was noted in signal level when compared to the multiplier Harmonic mixing configuration. This level was 8 db lower than expected, which was justified considering that no idler circuits were used in the multiplier chain.

3.4.4 IF Amplifier, Gated Amplifier and Discriminator

An IF amplifier was fabricated using 2N706 transistor. The gated amplifier was designed around a 3N35 tetrode transistor which was chosen for its excellent on to off RF signal rejection. However, initial tests indicated that only 25 db of isolation was obtained using these units. A high conduction diode (1N830A) was added in series with each output and these were switched concurrently with their associated transistors. Isolation was improved to 40 db with this configuration. Tests also indicated that the gain of these stages would be lower than predicted due to the losses in the diodes and the mutual loading effects. An additional amplifier was inserted at the input to the discriminator to raise the level and insure good limiting. A conventional Foster Seely discriminator is employed as a detector. Back to back diodes placed across the collector tank serve as the limiter.

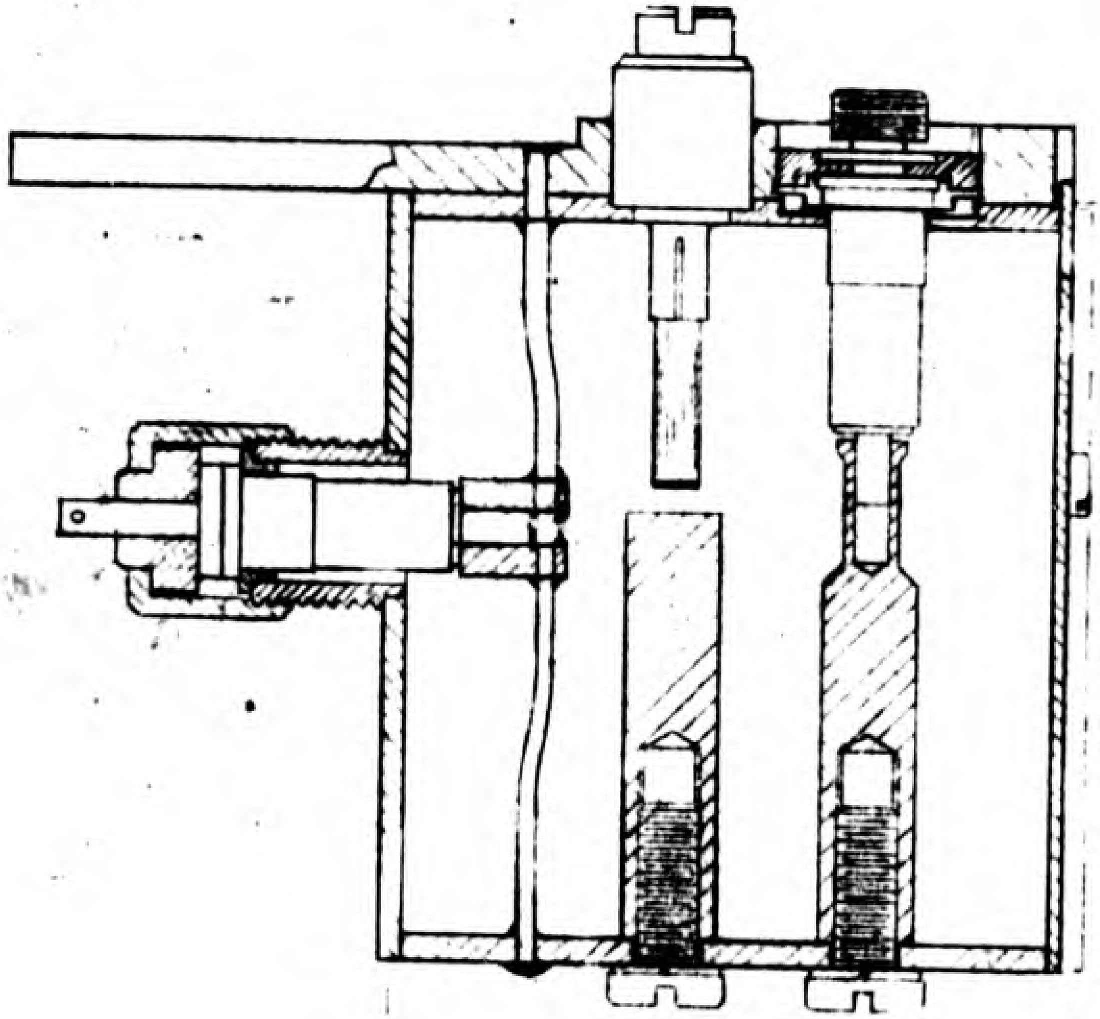


FIG. 15
HARMONIC GENERATOR, MIXER CAVITY

In order to conserve space and reduce stray capacity, circuit boards are mounted directly on the discriminator coil form. The diodes and associated components are mounted on the boards which also serve to form the spacing between primary and secondary winding of the coil. Although the discriminator was wide enough to encompass the entire band a tuning adjustment was provided so that the crossover frequency would correspond to the reference frequency. This addition had several advantages.

1. Equal lock range is obtained across the band.
2. In the absence of the transmitter signal the output of the discriminator will be near zero, thus limiting the amount of error signal and preventing a runaway loop.

A simple method was formulated to make this adjustment and is described in other sections of this report.

The output of the discriminator is approximately $\frac{0.1 \text{ volts}}{\text{Mcps}}$ with a bandwidth of 6 to 8 Mcps. This bandwidth defines the lock and capture range and makes possible the simple transmitter tuning procedure the equipment possesses. Environmental tests were conducted on this unit over a temperature range of -50 to + 105°C. The change in discriminator slope was less than two to one over these extremes.

3.4.5 Switch Oscillator, Amplifier and Phase Detector

In conjunction with the development of the RF Section of the AFC loop, work proceeded on the design and fabrication of the additional required stages. This included a multivibrator which acted as the switch oscillator, low frequency amplifier and phase detector. A schematic of this section is shown in Figure (16).

The output of the discriminator, which was balanced to ground, was fed to a series of push pull amplifiers. A conventional multivibrator served as the switch source. The output from each emitter provided a 5 volt negative pulse, which was clamped to ground, to the Gated Amplifier. In addition, the reference for the phase detector was taken from a transformer connector across the collector of the multivibrator. A 4 diode gate circuit served as the phase detector. The reference square wave alternately gated on two diodes of the bridge at a time which caused full wave rectification of the error signal. Depending upon the phase relationships between the error and reference signal a positive or negative voltage proportional to the error frequency resulted. A low pass filter was inserted at the output to strip the unwanted modulation off the DC which was to be fed through an emitter follower to the AFC Varactor. Temperature tests were performed on this unit and satisfactory operation was noted over the extremes. It was at this point in the development that tests indicated extreme problems in using varactors for frequency modulation. The data acquired for achieving a $\pm$ 500 KC deviation on the Power Oscillator indicated that a lock range of $\pm$ 2 MC for the AFC loop was beyond the "State of the Art" with the use of varactors. As such, the existing low frequency section of the loop was scraped in favor of a servo system. A set of cams, mounted on a common shaft, varies the depth of the plate and

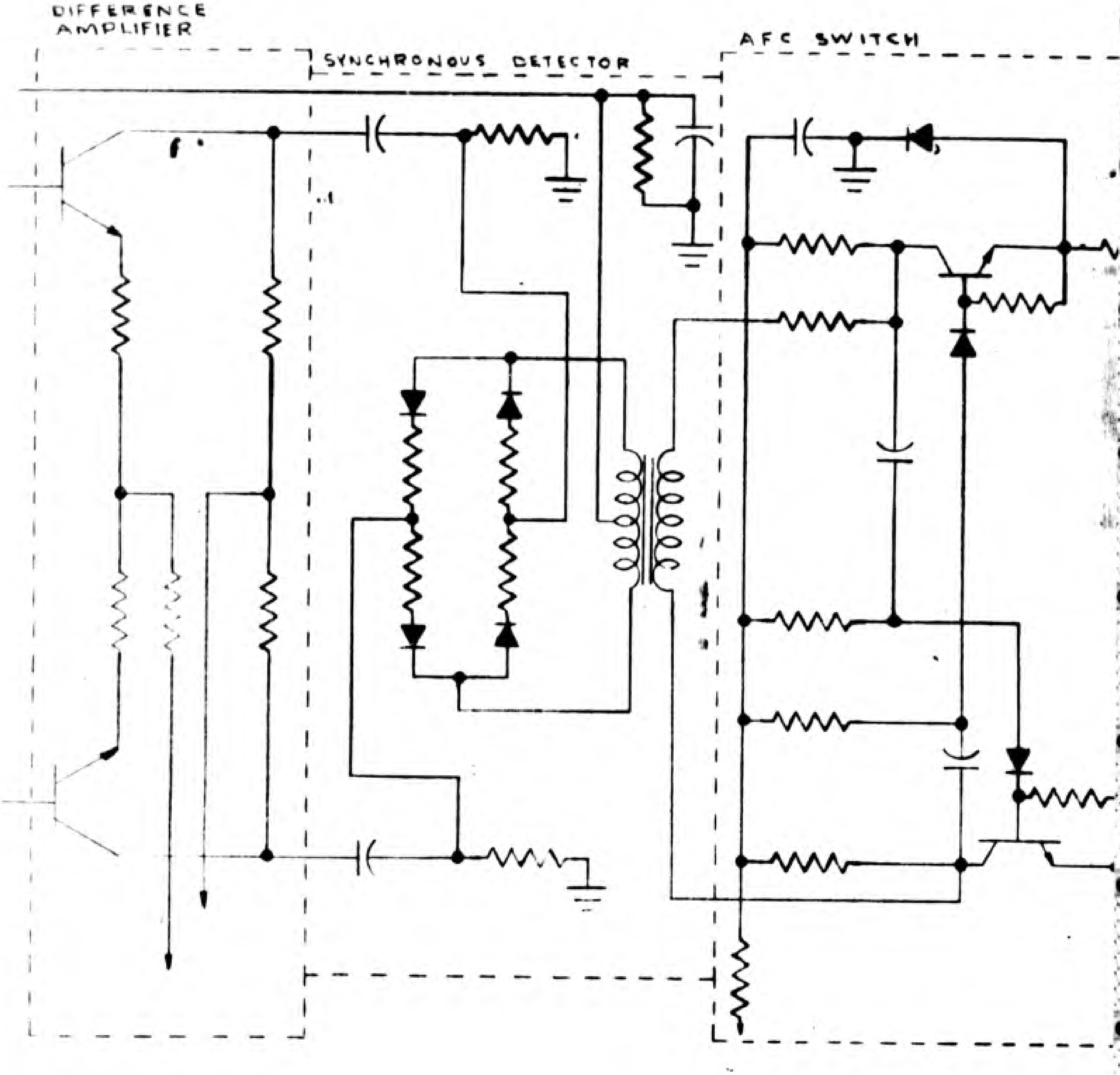
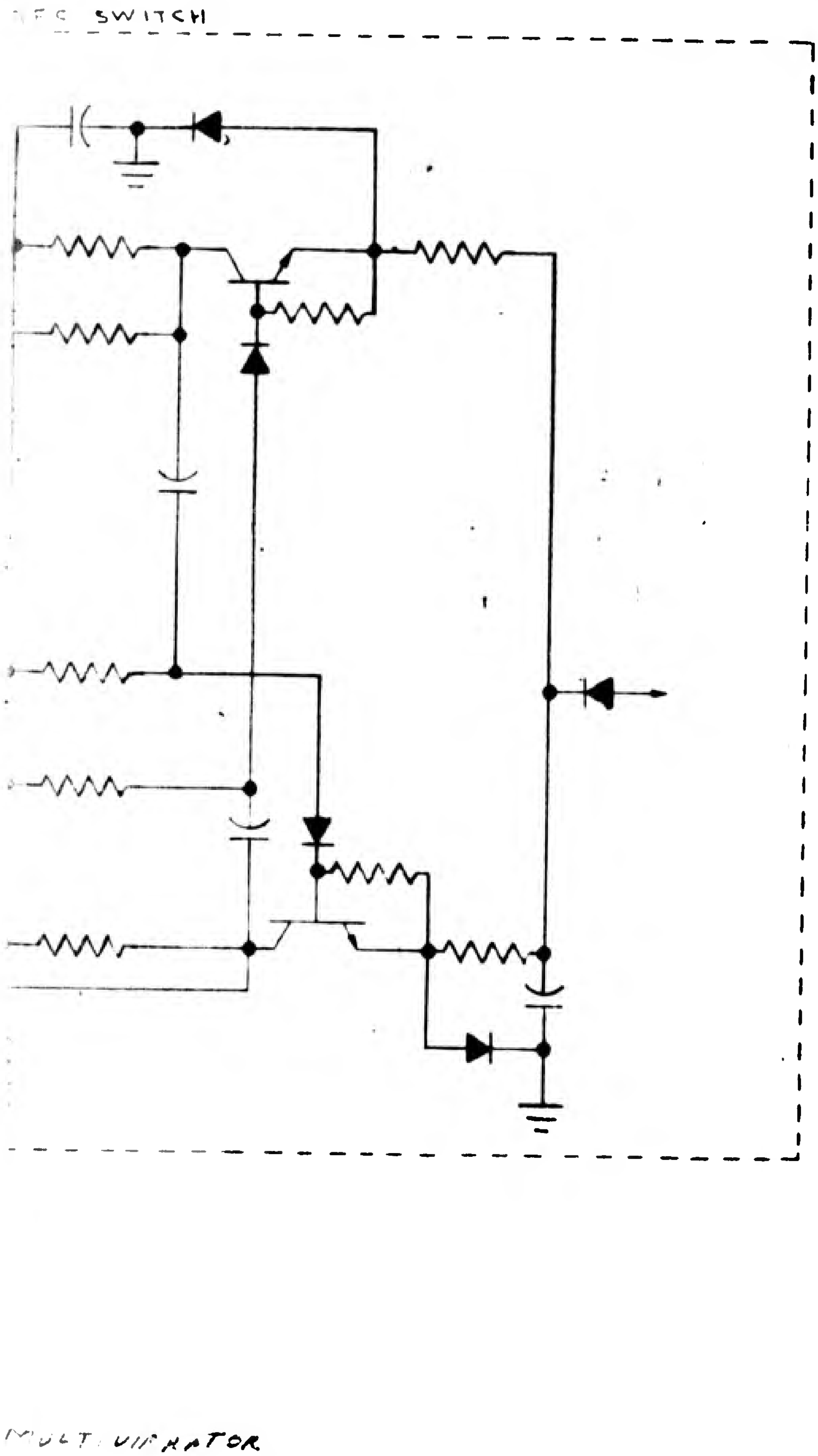


FIG 16

PHASE DETECTOR AND MULTIVIBRATOR



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cathode tuning plungers and thus the frequency of the Power Oscillator. It was originally intended that this would serve as the course frequency adjustment for the transmitter. The linkage to the cams was not modified so that it could be driven by a small servo motor through a differential gear arrangement. A more complete description of this configuration is given in later sections of this report.

A size 8 servo meter was chosen which delivered a torque of 0.22 oz in. It required 400 cps excitation with 3.1 watts for the reference phase and 2.7 watts for the control phase. The direction if rotation is governed by the phase relationship between reference and control voltages. This eliminated the need for a phase detector. The original multivibrator circuit could not supply the necessary reference power for the motor. In its place a DC to AC converter operating at a 360 cps rate was designed. The unit is conventional in design and supplied the necessary switch voltages for the gated amplifier as well as the required reference voltage. A six stage amplifier, with Class B push pull output, drove the control phase of the motor. The response of the motor to frequencies higher than 400 cps is considerably reduced so that adequate attenuation of high frequency modulation component is accomplished with a single RC low pass filter. Small signal gain of the unit is in the order of 400 with the saturated output voltage in the order of 25 VRMS. A tuning capacitor is added across the signal winding to resonate with the control winding inductance so as to increase its efficiency.

A schematic of the circuit is shown in Figure (22). In order to prevent runaway of the motor during transmitter warm up the AFC loop is disabled. The same bias voltage used to keep the power oscillator off is used to back bias the switch voltage. A +40 volt potential is placed on the center tap of Transformer T304 (Pin 5) back biasing diodes CR311 and CR312. As a result of this no switch voltage is present on the gate amplifier until the expiration of the delay time. After delay this point goes to ground and the AFC loop is closed.

3.5 Modulator

The development of the modulator was broken down into two parts, the modulation amplifier and the varactor transducer. The ultimate gain of the amplifier would of course be determined by the Varactor - Oscillator modulation sensitivity. Preliminary calculation indicated a gain of five with a 14 volt output swing would meet requirements.

3.5.1 Modulation Amplifier

The main problem concerning the design of the amplifier was its high input impedance requirements. Since transistors are relatively low impedance devices, feedback techniques have to be employed to raise its level. However, in all cases the absolute limit on the input impedance was the effective capacity seen at the input due to the transistor, cables, connectors and stray components. It was anticipated that where a high input impedance was required the information source impedance would also be large. If the information was in the form of pulse, as in PCM-FM, severe distortion of the pulse could be expected. These conclusions were verified in tests on an emitter follower amplifier as well as a negative feedback amplifier. In view of obtaining the best results possible an investigation was initiated with an aim of minimizing the capacity effect.

A scheme which employed both positive and negative feedback was breadboarded. The basic circuit configuration is shown in Figure (6).

The total input admittance can be shown to be

$$Y_{in} = \frac{1}{B_1 B_2 R_1} + Y_L - KY_F$$

Where:

B_1 and B_2 are the small signal current gain of transistors one and two

R_1 is the emitter load on the first transistor

Y_L = admittance to the stray and transistor capacity and transistor collector loading

K = feedback coefficient

Y_F = admittance of the feedback network (R_F & C_F)

The first term of the expression is the input impedance of the amplifier due to negative feedback which is then shunted with the input loading (Y_L). The term KY_F is adjusted so that it is equal to Y_L.

Note that in the event the term KY_F is greater than $Y_L + \frac{1}{B_1 B_2 R_1}$ the admittance is negative and the system is unstable. To accommodate working with a high source impedance generator, which is connected to the modulator with a coaxial cable, the admittance of Y_F is varied. The circuit is normally adjusted for the generator and cable length to be used with the transmitter.

A schematic of the final form of the modulator is shown in Figure (22). An emitter follower was added to the basic amplifier and served as a low impedance source for the modulation varactor. The modulation gain control is in the form of a potentiometer which is the load for the emitter follower. C357 varies the feedback admittance and K, the feedback factor, is the ratio of R345 to R346. Total gain of the unit is 5 with a 5 VRMS saturation level. The input impedance measured at 400 Kcps is in the order of 530 K and the circuit can compensate for up to 60 μμfd stray cable capacity. Distortion at full output (5 VRMS) is 0.6% while the 3 db bandwidth is in excess of 400 Kcps. Temperature tests on the unit were run over a range of -50°C to 105°C with little changes in gain or input impedance. This being due to the excellent bias stability and negative feedback circuitry.

3.5.2 Modulation Varactor

After initial investigation, the development of the varactor circuitry was delayed due to the problem encountered in the power oscillator circuit. When it was indicated that the power oscillator would operate at a 15 watt level, even more strain was placed on the varactor. At this time information on varactors in general was a minimum and in particular, information on high frequency high power application did not exist. The transmission line amplification concept, (derived in section 2.2) was used as the basic design approach.

The initial varactor mount was constructed in a coaxial section of transmission line. A commercial line stretcher was inserted between the mount and the cavity. Provisions were made for varying the back bias on the varactor and adjusting the coupling into the cavity. Initially the varactor circuit was coupled into the plate line with an inductive loop. Moderate success was obtained in that a $\pm$ 200 KC peak carrier deviation was noted. However, the oscillator was detuned due to the quiescent capacity inserted by the varactor circuit which resulted in unstable operation. In addition, tuning of the entire circuit was extremely critical. When the coupling was increased in an effort to acquire a larger carrier deviation, this effect became more pronounced. In addition, several varactors were destroyed even though great care was taken to limit forward and reverse current. The initial varactors employed were of the MA4287 type, which were 30 V, 1 watt units with a 10 KMC cutoff frequency. These were the best varactors available at the time, but analysis showed that they were not suitable for our purpose. In order to ascertain the necessary varactor parameter, an investigation was performed to determine the power dissipation capability the unit must possess (see appendix). The varactor dissipation is given by:

$$Pd = PoQ_L \quad \frac{2\delta f}{fo} \cdot \frac{1}{Q^2} \cdot \frac{C_2}{C_2 - 1}$$

where:

Po = Output power of oscillator

Q = Loaded Q of the oscillator cavity

$\frac{2\delta f}{fo}$ = ratio of total carrier deviation to carrier frequency

$C_2 - C_1$ = Change in capacity of varactor

C_2 = maximum value of capacity

C_1 = minimum value of capacity

$Q_2 = \frac{1}{WC_2R_s}$

R_s = equivalent series resistance

It is seen that the term $P_oQ_L \frac{2\delta f}{fo}$ is dependent upon circuit requirements while the term $\frac{1}{Q_2} \cdot \frac{C_2}{C_2 - C_1}$ is a function of the varactor parameter. Note that for low power dissipation the varactor should have a high Q and a large ratio of minimum to maximum capacity. Various manufacturers were consulted in relationship to the desired parameter. Information obtained at these consultations indicated that our requirements were beyond the then "State of the Art". However, two manufacturers (Microwave Associates and Microstate Electronics) indicated developmental progress in their varactors that would make them fit our needs. Agreement was made to purchase these units as soon as they were available. In the interim, a new modulation system was investigated.

The new system utilized a high frequency diode placed in series with a fixed reactance. The circuit was then coupled to the oscillator through a small section of transmission line. The modulation voltage was used to vary the conduction period of the diode and thus vary the reactance to the tank. A low frequency breadboard was designed to prove the concept with excellent results obtained at 20 mcps. However, when an equivalent was tried at "S" Band, the oscillator became extremely unstable. In addition, although large deviation was noted, the detected output showed appreciable distortion. It was recognized that the problems encountered in this configuration would require a great deal of time to solve and work was therefore stopped.

Late in 1961, varactors were procured from Microstate and Microwave Associates. The Microstate unit was a Gallium Arsenide varactor with 1 watt dissipation and a maximum series resistance of 1.0 ohms. The use of Gallium Arsenide was comparatively new in varactor diodes but did offer excellent temperature characteristics. A silicon varactor diode was obtained from Microwave Associates. This unit was rated for 120 volts and 3.5 watts dissipation. The maximum series resistance was 1.7 ohms with a maximum to minimum capacity ratio of 10. In addition to the new varactors, the coupling arrangement was modified to work into the cathode cavity instead of the plate cavity. Considerable success was obtained with both varactors. Peak frequency deviation of $\pm$ 400 KC was noted and the modulation linearity was good.

Although power pulling was reduced considerably, the cathode cavity did require re-adjustment. In addition, some trouble was experienced at the low end of the band (2150 MC) in achieving maximum deviation while holding the output power to 15 watts. It was concluded that the feedback circuit would have to be modified to compensate for the power lost to the varactor by the cathode cavity. At this point in the development, a set of cams had been fabricated for the plate and cathode cavities. Data indicated that these would have to be modified to compensate for the varactor capacity. It was recognized that the varactor multiplication scheme would be frequency sensitive. Experiments were conducted with view of using stubs to broadband the device. However, data showed that a device which possessed a negative slope of frequency vs reactance would be required. It was impossible to do this, so that a tuning adjustment had to be devised. Since it was completely impractical to provide a line stretcher to compensate for frequency change a stub was added to perform the function. In effect, the stub was a variable capacity which loaded the end of the transmission line thus varying its line length. For values of impedance with small resistance components, the effect of changing the capacity is almost the same as changing the line length. Modifications were performed on the cavity feedback loop and cams to accomodate the varactor. Additional work was performed on the stub and varactor mount so that it would track over tuning band. A schematic drawing of the varactor modulator is shown in Figure (17).

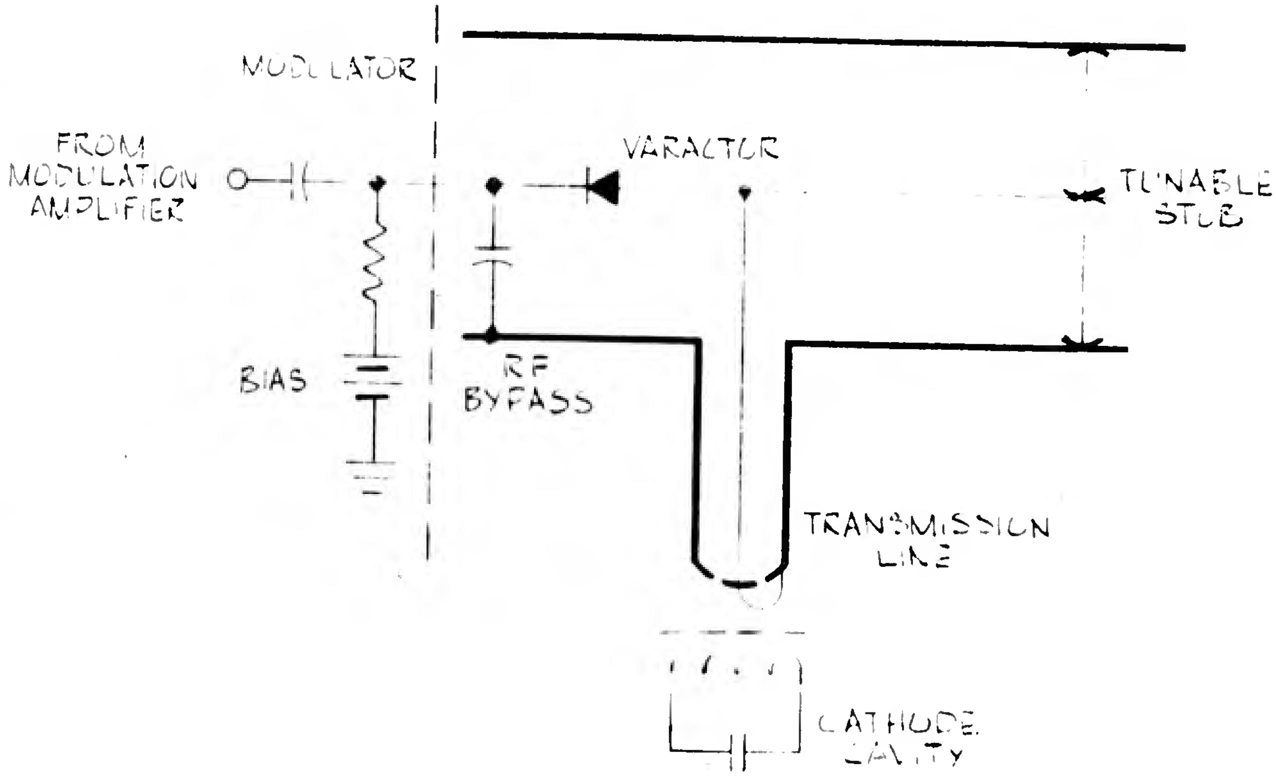


FIGURE 17 VARACTOR MODULATOR DIAGRAM

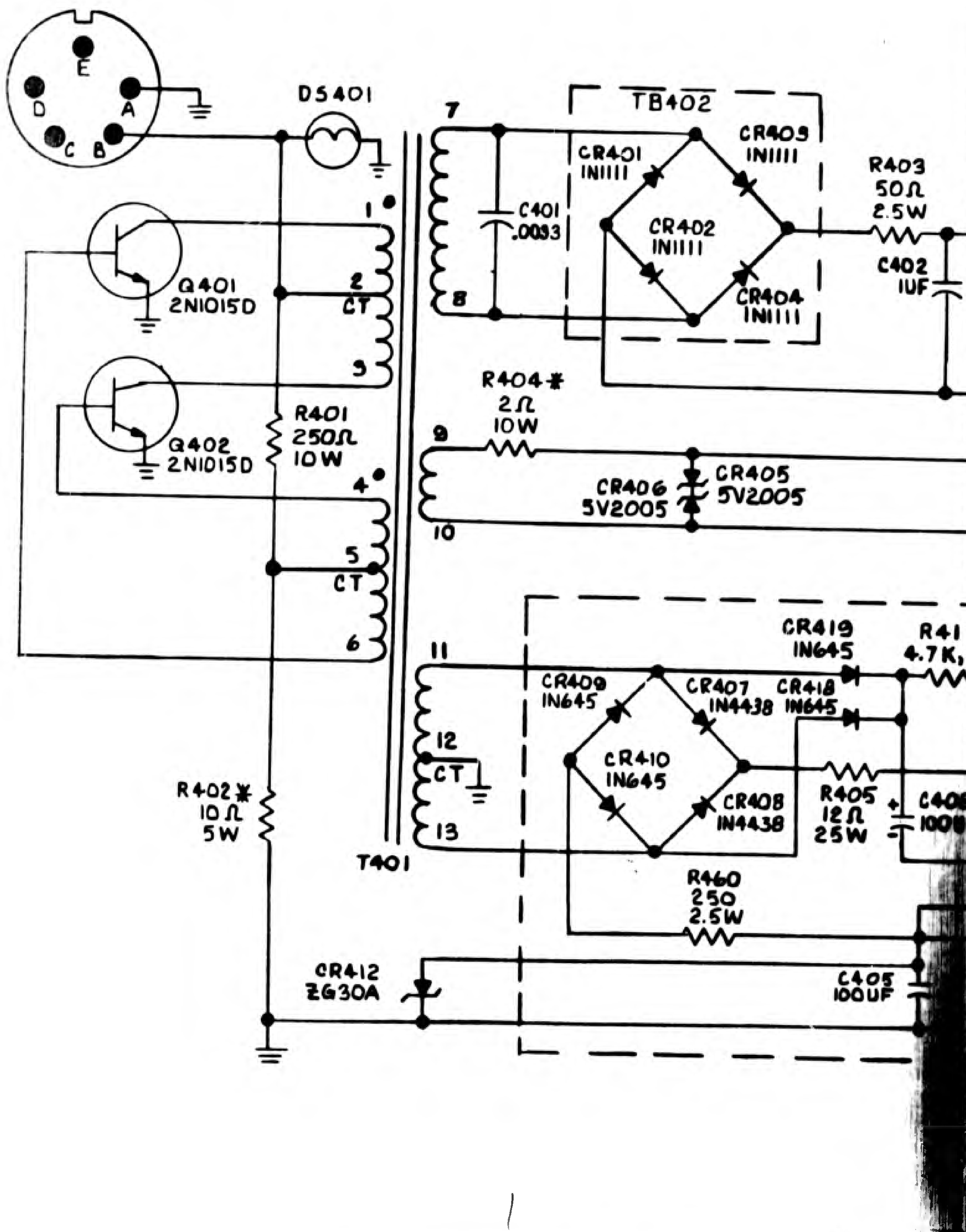
Excellent results were obtained with the system. At 2200 MC a ± 500 KC deviation was obtained with a minimum output power of 15 watts. At 2300 MC over a ± 1 MC deviation was recorded with 15 watts output power. This change in modulation sensitivity was due to the effective tuning capacity across the plate cavity. At 2300 MC very little external capacity is used to resonant the plate line and thus the effective change caused by this varactor. has greater effect. Detail tests on modulation linearity and distortion are shown in later section of the report.

3.6 Power Supplies

The equipment was designed to operate with either of two power inputs: 115 V at 400 cps, or 26 Volts DC. Both units were to be identical in size and capable of operation with or without the control box. In addition to supplying the necessary power for the transmitter, provision had to be made for a delay circuit to keep the power oscillator tube off while the filament warmed up. Of prime importance in the design was a low percentage of ripple which would minimize residual FM Carrier output.

3.6.1 DC Supply

A schematic of the DC supply is shown in Figure (18). It consisted of a conventional DC to DC convertor with rectification and filtering accomplished in the secondary circuitry. In order to reduce filtering requirements, the converter is designed to operate at a frequency of 2500 cps which yields a secondary


E
D
A
C
B
DS401
Q401
2N1015D
Q402
2N1015D
R401
250Ω
10W
R402*
10Ω
5W
T401
1
2
CT
3
4
5
CT
6
7
8
9
10
11
12
CT
13
TB402
C401
.0033
CR401
IN1111
CR402
IN1111
CR403
IN1111
CR404
IN1111
R403
50Ω
2.5W
C402
1UF
R404*
2Ω
10W
CR405
5V2005
CR406
5V2005
CR419
IN645
R41
4.7K,
CR409
IN645
CR407
IN4438
CR418
IN645
CR410
IN645
CR408
IN4438
R405
12Ω
25W
R460
250
2.5W
CR412
ZG30A
C405
100UF

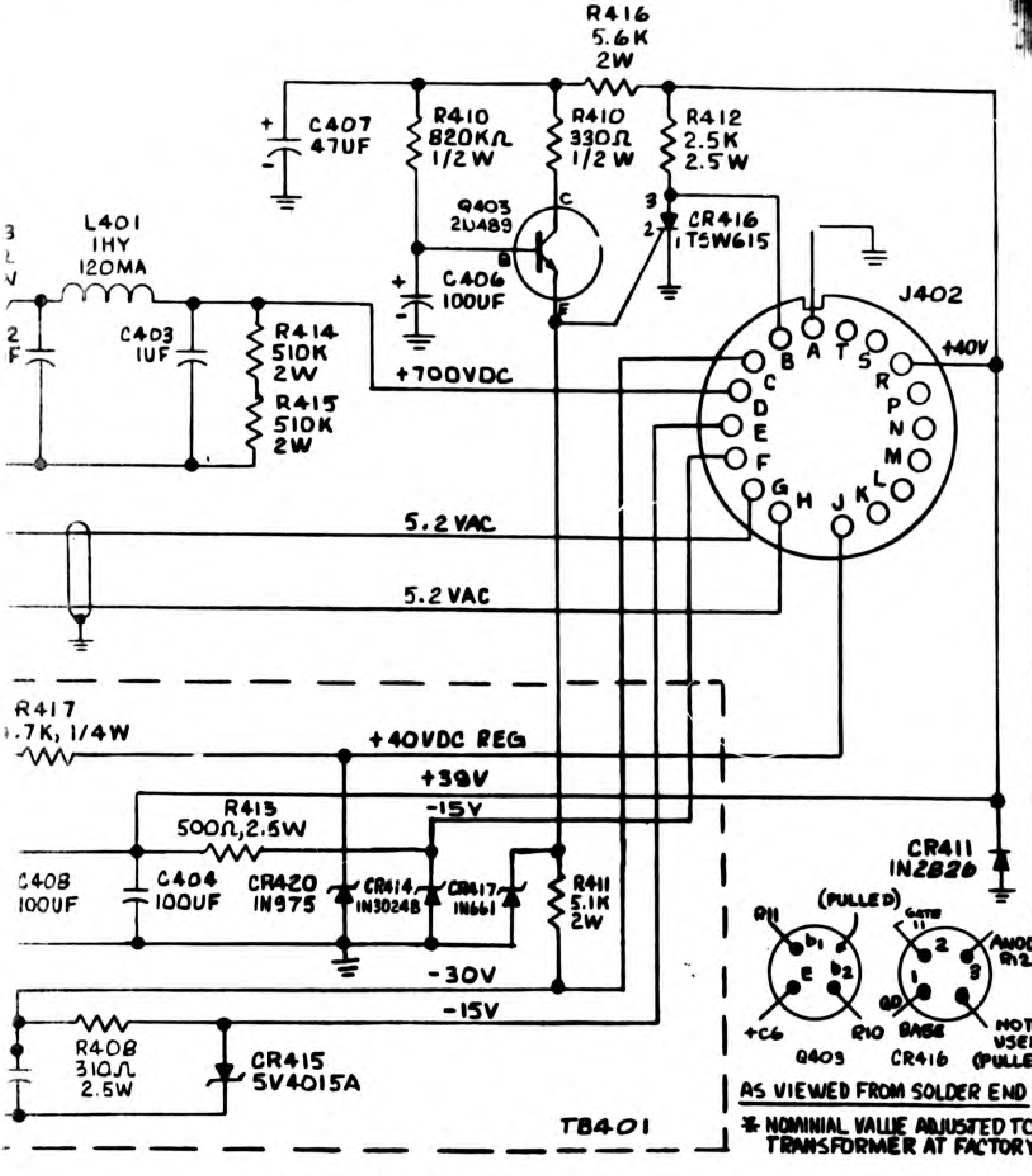
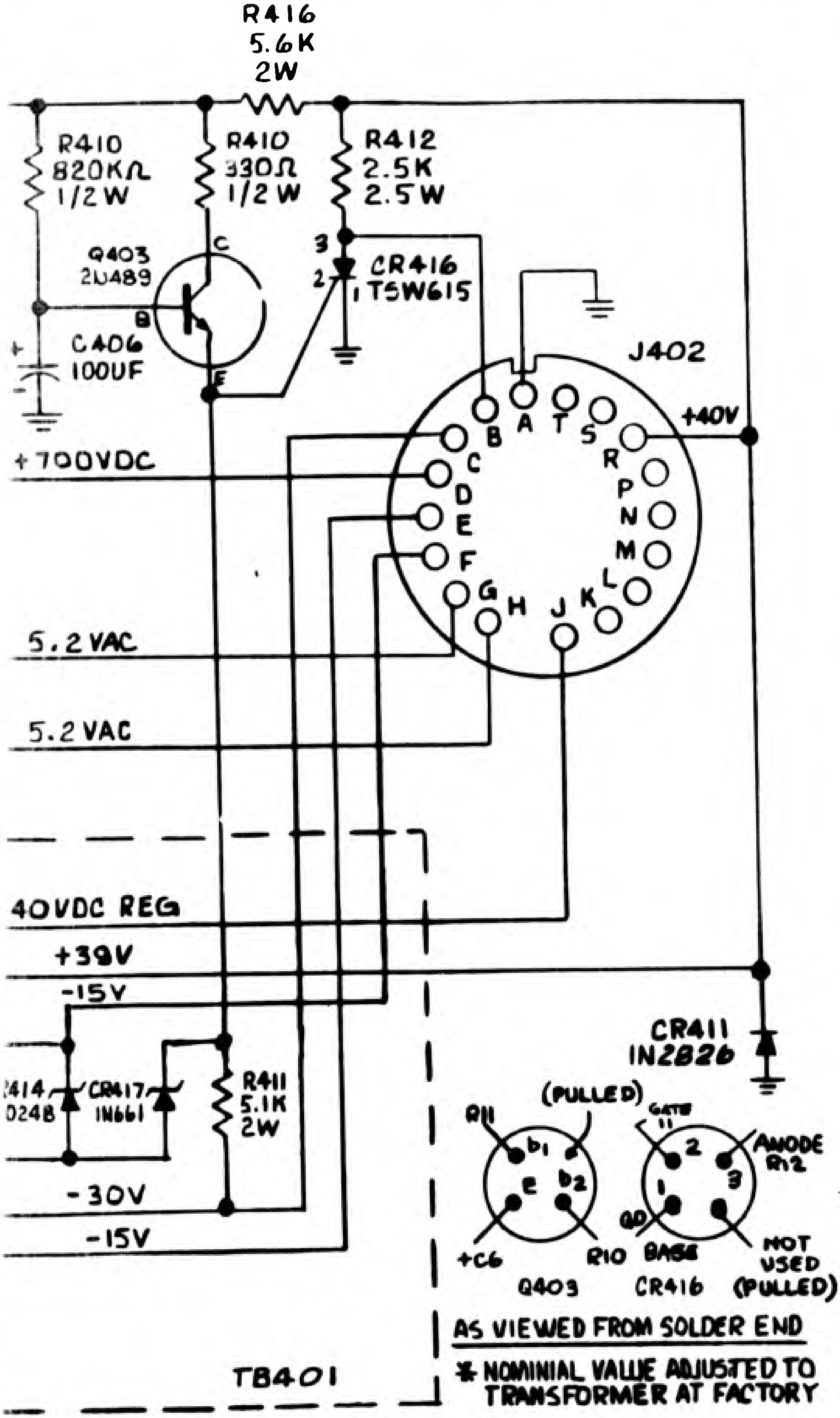


FIGURE 18. 28V POWER SUPPLY, SCHEMATIC DIAGRAM.

2



3. 28V POWER SUPPLY, SCHEMATIC DIAGRAM.

ripple frequency of 5000 cps. Silicon Zener Diodes provide regulation for the low level voltages. The plate supply for the power oscillator is derived from a full wave bridge and pie filtering network. While this circuit is unregulated, the minimum output power will be delivered by the transmitter at low line levels.

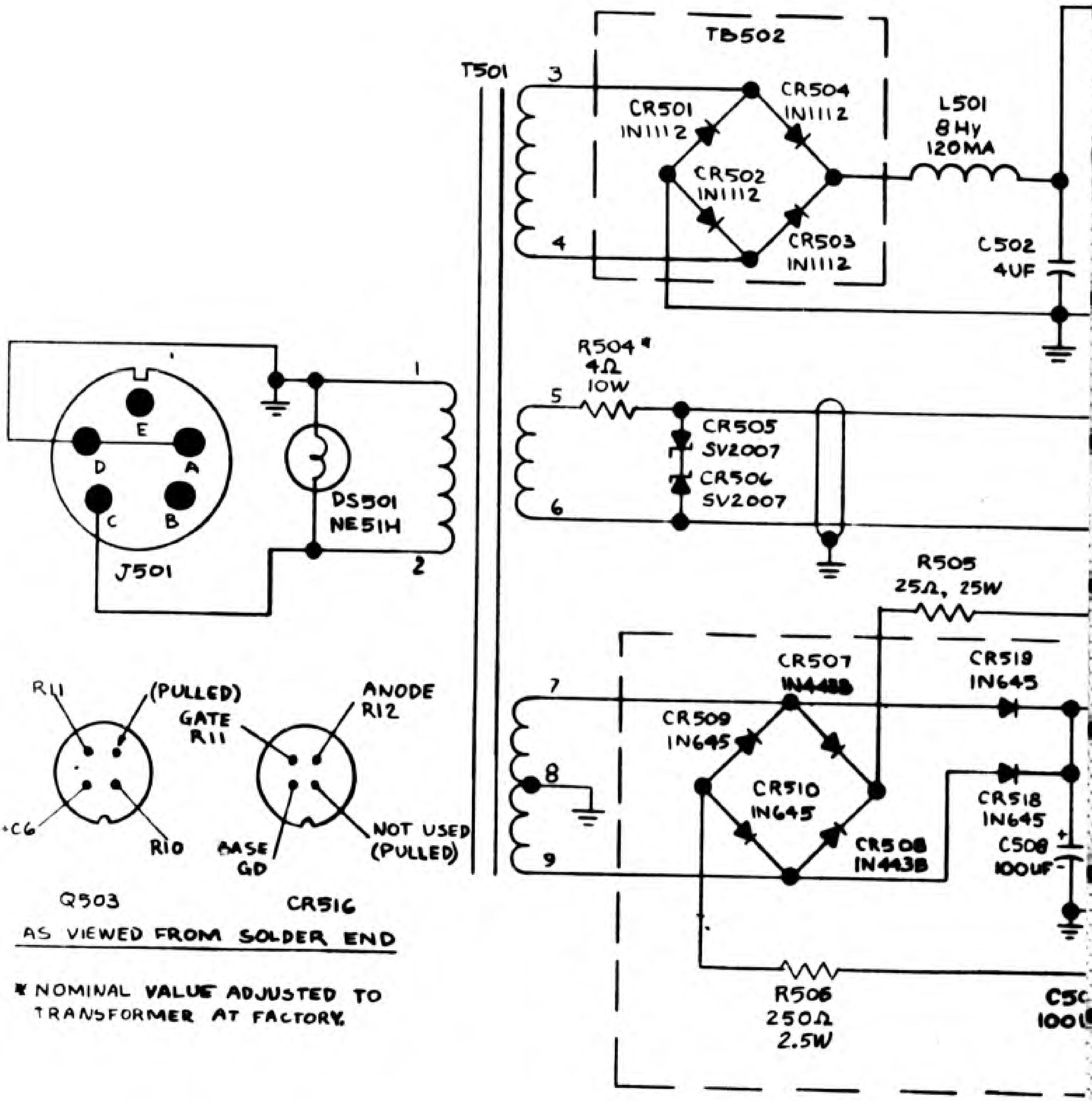
Data supplied by the manufacturer indicated that a 1 minute warm up period would be necessary before the tube would be allowed to draw current. Conventional time delay relays would not withstand missile environment and in addition would require additional power. The time delay system developed is an electronic unit which biases the tube off for a period of 1 to 1 1/2 minutes. It consists of a unijunction transistor operating as a relaxation oscillator. The output is used to control the gate of a silicon controlled rectifier. Once the SCR fires, the gate loses control of the anode current and can only be turned off by turning off the unit. When the SCR is open, +40 volts are applied to the cathode of the transmitter tube biasing it off. This same voltage back biases the switch diodes, thereby disabling the servo loop. When the SCR fires the anode goes to near ground potential, which allows the tube to draw current and activates the servo loop.

Environmental data was performed on power supply and results indicated satisfactory operation up to 105°C.

3.6.2 AC Supply

A schematic of the AC power supply is shown in Figure (19). The circuit is conventional and very similar to the DC supply.

REPRODUCED AT GOVERNMENT EXPENSE



* NOMINAL VALUE ADJUSTED TO TRANSFORMER AT FACTORY.

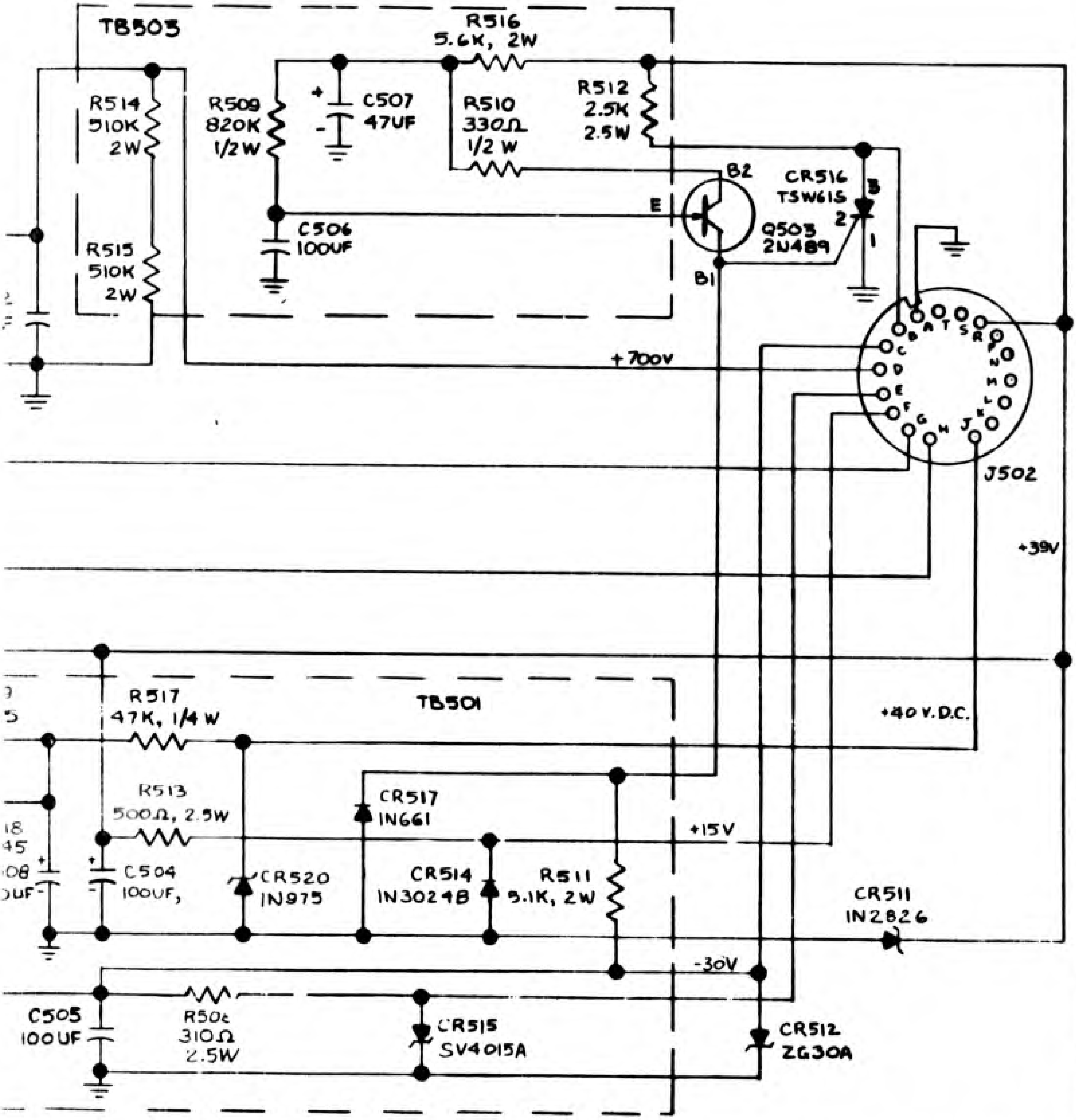
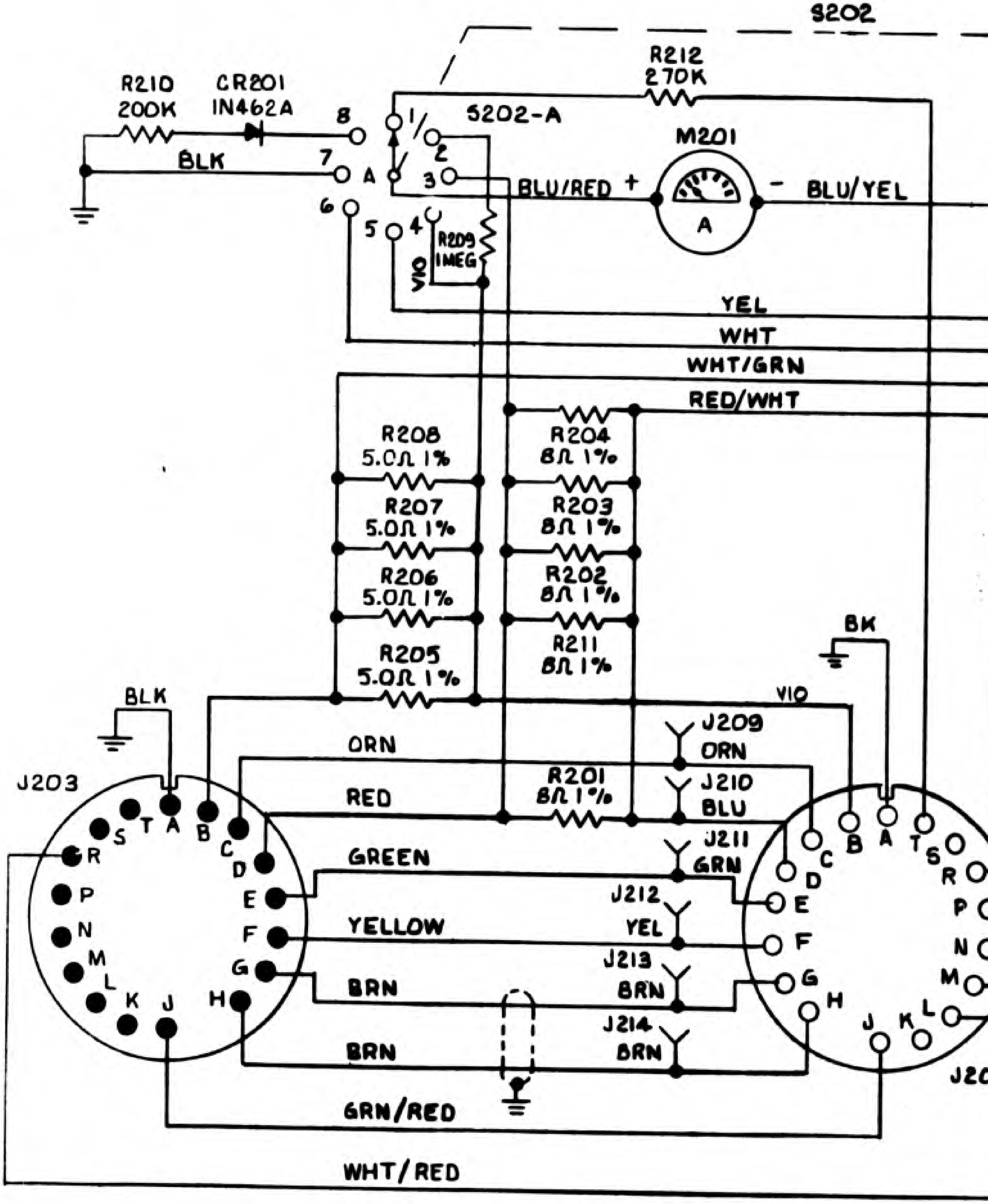
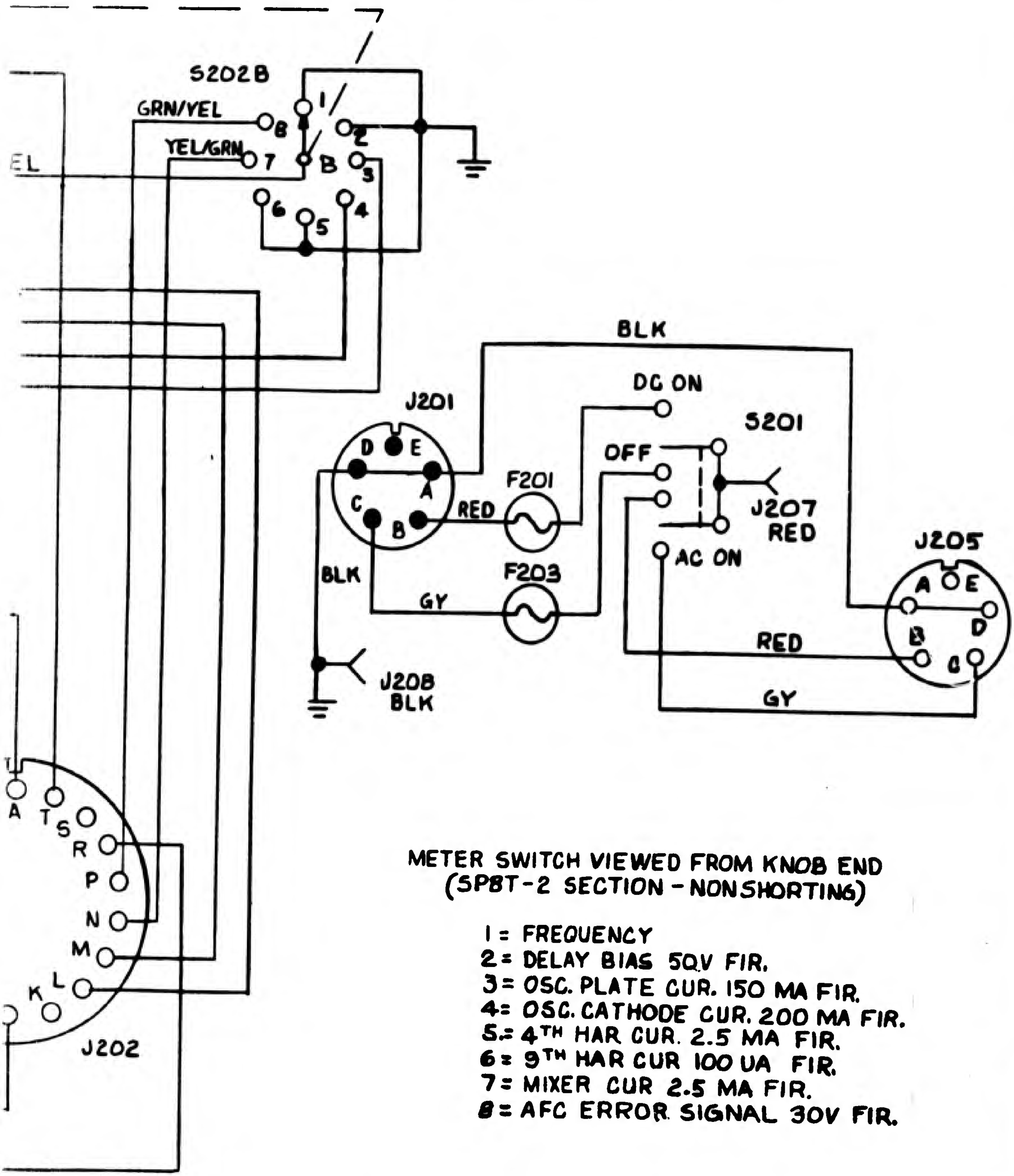


FIGURE 19. 400 CPS POWER SUPPLY, SCHEMATIC DIAGRAM.


S202
R212
270K
R210
200K
CR201
IN462A
S202-A
M201
BLK
BLU/RED
BLU/YEL
A
R209
IMEG
VIO
YEL
WHT
WHT/GRN
RED/WHT
R208
5.0Ω 1%
R207
5.0Ω 1%
R206
5.0Ω 1%
R205
5.0Ω 1%
R204
8Ω 1%
R203
8Ω 1%
R202
8Ω 1%
R211
8Ω 1%
BK
BLK
VIO
J209
ORN
J210
BLU
J211
GRN
J212
YEL
J213
BRN
J214
BRN
J203
ORN
RED
R201
8Ω 1%
GREEN
YELLOW
BRN
BRN
GRN/RED
WHT/RED



METER SWITCH VIEWED FROM KNOB END
(SPBT-2 SECTION - NONSHORTING)

1 = FREQUENCY
2 = DELAY BIAS 50 V FIR.
3 = OSC. PLATE CUR. 150 MA FIR.
4 = OSC. CATHODE CUR. 200 MA FIR.
5 = 4TH HAR CUR. 2.5 MA FIR.
6 = 9TH HAR CUR 100 UA FIR.
7 = MIXER CUR 2.5 MA FIR.
8 = AFC ERROR SIGNAL 30V FIR.

FIGURE 20. CONTROL UNIT, SCHEMATIC DIAGRAM

2

for transmitter frequency. Position 1 of the meter switch monitors this voltage and is calibrated for each equipment in terms of frequency. In this position of the switch, each division is equal to approximately 6 MC so that the transmitter can be brought into the AFC lock range without the use of auxiliary equipment. Detailed tuning procedures are given in the equipment maintenance manual.

3.8 Tuning and Maintenance Provisions

One of the unique features of the equipment is the relative ease of tuning and internal checking provisions provided. An overall Block Diagram of the transmitter is shown in Figure (21) and overall schematic in Figure (22). The heart of the tuning and maintenance procedures is the AFC function switch and the metering circuitry located in the control box. With reference to Figure (21), the AFC function switch controls the gate voltages to the switch amplifiers as well as the error signals to the control winding of the servo motor. In position 1 of the switch, the AFC loop is closed and both reference and transmitter channels receive gate voltage. In position 2, the gate voltage is removed from the transmitter channel and the signal winding of the servo motor is opened. However, the error signal voltage diode circuitry still records the output of the servo amplifier. Thus by offsetting the discriminator crossover point an error signal is fed through the Servo amplifier checking its operation. Now if the mixer is detuned and the discriminator is offset, when the AFC function switch is brought back to position 1, a false error signal is created and the servo motor will turn. In changing

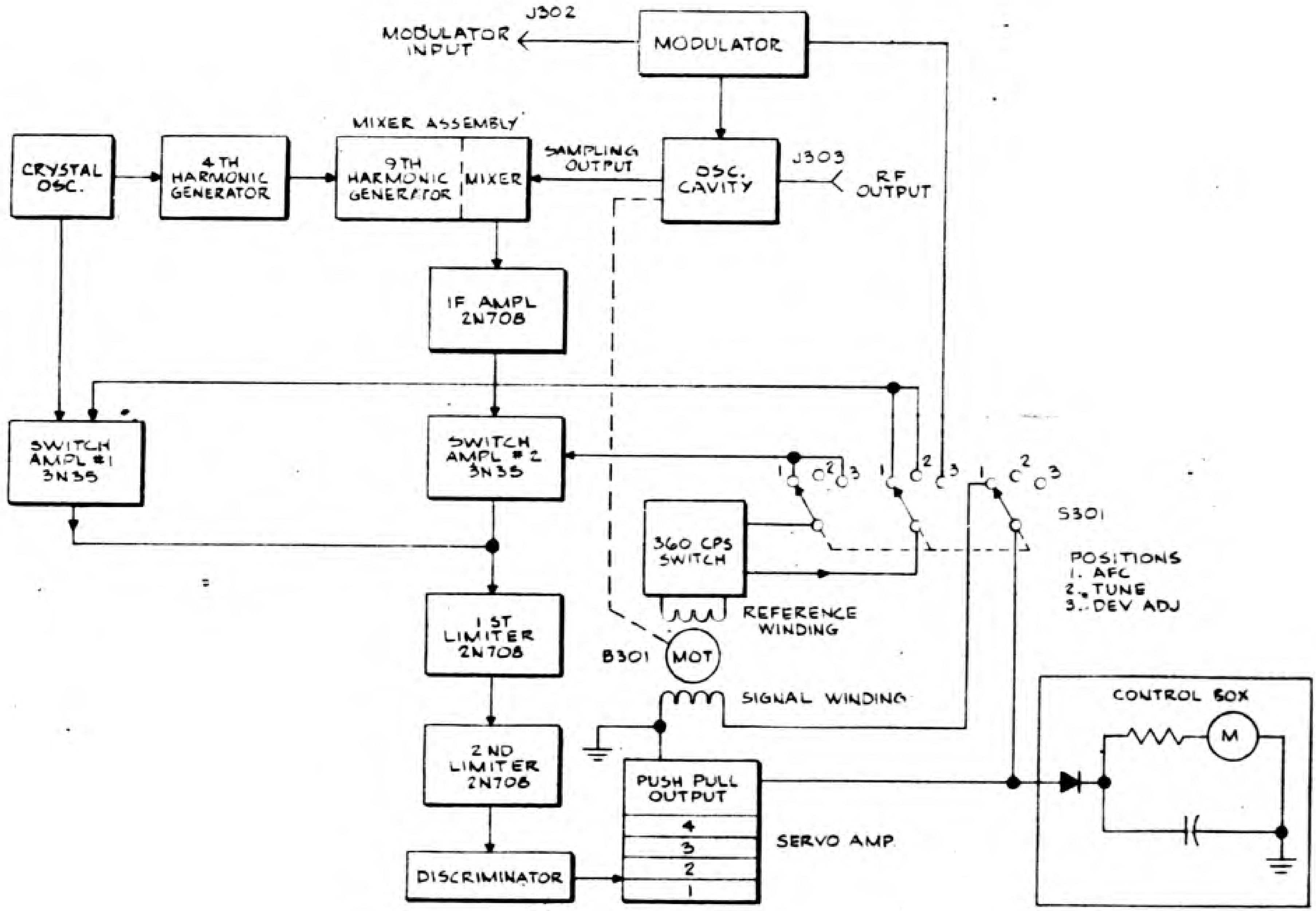
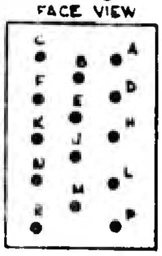


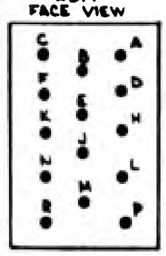
FIGURE 21 AFC CIRCUIT, BLOCK DIAGRAM

J315
FACE VIEW



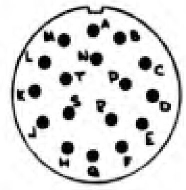
A - 40VDC CONTROLED
B - -15V PIN E, J101
C - GND
D - PIN K, J314
E - PIN L, J314
F - 4th HAR CUR 2.5 MA FIR
H - 9th HAR CUR 100 UA FIR
J - MIXER CUR 2.5 MA FIR
K - +15VDC
L - -30VDC (MOD SUP)
M - AFC - IN
N - AFC - OUT
P - 40V CONTROL
R - +15V OUT

J314
FACE VIEW



A -
B - GND
C - +40V CONTROL
D - AFC
E - REF
F - REF
H - +40V DELAY BIAS
J - AFC ERROR SIGNAL
K - PIN D, J315
L - PIN E, J315
M -
N - 40VDC
P -
R -

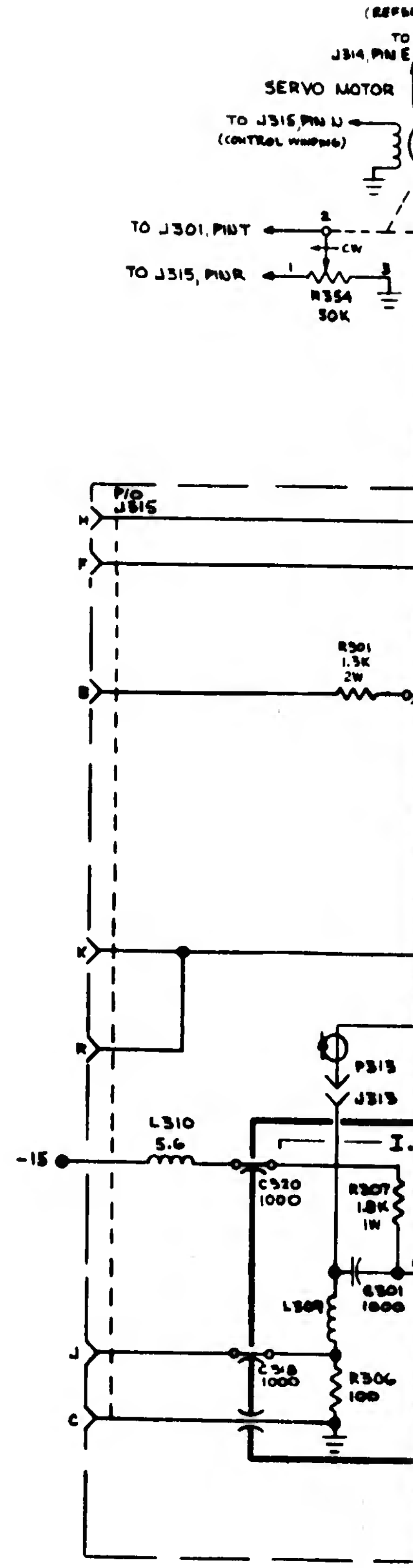
J301
FACE VIEW

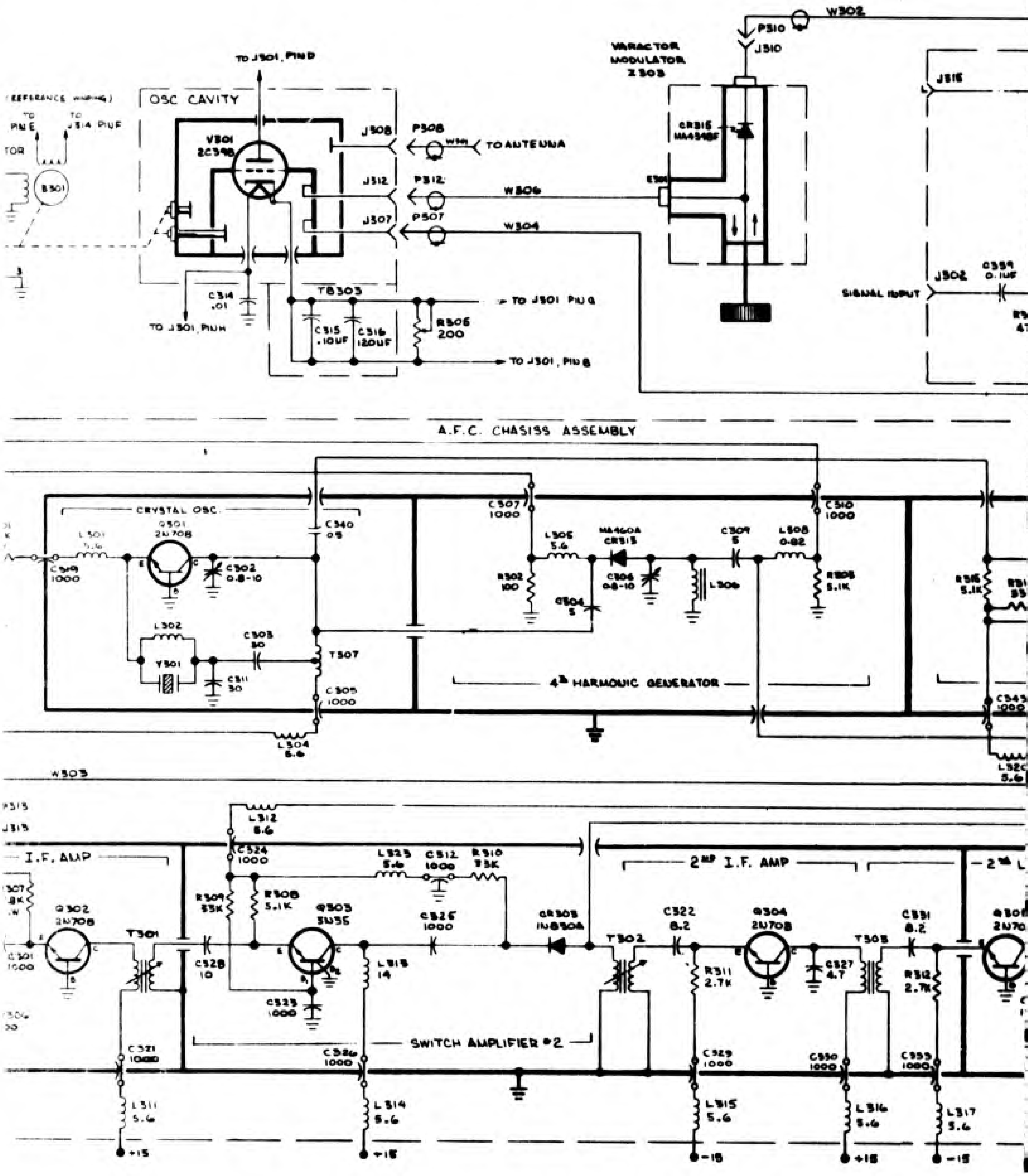


A - GROUND
B - +40V DELAY BIAS
C - -30VDC (MOD SUP)
D - +700VDC (OSC B+)
E - -15VDC
F - +15VDC
G - 5VAC
H - 5VAC
J - +40VDC CONTROLLED
K -
L - 4th HAR CUR 2.5 MA FIR
M - 9th HAR CUR 100 UA FIR
N - MIXER CUR 2.5 MA FIR
P - AFC ERROR SIGNAL 30V FIR
R - +40VDC
S -
T - FREQUENCY

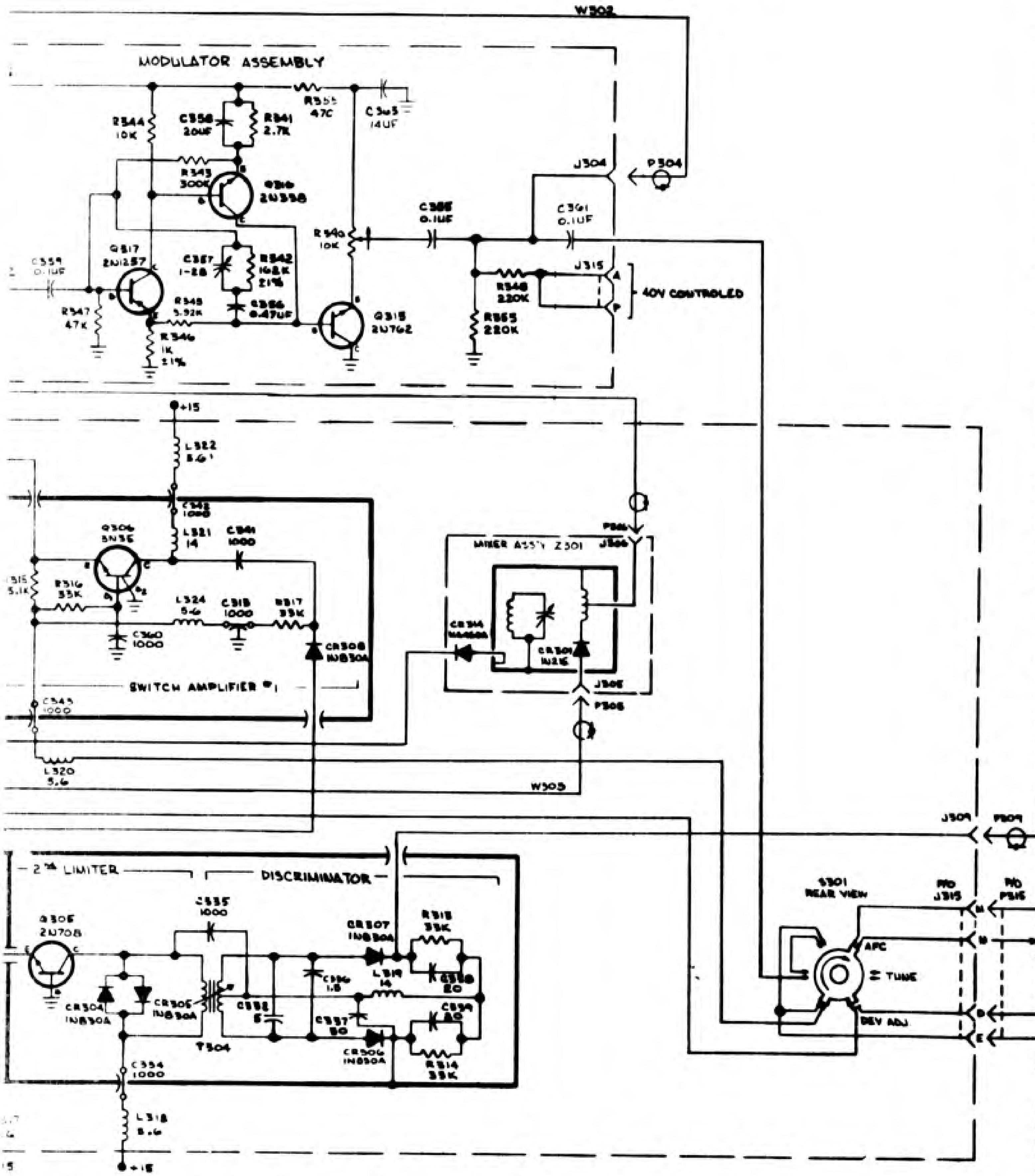
NOTES:

1. UNLESS OTHERWISE SPECIFIED:
   ALL RESISTORS ARE OHMS 1/4 WATT, 5% TOLERANCE.
   ALL CAPACITOR ARE IN UUF.
   ALL COILS & RF CHOKES ARE IN UH.
2. CONNECTOR J314, CONTINENTAL MM14-22 PGSD, MATES WITH P314, CONTINENTAL MM14-22 SDGSK.
3. CONNECTOR J315, CONTINENTAL MM11-14 PDGS, MATES WITH P315, CONTINENTAL MM11-14 SGDSK.

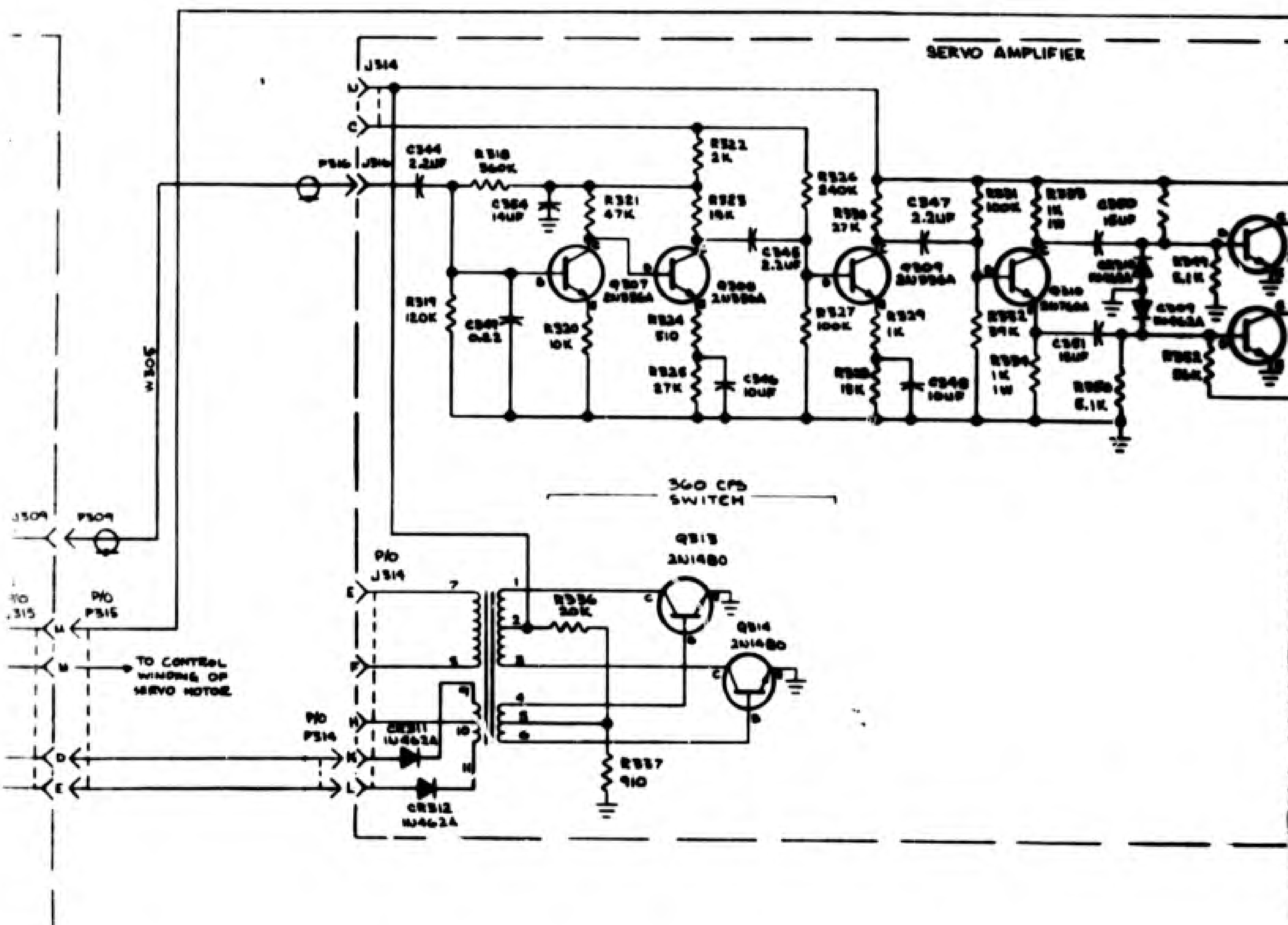



OSC CAVITY
VARACTOR MODULATOR Z303
A.F.C. CHASISS ASSEMBLY
CRYSTAL OSC.
4th HARMONIC GENERATOR
I.F. AMP
2ND I.F. AMP
SWITCH AMPLIFIER #2
SIGNAL INPUT
TO ANTENNA

2


W302
MODULATOR ASSEMBLY
R344 10K
C358 20UF
R341 2.7K
R355 47C
C363 14UF
R343 300K
Q316 2N338
J304
P304
R340 10K
C386 0.1UF
C361 0.1UF
Q317 2N1257
C359 0.1UF
C357 1-28
R342 162K 1%
R345 5.92K
C356 0.47UF
R347 47K
R346 1K 1%
Q315 2N762
J315
R348 220K
R365 220K
40V CONTROLLED
+15
L322 5.6
C342 1000
Q306 3N35
L321 14
C341 1000
R316 33K
L324 5.6
C318 1000
R317 33K
C360 1000
CR308 1N830A
MIXER ASS'Y Z301
CR314
CR301 1N21E
J306
P306
J305
P305
SWITCH AMPLIFIER #1
C343 1000
L320 5.6
W303
J309
P309
2ND LIMITER
DISCRIMINATOR
Q305 2N708
C335 1000
CR307 1N830A
R313 33K
L319 14
C336 1.8
C338 20
CR304 1N830A
CR305 1N830A
C332 5
C337 30
C339 20
CR306 1N830A
R314 33K
T304
C334 1000
L318 5.6
+15
S301 REAR VIEW
APC
TUNE
DEV ADJ

3


SERVO AMPLIFIER
360 CPS SWITCH
TO CONTROL WINDING OF SERVO MOTOR
J314
P316
J316
W306
J309
P309
P/O J315
P/O P315
P/O J314
P/O P314
C344 2.2UF
R318 560K
R321 47K
R322 2K
R323 15K
R324 510
R326 27K
R319 120K
R320 10K
R327 100K
R329 1K
C347 2.2UF
C346 2.2UF
Q313 2N1480
Q314 2N1480
R336 20K
R337 910
CR311 1N462A
CR312 1N462A

FIGU

4

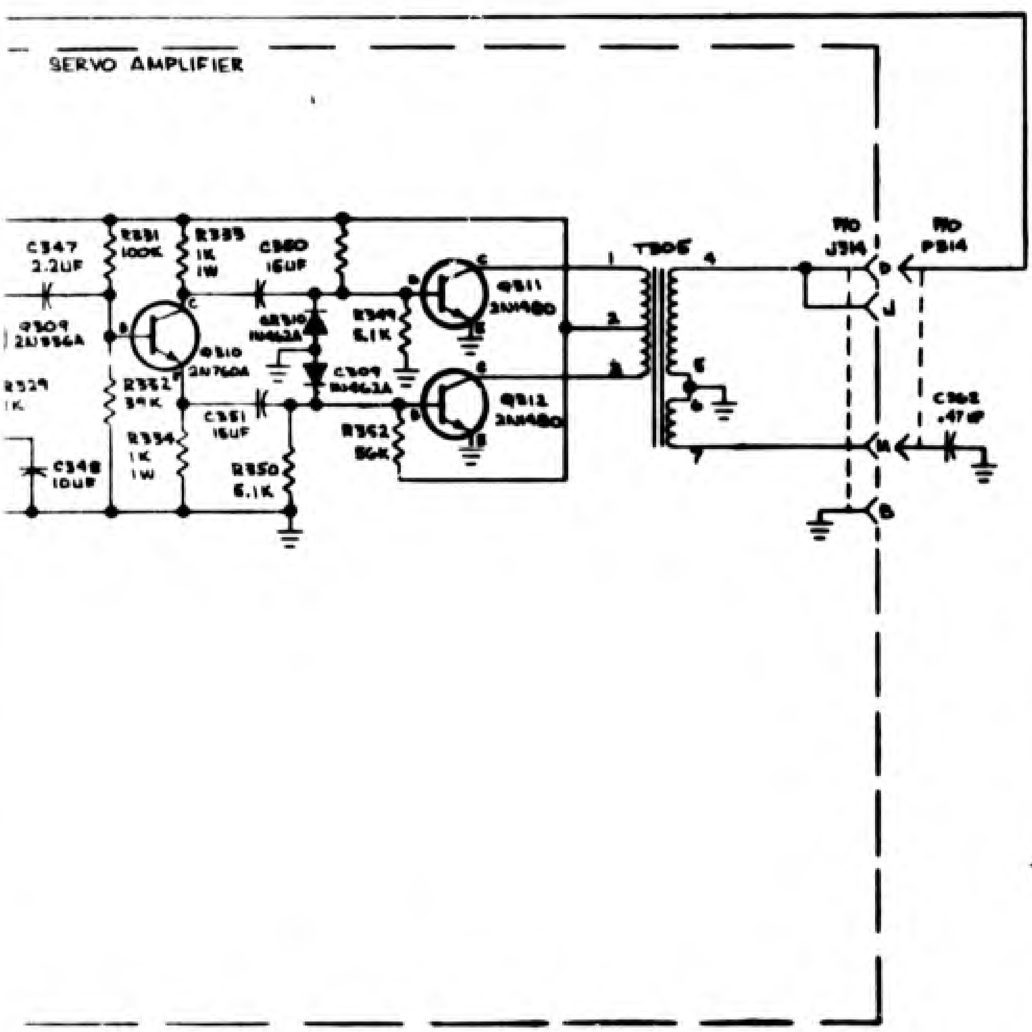


FIGURE 22. TRANSMITTER GROUP, SCHEMATIC DIAGRAM

5

transmitter frequency, a procedure similar to this is utilized. A false error signal is created and the transmitter channel is disabled. The loop is then closed and the servo motor is allowed to run until the frequency (as observed on the control box meter, position 1) is within lock range and then turned off. The multiplier chain and mixer are then tuned and the transmitter is then locked on frequency. In position 3 of the function switch the reference channel is disabled and the loop is opened. However, the transmitter channel is operating so that the output of the discriminator is proportional to the transmitter frequency. After the system is brought within lock range, the mixer is now tuned in this position for maximum output from the servo amplifier (as observed by meter on control box in position 8). In addition this serves as an excellent check on the transmitter channel of AFC circuit. The combination of the control box meter and test jacks in conjunction with the AFC function switch can isolate any malfunction of the equipment. For a more detailed tuning and maintenance procedure, reference is made to the maintenance manual written for the equipment.

3.9 Packaging

One of the major problems in this development was to provide for acceptable operating temperatures of both active and passive electrical components. This was difficult in that the volume of the transmitter and the power supply packages were limited to 460 cubic inches, and cooling was to be limited to radiation and natural convection into ambients of 100°F and 100,000 feet altitude. The size and shape of the transmitter's oscillator cavity, power supplies, transformers and filter capacitors restricted the overall package to the extent that a large flat package with an abundance of external surface area for cooling became infeasible. Therefore, special attention had to be given to "heat sinking" many electrical components so as to minimize the temperature drop between these components and the exterior cooling surfaces of these respective housings. This left only a few components which were able to operate in the 125°C sea level ambient predicted for the air internal to the packages.

One of the most unique solutions to the "heat sinking problem" involved the anode of the 2C39 type planar triode oscillator tube. Inasmuch as grounded grid operation became necessary, the "heat sink" had to operate across a potential of approximtely 1000 volts. The solution involved the use of a thin slab of sintered beryllium oxide which is far superior to the more conventional mica as it has a co-efficient of thermal conductivity approximating that of aluminum. A drawing showing this construction is shown in Figure (23).

Construction of the power supply castings is also rather unique in that they resemble an I-beam with closed ends. The web, in this case, serves to indirectly "heat sink" components to the exterior surfaces of the supply and removal of the covers provides access to components on either side of the web.

To facilitate replacement, all transistors are mounted in sockets. The planar triode, the modulation varactor, the servo motor, the AFC crystal, mixed diode, as well as all the transistors in the AFC are reached via convenient access plates. All

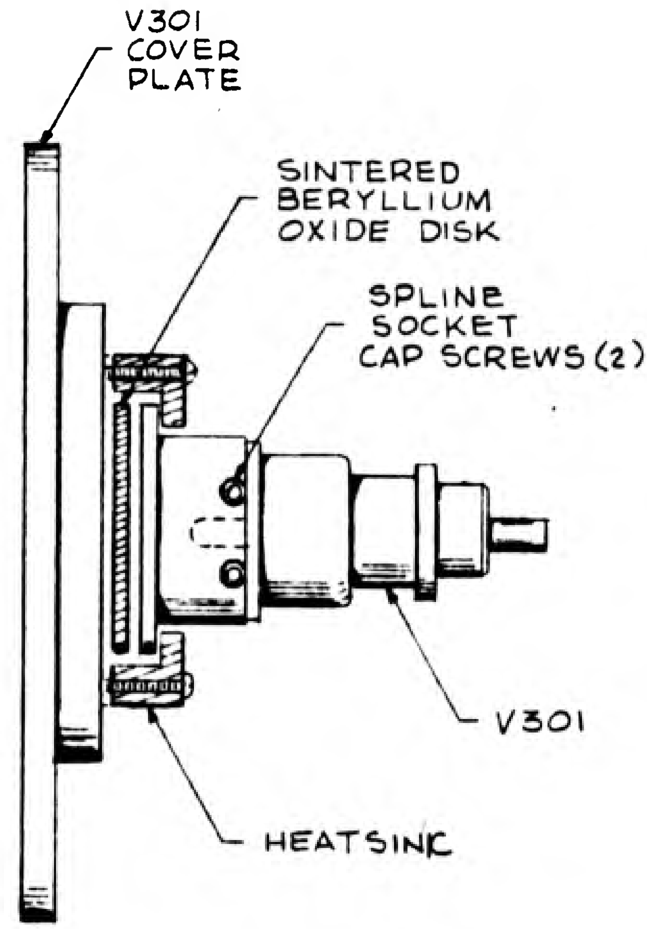


FIGURE 23 V301 AND COVER PLATE ASSEMBLY, CUT-AWAY VIEW.

tuning controls are readily accessible via a single convenient access plate, and tuning of the center frequency is accomplished electrically. An overall view of the equipment is shown in Figure (24).

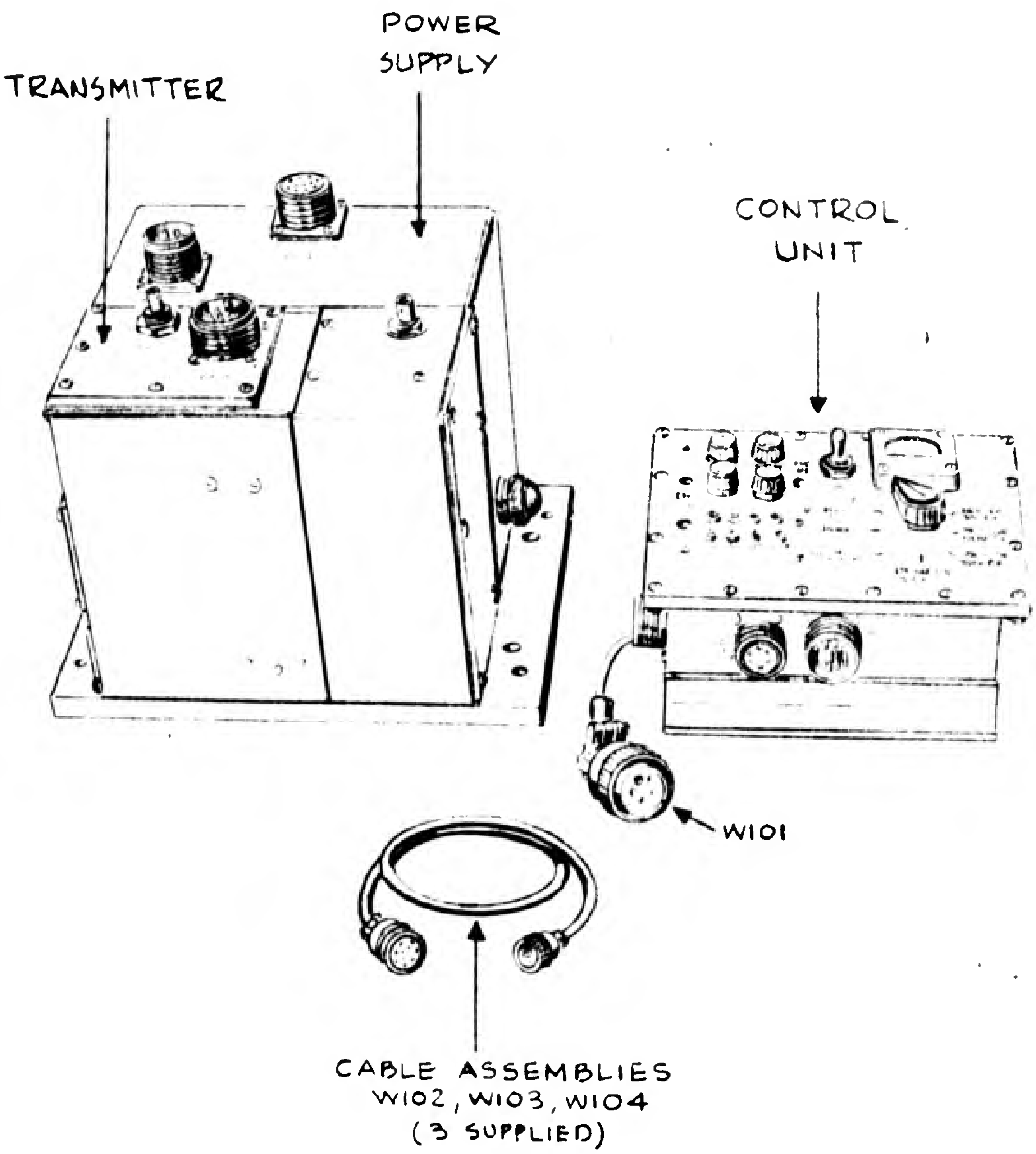


FIGURE 24 TRANSMITTER GROUP AN/AKA-( )

4.0 PERFORMANCE DATA

One of the critical problems associated with the development of this equipment was the difficulty encountered in procuring test equipment suitable for FM deviation, distortion and linearity measurements. No commercial equipment was available that would measure these parameters at "S" Band. The following procedures outline the methods and equipment that were evaluated to yield results with accuracy that was meaningful.

4.1 Electrical Test Data

The test setup and the equipment used for testing are shown in Figure (22) and Table (1) respectively. The electrical test data may be broken down into 3 main categories. They are:

1. Frequency Stability
2. Power Output
3. Modulation

4.1.1 Frequency Stability

The transmitter is tuned to the desired frequency. After turn-on, a 5 minute warm up period is allowed before stability measurements are made. With reference to Figure (23), a sampling of the transmitter frequency is fed to the transfer oscillator counter type frequency measuring setup,through a directional coupler. After the warm up period, a plot of carrier frequencv vs time is taken every minute for a period of not less than 30 minutes.

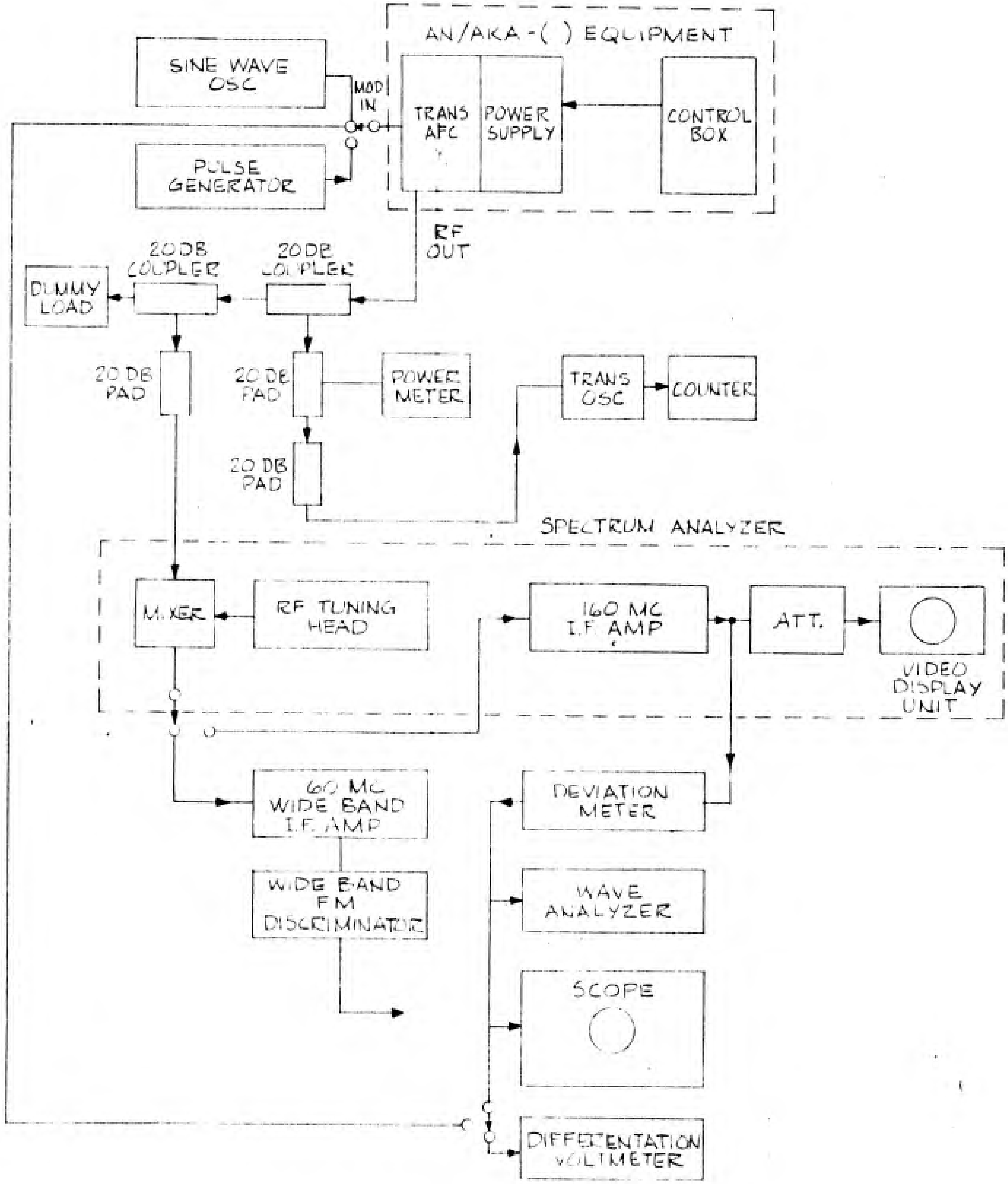


FIG. 25 TEST CONNECTION DIAGRAM

TABLE 1

LIST OF TEST EQUIPMENT

NOMENCLATURE	MANUFACTURER	MODEL #
Transfer Oscillator	Hewlett Packard	504B
Frequency Counter	Hewlett Packard	524D
Wave Analyzer	Hewlett Packard	302A
Audio Oscillator	Hewlett Packard	200C
Power Meter	Narda	440
Directional Couplers (20DB)	Narda	3003-20
Attenuator (20 DB)	Narda	757-20
Load Resistor	M.C. Jones	636NC
Differential Voltmeter	John Fluke	803
FM Deviation Meter	Marconi Instruments	TF928
Spectrum Analyzer	Lavoie Laboratories	LA-18M-133
Oscilloscope	Tektronix	541A
Pulse Generator	Electro-Pulse, Inc.	3450D
Wide Band I.F. * Amplifier and Discriminator	Collins Radio Co.	551A-1

* Part of R.F. Receiver Model #551A-1 manufactured by Collins Radio Co.

In addition, the line voltage is adjusted for both low and high setting (25 to 29 volts) and the carrier frequency recorded. Typical results obtained with the unit showed a stability of .002%.

4.1.2 Power Output

The Power Output is measured by sampling the transmitter power through 2 calibrated directional couplers. The output of the directional coupler is fed to a thermistor tvpe power meter and the resultant power read directly. The output power is recorded for both low and high line and, in addition, it is checked at low (2200), middle (2250), and high end (2300) of the frequency band. Results show that a minimum of 15 watts is obtained under the above stated condition.

4.1.3 Modulation

(a) General

The test setup for measuring linearity, deviation and distortion is shown in Figure (23). In all cases the measuring procedure involves converting the transmitter signal to a lower frequency, demodulation of the signal with a calibrated linear discriminator, and using the resultant waveform for testing. For deviation up to 500 KC's and modulation rates up to 125 KCP's, the Marconi Deviation Meter serves as the calibrated discriminator. Calibration and linearity of this equipment are within 0.1%.

For the narrow pulse width high bit rate modulation (PCM/FM), a wideband discriminator is used. The output of the unit has been calibrated with the Marconi Deviation Meter so that high bit rate 500 KC deviation can be measured.

4.1.4 Deviation Measurement

The demodulated output of the deviation meter is fed to an oscilloscope and accurate differential voltmeter. For deviation of 10 KC to 400 KC the deviation meter has a direct readout. For deviation below and above this range, additional calibration has been made. To check these results and to insure that the deviation meter is not working on a spurious harmonic, the spectrum analyzer is used to measure deviation by the well known lost carrier technique. Results show that the equipment is capable of 500 KC deviation with a maximum of 1.1 VRMS input.

4.1.5 Distortion Measurements

The input voltage to the modulator is adjusted for a specific carrier deviation. With this output level held fixed, the distortion of the modulation source is measured with the use of a Wave Analyzer. This instrument is essentially a tunable narrow band-pass filter followed by a low noise amplifier. The relative magnitude of the fundamental component of the input wave is then recorded with respect to the various harmonics. So that the % of distortion is given by:

$$D_1 = \% \text{ Distortion} = \frac{\sqrt{\sum_{N=2}^{\infty} E^2_N}}{E_1}$$

This method of measurement eliminates noise and 60 cps pickup, which is the limiting factor in most distortion measurement systems. The process is then repeated for the demodulated output from the deviation meter. The resulting distortion D_2 is measured as outlined above.

The actual distortion is then given by

$$D = \sqrt{D_2^2 - D_1^2}$$

The following data is typical of the unit's performance:

Deviation in $\pm$ KCPS	% Distortion
125	0.5%
160	1.0%
250	2.0%
400	4.5%
500	6.5%

4.1.6 Linearity Measurements

The output of the deviation meter is monitored bv a differential voltmeter. The deviation is adjusted to a specific value using the Marconi Deviation Meter. The voltmeter is then switched to measure the input voltage to the modulator and this process is repeated in 10 KC steps up to 150 KC. The accuracy of the differential voltmeter is 0.1%. Typical results of these tests are shown in Table 2. This data indicates a linearity of 1% when referred to the 150 KC deviation voltage.

4.1.7 Residual FM

The test system described for measuring carrier deviation is used to measure residual FM. Results taken with the DC to DC convertor power supply indicate that the unit has a maximum of $\pm$ 1 KCPS carrier deviation in absence of a modulation signal. It should be pointed out, however, that the 28 volts source used in the test had low ripple (less than 1V P to P). No tests were run with a 5 volt ripple on the input line. It is expected that such ripple could cause excessive residual FM.

4.1.8 PCM/FM, PDM/FM

A pulse generator with variable pulse width and pulse repetition rate capabilities is used as a modulator source. The bandwidth limitation of the Marconi deviation meter is 125 KCPS so that the high bit rate requirements of PCM/FM modulation cannot be used. In order to view this type modulation, a wideband FM discriminator is used. Bit rates of 800,000 and pulse width as small as 0.6 microsecond have been demonstrated. The rise time of the output pulse is limited by the 500 KC bandwidth of the modulator and is in the order of one microsecond. It should be noted that PCM information is AC coupled to the modulator so that it looses its DC reference. However, a "one" can be distinguished from a "zero" if the receiver discriminator characteristics are known.

TABLE 2

Carrier Deviation ± KCP's	Modulation Voltage in MV RMS
50	101
60	119
70	134
80	150
90	165
100	182
110	199
120	219
130	234
140	251
150	269
180	317
200	345
250	436
300	507
400	686
500	908

A "one" pulse will cause the carrier to deviate to a lower frequency while a "zero" will cause it to deviate to a higher frequency.

4.2 Environmental Tests

4.2.1 Temperature

The equipment was subjected to temperature variation of $-50^{\circ}C$ to $+85^{\circ}C$. Power Output, modulation characteristics and frequency stability measurements were made at these extremes, using the measuring techniques described previously. The result showed a frequency stability of $\pm$ 005%. A minimum of 15 watts output power is obtained and modulation characteristics were essentially the same.

4.2.2 Relative Humidity and Altitude

No tests were performed in the unit in regard to relative humidity and altitude. However, the unit is pressure sealed and no trouble is expected in this area. In addition, high altitude radiation cooling has been evaluated and appears satisfactory.

4.2.3 Vibration

The unit was subjected to the following vibration tests: 10 to 500 cps at a double amplitude of 0.036 inch and 500 to 2000 cps at 15 G applied acceleration in three mutually perpendicular planes. The data taken showed that the unit possesses no major resonances (magnification greater than 3:1). Output power and frequency stability are within the limits previously described. However, a large amount of residual FM was noted at various vibration excitation frequencies. The residual FM was in the order of ± 30 to 60 KCPS peak. Examination of the detected output showed the modulation rate of the residual FM to be at 10.7 KCPS. This condition was noted at various vibration frequencies across the band with more drastic results noted at higher vibration frequencies. It was determined that the disturbance was due to a resonance in the oscillator tube. The tube was mounted on a heavy plate with all power leads soldered directly to the contacts. The vibration source consisted of an audio oscillator driving a pair of headsets which were mounted on the same plate. The test setup is shown in Figure (24). By varying the frequency of the audio oscillator and observing the sampled plate current, it was noted that the tube exhibited a self resonance at 10.7 KCPS. The Q of resonance was extremely high and it was possible to excite the same resonance at various sub-harmonics of the resonant frequency. It was found that with less than 1G of applied acceleration, the peak change in plate current was in the order of 100 microamps. The bandwidth of this mechanical resonance was of the order of 10 cps ($Q \sim 1000$). In addition, a change in heater voltage resulted in a change in the resonant frequency. Contact was made with various tube manufacturers concerning the problem and other types as well as special tubes were tested for the same property. The following is an example of the data obtained:

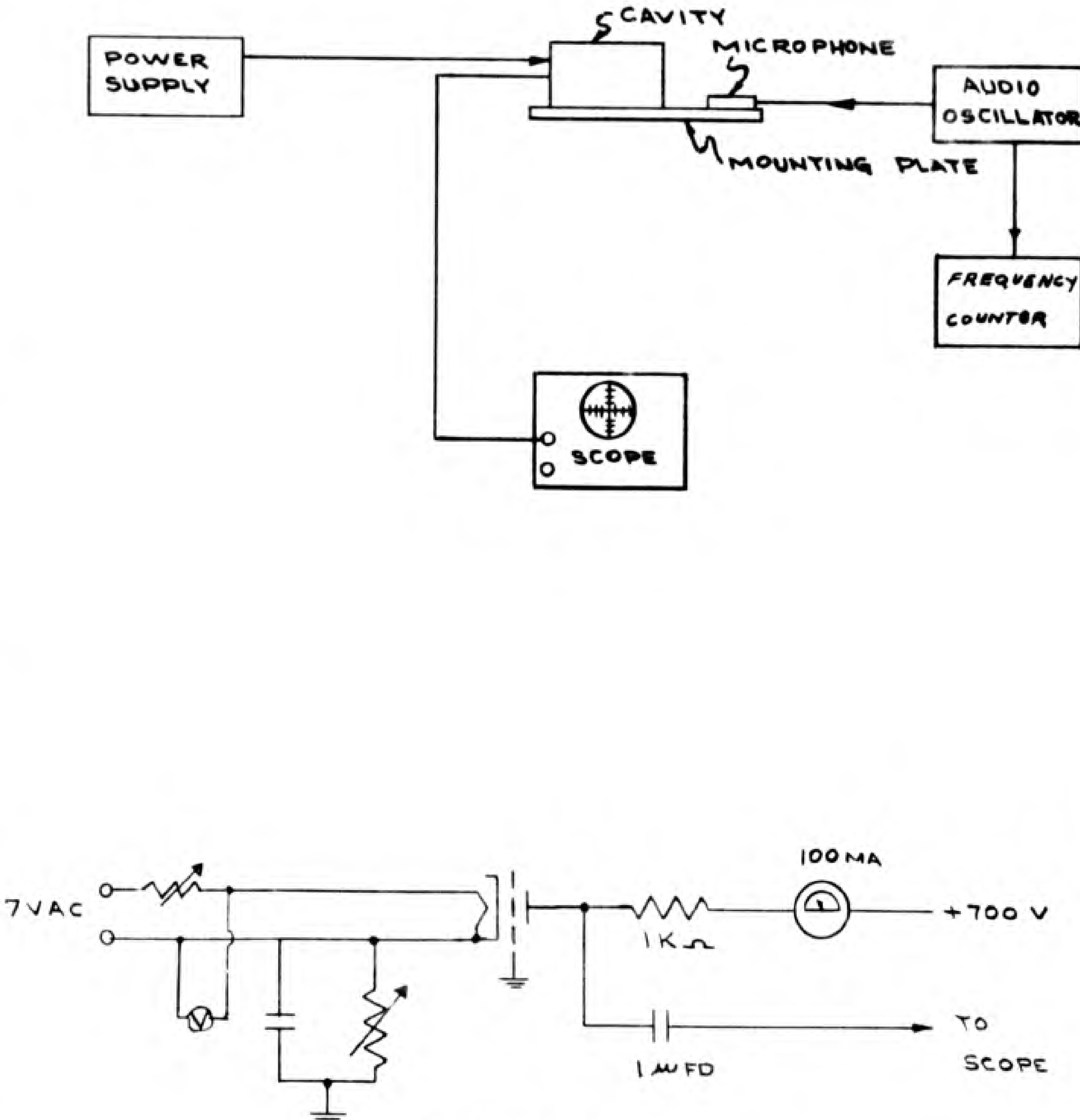


FIGURE 26
TUBE RESONANCE TEST SETUP

Tube Type	Resonant Freq.	Q	V out Across 1K
2C39B	10.7 KCPS	1000	100 MV
7289	8.0 48 KCPS	800	200 MV
6897	5.25 KCPS	1400	800 MV
3CX100A5	8.500 KCPS	____	100 MV
ML7211	4727	≈ 20,000	2 volts
ML7855	8947	≈ 800	300 MV
3CX100A5	9.7 KCPS	____	50 MV

It should be noted that tests indicated that the resonant frequency the tube exhibited was repeatable to within 10% for the same tube type and same manufacturer. By filtering out the unwanted 10.7 KCPS signal we were able to recover a clean modulation signal. In addition, a check was run on the intermodulation components with the resonant frequency and modulation frequency both present. Intermodulation components were found to be around 1% of the modulation signal at a deviation of $\pm$ 150 KCPS.

4.2.4 Shock

The unit was subjected to 50 G for 11 millisecond duration in three mutually perpendicular planes. Here again the resonant frequency exhibited by the tube caused excessive residual modulation.

4.2.5 Acoustic Noise

To date, no tests have been run for this property. However, it is felt that the same resonant property exhibited during vibration and shock would also show up in this test.

4.2.6 Acceleration

To date, no tests have been run for this property. However, no problem is forseen in meeting 40 G for one minute in three mutually perpendicular planes.

5.0 CONCLUSIONS AND RECOMMENDATIONS

5.1 Conclusions

The primary objective of the AN/AKA( ) program was the development of an airborne telemetry transmitter for the new 2200 to 2300 mcps telemetry band. Most of the design objectives outlined for the equipment have been designated as goals to be met by 1970[10]. The following objectives were met by the transmitter:

(a) High input impedance modulator which can effectively reduce the effect of stray wire and cable capacity.

(b) Capable of FM/FM, PDM/FM, and PCM/FM type modulation in accordance with MIL-STD-442.

(c) Maximum carrier deviation in excess of $\pm$ 500 K cps.

(d) A 1% modulation linearity with carrier deviations up to $\pm$ 150 K cps. Total distortion of 1% with + 125 K cps carrier deviation and 6% at $\pm$ 500 K cps carrier deviation.

(e) An output power of 15 watts, in any 100 mcps band over a range of 2150 to 2350 mcps, without the use of an auxiliary amplifier.

(f) Carrier stability of $\pm$ .005% over a temperature range of -50°C to $+85^\circ$C without the use of crystal oven.

(g) Total volume including power supply of 430 cu. inches.

In meeting the specifications it was felt that an increase in the "State of the Art" was accomplished in several areas. These include:

1. The use of varactors and its associated multiplier circuitry in modulating a high frequency, high power transmitter.

2. Development of a 15 watt power oscillator which is tunable over a 100 MC range at "S" Band.

In addition, the program revealed the resonance effect in the UHF planar triode grid structure and its effects on residual modulation when excited under shock and vibration. To the author's knowledge, there is no other published literature available on this phenomenon. It is also felt that the unit is unique in requiring a minimum of auxiliary equipment and time for the purpose of changing carrier frequency.

In conclusion, it is felt that the equipment represents a significant step forward in meeting the IRIG proposed goals for the new "S" Band range.

5.2 Recommendations

It is highly recommended that a program be instituted to minimize the grid resonance effects now present in UHF planar triode tubes. Unsolicited proposals were received from such companies as General Electric, Eitel-McCullough, and Machlett

Labs.,for programs which would modify the resonance effect. In addition, the basic military specification governing tubes (MIL-E-1D) should be modified to include vibration frequencies up to 2000 cps. The present tests are conducted at frequencies which are too low to indicate the presence of high frequency resonances.

It is also recommended that modifications be performed on the existing unit for operation at the new L band (1435 - 1535 mcps) telemetry band. Preliminary investigation has indicated that considerably more output power, at a greater efficiency could be obtained with minor modification of the power oscillator.

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APPENDIX

VARACTOR POWER DISSIPATION

By H. Cohen

Let ΔC = Effective change in varactor capacitance

δC = Effective change in cavity capacitance

$$\frac{\delta C}{\Delta C} = \frac{E_v^2}{E_t^2}$$

The ratio of $\frac{\delta C}{\Delta C}$ is, in effect, a coupling factor which relates the tank voltage to the varactor voltage. We also note that the differential change in frequency is related to the differential change in capacity by:

$$(1) \quad \frac{\delta f}{f_o} = 1/2 \; \frac{\delta C}{C_o}$$

Where δf = Change in osc. frequency

C_o = Nominal tank capacity

f_o = Center frequency of osc.

The voltage across the varactor is related to the tank voltage by:

$$(2) \quad E_v - E_t \left(\frac{\delta C}{\Delta C}\right)^{1/2}$$

The voltage across the tank is given by:

$$(3) \quad E_t = \left[(V.A.)\;(X_t)\right]^{1/2}$$

Where V.A. = Volt amperes in the tank circuit

X_t = Tank reactance

Sub (4) into (3) yields:

$$(4) \quad E_v = \left[(V.A.)\;(X_t)\;\frac{\delta C}{\Delta C}\right]^{1/2}$$

Now,

$$X_t = \frac{1}{\omega C_o} \qquad \text{and} \quad \frac{\delta C}{\omega C_o} \;-\; \frac{2\,\delta f}{f_o}$$

Therefore,

$$(5)\quad E_v = \left[(V.A.)\left(\frac{1}{\omega \Delta C}\right)\left(\frac{2\delta f}{f_o}\right)\right]^{1/2}$$

We now note that the volt amperes of a tank circuit is given by the product of the Power Output and the loaded Q of the tank. Therefore,

$$(6)\quad E_v = \left[P_o Q_L \cdot \frac{2\,\delta f}{f_o} \cdot \frac{1}{\omega \Delta C}\right]^{1/2}$$

The above is a general equation of the RMS voltage across the varactor. If we represent the varactor as a resistor in series with a variable capacitor, then a portion of the voltage will be developed across each. Before proceeding with the computation, it would be desirable to work with a parallel combination of R & C rather than a series.

It can be shown that:

$$C_p = \frac{C_s}{(\omega C_s R_s)^2 + 1} \qquad R_p = \frac{(\omega C_s R_s)^2 + 1}{\omega^2 C_s^2 R_s} = R_s(Q^2 + 1)$$

Where C_p = Equivalent parallel "C"

R_p = Equivalent parallel "R"

Also, we let

C_1 = Minimum value of series capacitor

C_2 = Maximum value of series capacitor so that

$$\Delta C = C_2 - C_1$$

$$(7)\quad \Delta C_p = \frac{C_2}{(\omega C_2 R_s)^2 + 1} - \frac{C_1}{(\omega C_1 R_s)^2 + 1} = \frac{C_2 Q_2^{\,2}}{1 + Q_2^{\,2}} - \frac{C_1 Q_1^{\,2}}{1 + Q_1^{\,2}}$$

From equation (6)

$$E_v = \left[P_o Q_L \cdot \frac{2\delta f}{f_o} \cdot \frac{1}{\omega \Delta C_p} \right]^{1/2}$$

Sub in equation (7)

The power dissipated in the varactor is given by:

$$\frac{E_v^{\,2}}{R_p} = P_o Q_L \cdot \frac{2\delta F}{f_o} \cdot \frac{\omega^2 C_s^{\,2} R_s}{(\omega C_s R_s)^2 + 1} \cdot \frac{1}{\omega \Delta C_p}$$

It is seen that this equation is a maximum for maximum C.
Therefore, we will compute the power using the value of C_2 and Q_2.

$$(8)\quad P_d = P_o Q_L \cdot \frac{2\delta F}{f_o} \cdot \frac{1}{R_s\,(Q_2^{\,2} + 1)} \cdot \frac{1}{\omega \Delta C_p}$$

Multiply numerator and denominator by C_2, yields:

$$(9)\quad P_d = P_o Q_L \cdot \frac{2\delta F}{f_o} \cdot \frac{1}{\omega \Delta C_p} \cdot \frac{C_2 Q_2}{Q_2^{\,2} + 1}$$

and then sub for ΔC_p

$$(10)\quad P_d = P_oQ_L \cdot \frac{2\delta F}{f_o} \cdot \frac{C_2Q_2}{Q_2^{\,2}+1} \cdot \frac{\dfrac{1}{C_2Q_2^{\,2}}}{1+Q_2^{\,2}} - \frac{C_1Q_1^{\,2}}{1+Q_1^{\,2}}$$

For most cases of interest,

$Q_1^{\,2} >> 1$

$Q_2^{\,2} >> 1$

So that

$$P_d = P_oQ_L \cdot \frac{2\delta F}{f_o} \cdot \frac{C_2}{Q_2} \cdot \frac{1}{C_2 - C_1}$$

$$(11)\quad P_d = P_oQ_L \cdot \frac{2\delta F}{f_o} \cdot \frac{1}{Q_2} \cdot \frac{C_2}{C_2 - C_1}$$

The final equation for the varactor dissipation can be broken up into two parts. The term $P_oQ_L \frac{2\delta F}{f_o}$ is fixed by circuit requirements. The term $\frac{C_2}{C_2 - C_1} \cdot \frac{1}{Q_2}$ is completely related to the varactor characteristics and, so to speak, is the "Figure of Merit" of the unit for this type application. We define:

$$FM = Q_2 \cdot \left[1 - \frac{C_1}{C_2}\right]$$

so that,

$$P_d = P_oQ_2 \cdot \frac{2\delta F}{f_o} \cdot \frac{1}{FM}$$

It is evident that in picking a varactor for this type application, the varactor should have high ratio of maximum and minimum capacity and a high Q at the maximum capacity point. The approximation made in the above derivation included neglecting the series inductance as well as the parallel capacity of the varactor. It is felt that these two parameters are held to a minimum for the most part, particularly in the new Pill Type Package. The approximation involved for C_p is valid as long as the Q of the varactor is four or greater at the maximum capacity value.

Aeronautical Systems Division, AF Avionics Lab, Wright-Patterson AFB, Ohio.
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(2) ± 500 kc carrier deviation with good

(over)

modulation linearity and low distortion
(3) All solid-state except for power oscillator stage.

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2. Microwave transmitters
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