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RADC-TDR-64-428
Final Report

MODIFICATIONS MADE TO THE USAF TRANSPORTABLE PASSIVE SATELLITE COMMUNICATIONS TERMINAL

TECHNICAL DOCUMENTARY REPORT NO. RADC-TDR-64-428

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Communications Applications Branch

Rome Air Development Center
Research and Technology Division
Air Force Systems Command
Griffiss Air Force Base, New York

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(Prepared under Contract No. AF 30(602)-2944 by Sylvania Electronic Systems, a division of Sylvania Electric Products Inc., Wehrle Drive and Cayuga Road, Williamsville, N.Y.)

FOREWORD

This report was prepared by Sylvania Electronic Systems, a Division of Sylvania Electric Products, Inc., Buffalo, New York under Contract AF 30(602)-2944. The work was performed for the Rome Air Development Center under AFSC Project Number 4519, Task Number 451901. The contract started on 31 January 1963 and ended 31 August 1964. This report was submitted on 31 August 1964. Special acknowledgement is made of the substantial contributions of the following persons: Henry Hyams, John Bartnik, James Kellett, Thomas Leonard, Robert McIntyre, Joseph Casserta, Bernard Paters, Robert Bennett and Roger Brockington.

AIR FORCE, GAFB, N.Y., 6 JAN 65-122

ABSTRACT

The purpose of this effort was to increase the capabilities of the Transportable Passive Satellite Communications Terminal. Modifications made include a new "X" band feed structure to permit transmission and reception on either of two orthogonal linear polarizations, adding a diplexer to permit reception of an "X" band signal while simultaneously transmitting a 10 kw "X" band signal, and adding single sideband and double sideband suppressed carrier modulation capability to the existing frequency modulated transmitter. In addition, a stable frequency source was installed which has a stability and accuracy of better than one part in 10^9. Necessary frequency synthesizers were added to permit changing the transmit frequency from 7630 Mc to 8330 Mc and the antenna feed line was modified to permit operation at the new frequency of 8330 Mc.

PUBLICATION REVIEW

This report has been reviewed and is approved. For further technical information on this project, contact James W. Bailey, EMCRR, extension 3185.

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INTRODUCTION

The TPSC Terminal as originally designed consisted of a 10 Kw Cw X-band transmitter operating at 7630 Mc and capable of being frequency modulated at baseband frequencies between 30 cps and 30 kcps and an S-band monopulse receiving system capable of providing automatic steering data to the antenna system as well as recovering the information traffic.

The antenna system consists of a 30 ft. diam. paraboloid and a 5 ft. diam. cassegrainian sub-reflector on an azimuth and elevation mount which in turn is mounted on a trailer bed.

The modification program reported herein was intended to provide the terminal with an X-band diplexer, a very stable frequency standard and synthesizer and the capability of transmitting at X-band using single sideband and double sideband suppressed carrier modulation techniques. A complicating factor in the implementation of the proposed terminal modifications was the future shift of transmitter frequency from 7630 Mc to 8330 Mc. This meant that the synthesizer would have to supply two sets of outputs, one for each radiated frequency. Because the X-band receiver installation was not scheduled until after the change in transmitter frequency, it was not necessary to design the diplexer for operation at 7630 Mc but a simple means of bypassing it was necessary. An additional flexibility feature of the modified terminal was to be the ability to transmit at X-band using either horizontal or vertical polarization and to receive at X-band using either horizontal or vertical polarization, the choice of the latter to be independent of the transmitter polarization in use at a particular time. In order to obtain this flexibility in polarization it was necessary to modify also, the X-band antenna feed assembly.

The transmission line used to transfer power from the X-band transmitter to the antenna feed point contains several sections of three inch circular waveguide propagating the TE_{01} circular mode. Six mode transducers are employed to effect the transition from circular to rectangular guide. Reference to the manufacturer's data and examination of the mode transducers disclosed that the original transducers would not perform properly at the new transmitter frequency of 8330 Mc. Engineering Change "A" altered the contract to call for the design, fabrication and installation of six mode transducers having a bandwidth of 6 to 10% at 8330 Mc and an insertion loss of 0.025 db or less.

1.0 ANTENNA AND TRANSMISSION LINE SYSTEM

The transmission line which transfers the 10 Kw X-band output of the transmitter to the X-band antenna feed element contains two rotary joints and three different waveguide sizes. The terminal was originally designed to transmit at 7630 Mc. At this frequency WR112 waveguide has a minimum attenuation of 2.5 db per 100 ft. and WR137 waveguide has a minimum attenuation of 1.6 db per 100 ft. This loss is significant as is the difference in the losses of the two guides. The VA823 klystron amplifiers and the associated waveguide components inside the equipment van employ WR112 (RG/51U) waveguide. Between the van and the feed elements WR137 (RG/50) waveguide is employed wherever rectangular guide must be used. If rectangular waveguide were used for the entire run, the losses would be excessive even with WR137 guide. In order to keep the overall transmission loss low 3.0" ID circular waveguide has been used wherever practical. The low loss properties of the circular guide are realized only when it is propagating the circular TE_{01} mode. Any circular waveguide which will propagate a TE_{01} mode at a given frequency will also propagate several non-circular modes including the dominant $TE_{1,1}$. In general, any disturbance in the guide which destroys its circular symmetry or alters the phase velocity of one part of the guide with respect to another will cause some of the energy in the desired TE_{01} mode to couple to other non-circular modes. A feature of the $TE_{0,1}$ mode is that it produces no wall currents having components in the axial (Z) direction. As a result of this rotary joints can be formed by simply butting two guide sections end-to-end with a small gap between. The $TE_{0,1}$ mode will not excite the gap and radiate from it. The same is not true of the degenerate non-circular modes. These observations point to the urgent need of preserving a high degree of $TE_{0,1}$ mode purity in the circular guide. For this reason no bends can be formed in the circular guide. It is necessary to transform to a more stable mode where bends must be incorporated. Because of the way in which the waveguide is laid out six mode transducers are required in the overall transmission line. The mode transducers installed in the original line are described in the Technical Report RADC TDR-62-502 on Page 62. The VSWR versus frequency of these transducers is given on Page 95, (Figure 48) of the same document. Because of the relatively narrow bandwidth of these transducers it was necessary to replace them in order to accommodate a transmitter frequency of 8330 Mc. The mode transducers installed under Contract AF30(602)-2944 employ a folded series Tee to split the power incident on the WR137 part into two equal amplitude parts 180° out of phase. These two parts emerge in one-half height WR137 guide. They then pass into a tapered transformer section which first transforms the $TE_{0,1}$ mode in the half height WR137 guide to a $TE_{0,1}$ sector mode and then gradually expands the sector angles until the energy is transformed to a full circular $TE_{0,1}$ mode at which point the guide cross section becomes a 3 inch ID circular pipe.

Except for the mode transducers, all other modifications to the Antenna - Transmission Line System are located in the equipment box, the feed box and the feed dome atop the antenna pedestal. Figure 1-1, which is a reproduction of the feed system assembly drawing DC-16-3019, shows the additions and changes to the antenna mounted equipment. The principal items are the diplexer 15 , the receiver polarization switch 14 , the dual mode transducer 9 , the feed horn assembly DC-16-3015-8, the receiver bandpass filter (not shown), and the local oscillator multiplier assembly 27 .

1.1 Diplexing System

The requirements placed upon the diplexing system are as follows:

A. The system must be capable of radiating a minimum transmitter power of 10 Kw CW at either 7.630 Gc or 8.330 Gc.

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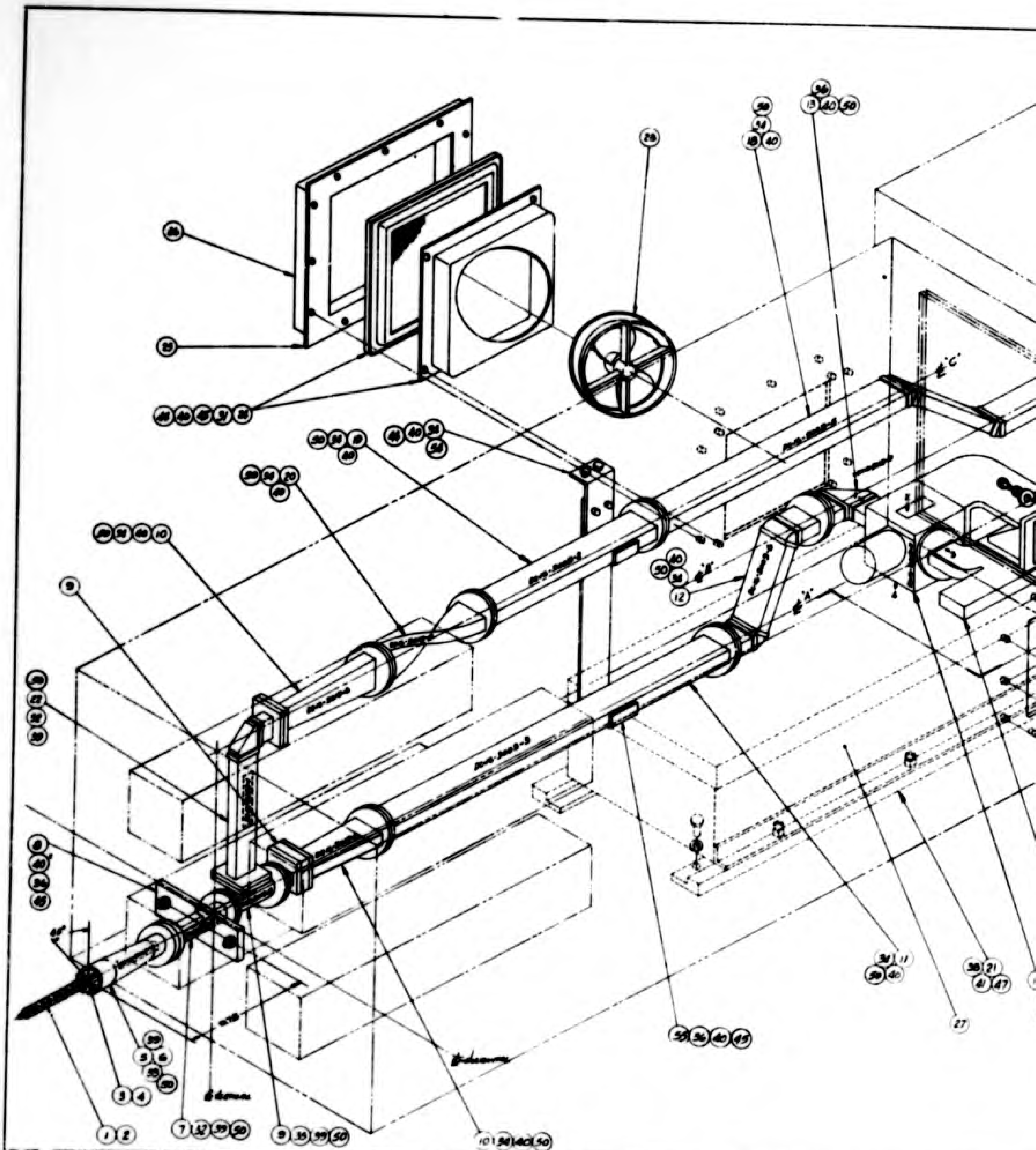


A

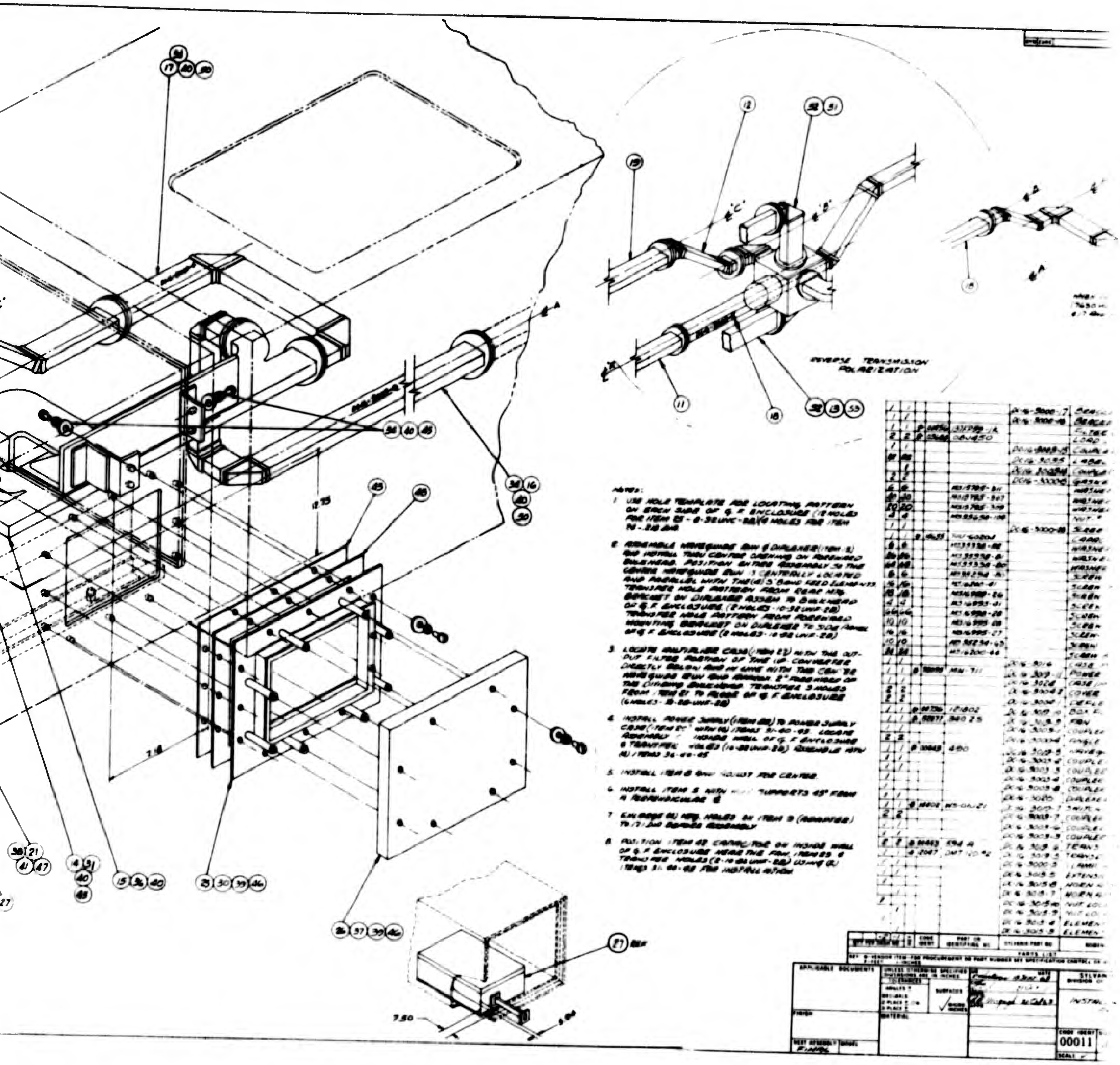
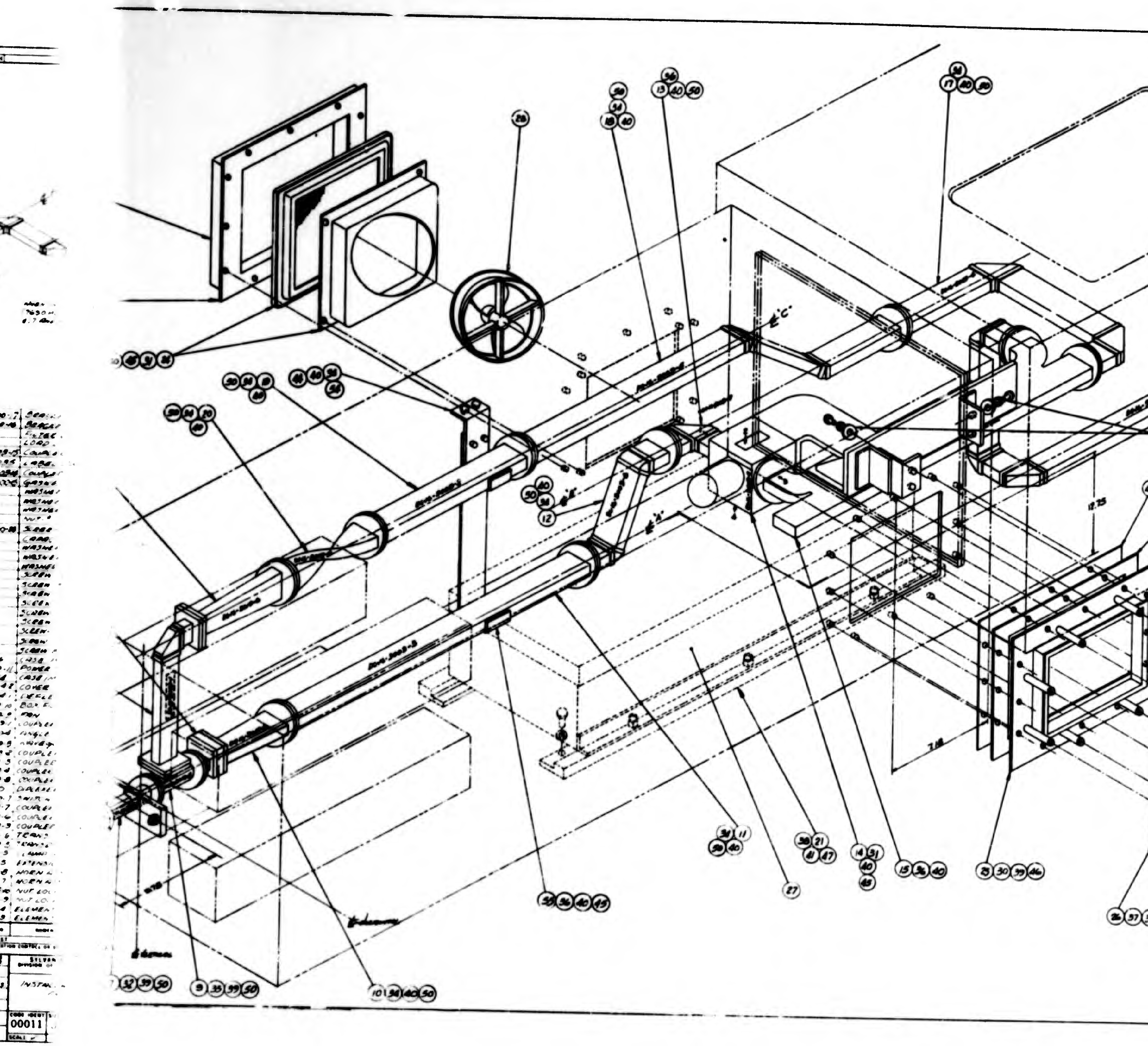


Figure 1-1. Installation Assembly, Antenna Microwave Equ



e Equ

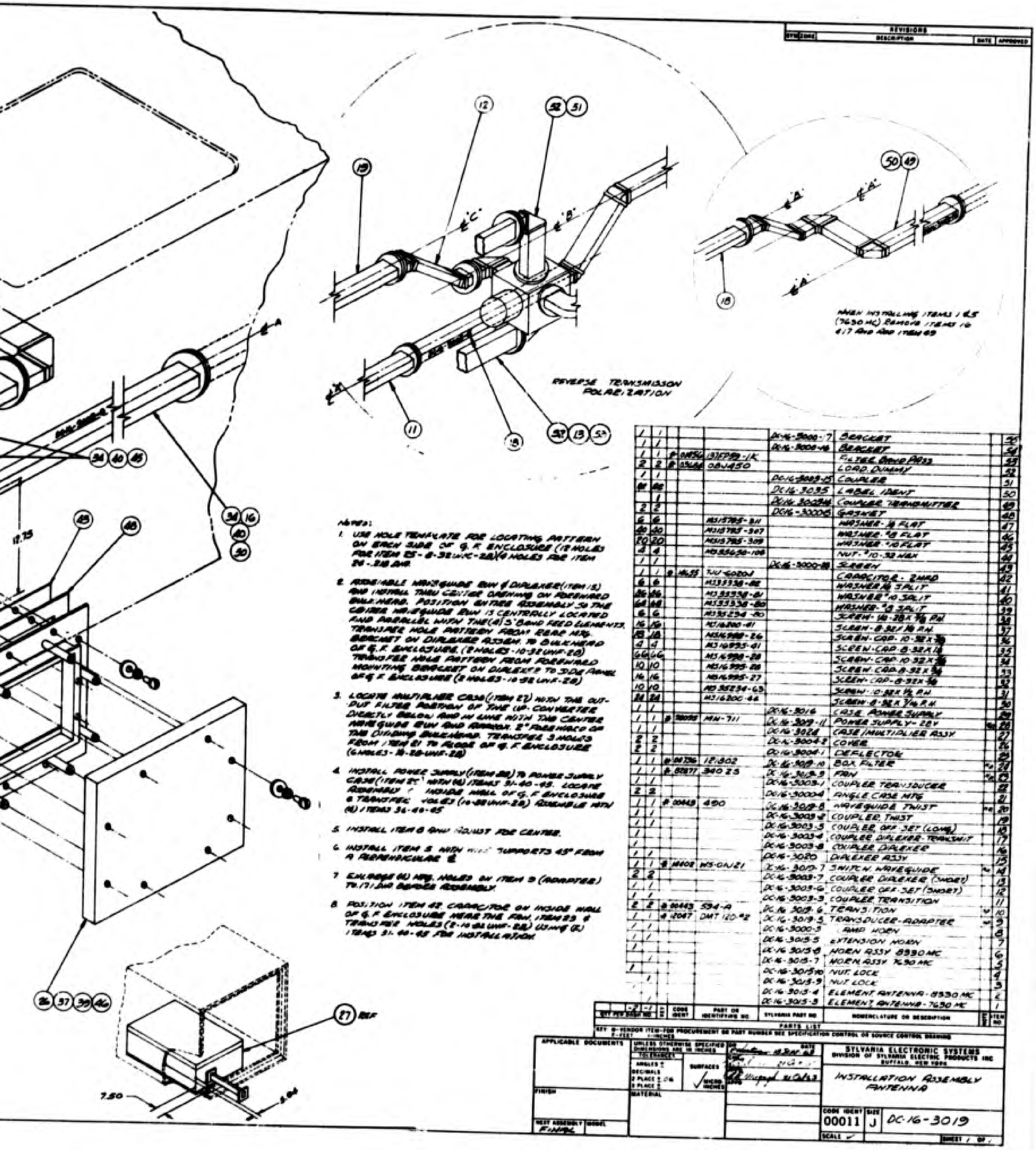


Figure 1-1. Installation Assembly, Antenna Microwave Equipment

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B. The system must be capable of receiving frequencies in the range from 7.690 Gc to 7.790 Gc while simultaneously transmitting at a frequency of 8.330 Gc.

C. The system must be capable of radiating either vertically or horizontally polarized waves.

D. The system must be capable of receiving vertically or horizontally polarized waves independently of the polarization being radiated.

E. The system must not contribute more than 0.5 db additional loss in the transmission path over that existing prior to this modification.

F. The insertion loss of the reception path from the feed horn assembly to the receiver input port must not exceed 0.75 db over the receive pass band 7.690 Gc to 7.790 Gc.

G. The system must attenuate power at the transmission frequency to a level less than -30 dbm at the receiver input port.

H. The overall system must attenuate portions of the transmitter spectrum lying within a band of plus or minus 10 megacycles of the receive frequency (7.730 Gc) to a power level less than -130 dbm at the receiver input port.

The diplexer satisfies these requirements by providing five possible signal flow arrangements. Each of these will be considered below.

Transmitter On 7.630 Gc - No Receiver

In this set up waveguide section DC-16-3003-8 is removed and section DC-16-3003-14 is substituted. This bypasses the diplexer and connects the transmitter directly to the dual mode transducer.

Transmitter on 8.330 Gc - Transmitter and Receiver Vertically Polarized

In this configuration the transmitter signal passes through the diplexer (0.05 db insertion loss) directly to the vertical port of the dual mode transducer. The received signal passes through the diplexer (0.15 db insertion loss) and the receiver bandpass filter (0.3 db insertion loss) to the receiver input port. The horizontal port of the dual mode transducer is connected through the waveguide switch to a waveguide load. Isolation of the diplexer at 8.330 Gc is greater than 50 db. The receiver bandpass filter attenuation to 8.330 Gc is greater than 60 db so that the total isolation is more than 110 db. The transmitter power level is +70 dbm. The transmit frequency power level to the receiver is -40 dbm or less.

Transmitter On 8.330 Gc - Transmitter and Receiver Horizontally Polarized

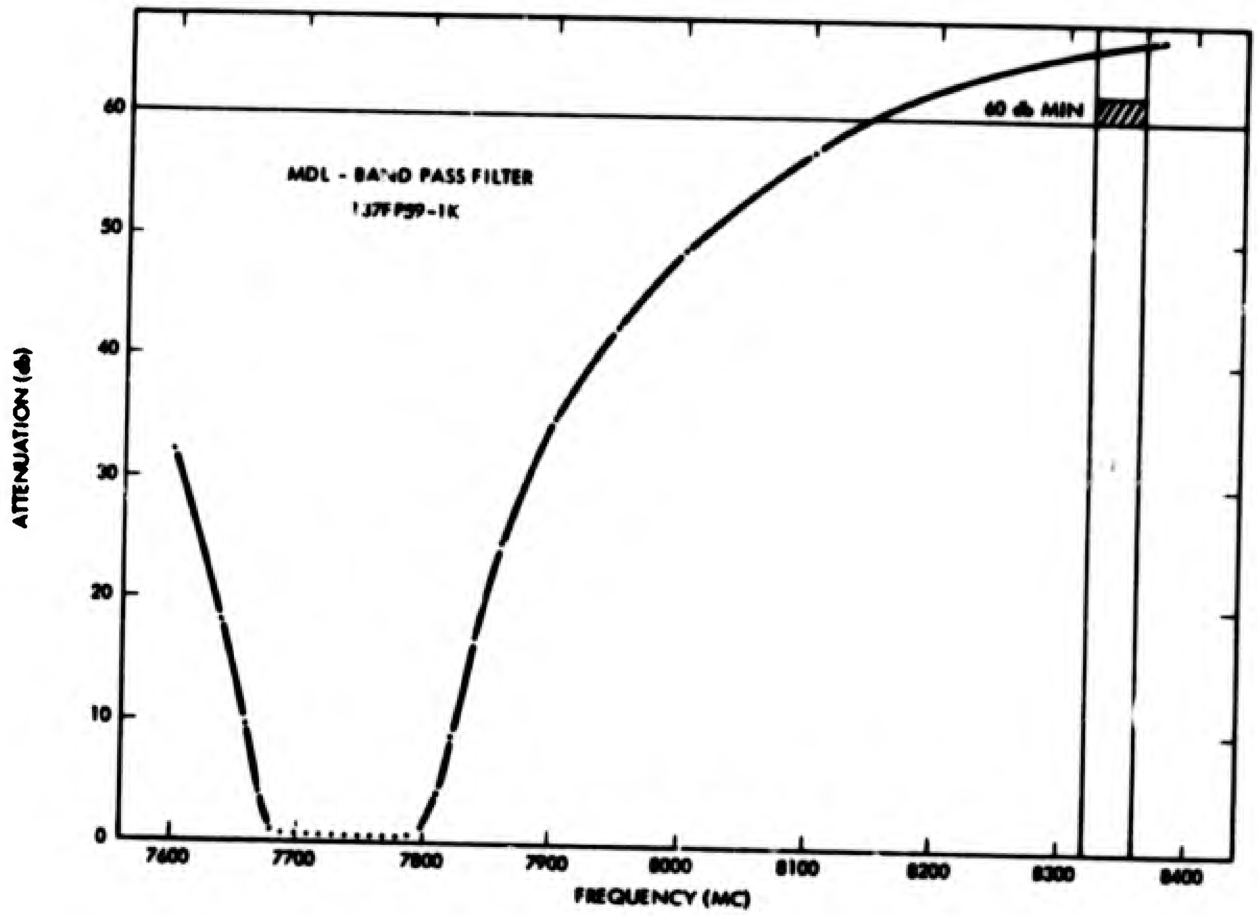
This configuration is identical to the one described above except that the diplexer output connects to the horizontal port of the dual mode transducer.

Transmitter On 8.330 Gc - Transmitter Vertically Polarized - Receiver Horizontally Polarized

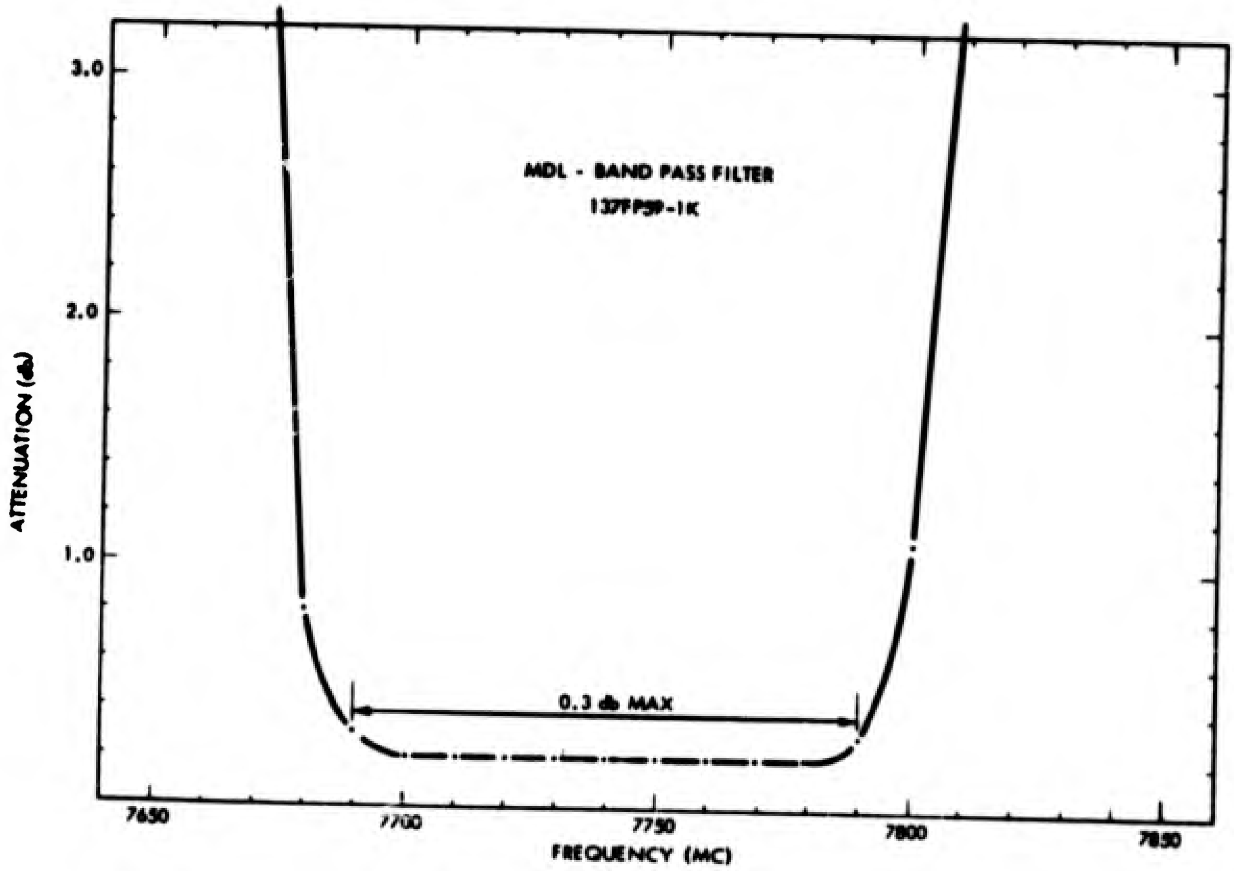
For this condition the transmitter signal passes through the diplexer (0.05 db insertion loss) to the vertical port of the dual mode transducer. The receiver port of the diplexer is terminated in a load via the waveguide switch. The received signal is taken from the horizontal port of the dual mode transducer (44 db isolation), passes through the receiver bandpass filter (60 db attenuation to 8.330 Gc) via the waveguide switch to the receiver input port. Total isolation at the transmitter frequency is 104 db.

Transmitter On 8.330 Gc - Transmitter Horizontally Polarized - Receiver Vertically Polarized

This condition is identical to the one above except that the horizontal and vertical ports of the dual mode transducer are interchanged and the waveguide switch is in the alternate position. Measured characteristics of the receiver bandpass filter are given in Figure 1-2. Details of the remaining portions of the diplexing system are described below.



a) Overall Characteristics



b) Details of Passband

Figure 1-2. Receiver Bandpass Filter Response Characteristics

1.1.1 Diplexer

The diplexer permits the antenna feed horn assembly to be used simultaneously both for transmission and reception at X-band. It is designed to allow transmission in a band extending from 8320 Mc to 8360 Mc and reception in the band between 7690 Mc and 7790 Mc. The maximum insertion loss in the transmission path (transmitter to antenna) is less than 0.15 db. Insertion loss in the received signal path is less than 0.3 db. A schematic representation of the TPSC Terminal diplexer is given in Figure 1-3.

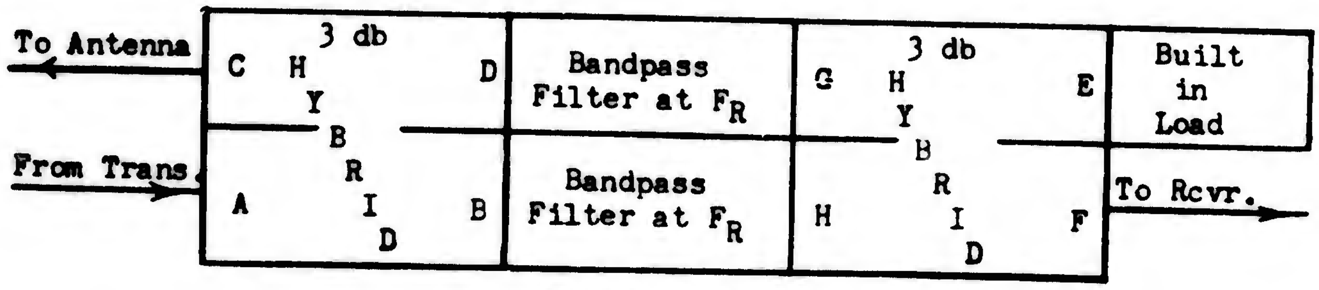


Figure 1-3

In this configuration, the pass band of the filters is centered on the received frequency. The filters are designed to reject the transmitter frequency. Being non-dissipative, the filters reflect power at frequencies outside of their pass band. Power from the transmitter enters the short slot sidewall hybrid at "A". The two bandpass filters at "B" and "D" appear as short circuits. The action of the hybrid is such that the voltage at "B" is advanced in phase - 45° beyond the value it would have were the slot not present. The voltage at "D" leads the voltage at "B" by 90° or it has been advanced a total of 135°. Any energy reflected from "D" which couples back through the slot to "A" will have been advanced a total of 270°. Any energy reflected from "B" which appears at "A" will be advanced 90° and will be in exact phase opposition to the energy coupled back from "D". Energy reflected from "D" which does not couple back will appear at "C" with a total phase advance of 135° +45° or 180°. Any energy reflected from B and coupled through the slot will appear at C with a total phase advance of 45° + 45° + 90° or 180° and will add to the component reflected from D. Therefore, all the energy entering port A exits via Port C.

Received energy enters Port C. One half of it couples through the slot and appears at B. The remaining half of the power appears at D. The bandpass filters present a matched load impedance at the received frequency and the energy at D and B passes on to G and H. Because of the hybrid action, the voltage at B leads the voltage at D by 90°. Also, the voltage at H leads that at G by 90°.

Voltage from G appearing at F will be advanced 135°. Voltage at H is advanced 90° over G. Voltage at F from H will be advanced 90° + 45° or 135° over G and will add in phase to that from G. Voltage at E from G will be advanced 45°. Voltage at E from H will be advanced over G by 90° + 135° = 225°. Therefore, no energy exists at E and all energy exits at F.

The diplexer used in the TPSC Terminal modification was obtained from Microwave Development Laboratories. Specified and measured performance data on this unit are tabulated in Figure 1-4.

<u>Antenna Feed Horn Assembly</u>

The feed horn assembly is located inside the plastic feed dome between the four S-band feed elements. It consists of a right circular conic horn having a mouth diameter of 1.55 inches and a disc-on-rod slow wave element. Some of the energy radiates directly from the horn and propagates forward at the velocity of light. The remaining energy becomes trapped on the disc-on-rod element and propagates forward at a velocity

TEST INFORMATION - DIPLEXER UNIT

156-001

Per Spec. 111-002-1 - Rev. 1 - 12 June 1963

REQUIREMENTS

Transmitter Link: Insertion loss; less than 0.15 db from 8.320-8.360 Gc

VSWR; less than 1.10:1 from 8.320-8.360 Gc

Receiver Link: Insertion loss; less than 0.3 db from 7690-7790 Gc

Isolation: Transmitter to receiver link; greater than 40 db from 8.320-8.360 Gc

TEST DATA

Insertion Loss Transmitter Link:	8.320	0.05 db
	8.340	Less than 0.05 db
	8.360	0.10 db
Insertion Loss Receiver Link:	7.690	0.15 db
	7.740	0.10 db
	7.790	0.20 db
VSWR Transmitter Link:	8.320	1.03:1
	8.340	1.035:1
	8.360	1.05:1
Isolation Transmitter to Receiver:	8.320	Greater than 50 db
	8.340	Greater than 50 db
	8.360	Greater than 50 db

Figure 1.4

less than the velocity of light. The vector sum of the direct radiation and the phase retarded radiation from the slow wave element is a narrow focused beam. The wave fronts are retarded near the axis of the horn causing the rays to tilt toward the axis. Figure 1-5 is a radiation pattern obtained using the horn only. Figure 1-6 is the pattern obtained when the disc-on-rod element was inserted. Both the horn alone and the disc-on-rod element have broader bandwidths than the complete assembly. The overall bandwidth is narrowed primarily because of the resonant nature of the rod supports. It is possible to obtain a small amount of tuning in order to trim to a low VSWR by axially positioning the disc-on-rod element and by moving the tip loading nut inside the horn mouth. This is not a recommended field adjustment as the adjustment is critical.

The feed horn assembly is symmetrical about the center. It appears exactly the same to a horizontally polarized wavefront as it does to a vertically polarized wave. The feed horn itself is equally adaptable to vertical, horizontal or circular polarization. The mode pattern in the throat of the horn and in the circular guide behind it is the TE_{11} which resembles the $TE_{1,0}$ rectangular guide mode. (See Figure 1-7.)

1.3 Dual Mode Transducer

The dual mode transducer is the three-port waveguide junction located just inside the forward bulkhead of the feed box and directly behing the 1.046" diameter circular guide to the feed horn. The basic property of this junction is that it allows the feed horn to be driven with both horizontally and vertically polarized waves simultaneously without interaction of the sources. In this application it is possible to be transmitting through one rectangular guide on a given polarization and to simultaneously receive the perpendicular polarization through the remaining rectangular guide. When the circular guide is terminated in a matched load, the isolation

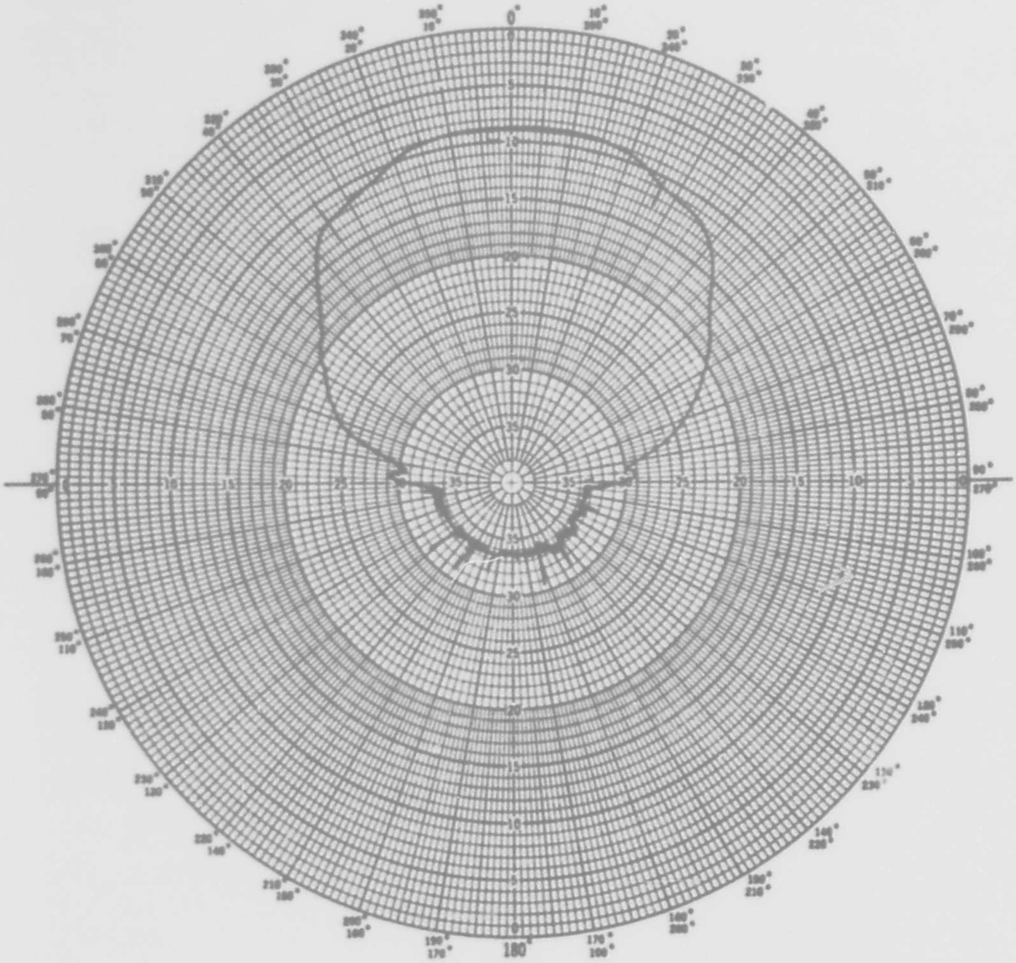


Figure 1-5. Radiation Pattern of Feed Horn Alone

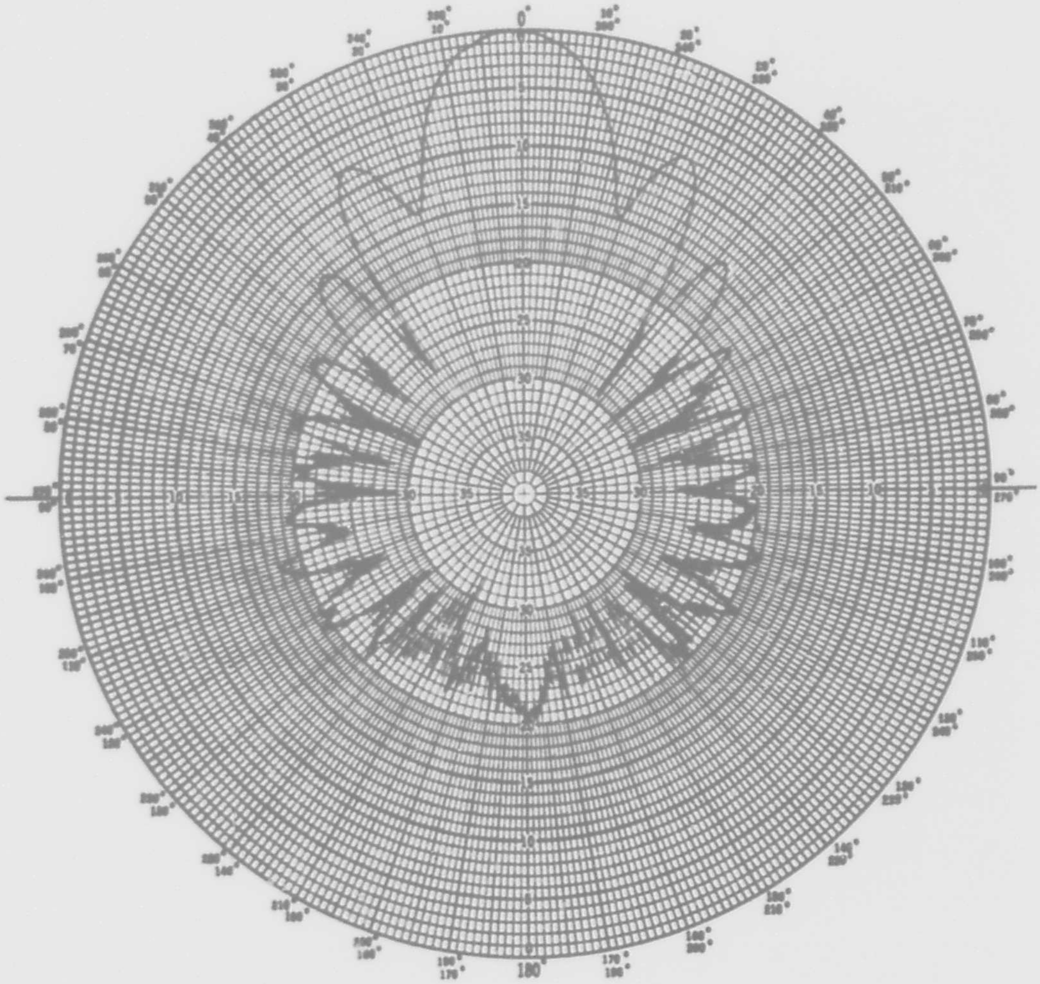


Figure 1-6. Radiation Pattern of Feed Horn with Disc-on-Rod Element

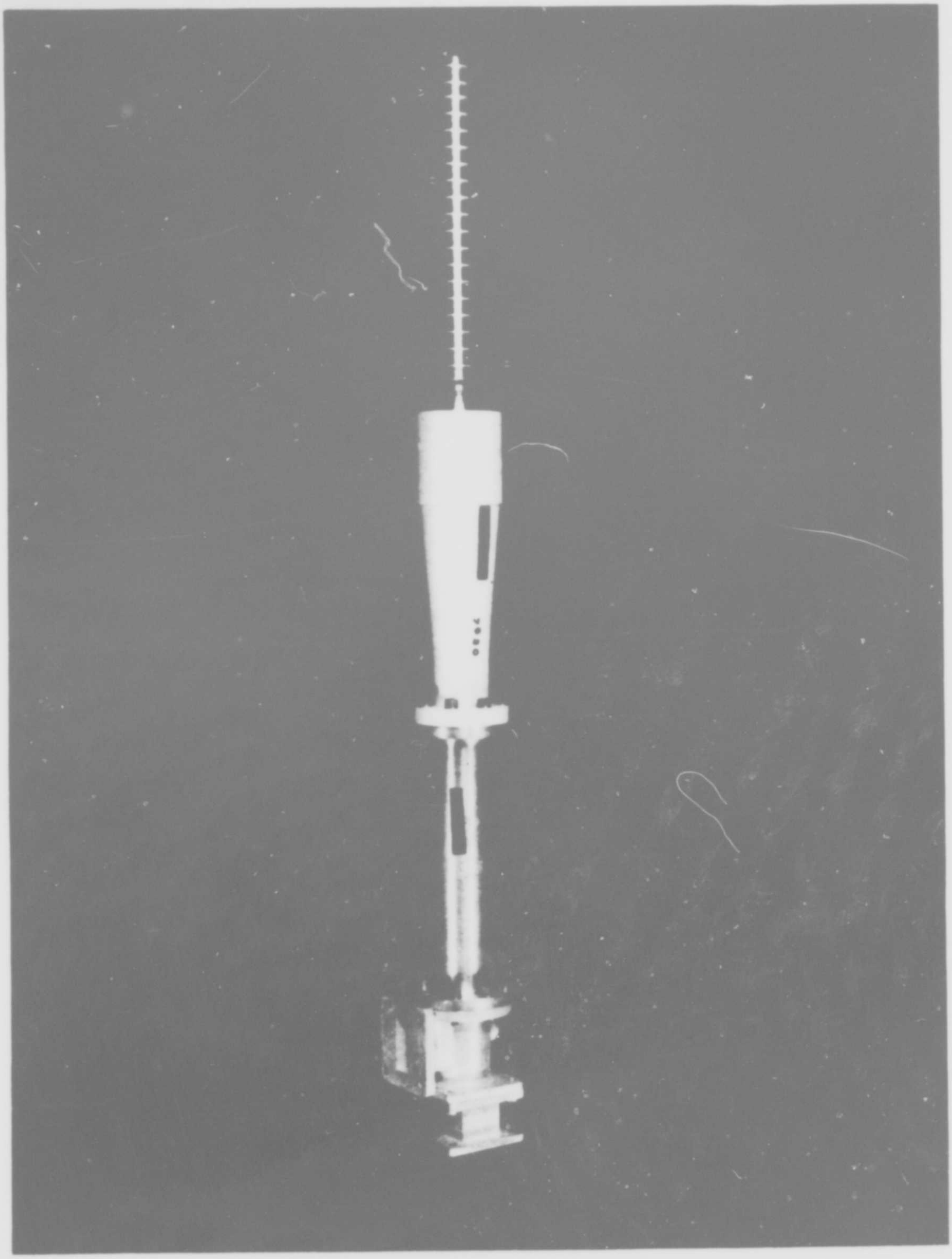


Figure 1-7. Completed Feed Assembly Showing Horn, Disc-On-Rod Element and Dual Mode Transducer

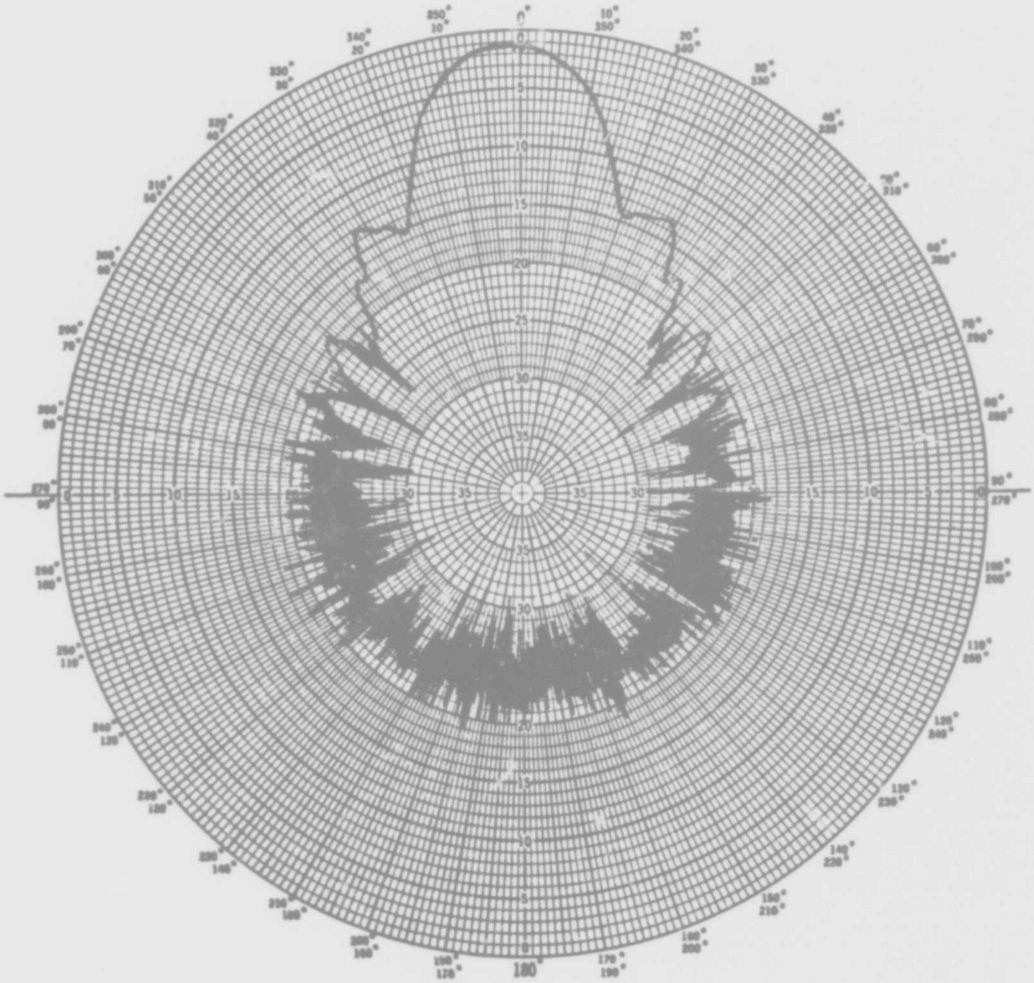


Figure 1-8. Radiation Pattern of Completed 7630 Mc Feed Assembly

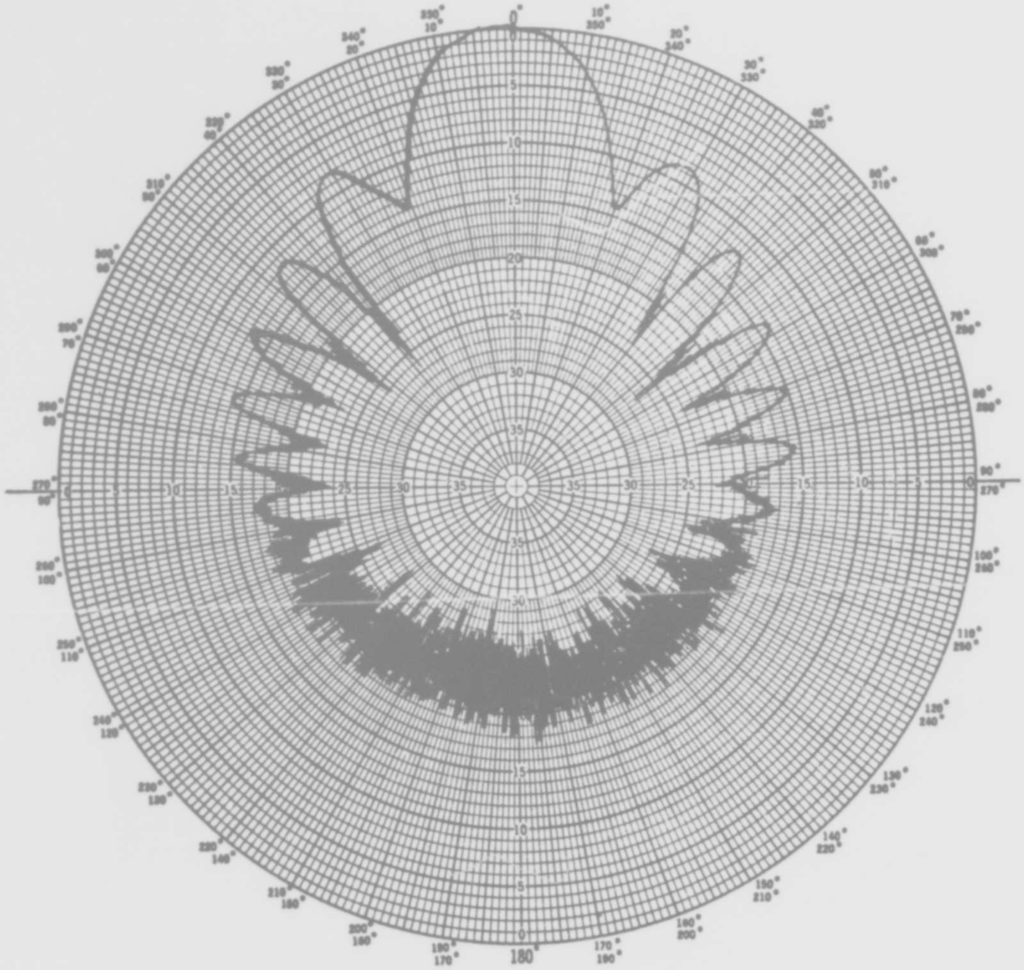


Figure 1-9. Radiation Pattern of Completed 8330 Mc Feed Assembly

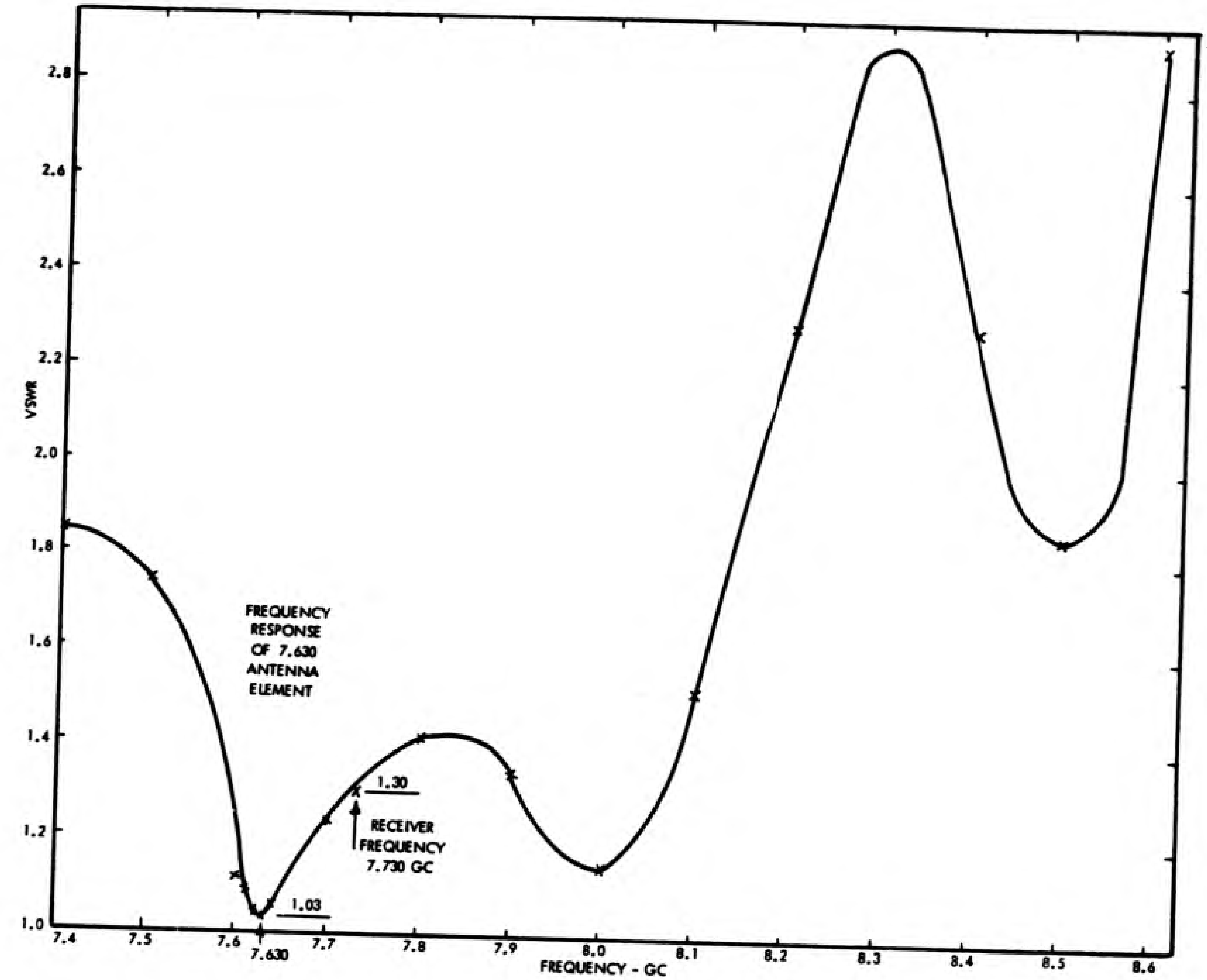


Figure 1-10. Standing Wave Ratio vs. Frequency for Completed 7630 Mc Feed Assembly

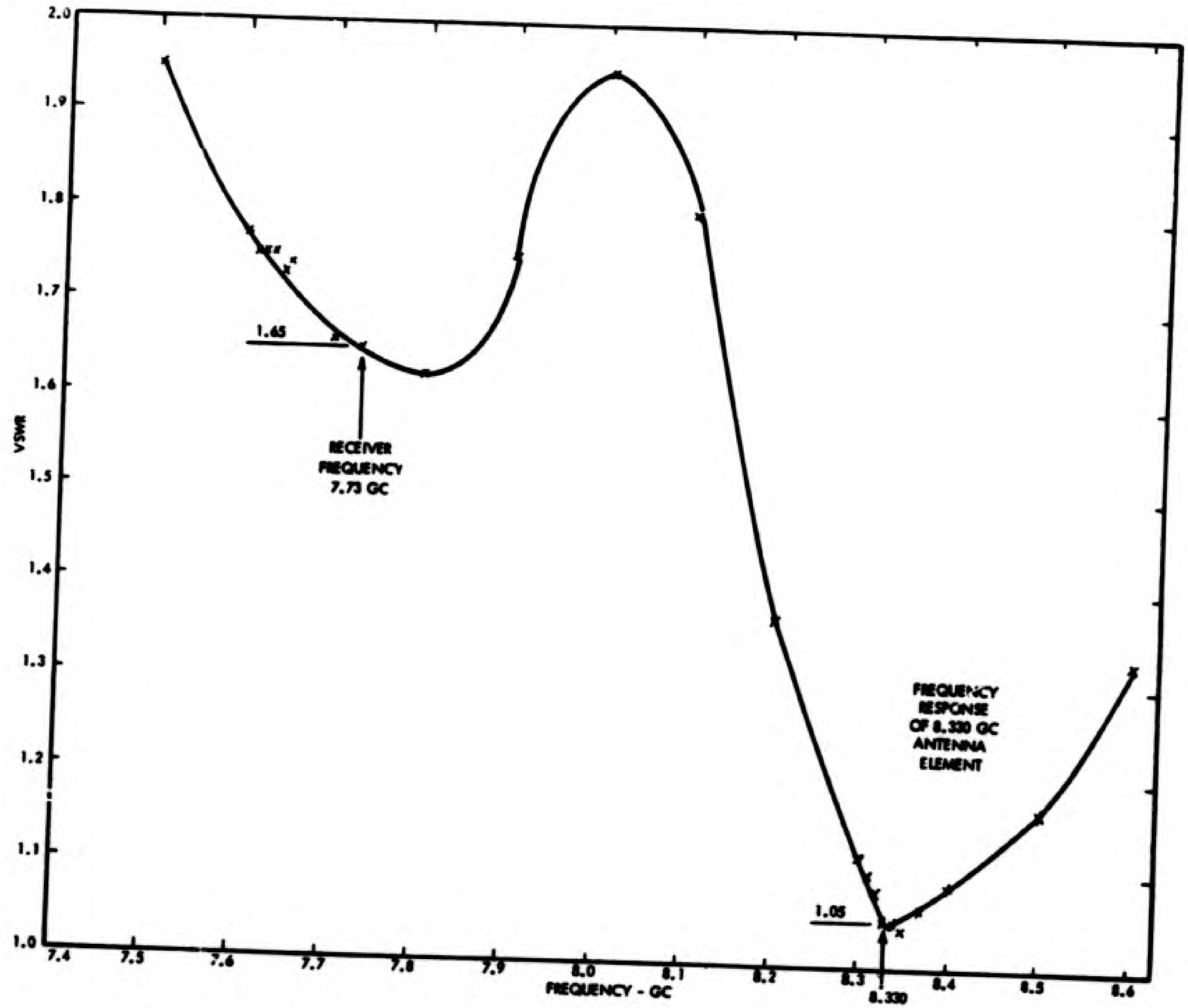


Figure 1-11. Standing Wave Ratio vs. Frequency for Completed 8330 Mc Feed Assembly

between the rectangular guide ports is over 44 db. The dual mode transducer permits the selection of either of two polarizations at right angles to each other and in the case of transmission and reception on crossed polarizations, it performs a diplexing function.

1.4 Receiver Polarization Selector Switch

This waveguide switch permits the X-band receiver to be connected either to the receiver port of the diplexer, in which case reception is accomplished with the same polarization as the transmitter, or to the cross polarized port of the dual mode transducer, in which case reception is on a polarization perpendicular to the transmitter polarization. This switch is remotely controlled by a toggle switch on the control panel in the equipment van. Internal indicator contacts in conjunction with a panel lamp give the operator a positive indication of the waveguide switch operation. This is a four port transfer switch. It will be noted that either the diplexer receiver port or the unused arm of the dual mode transducer may be used for reception but not both at once. The unused port is connected via the waveguide switch to a matched dummy load. The waveguide transfer switch permits the operator to instantaneously select a receiving polarization identical to that of the transmitter or one at right angles to it. It does not permit changing the transmitter polarization. The transmitter polarization can be changed by manually shifting the two waveguide pieces (DC-16-3003-6 and DC-16-3003-5) as indicated in Figure 1-1. This operation reverses the sense of the receiver polarization with respect to the waveguide switch position. For this reason the control panel has not been engraved with an absolute indication of receiver polarization.

1.5 Antenna Equipment Box Fan

A blower has been added to the equipment box to provide an increased air flow through the box when the internal temperature in the vicinity of the diplexer exceeds 100°F. The blower motor is a two phase induction motor. A 2.0 microfarad oil capacitor provides the required phase shift for the second winding. The ventilator to which the blower is attached contains an air filter which should be inspected periodically and cleaned as necessary.

THE MODULATION SYSTEM

This section will deal only with the single sideband and double sideband suppressed carrier modulation equipment and auxiliary devices made necessary by its addition. It will not deal with the FM circuitry as this has not been altered and is described in detail RADC-TDR-62-502.

Figure 2-1 is a reproduction of the exciter block diagram from Page 49 of the TDR-62-502. The modulators added under this Contract (AF-30(602)-2944) are inserted in the signal path immediately following the X128 varactor multiplier. They operate upon the 7630 Mc carrier from the existing Modulator-Exciter. When they are in use, frequency modulation of the carrier is prevented by removing the modulation voltage from the input to the operational amplifier in the exciter. Front panel controls are provided by which the operator may bypass the SSB and DSSC modulators for straight FM operation. Also provided are a switch for selecting SSB or DSSC modulations and a gain control for setting the modulator output power level.

Figure 2-2 is a photograph of the left hand transmitter cabinet prior to the modification and Figure 2-3 shows the same area after the modifications were completed. In Figure 2-2 the box in the upper left hand corner of the cabinet is the X128 multiplier. The horizontal waveguide run at the top of the cabinet opening is the connection between the X128

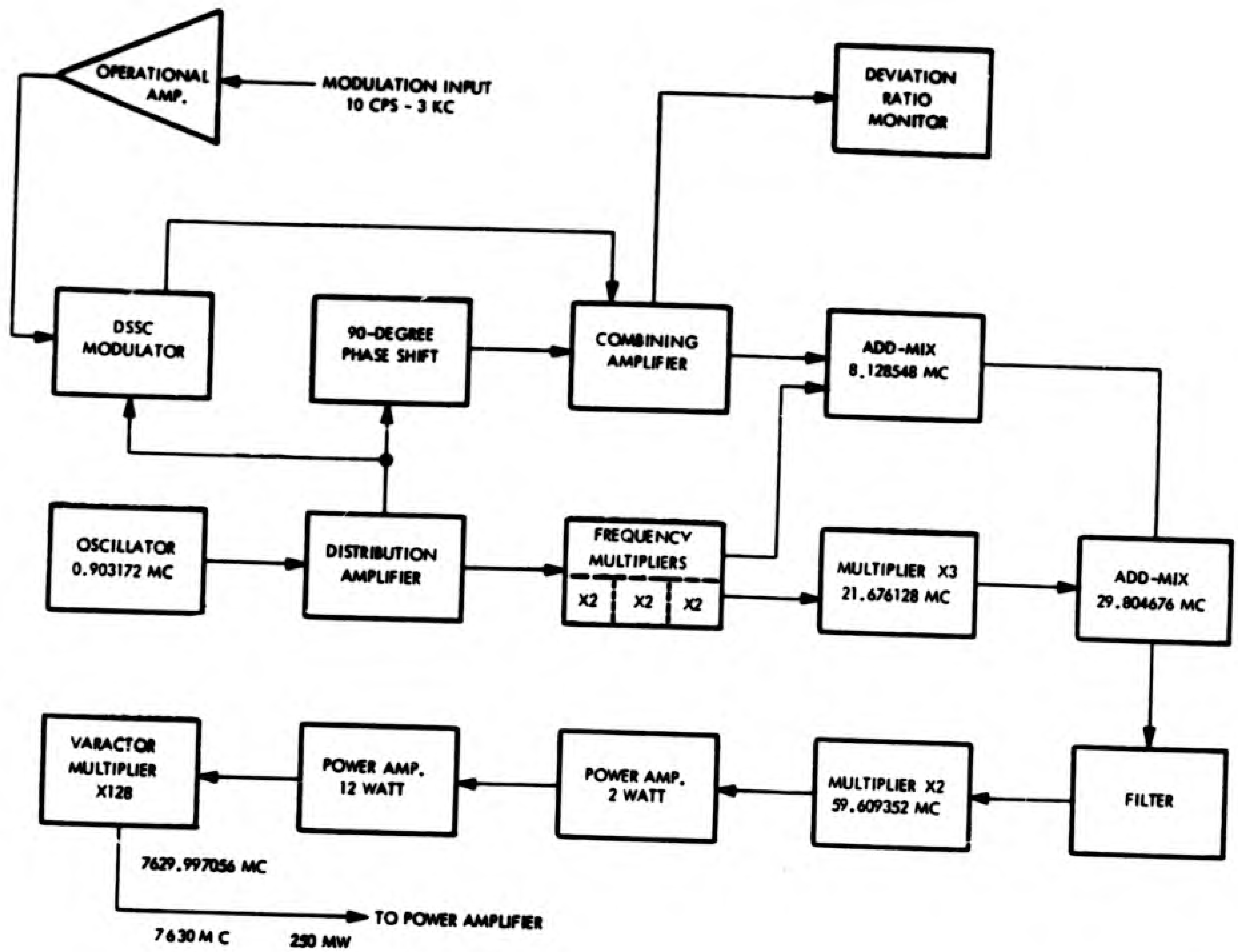


Figure 2-1. Functional Block Diagram of Exciter

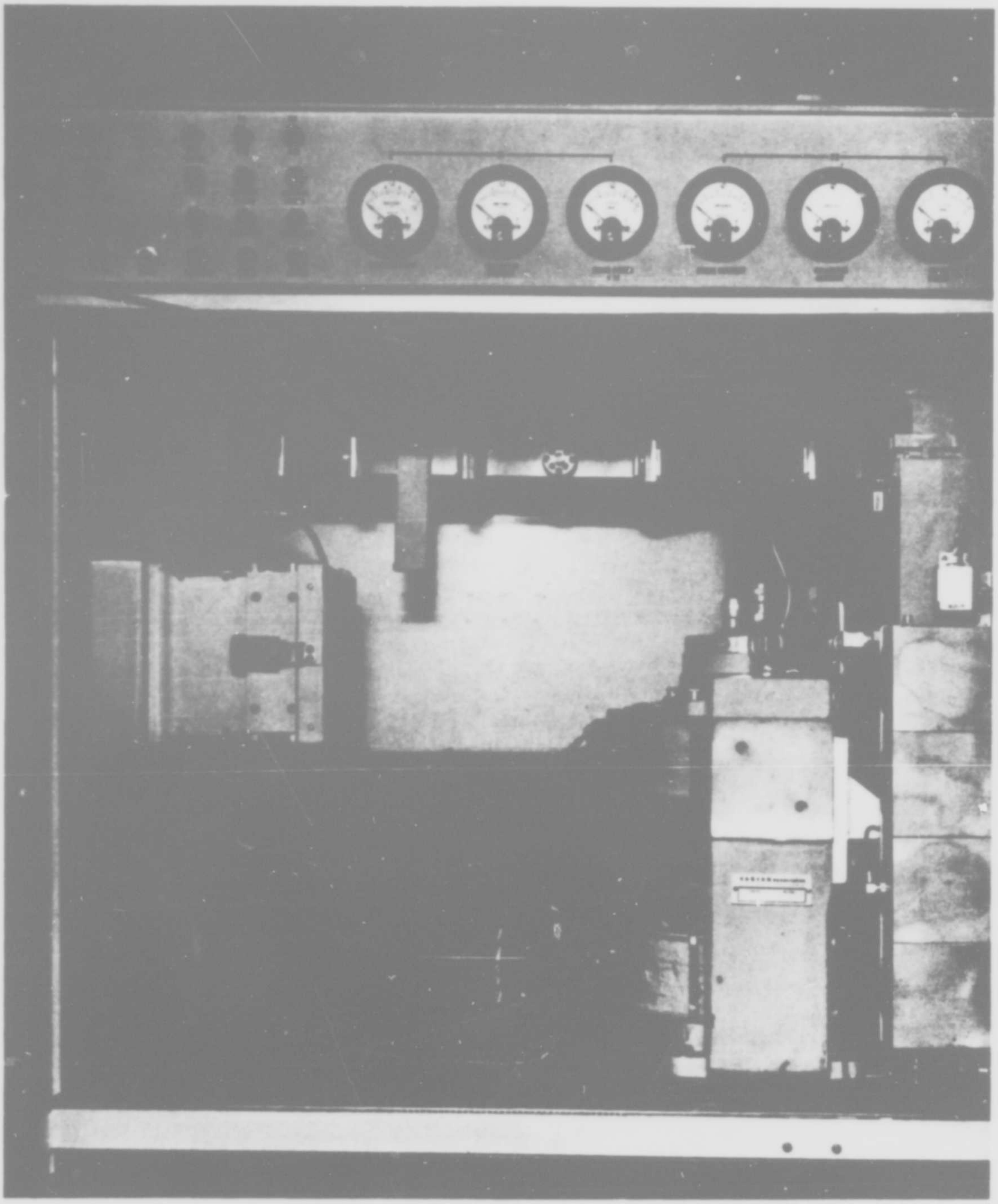


Figure 2-2. Left Hand Transmitter Cabinet Before Modification

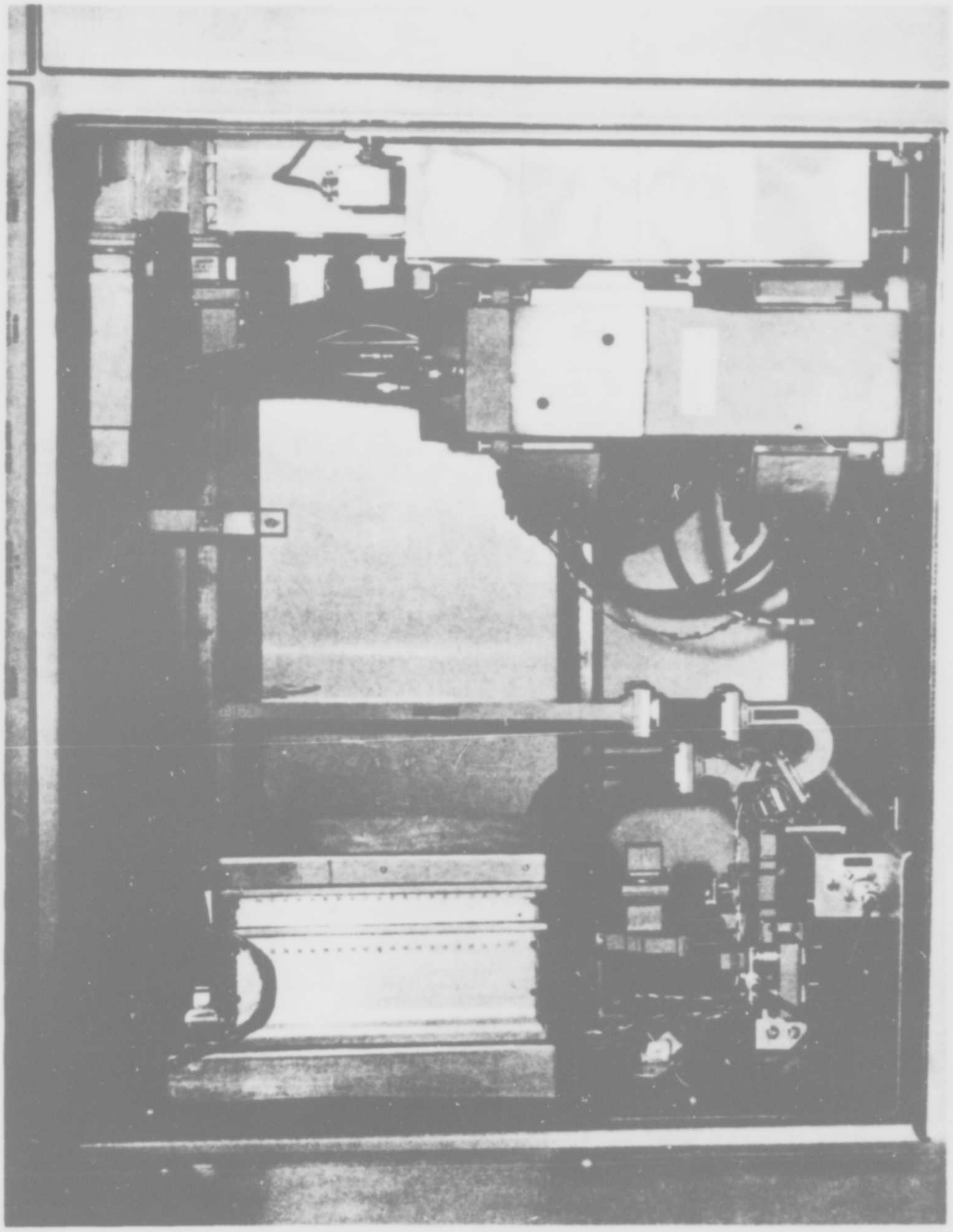


Figure 2-3. Left Hand Transmitter Cabinet After Modification

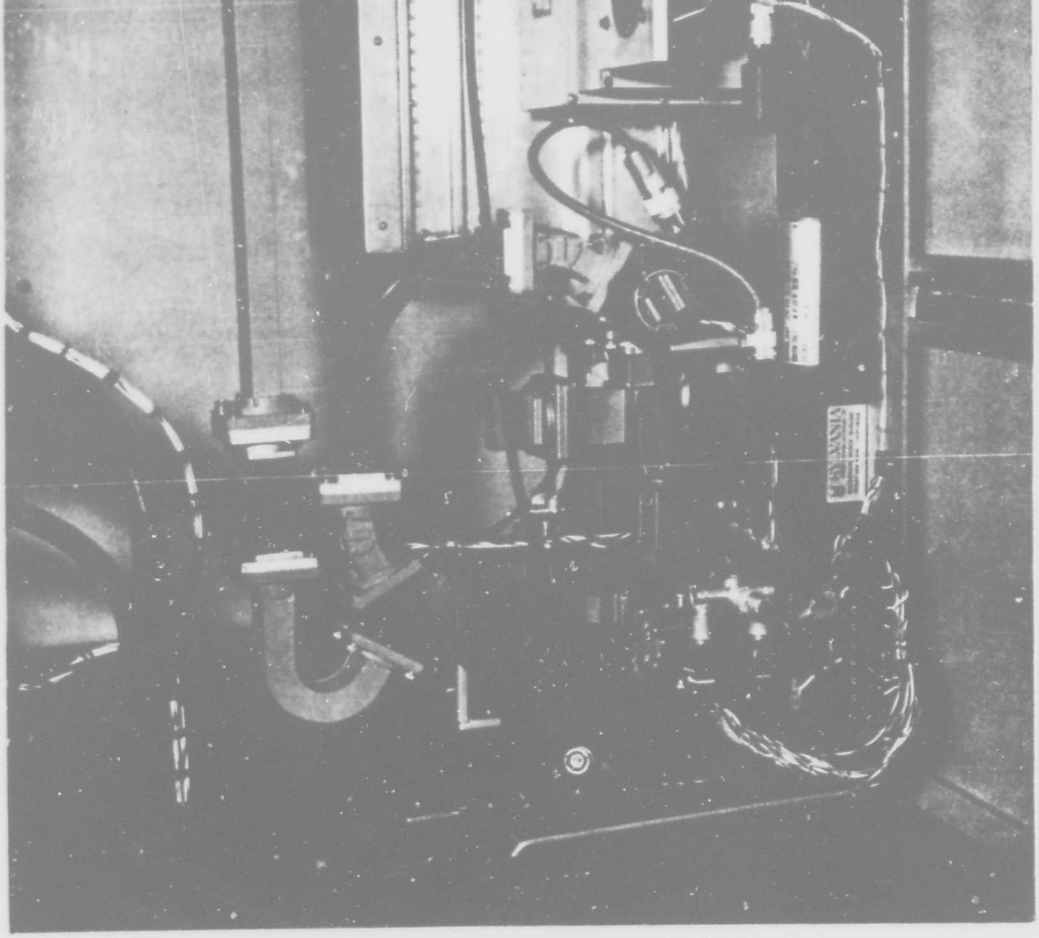


Figure 2-4. Rear View of Modulator Assembly

multiplier and the power amplifier. In Figure 2-3 the X128 multiplier box has been rotated 90°. The horizontal waveguide at the top of the photograph is the output connection between the modulator and the power amplifier. The vertical waveguide near the center of the photograph connects the output of the X128 multiplier to the modulator assembly. Figure 2-4 is a rear view of the modulator assembly showing somewhat more detail. The signal flow and switching arrangement can be traced on this photograph. When the panel controls are set to "FM", the signal leaving the X128 multiplier at the upper right of the photo passes through the 90° E-plane bend, flexible coupler and "J" bend to the pole of the solenoid operated waveguide switch, (triangular piece at lower left). It passes through the switch, the 30° E-plane bend, the 90° E-plane flexible coupler and another 30° E-plane bend to port 2 of a second waveguide switch. From the pole of the second switch, the signal goes to a waveguide-to-coax adaptor to the input of the traveling wave tube. The output of the tube is coupled through a coaxial cable and another adaptor to the horizontal output waveguide at the top of the photo. When the panel control is set to "AM" the path between the two waveguide switches is via the DSC-SSB modulator assembly at lower right center of the photo. Visible just above the TWT is a small blower used to cool the collector.

2.1 <u>Operation - Design Notes</u>

The need to maintain the transmitter center frequency accurate to within one part in ten to the ninth or better, the non-linear amplitude response of varactor frequency multipliers and the desire to employ the existing equipment to the fullest practical extent led to the modulator configuration shown in simplified block diagram form in Figure 2-5.

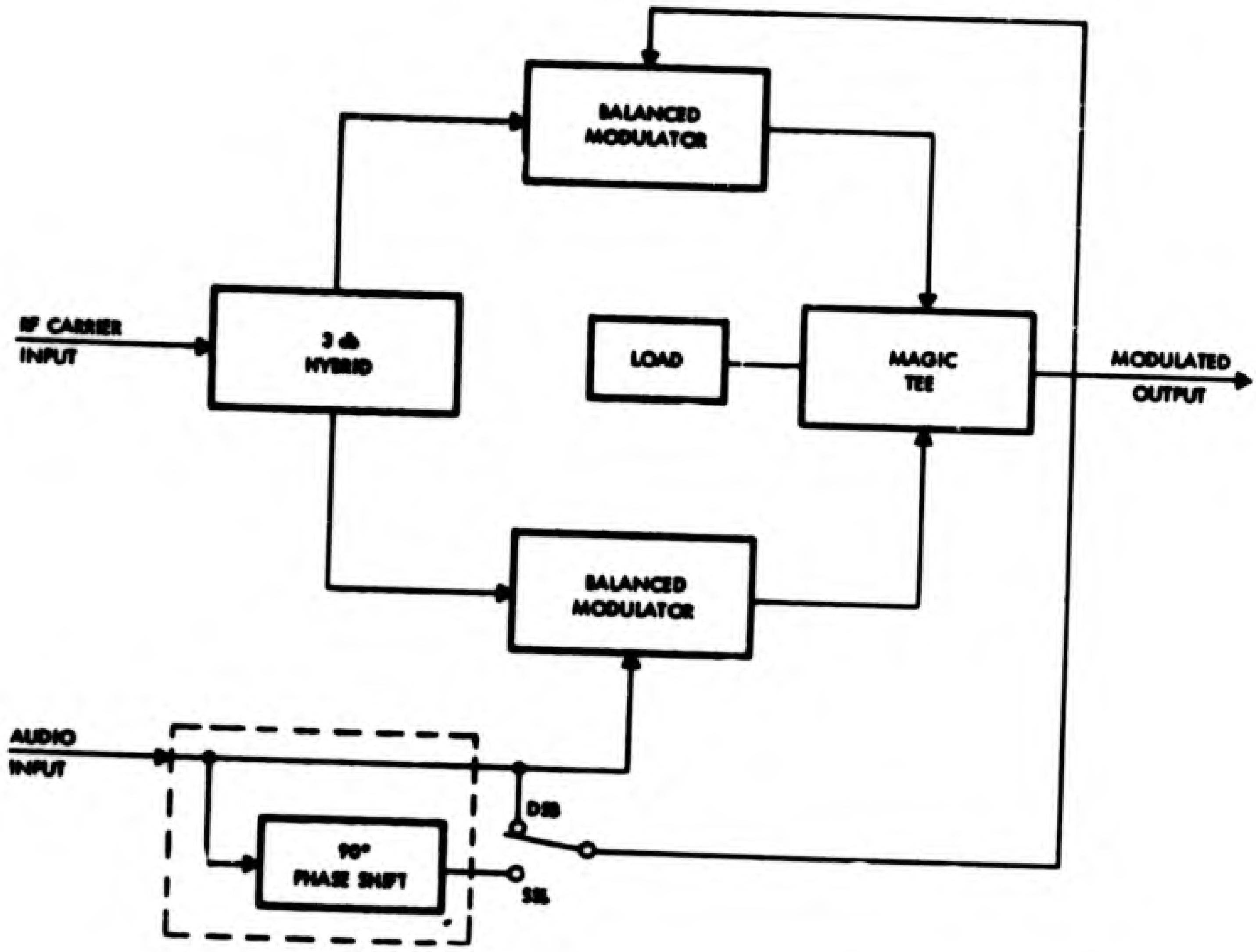
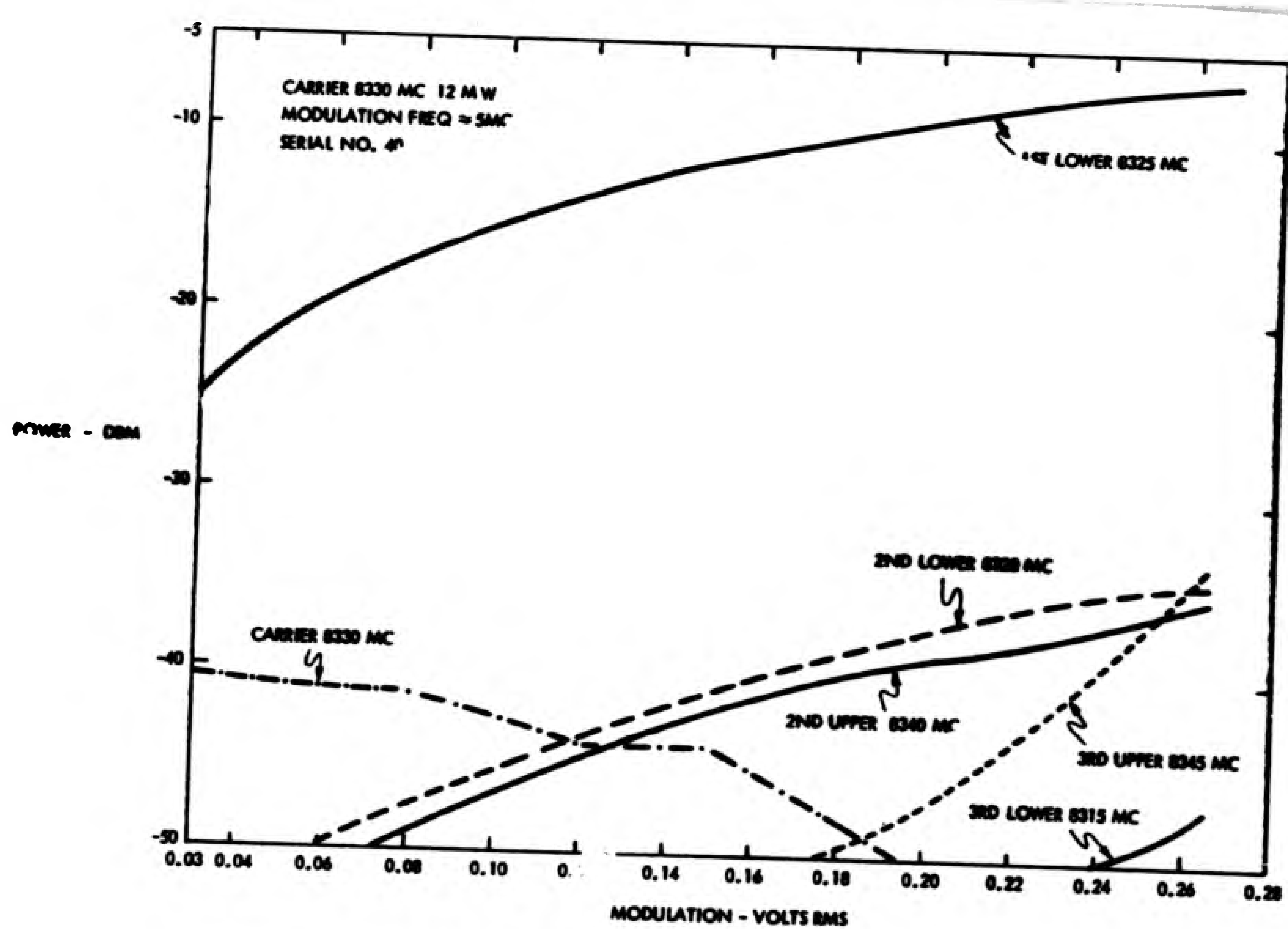


Figure 2-5. Implementation of Phase Discrimination SSB Modulation Technique

This system employs the same time base and the same multiplier chain as the FM modulator and operates on the carrier at the X-band transmission frequency. Because both the X128 multiplier output and the power amplifier input were in WR112 waveguide it was decided to build the modulator in this size waveguide also.

The phase discrimination method of single sideband modulation is employed in this system. In this method the outputs of two balanced modulators are combined in such a way that one sideband frequency is cancelled. In order to accomplish this cancellation it is necessary that the two balanced modulator outputs be equal and that both the carrier voltage and the modulating voltage of one balanced modulator be 90° out of phase (in phase quadrature) with those of the other balanced modulator. Under these conditions one of the sidebands of each modulator will be in-phase with its counterpart from the other modulator whereas the other sideband outputs will be 180° out of phase (phase opposition). Summing the outputs of the two balanced modulators produces the required single sideband output. The output of the balanced modulators contains no carrier component if the balance is perfect. Division of the carrier frequency power into two equal amplitide parts 90° apart in phase is accomplished using a 3 db short slot sidewall hybrid junction. Two magic tee balanced mixers are used as balanced modulators. Summation of the outputs of the two balanced modulators occurs in a third magic tee junction. The degree of carrier suppression obtainable with the modulator depends upon how well each balanced modulator is balanced. This balance is influenced mostly by the crystal holders and crystals. Merely inserting "matched crystal pairs" will not produce good carrier suppression. In order to be able to compensate for crystal and holder irregularities four tuning screws have been added to a very short guide section ahead of each balanced mixer.



THE TUNING SCREWS IN FLA[illegible]S BETWEEN THE MIXER CRYSTAL MOUNTS AND HYBRIDS WERE ADJUSTED FOR BEST CARRIER AND SIDE[illegible]D SUPPRESSION AT 0.265 MODULATION VOLTS. THESE WERE NOT TOUCHED FOR REST OF MODULATI[illegible]LTAGES. THE MODULATION VOLTAGE IMPRESSED ON THE MIXER WAS BALANCED FOR BEST [illegible] SUPPRESSION AT EACH MODULATION VOLTAGE.

THE 1ST ORDE[illegible]AND WAS BELOW -50 DBM FOR EACH READING EXCEPT AT 0.180V WHERE IT WAS -48 DBM. [illegible]R ORDER SIDEBANDS NOT NOTED WERE BELOW -50 DBM.

Figure 2-6. Performance Characteristics of MDL Model XL SSB Modulator

2.2 Balancing the Modulator

It may become necessary to readjust the modulator balance if diodes are replaced or if the carrier frequency is changed. Two problems are encountered if an attempt is made to balance the modulator directly using modulation frequencies in the audio range. These concern the difficulties encountered in obtaining a 90° phase shift at audio frequencies and in observing sidebands only a few thousand cycles away from an 8000 megacycle carrier. Because the mixers are very broadband with respect to the normal baseband it is permissible to do the balancing at a modulation frequency of a few megacycles. This permits observation of the first several sidebands on generally available spectrum analyzers and allows the use of an electrical quarter wave-length coaxial cable as a phase shifter. Phase trimming can be accomplished by varying the modulation frequency. Using the test set-up shown in Figure 2-8 proceed according to the following sequence of steps.

1. If the modulator diodes have been removed, make sure that each balanced mixer contains a forward and reversed crystal in the corresponding holders. It will be impossible to effect a balance unless each mixer has a forward and a reversed crystal. Inter-changing a forward and reverse diode of a pair will result in production of the opposibe sideband and suppression of the desired one. The same is true of interchanging the modulation leads to the mixers.

2. Remove waveguide sections below the power set attenuator AT 1 and attach a matched X-band power meter padded to measure from 1 to 200 milliwatts. Adjust AT 1 to get a power level between 10 and 15 milliwatts. Remove the power meter and replace the waveguide section(s) that were removed.

3. Set the modulation switches to "AM" and "SSB" respectively. Connect the TWT output through suitable matched pads to the spectrum analyzer. Enough padding must be provided to prevent overloading the analyzer when the TWT output is 1.0 watt. Note that overloading the analyzer may result in a presentation of acceptable appearance but one in which additional spectral components are present due to intermodulation products produced in the analyzer front end. Use only enough power to the analyzer to obtain a reasonable signal to noise ratic in the presentation.

4. With the normal audio cables to the modulator in place but with no modulating voltage applied, locate the carrier on the analyzer screen. Adjust the modulator tuning screws to minimize the carrier frequency output.

5. Remove the audio cables to the modulator and connect the 5 Mc signal generator, the two adjustable attenuators and the 90° phase shift line. Using a good RF voltmeter, adjust the RMS voltage at each mixer to 0.25 volts. Maintain this level. Observe the appearance of sidebands 5 Mc above and below the carrier. Adjust the modulation frequency to minimize the unwanted sideband while maintaining equal voltages at the two mixers. Adjust (retouch) the modulator tuning screws to minimize the carrier and the unwanted sideband -- as well as the higher order sidebands. With careful adjustment, it is possible to obtain a suppression of all responses of at least 30 db below the desired sideband.

6. Either the upper or lower sideband can be suppressed. The relative phasing of the carrier and modulation frequencies to the two mixers determines which sideband is suppressed. Reversing the modulation

cables will interchange sidebands. The selection of the sideband to be suppressed must be compatible with the receiver to be used to receive the signal. Use of the wrong sideband will invert the received spectrum.

7. Return the equipment to the normal configuration.

2.2.1 Changes - Alterations

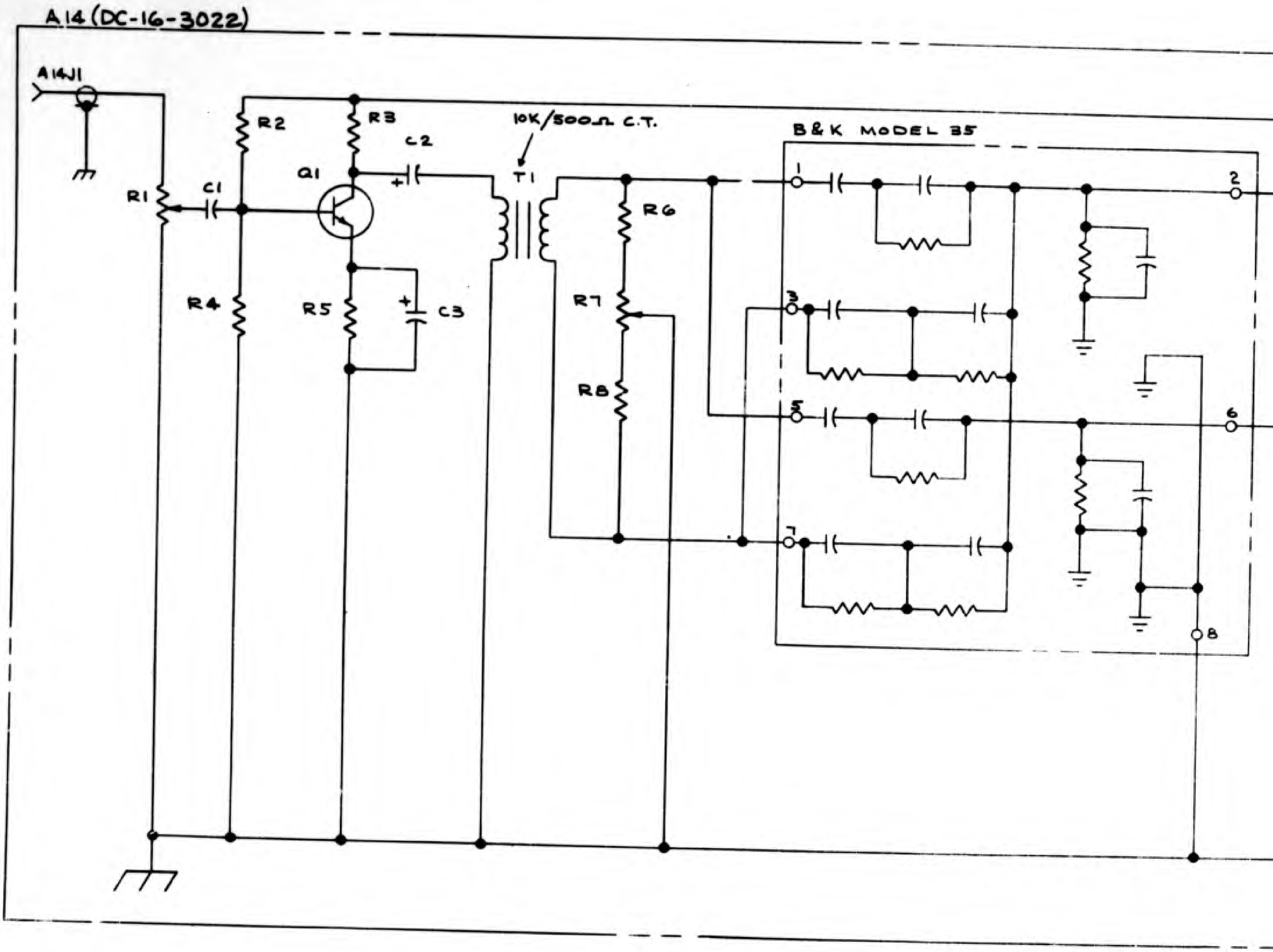
After the photograph of Figure 2-4 was prepared, two waveguide attenuators were added to the modulator assembly. These are Waveline Model 513, 20 db power set adjustable attenuators made in WR112 size guide. One unit has been installed in the vertical guide at the left of the photo just above the short piece of flexible guide. It provides a means of adjusting the power level to the modulator. The second unit is installed where the curved section of flexible guide is shown in the photo. It provides a means of adjusting the attenuation of the modulator bypass path so that it may be made approximately equal to the modulator insertion loss. The input power set attenuator is designated AT-1, the bypass path attenuator is AT-2.

2.3 Audio Phase Splitter (Quadrature Amplifier) DC-16-3022

This unit splits the audio information signal into two parts having equal amplitudes and which are 90° apart in phase. It does this for all frequencies in the band from 300 to 3000 cps. Balancing and gain adjustments have been provided. Readjustment of the balance control (R-7) in the field is not recommended. R-23 permits equalization of the two output voltages. R-1 is an overall gain set adjustment. (See Figure 2-7.)

2.4 Traveling Wave Tube Amplifier

The insertion loss or (conversion loss) of the single sideband modulator is approximately 10 db. The spectral composition of the output is best


A14 (DC-16-3022)
A14J1
R1
C1
R2
R3
Q1
C2
R4
R5
C3
10K/500Ω C.T.
T1
R6
R7
R8
B&K MODEL 35
1
2
3
5
6
7
8

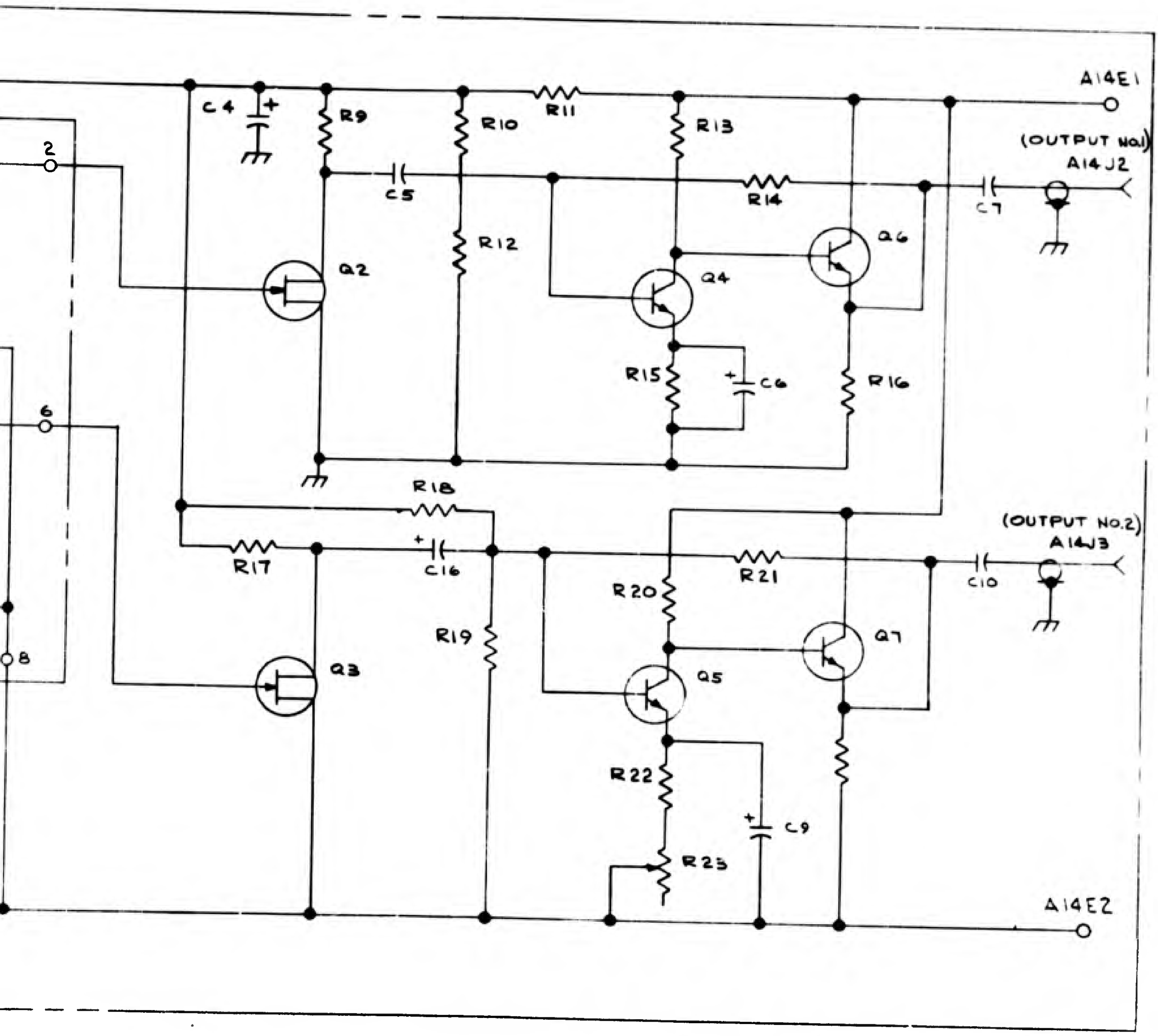


Figure 2-7. Quadrature Amplifier Schematic Diagram

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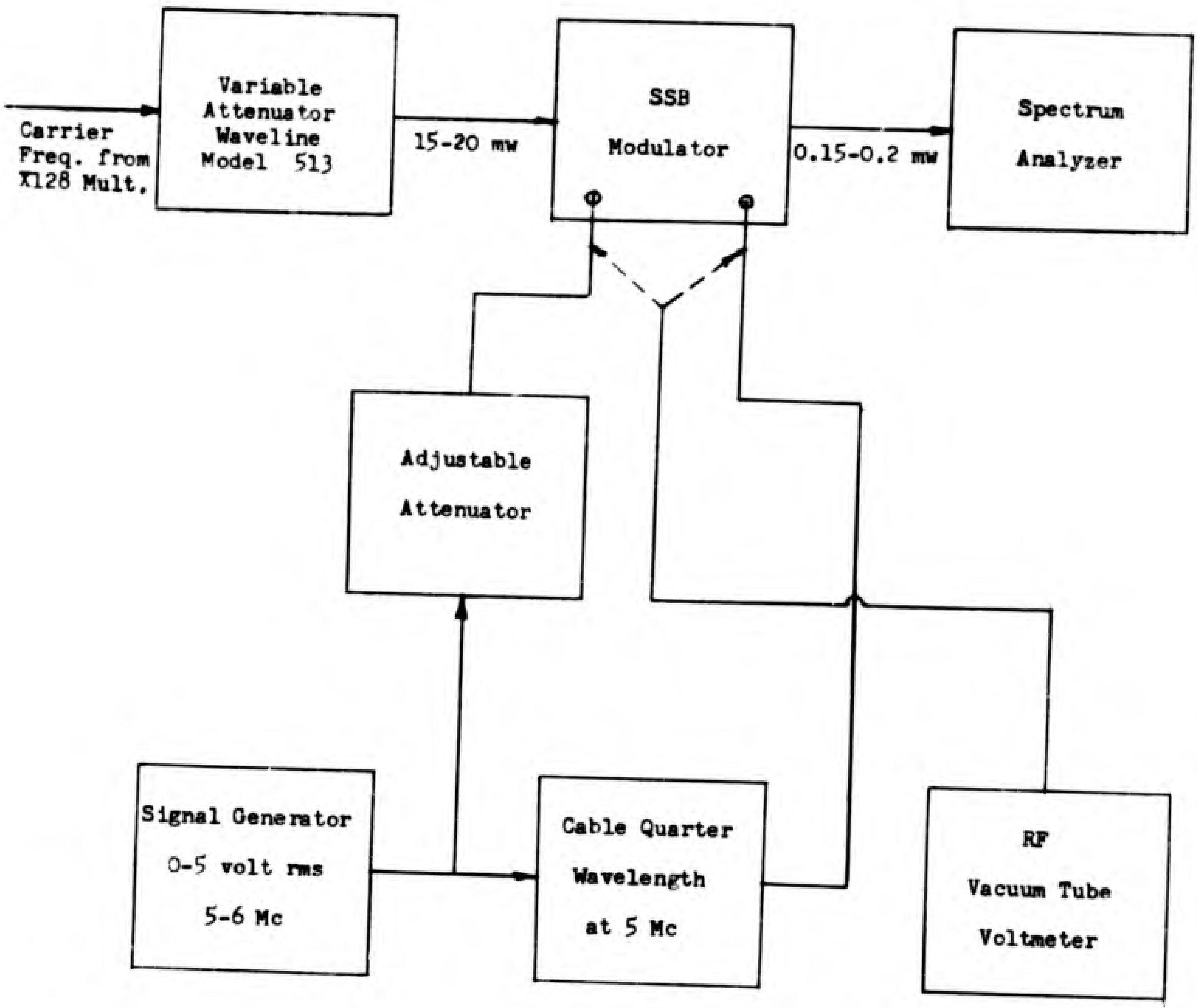

Test-Set-Up For Modulator Adjustment

Figure 2-8

when the carrier power level is between 10 and 15 milliwatts. Under these conditions, the modulator output is typically between 0.2 and 1.0 milliwatt.

The power amplifier is designed to operate from a 200 mw driver. The modulator requires an amplifier having a nominal gain of 30 db. The TWT provides this gain.

The tube used is a Sylvania SYT-4273C. It has a nominal 1.0 watt output and over 30 db gain in the band from 7.0 to 12.4 Gc. Each tube is marked with the lead identification and the proper operating voltages. Operating voltages for serial number 51576 are:

Heater	6.3 V	at	1.06	A
Grid 1	+47.0 V	at	0.52	ma
Helix	+2475 V	at	2.58	ma
Collector	+2475 V	at	15	ma

The power supply for the traveling wave tube is a commercial unit (EH Research Laboratories, Inc. Model 510). It is located in the auxiliary equipment rack No. 18. Front panel controls are self-explanatory. The RF gain control adjusts the grid 1 voltage and directly affects the cathode current. Cathode current for the SYT-4273G should not exceed 16 milliamperes. The vendor's instruction manual has been supplied under separate cover and should be consulted for servicing information. The collector of the tube is cooled by a small blower mounted directly over it. The blower motor is wired through the traveling wave tube power supply switch. It will run whenever the power supply is "on".

The traveling wave tube should be operated unsaturated whenever single sideband or double sideband signals are present to avoid severe distortion. Saturation input power will vary from tube to tube. The input power should always be less than one milliwatt for linear operation.

3.0 THE TIME BASE SYSTEM

The Time Base System (Figure 3-1) consists of the stable station clock, the very low frequency phase comparison receiver, the 0.903172 Mc synthesizer, the X-band receiver local oscillator synthesizer, the exciter offset frequency generator, and the klystron phase servo-reference signal generator.

3.1 Stable Clock (Frequency Standard)

The heart of the Time Base System is the stable clock and the very low frequency phase comparison receiver. The stable clock employed in the TPSC Terminal is a crystal controlled standard made by Frequency Electronics, Inc. and designated as Model FE-1000A. This unit has a drift rate of 1 or 2 parts in 10^{10} per day. The short term stability (under one second) is 5 parts in 10^{11}. Outputs at 100 Kc, 1.0 Mc and 5.0 Mc are available at a level of one volt RMS across 50 ohms. The clock is powered by an internal DC power supply and an internal battery pack. It will automatically continue to operate on the internal battery pack for 20 hours following interruption of the 115 volt power. No frequency change or phase discontinuity occurs at switchover.

A front panel frequency trimming adjustment is provided. A digital dial on this adjustment permits setting frequency increments of one part in 10^{10} into the clock. Dial accuracy deteriorates at either end of its range.

The vendor's operating and service data for this unit are being supplied under separate cover.

3.2 VLF Phase Comparison Receiver

In order to maintain the absolute frequency accuracy of the frequency standard to one part in ten to the ninth over periods greater than one

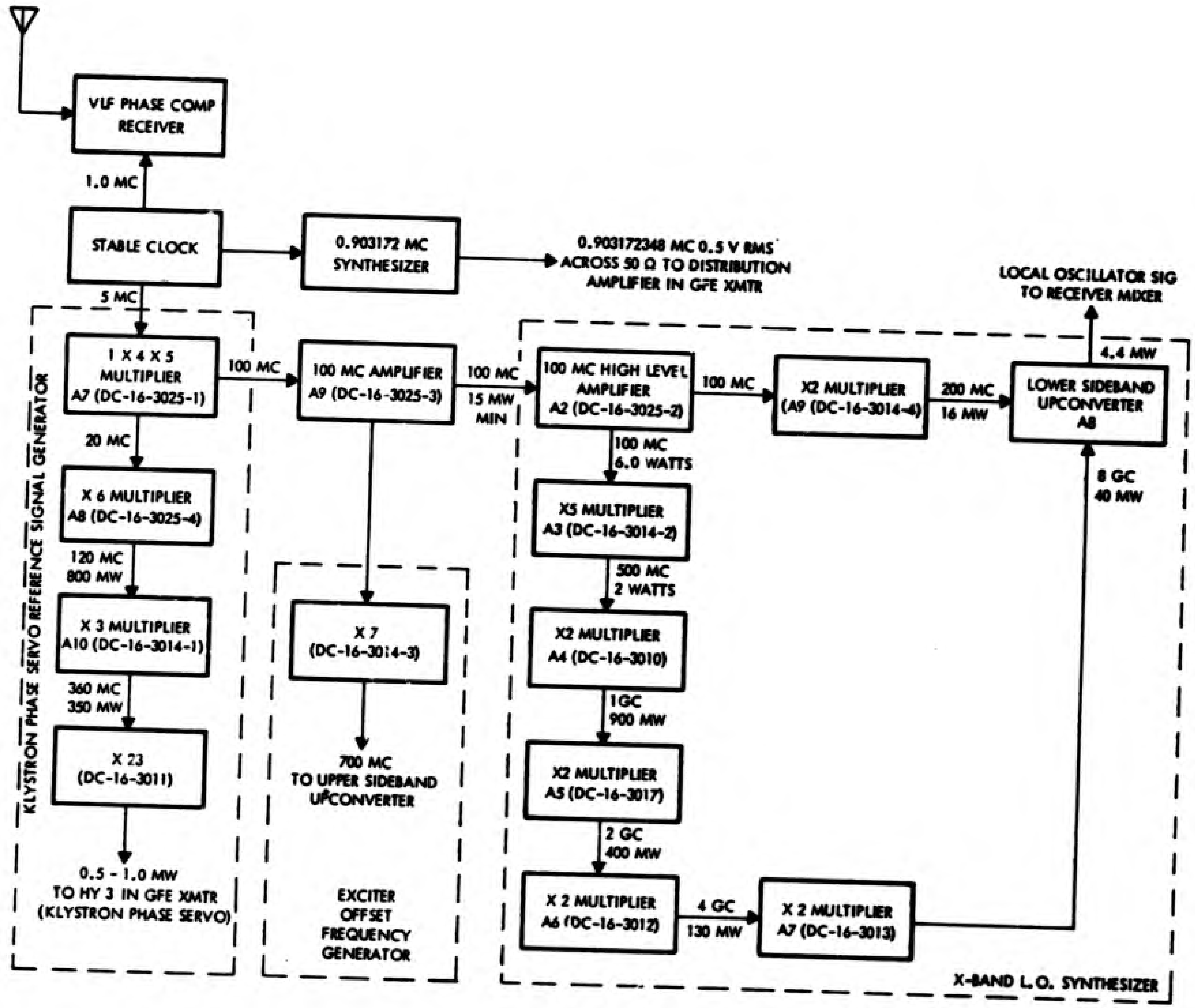

Figure 3-1 Time Base System Block Diagram

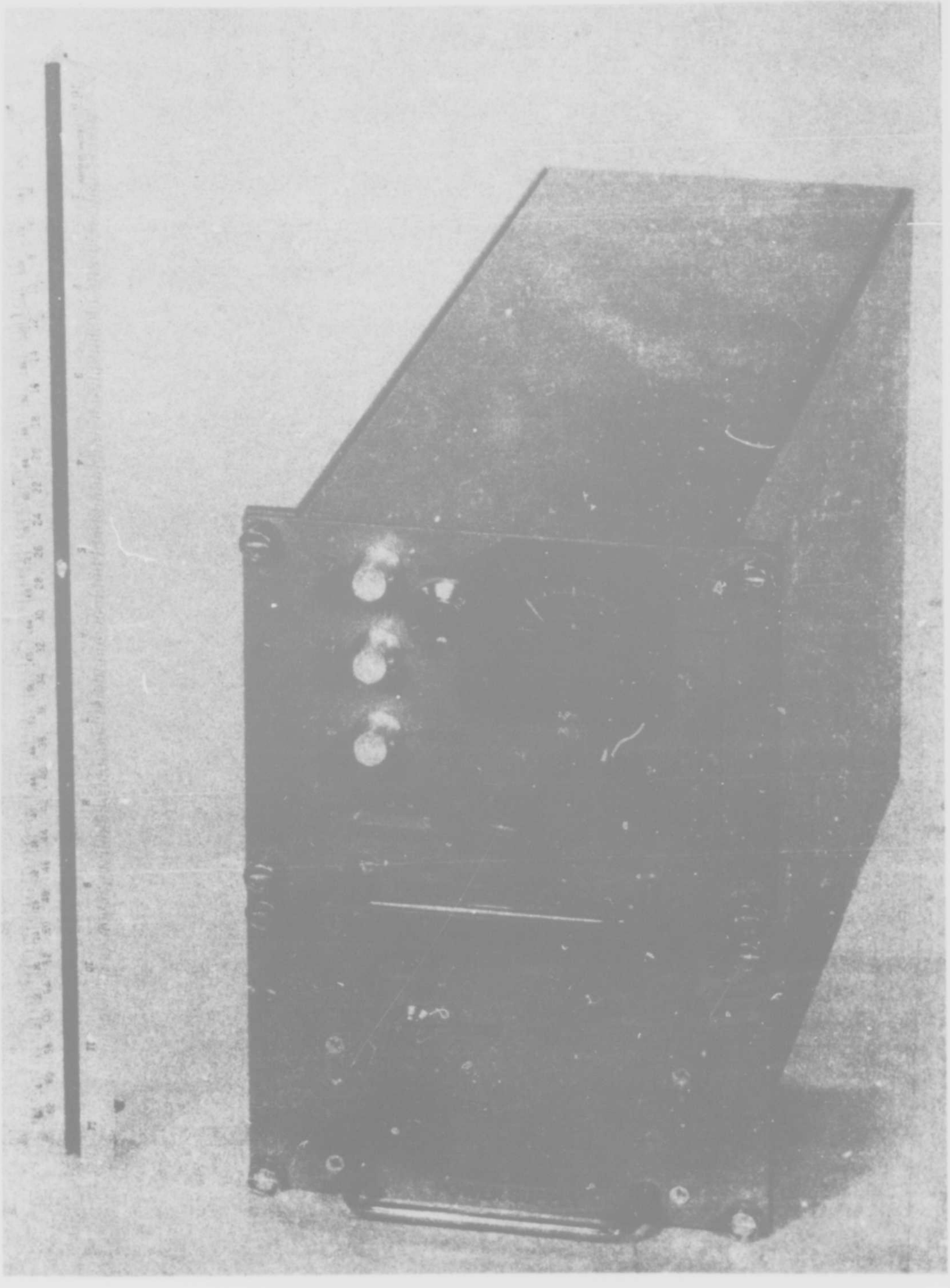

Figure 3-2. FE 1000A Frequency Standard

week, it is necessary to compare it with some primary standard and make periodic corrections. Ordinarily, the drift rate decreases with aging so that corrections are required less frequently as the standard ages.

The VLF comparison receiver permits a continuous comparison of the standard frequency against VLF standard frequency stations operated by various agencies. The receiver installed at the TPSC Terminal is a single channel receiver tuned to NPG/NLK Jim Creek, Washington (18.6 Kc). Three more channels can be added by purchasing additional plug-in RF heads, if necessary. With additional RF heads installed, any one of four channels can be selected by means of a front panel selector switch.

At 18 Kc one part in 10^9 is 18 (10^{-6}) cycles per second (18 millionths of a cycle per second). This is equivalent to 6.48 (10^{-3}) degrees per second. These magnitudes indicate the very small increments which must be measured in order to compare frequencies to a precision of one part in 10^9.

The TPSC Terminal VLF comparison receiver continuously compares the phase of the VLF reference station with that of the local frequency standard. The difference in phase is presented as a lateral displacement on a strip chart recorder built into the receiver. The instantaneous position of the trace represents the instantaneous phase difference. The displacement of the trace with time represents the total phase change over the interval. The slope of the trace, slope $= \frac{\Delta \phi}{\Delta t} = \Delta f$, is the average frequency difference. The sensitivities and scale factors are such that a frequency offset of one part in 10^{10} can be easily read from the recording.

During the periods when the propagation path to the VLF reference station is passing from daylight into darkness and vice versa, the propagation path length is undergoing a change due to the change in altitude of the reflective layers in the ionosphere. This produces a doppler shift in the received frequency of the VLF station. The shift so produced may be a few parts in 10^9; consequently, it is not desirable to make frequency comparisons during the hours of the diurnal shift in the ionosphere altitude.

It is entirely possible to design a closed loop automatic frequency correcting system. Such a system would, however, produce unwarranted perturbations in the standard frequency due to the diurnal shift and/or other temporary disturbances. It is preferable, therefore, to manually interpret the recordings and to make periodic manual corrections to the standard.

Operating instructions, servicing data and parts lists are contained in the vendor's manual supplied separately.

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U. S. NAVAL OBSERVATORY
WASHINGTON 25, D.C.

SCHEDULE OF CONSTANT FREQUENCY TRANSMISSIONS
ON VLF FROM U. S. NAVAL RADIO STATIONS

Call	Location	Frequency, kc/s	Radiated Power, Kw	Maintenance	Special Trans.; minutes of each hour
NAA	Cutler, Maine	17.8 (with FSK)	1000	As required	30 to 40 (1)*(4)
NBA	Balboa, C.Z.	24.0	30	(1300 to 2100 UT) (each Wed.)	4 to 10 (2)
NPG/NLK	Jim Creek, Wash.	18.6	250	(1600 to 2400 UT) (each Thurs.)	40 to 50 (1)
NPM	Lualualei, Hawaii	19.8	100	(0900 to 1700 UT) (each Wed.)	50 to 60 (3)
NSS	Annapolis, Md.	21.4	100	As required	50 to 60 (1)

*(1) First 4 minutes, one 200-millisecond pulse every 2 seconds.
Next 3 minutes, key down.
Next 3 minutes, key up.

(2) First 3 minutes, key down.
Next 3 minutes, key up.

(3) First 4 minutes, one 100-millisecond pulse every second.
Next 3 minutes, key down.
Next 3 minutes, key up.

(4) FSK for 1/2 hour ending at the odd hour.

Figure 3-3A

NOTES

1. The carrier frequencies of the 5 U. S. Navy VLF transmitters listed in the above schedule have been stabilized with high precision and may be used for calibration of frequency to about 1 part in 10^{10} in a 24-hour interval.

2. The modulation is normally, CW. However, other types may be used at times.

3. Time pulses of high precision are transmitted continuously from NBA. Pulses transmitted from other stations are provided for special tests; these are not time signal pulses.

4. The radiated powers listed are the nominal values. Variations may occur because of operating conditions or special tests.

5. Changes in the schedule may be made from time to time. Notice of these changes is sent via Time Service Announcements as rapidly as is practicable.

T. S. Baskett
Superintendent

PROPOSED SCHEDULE OF NBS VLF AND LF BROADCASTS

FROM SITE NEAR FT. COLLINS, COLORADO

	WWVB	WWVL
Carrier Frequency	60 Kc/s	20 Kc/s
Coordinates	40° 40' 29" N 105° 02' 39" W	40° 40' 51" N 105° 03' 02" W
Carrier Frequency Offset from C_s	-130×10^{-10} 1963 -150×10^{-10} 1964	-130×10^{-10} 1963 -150×10^{-10} 1964
Program	Time Signal Modulation (DSB): Second markers - A pulse of 5 cycles of 1 Kc/s. Minute markers - A pulse of 1 Kc/s for 1/4 sec. duration. Hour markers - A pulse of 1 Kc/s for 3/4 sec. duration.	1 sec. alternate transmission of 20 Kc and experimental sideband
Schedule of Operation	Continuous	Continuous (except to permit experimental changes)
Station Identification	Keyed 600 c/s modulation, using International Morse Code, of call letters, repeated three times on the hour and every 20 minutes thereafter.	Keyed carrier, using International Morse Code, of call letters, repeated three times on the hour and every 20 minutes thereafter.
Frequency Offset Identification	Keyed 600 c/s modulation, using International Morse Code, of frequency offset, to immediately follow the station identification keying. The letter M will be used for minus sign.	Keyed carrier, using International Morse Code, of frequency offset, to immediately follow the station identification keying. The letter M will be used for minus sign.

Figure 3-3B

BRITISH GOVERNMENT LOW FREQUENCY
STANDARD FREQUENCY STATIONS

Station	GBR 16 kc/s	MSF 60 kc/s	Droitwich 200 kc/s
Location	52°22'N 1°11'W		52°18'N 2°06'W
Carrier power to antenna (kw)	300 (40*)	10	400
Daily period of operation (hours)	22-1/2(1)	1(2)	18-20
Duration of time signal transmission (minutes)	4 x 5(3) per day	60(4) per day	-----
Accuracy of frequency (parts in 10^{10})	±5(5)		±10(5)

* Estimated radiated power.

Notes

(1) The maintenance period is from 1300 to 1430 UT, approximately.

(2) From 1429 to 1530 UT.

(3) Telegraphy time signals for 5 minutes preceding 0300, 0900, 1500 and 2300 UT.

(4) Modulated time signals consisting of 5 cycles of 1000 c/s modulation, the minute signal is lengthened to 100 milliseconds.

(5) These are statutory limits and adjustment is usually much closer. For MSF/GBR the accuracy is with respect to the offset value.

Note: The offset frequency for 1964 is -150 parts in 10^{10}.

Figure 3-3C

3.3 0.903172 Mc Frequency Synthesizer

3.3.1 Specifications

Input:	0.3 to 1.0 volts rms, 1.0 megacycle.
Output:	0.5 volts rms, 0.90317248 megacycles into a 50 ohm load.
Stability:	Equal to that of the 1.0 megacycle source.
Spurious:	All non-harmonically related signals 120 db below desired output.
Power Supply:	115 volts ±10%, 50 - 60 cycles approximately 30 watts.

3.3.2 Introduction

The frequency synthesis process is one in which a frequency or set of frequencies is derived from a fixed frequency, usually from a very stable oscillator. The equipment described in this manual is designed to directly derive 7630/8448 Mc from a 1.0 megacycle source which has a stability of one part in 10^{10} per day. Since the 7630/8448 Mc/s is directly derived, the source stability is retained in the output.

3.3.3 Theory of Operation

The fraction 7630/8448 can be factored to arrive at the following set of fractions:

$$\frac{7630}{8448} = \frac{(5)(7)(109)}{(2^7)(3)(11)} = \frac{(5)(7)(109)}{(11)(8)(48)} = \frac{(5)\ (7)}{(11)(8)} \times \frac{(64 + 45)}{(16)(3)}$$

$$= 7/4 \quad 1/2 \quad 5/11 \quad (4/3 + 3/4 \quad 5/4)$$

$$= 5 \quad 1/2 \quad 7/4 \quad (4/3 + 3/4 \quad 5/4) \quad 1/11$$

$$= 5 \quad 1/2 \quad 7/4 \quad (2/3 \quad 2 + 1/4 \quad 5.3/4) \quad 1/11$$

This set of fractions, which is one of many possible sets, was employed in the synthesizer design. A block diagram of the equipment necessary to perform this function is shown in Figure 3-4.

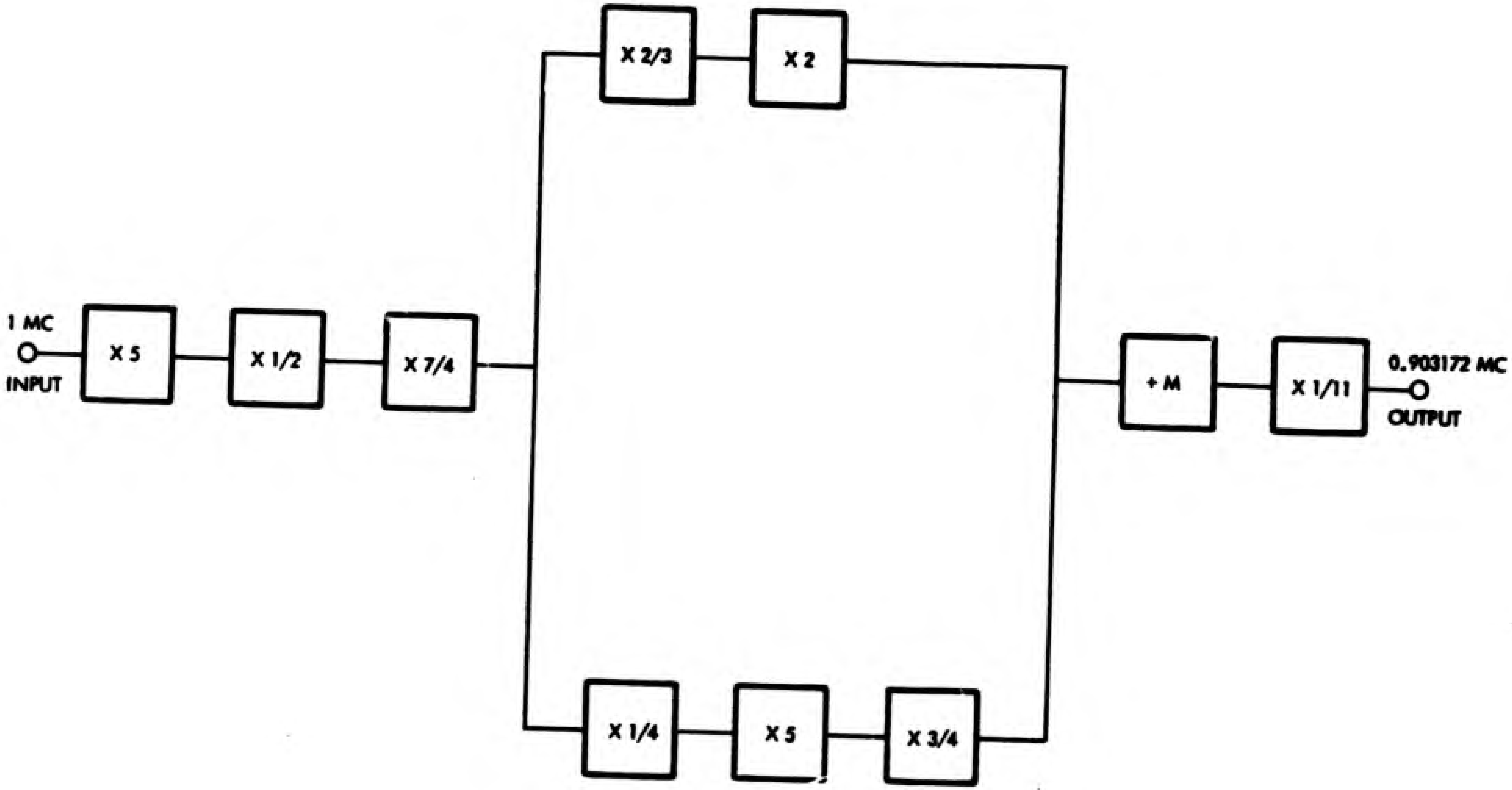

Figure 3-4. 0.903172 Mc Synthesizer Block Diagram

Although it is not readily apparent from the block diagram, this design was selected for the following reasons:

1. The lowest frequency employed is 625 Kcs and therefore coils for tuned tank circuits can be made of reasonable size;
2. The highest frequency is 9.934 Mcs which not only avoids the necessity for high-cost high-frequency transistors but also permits the use of simple ring mixers;
3. The highest multiplication necessary is 5 which precludes the requirements for very high Q circuits which are temperature sensitive;
4. Four of the operations are possible with nearly identical circuits (i.e., times 7/4, times 2/3, times 1/4 and times 3/4); and
5. Aside from the mixing operations necessary in the dividers only one additional mixer is necessary.

The operations which are indicated by the fractions in the block diagram such as 7/4, 1/4, 3/4, and 2/3 are performed with regenerative frequency dividers (RFD). The normal output frequencies of an RFD are $\frac{1}{n}$, $\frac{n-1}{n}$, and $\frac{2n-1}{n}$ where n is the divisor such as n = 4 in a divide - by - four. Therefore it is possible to obtain $1/4 = 1/n$, $3/4 = \frac{n-1}{n}$ and $7/4 = \frac{2n-1}{n}$ directly from one RFD.

The complete theory and operation of RFD's is readily available from other sources and only a simple explanation is included in this manual.

Operation of an RFD is most easily explained with the aid of a block diagram, Figure 3-5.

Where M = a mixer in which the lower sideband is extracted,

F = filter to extract desired sideband from mixer,

A = amplifier to achieve unity loop gain, and

m = multiplier = (n - 1)

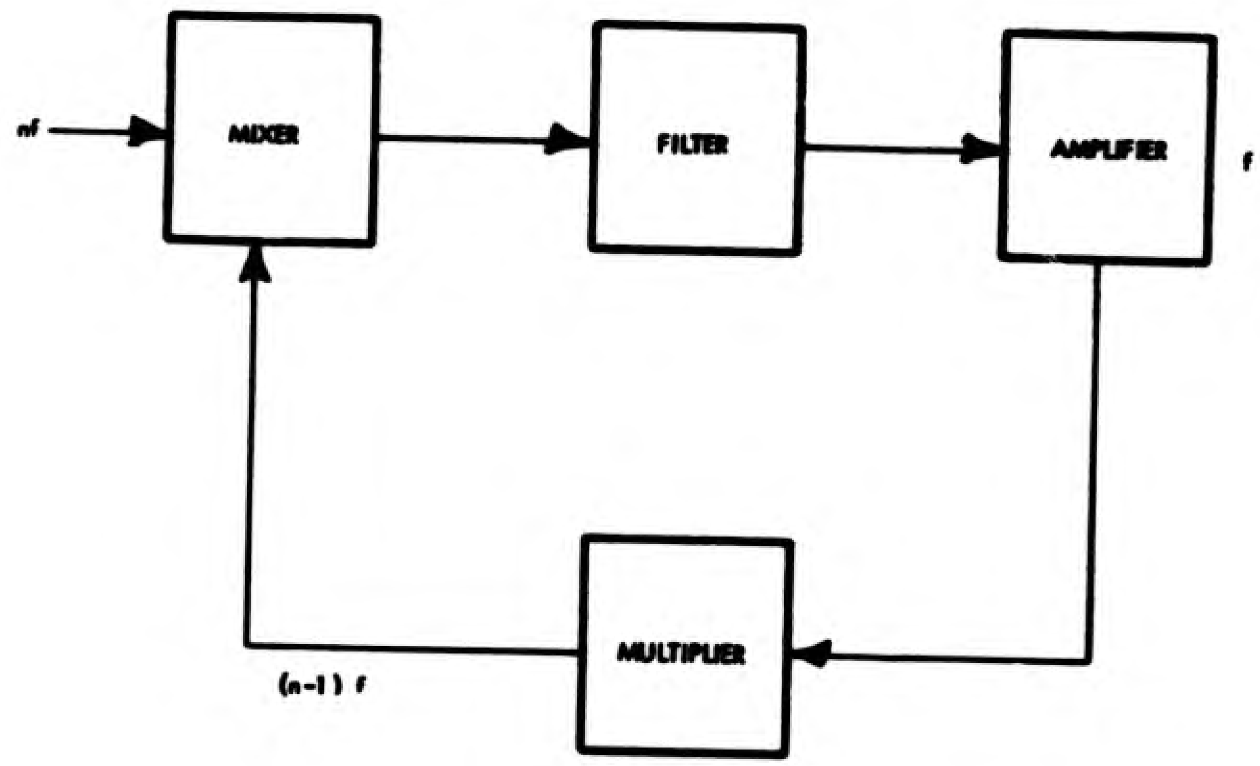

Figure 3-5. Regenerative Frequency Divider Block Diagram

If a signal of frequency nf is applied to the mixer and the loop has sufficient gain (≥ 1), the RFD will build-up from noise (similar to the build up in an oscillator) and the desired output will be achieved. Once the loop has started the operation is as follows. The input frequency, nf, mixes with the output frequency of the multiplier (n - 1) f, and produces a signal f. This signal f is then multiplied to produce (n - 1) f which again mixes with the incoming signal.

Since an RFD is a closed-loop system it is evident that some component in the system must saturate if the output level is to remain constant with changes in input level. The RFD design employed has essentially two saturating devices; these are:

1. A ring mixer, whose output level tends to remain relatively constant after the input has reached a level which is sufficient to saturate the diodes; and
2. A multiplier which consists of a diode directly connected to the base of a transistor. The direct connection limits the transistor gain and therefore the multiplier output and a type of saturation is achieved.

With the above saturating devices, it is possible to vary the input signal level at least 10 db, and minor changes in levels do not affect the circuit operation.

3.4 The X-Band Receiver Local Oscillator Synthesizer

The modules which comprise the Local Oscillator Synthesizer are indicated in Figure 3-6. Because the X-Band receiver front end including the first mixer will be located in the antenna equipment box or the feed box, the LO source must terminate there also. One hundred megacycles has been selected as the frequency to be transmitted from the van to the equipment

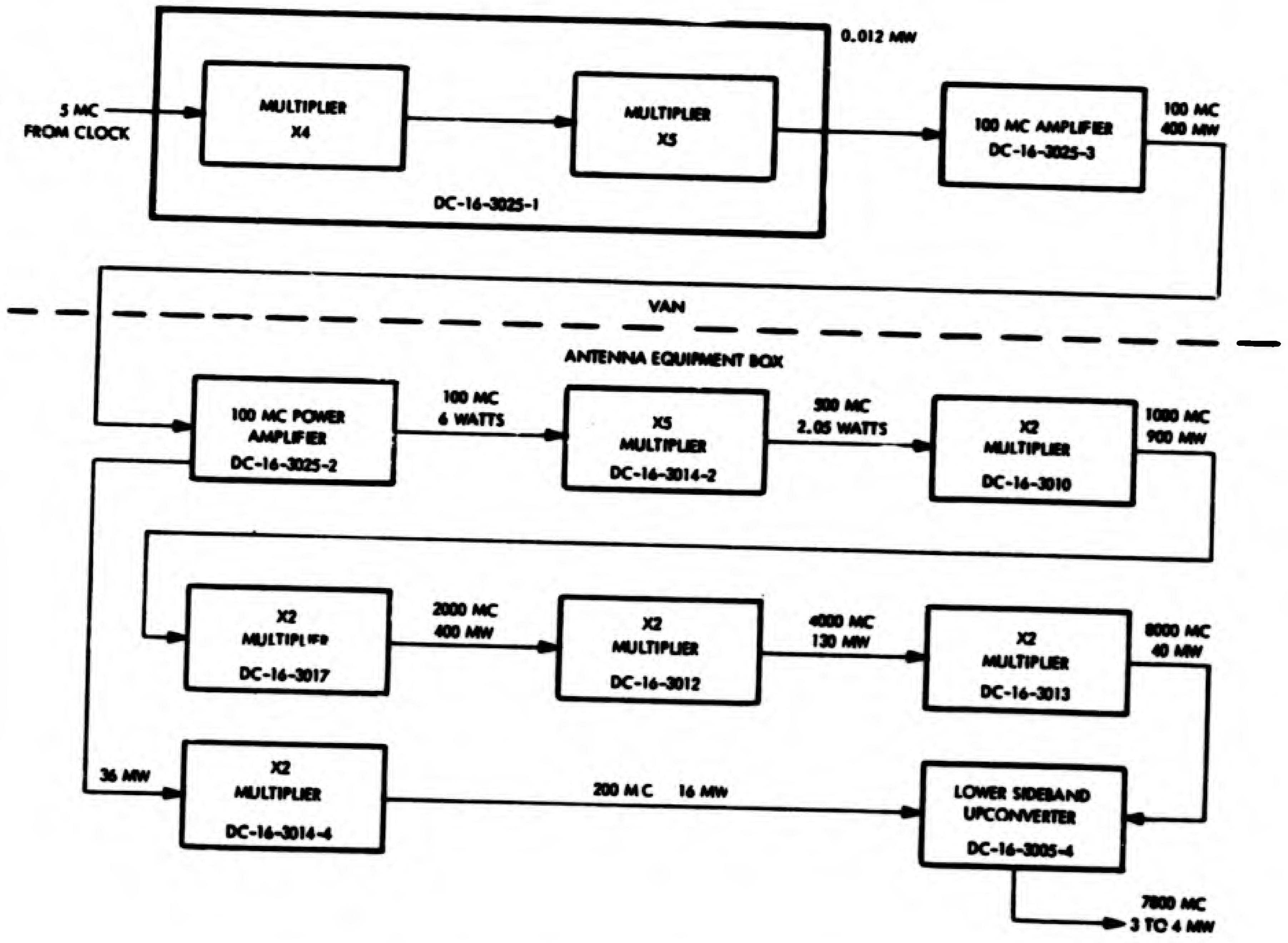

Figure 3-6. Receiver Local Oscillator Synthesizer Block Diagram

box. This is the highest frequency present in the system which can be transmitted over a coaxial cable with low loss and still be economically amplified at the antenna.

3.4.1 Transistor Frequency Multiplier X4X5 (DC-16-3025-1)

3.4.1.1 Description

The block diagram and schematic diagram for the X4X5 multiplier module are shown in Figures 3-7 and 3-8. A 5 Mc clock output is amplified in the first stage (Q1) which acts as a buffer between the stable clock and the quadrupler. The output of the quadrupler is amplified by Q2 and filtered by a 200 Mc crystal filter. The filtered output is divided. A portion of the 20 Mc signal is diverted to the X6 Transistor Frequency Multiplier and the remainder is amplified in the 20 Mc amplifier (Q3). The output of Q3 is multiplied to 100 Mc and amplified in the 100 Mc amplifier (Q4).

3.4.1.2 Specifications

The design specifications established for the X4X5 Transistor Frequency Multiplier are listed below:

Temperature Range:	15°C to 44°C
Output Power Variation:	±0.1 db
100 Mc Power Output:	0.01 mw
20 Mc Power Output:	4 mw
5 Mc Input Voltage:	1.0 volt rms ±20%
Spurious Output:	-90 db below 20 Mc output -40 db below 100 Mc output

3.4.1.3 Design Considerations

A diode harmonic generator is employed to generate harmonics of the input frequency. The desired harmonic is then extracted by a tuned circuit. This procedure enables the transistor amplifiers which follow to operate at a

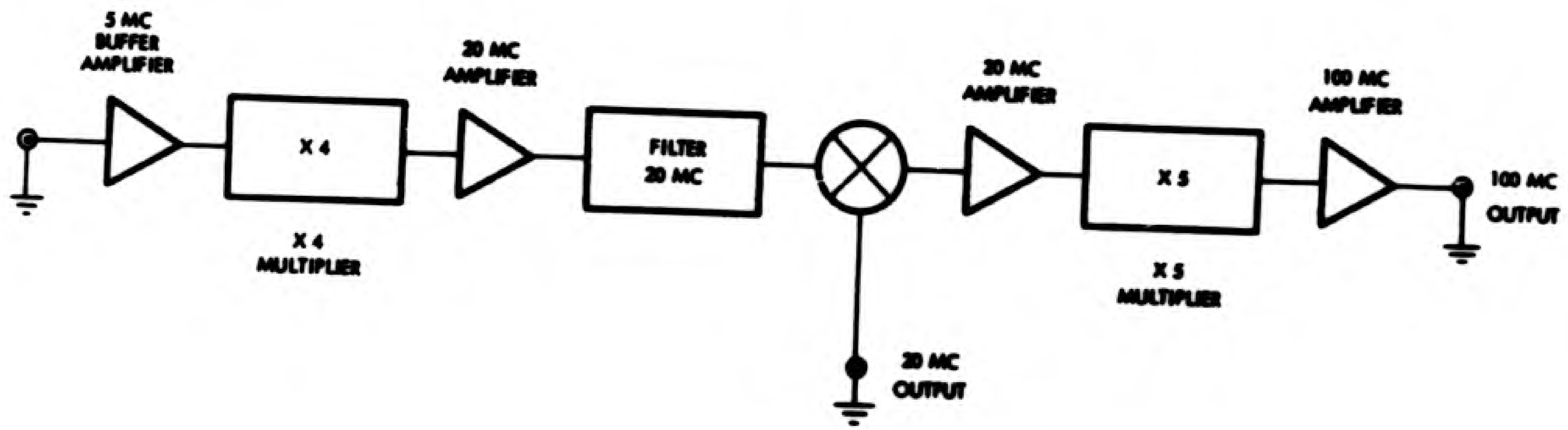

Figure 3-7. X4X5 Transistor Multiplier Block Diagram

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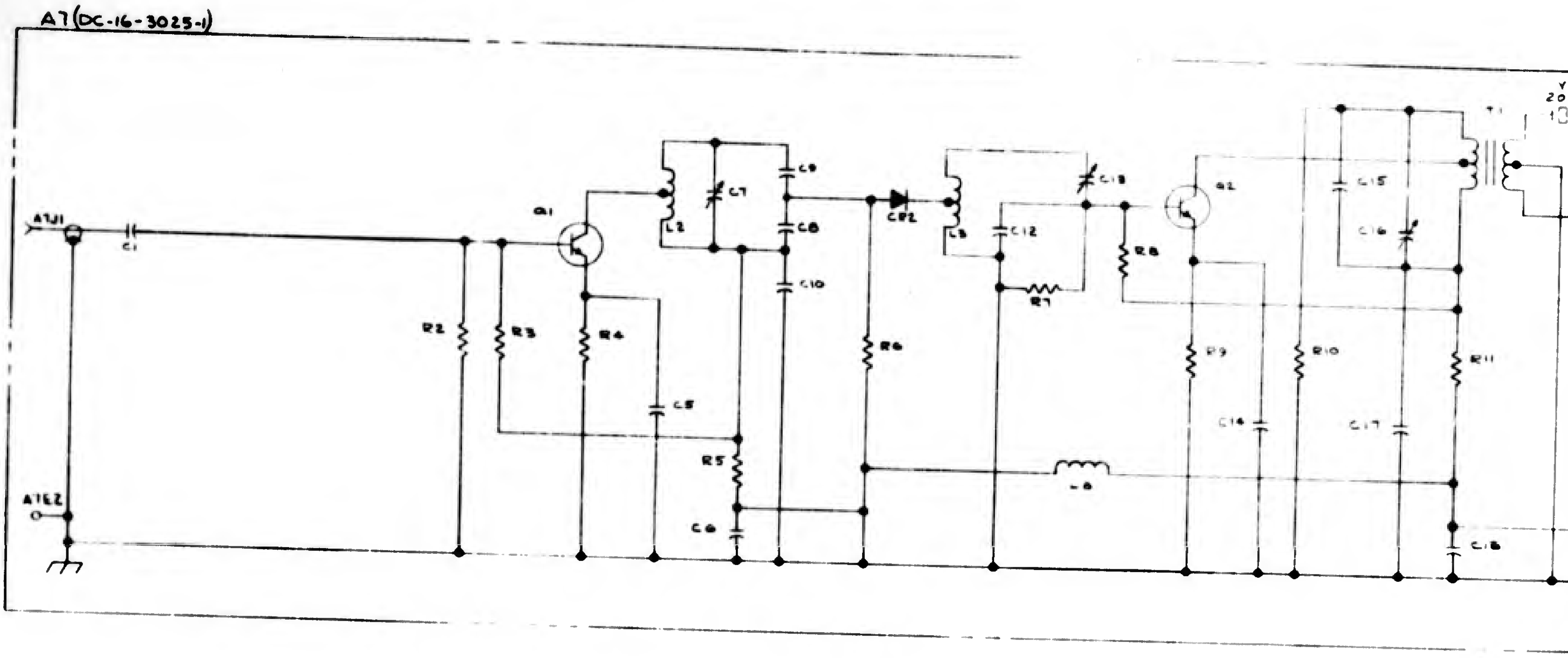

A7 (DC-16-3025-1)
A7J1
C1
A7E2
R2
R3
Q1
R4
C5
R5
C6
L2
C7
C9
C8
C10
CR2
R6
L3
C12
C13
R7
Q2
R8
R9
C14
R10
C15
C16
C17
R11
C18
T1

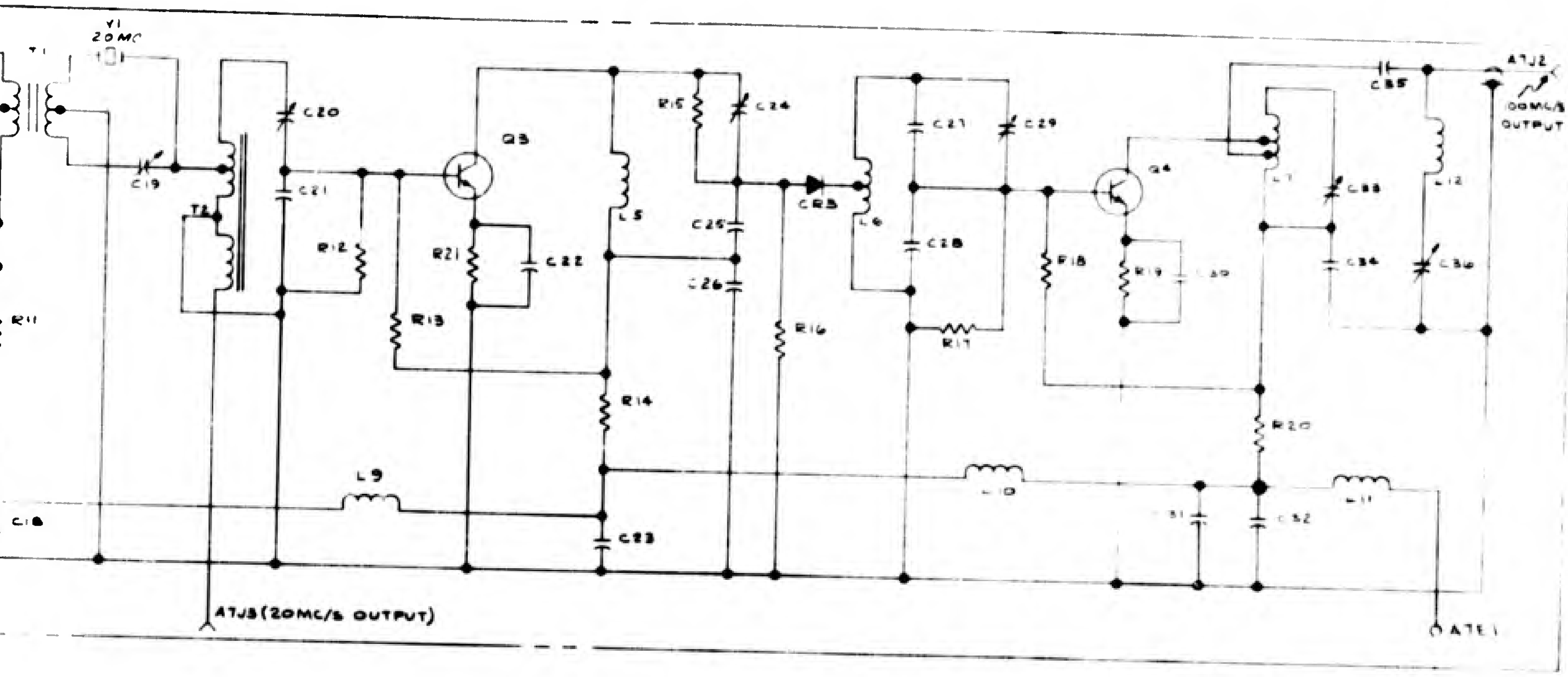

Figure 3-8. X4X5 Transistor Multiplier Schematic Diagram

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bias point yielding maximum gain. This method of operation provides a fixed input impedance to Q2 and simplifies the design of the tuned circuits.

The crystal filter employed is a balanced bridge circuit which provides a narrow bandwidth and a nominal 60 db rejection when it is properly balanced. The balancing is accomplished by means of the variable ceramic capacitor C19, which neutralizes the effect of the shunt capacitance of the crystal holder. The signal voltages applied to the capacitor and the crystal are equal but 180° out of phase; thus, when they are connected to a common point, there is a cancellation of all signals except the desired frequency which is passed by virtue of the series resonant mode of the crystal. The crystal filter rejects spurious frequencies close to the carrier in addition to supressing the 25 and 15 Mc components which result from the multiplication process. Transformer T2 has two secondary windings which divide the signal from the crystal filter. One winding provides a 50 ohm output to the X6 transistor frequency multiplier. The other winding forms part of the input circuit to the 20 Mc amplifier (Q3). This frequency is multiplied by five, further amplified, and filtered in a single tuned 100 Mc amplifier. This amplifier employs 2N708 silicon transistors in a common emitter configuration. The output tank circuit capacitance is chosen to provide a loaded Q that will provide 10 db of rejection at a normalized bandwidth (Bn') of 4. The normalized bandwidth is defined:

$Bn' = Bn/F_o$ where Bn is the bandwidth between points on the response curve at which the power response is $\left(\frac{1}{1+n}\right)$ times the response at the band center frequency and Fo = band center frequency.

The inductance required is determined by:

$$L = \frac{Bn'}{4\,Fo\,Qo}\;\frac{Qo - \sqrt{n}}{g_{o1}\sqrt{n}}$$

where

Qo = unloaded circuit Q

g_{o1} = equivalent parallel loss conductance at the transistor output

The value of capacitance required to resonate with this inductance is modified by the output reactance of the transistor, heat sink capacity, unwanted couplings and other stray capacitance effects. The amplifiers in this module are designed to provide 10 db rejection for a Bn' of 0.4. The outputs provided are 4.0 mw at 20 Mc and 0.01 mw at 100 Mc. The spurious response is held to approximately -87 db and -42 db respectively. The -42 db spurious level associated with the 100 Mc signal is further reduced in the 100 Mc amplifier module.

3.4.2 <u>100 Mc Amplifier (DC-16-3025-3)</u>

The 100 Mc amplifier is comprised of eight stages of amplification to provide separate outputs of 400 mw and 100 mws from a .01 mw input. The first four stages employ 2N708 transistor and the last four employ 2N914 which are more suitable for the higher power levels encountered. The 400 mw output which drives the high power 100 Mc amplifier is obtained by operating two transistors in parallel. Part of the output of the parallel amplifiers is also used to drive two single ended amplifiers which provide a 100 mw output to drive the X7 multiplier.

3.4.2.1 <u>Specifications</u>

The design specifications are delineated below for the 100 Mc amplifier:

Temperature Range:	15°C to 44°C
Output Power Variation:	±.2 db

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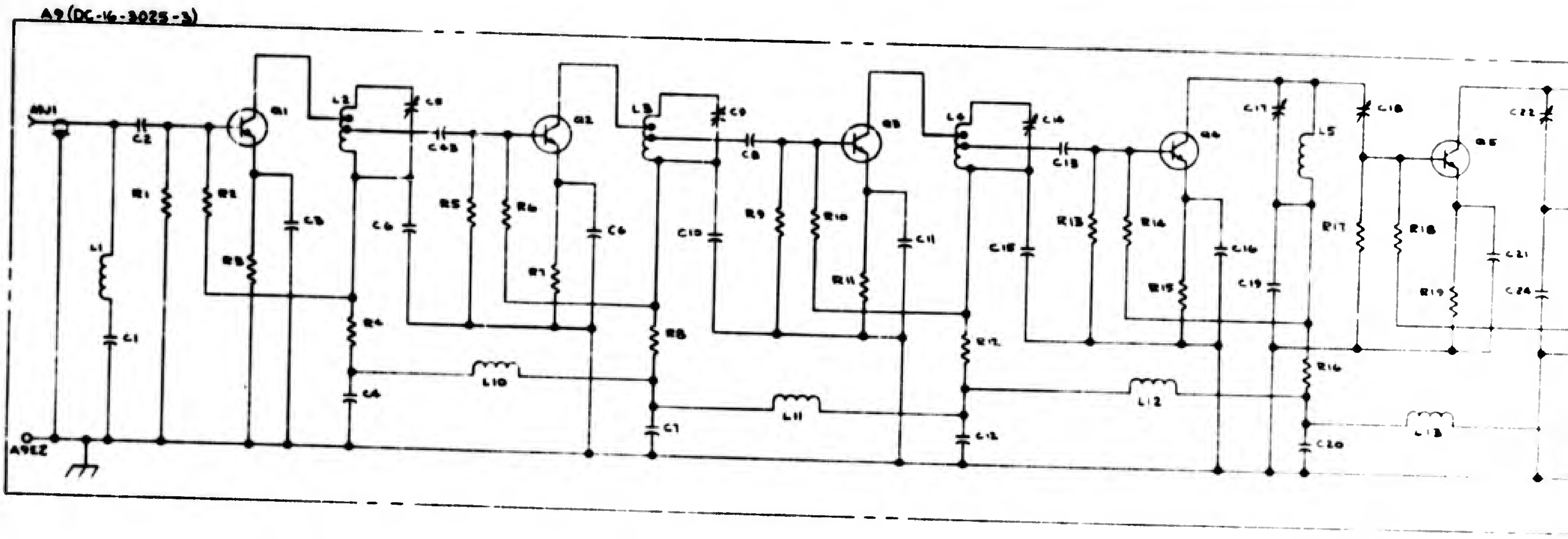
A9 (DC-16-3025-3)
A9J1
A9E2
Q1
Q2
Q3
Q4
Q5
L1
L2
L3
L4
L5
L10
L11
L12
L13
C1
C2
C3
C4
C6
C7
C8
C9
C10
C11
C12
C13
C14
C15
C16
C17
C18
C19
C20
C21
C22
C24
R1
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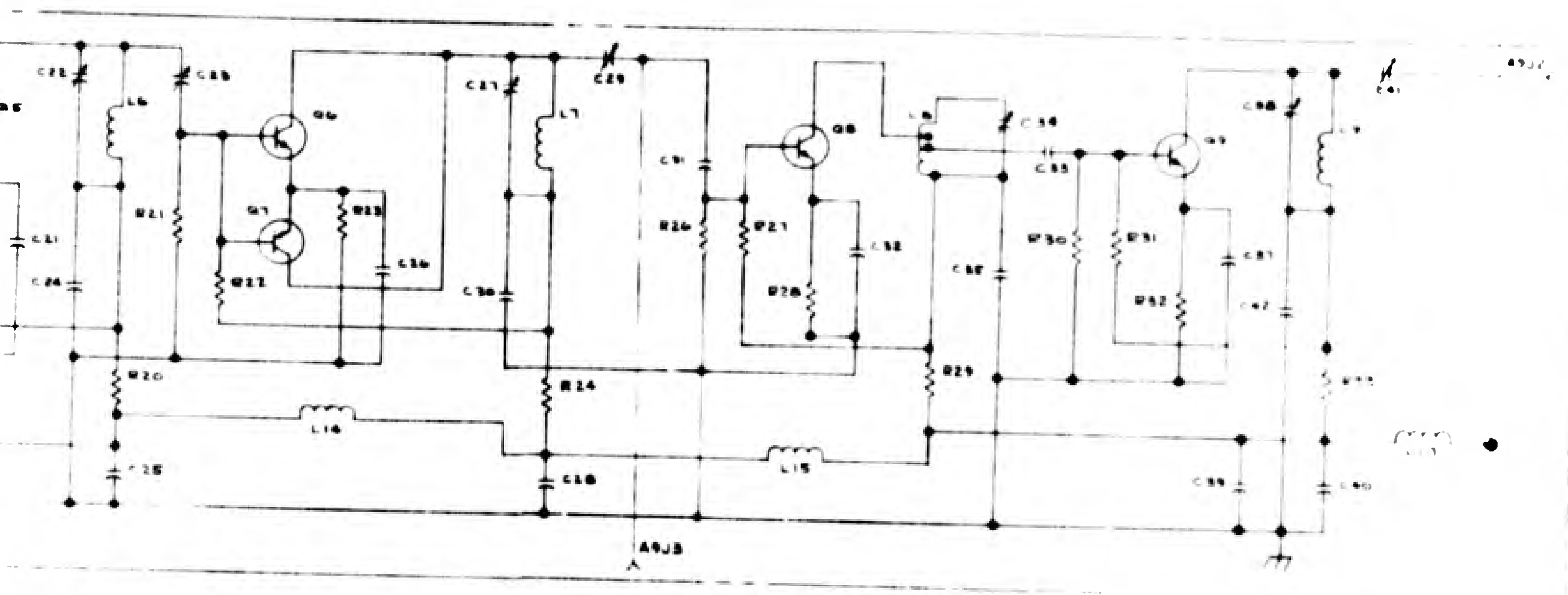

Figure 3-9. 100 Mc Amplifier Schematic Diagram

Power Output:	A) 100 mw B) 400 mw
Power Input:	.01 mw
Spurious Response (In Band):	90 db down

3.4.2.2 Design Considerations

The basic amplifier design used is similar to that described in Section 3.4.1. The first five stages employ 2N708 transistors in a single ended configuration. The sixth stage uses two 2N914 transistors in parallel to obtain a nominal 400 mw output. The output of the push-pull stage also supplies drive to two 2N914's in cascade which provides the 100 mw output to the X's7 multiplier.

The primary spurious signals of major concern are the 80 Mc and 120 Mc signals which are present as a result of the multiplication process in the multiplier/100 Mc amplifier module. The required suppression is obtained through the use of two traps and the bandpass characteristic of the amplifier. The traps are tuned to 80 and 120 Mc and the spurious are >90 db down. The actual rejection could not be measured due to the limitations of the test equipment available.

3.4.3 100 Mc Power Amplifier (DC-16-3025-2)

3.4.3.1 Description

The 100 Mc Power Amplifier, located in the antenna equipment box, is supplied with a 400 mw input via approximately 50 feet of RG-9 from the 100 Mc amplifier DC-16-3025-3 located in the equipment van. This input is divided into two channels and amplified to the required levels. The schematic diagram of the amplifier is shown in Figure 3-10.

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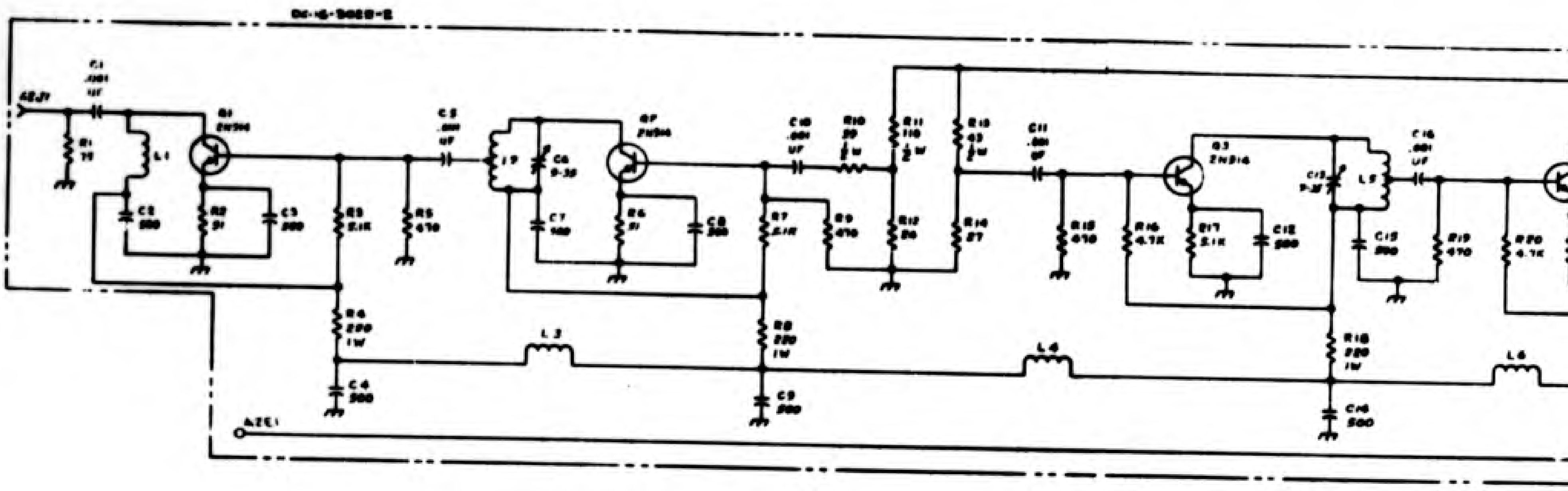

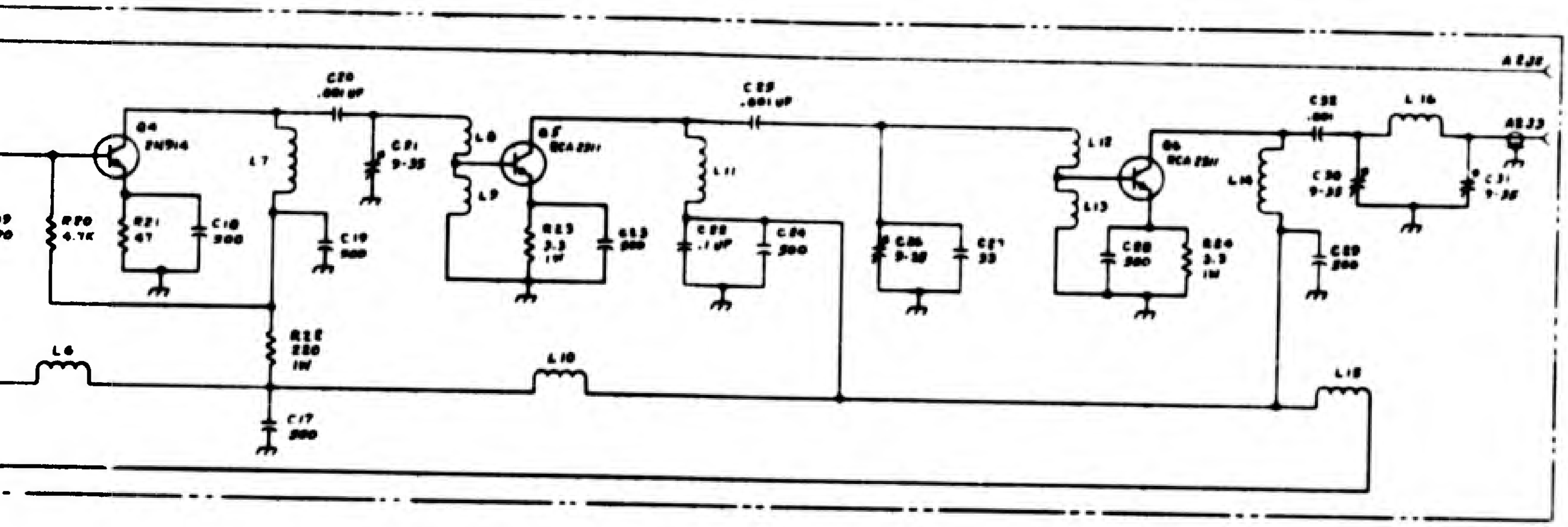

Figure 3-10. 100 Mc Power Amplifier Schematic Diagram

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3.4.3.2 Specifications

Temperature range:	$-30^{o}C$ to $+60^{o}C$
Power Output:	Channel A 10 mw Channel B 6.0 watts
Power Output Variation:	less than ± 0.5 db
Isolation Between Channels:	30 db
Power Input:	200 mw or greater

3.4.3.3 Design Considerations

The resistive divider network (R10 through R14) provides the dual function of power division and isolation. Channel B utilizes four amplifier stages Q3, Q4, Q5 and Q6 to obtain a 6.0 watt output. The stage Q4 is operated in the saturated region and so insures a constant output power over the anticipated range of input power variations. Q5 and Q6 are typical power amplifier stages operating in the Class C region for high efficiency. Channel A employs two stages (Q1 and Q2) to obtain a 10 mw output. The output of each channel is isolated to load variations on the other channel by at least 30 db. This isolation is achieved through the action of the resistive divider and the unilateral characteristics of each amplifier. Output power of each channel is stable to within ± 0.5 db over a temperature range from $-30^{o}C$ to $+65^{o}C$.

3.4.4 X-5 Varactor Multiplier (100/500 Mc) DC-16-3014-2

The operation of the X5 multiplier is similar to that of the X3 (DC-16-3014-1) and X7 (DC-16-3014-3). In the case of the times 5, the third and second harmonics are provided with idler circulation paths. It will be noted that second harmonic distortion is unsymmetrical with respect to the positive and negative half cycles of the fundamental, and, hence, can only be generated by some non-linear characteristic which is non-symmetrical about the operating point. Therefore, if two diodes have identical characteristics and both are operated at the same AC level and bias but one is reversed with respect to the other, then any second harmonic voltage developed across one will be opposite in phase to that across the other. The symmetrical bridge will always idle the second harmonic within itself.

3.4.4.1 Performance

Power Input at 100 Mc:	6.0 watts
Power Output at 500 Mc:	2.05 watts

3.4.4.2 Adjustment; Retuning

Reference to Figure 3-6 will show that this multiplier is the first in a chain of five cascaded varactor frequency multipliers. Each coaxial multiplier has been designed and adjusted to operate from a 50 ohm source and into a 50 ohm load. Input and output impedance levels, however, depend upon the input power level and the varactor bias. If one multiplier in the string becomes defective or is disturbed, it will present an unmatched load to the unit driving it and improper drive to the unit following it. Only a single multiplier should be adjusted at one time and it should be operated into a pure 50 ohm resistive load during adjustment. A spectrum analyzer is a necessity when adjusting varactor multipliers because of the numerous responses that are possible under circumstances of mis-adjustment.

Figure 3-11 is the electrical schematic of the X5 multiplier. C1 and C2 can be used to set the input impedance by increasing one capacitor while decreasing the other. C3 is tuned to resonate with L1 at the input frequency. C4 and C5 allow residual unbalance between the two varactors to be compensated. They are adjusted to minimize second harmonic in the output. L2 and C6 are tuned to block, i.e. minimize, the fundamental (input) frequency present in the output. L4 and C7 are tuned to permit

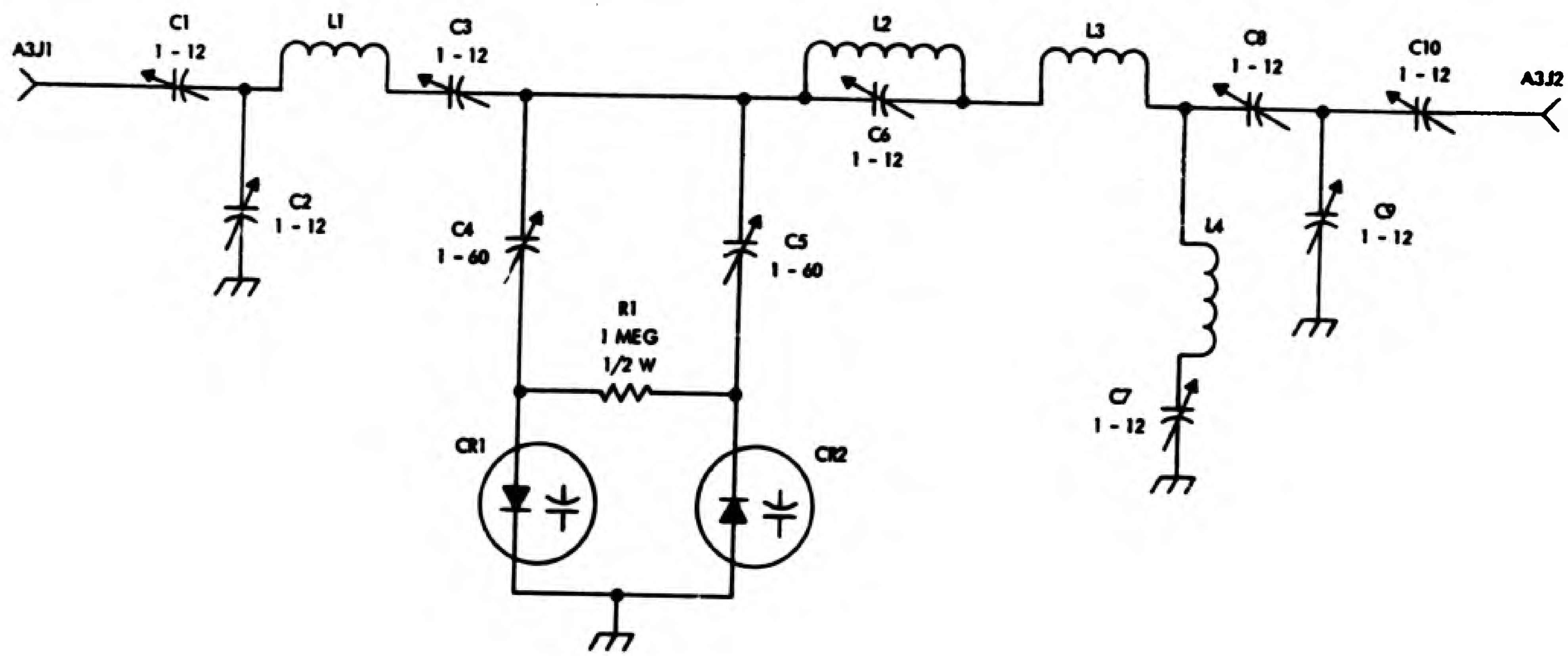


Figure 3-11. X-5 Varactor Multiplier (100/500 Mc)
Schematic Diagram

circulation of the third harmonic. They are adjusted for minimum third harmonic output. There is interaction between the various adjustments and it is necessary to recheck each setting several times.

3.4.5 X-2 Varactor Multiplier (500/1000 Mc) DC-16-3010

The schematic diagram of Figure 3-12 is applicable.

This multiplier employs lumped parameter input circuits and distributed parameter output circuits. C2 and L1 are series resonant at the fundamental frequency. C1 and C3, in conjunction with C2, L1, permit impedance matching the source and varactor while maintaining a low impedance to the fundamental and blocking the flow of second harmonic currents in the input circuitry. L2 is a choke to provide a DC return for the bias circuitry. R1 is the self biasing resistor whereas C4 is the bias holding capacitor. The coaxial capacitor presents a low impedance to second harmonic currents while providing only a minor shunting affect upon the fundamental currents. Shorted stub S-201 allows tuning of the output impedance match. Cavity T-201 is a single tuned filter at the second harmonic output frequency. T-201 is tunable by adjusting the end loading capacitance.

Power Input at 500 Mc:	2.0 watts
Power Output at 1000 Mc:	900 mw

3.4.6 X-2 Varactor Multiplier (1000/2000 Mc) DC-16-3017

The schematic diagram of Figure 3-13 is applicable.

This unit is constructed using coaxial circuit techniques throughout. Quantatron Inc. "Rotostub" double stub tuners are employed for input and output impedance matching.

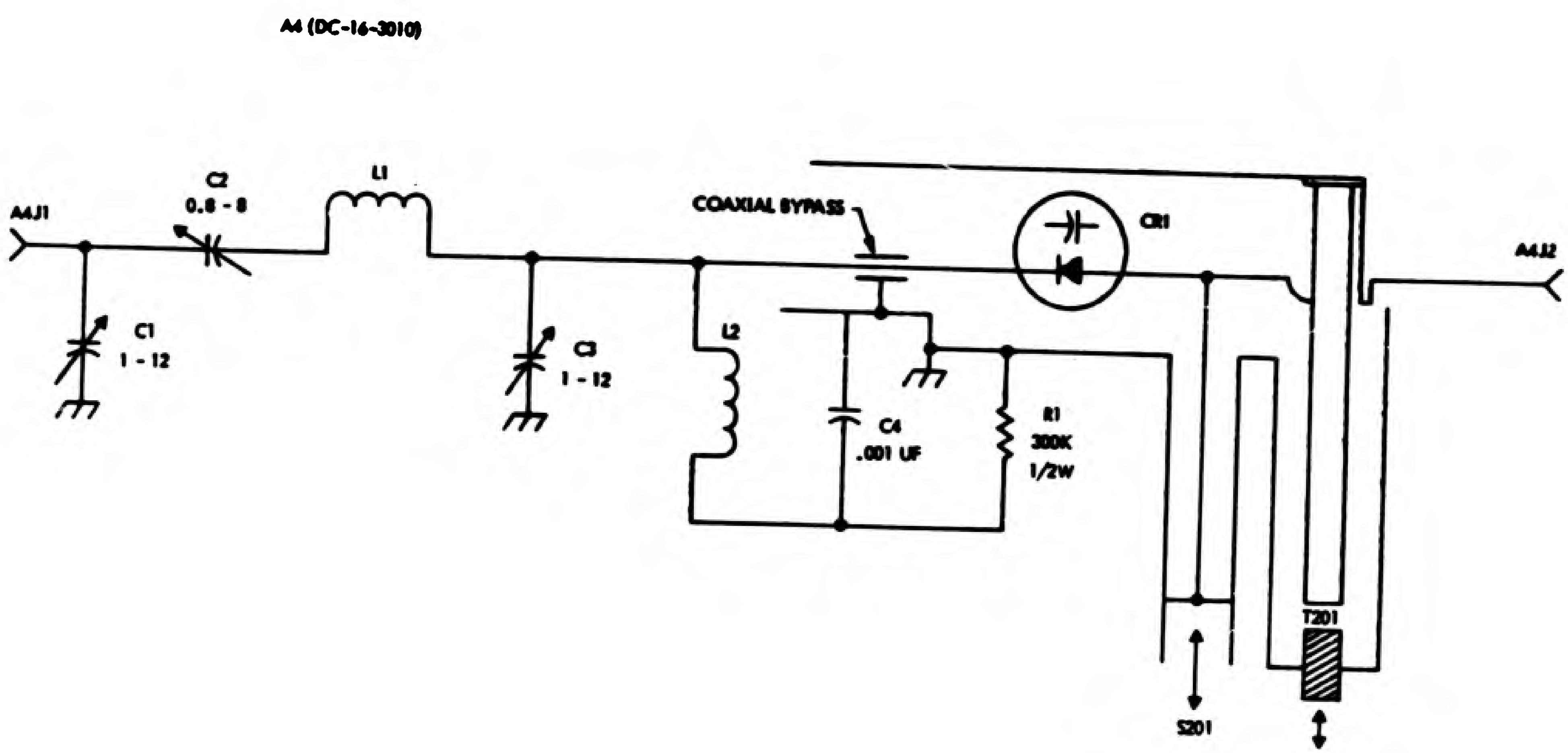


Figure 3-12. X-2 Varactor Multiplier (500/1000 Mc)
Schematic Diagram

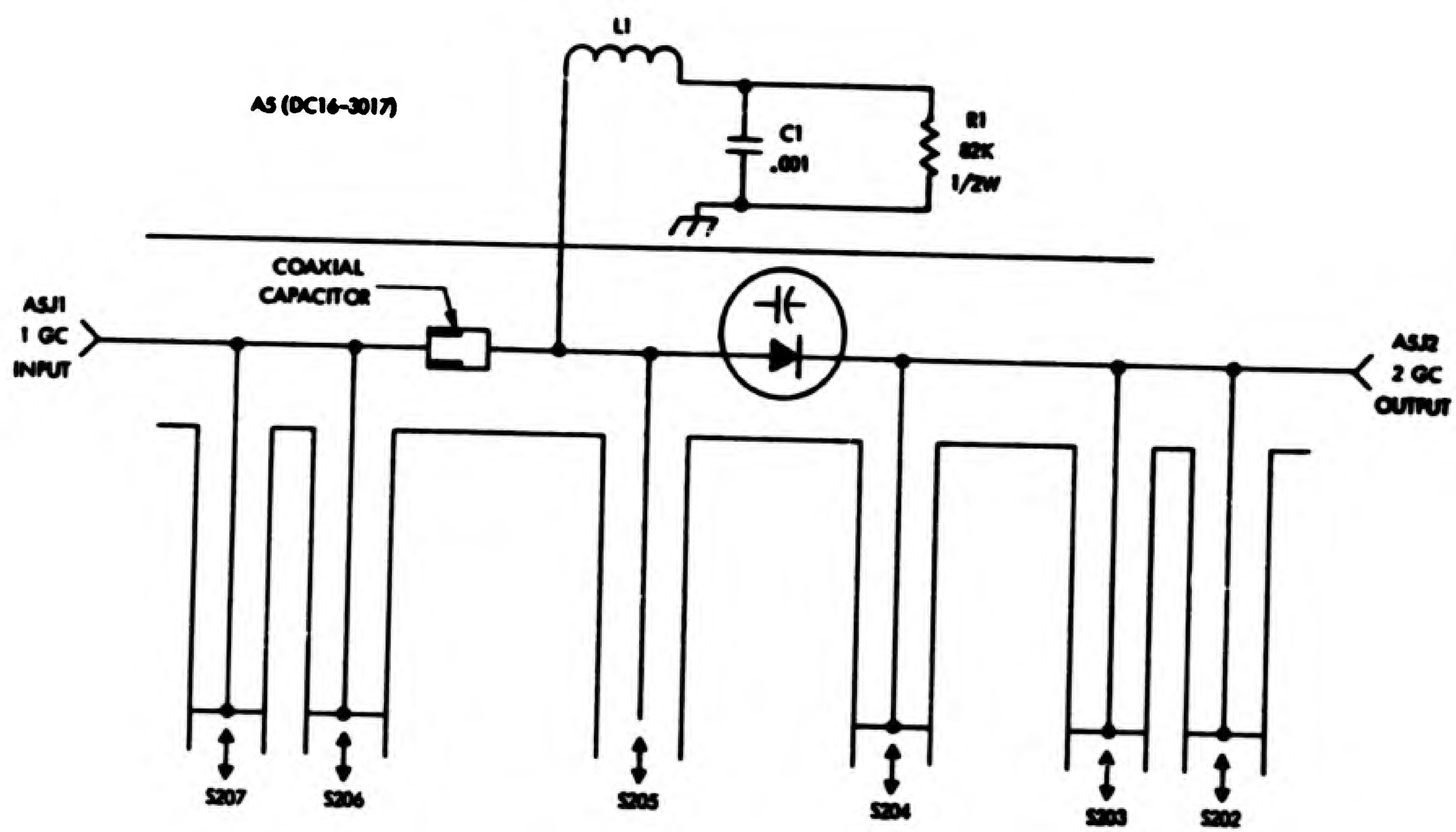


Figure 3-13. X-2 Varactor Multiplier (1000/2000 Mc)
Schematic Diagram

Open circuited stub line S-205 appears as a short to the second harmonic frequency at the varactor. Short circuited stub S-204 represents a virtual short to the fundamental frequency on the output side of the varactor. L1, C1 and R1 comprise the self-biasing circuitry.

Power Input at 1000 Mc:	900 mw
Power Output at 2000 Mc:	400 mw

Because of the interaction between this unit and DC-16-3012 which follows it, a ferrite isolator, M-H Microwave Inc. Model CT-330-TT, was inserted in the signal path between these two units. Performance data on the isolator are:

VSWR at 2.0 gc:	port 1 1.04 port 2 1.03
Insertion Loss:	approximately 0.2 db
Isolation at 2.0 gc:	35 db

3.4.7 <u>X-2 Varactor Multiplier (2000/4000 Mc) DC-16-3012</u>

Schematically, this unit, Figure 3-14, is identical to DC-16-3017. Physically, it differs in having the input and output matching stubs built into the body of the multiplier assembly. Normal power input at 2000 Mc is 400 milliwatts which produces an output power of 130 milliwatts at 4000 Mc.

3.4.8 <u>X-2 Varactor Multiplier (4000/8000 Mc) DC-16-3013</u>

This unit (Figure 3-15) employs coaxial input circuitry and a waveguide output. The three stub tuner permits impedance matching at the fundamental frequency and rejection of the second harmonic on the input side of the varactor. Adjustable shorted stub S217 allows adjustment of the phase of the second harmonic standing wave on the coaxial system to permit optimum coupling to the waveguide. The waveguide short also allows adjustment of

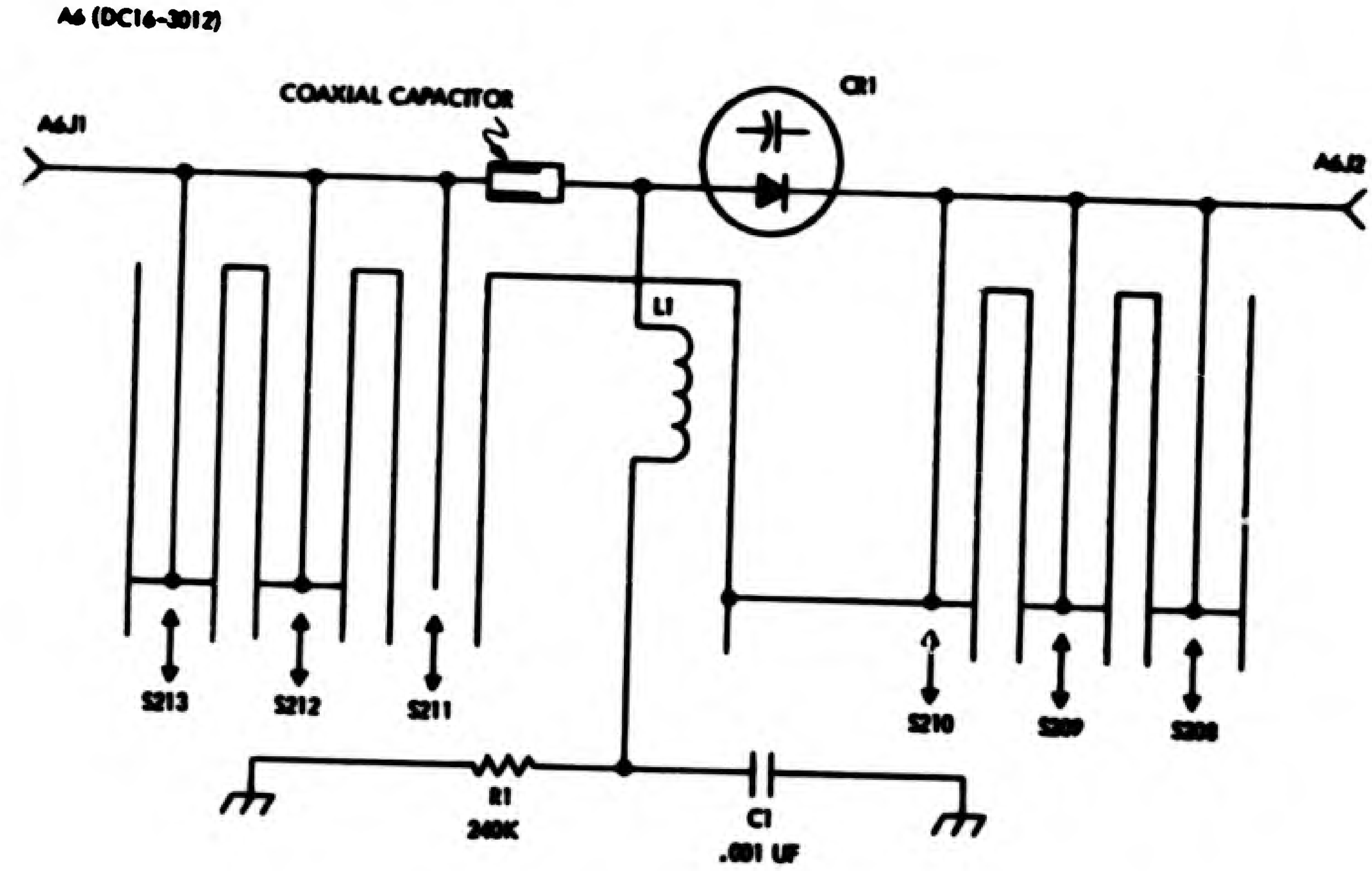


Figure 3-14. X-2 Varactor Multiplier (2000/4000 Mc)
Schematic Diagram

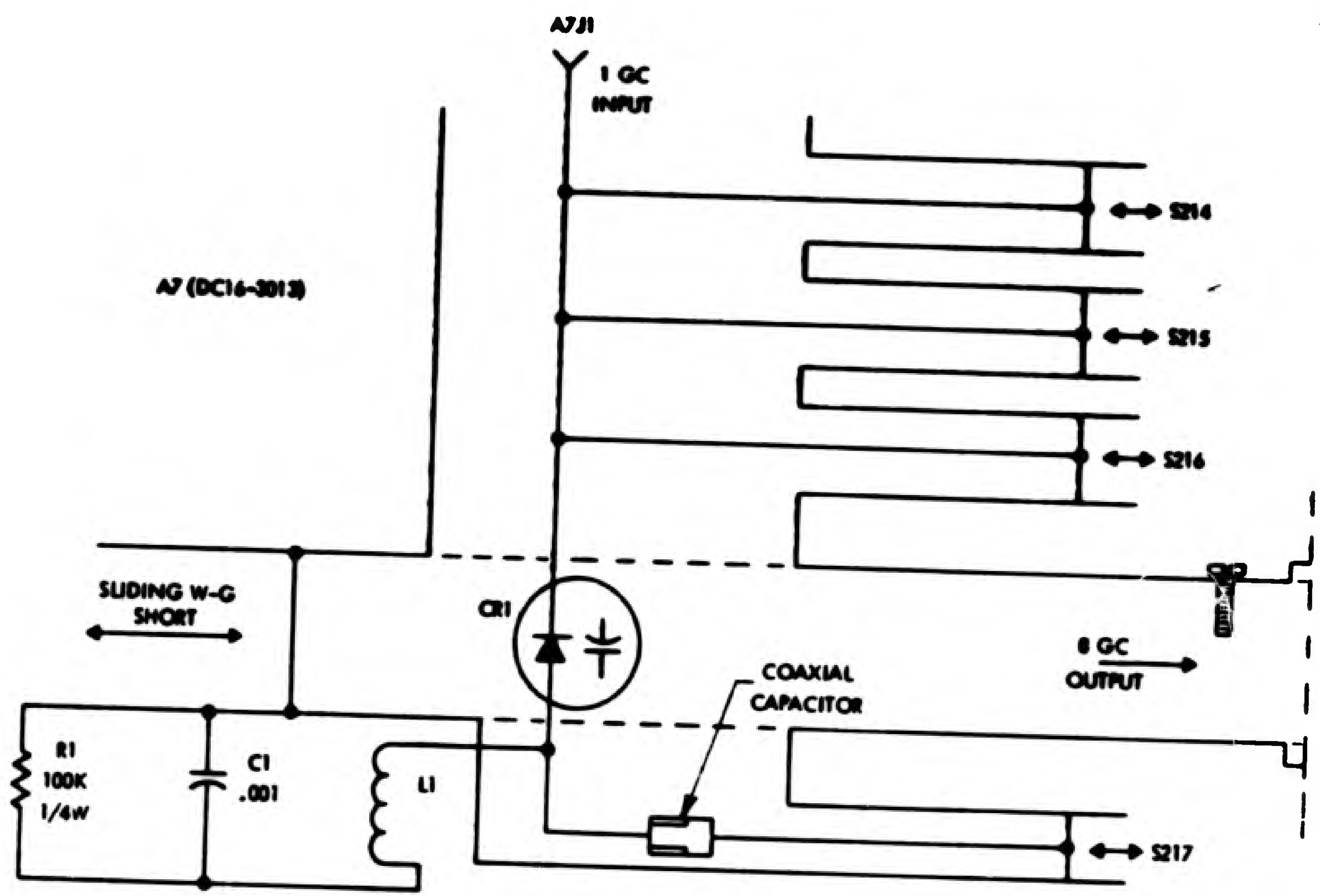


Figure 3-15. X-2 Varactor Multiplier (4000/8000 Mc)
Schematic Diagram

the diode to waveguide coupling.

Power Input at 4000 Mc: 130 mw

Power Output at 8000 Mc: 40 mw

3.4.9 X-2 Varactor Multiplier (100/200 Mc) DC-16-3014-4

This circuit (Figure 3-16) bears some resemblance to the odd harmonic generators described elsewhere. It is, however, significantly different. C7 and C8 permit input impedance matching. T1 provides a fundamental frequency voltage balanced above ground to drive the varactor bridge circuit. C6 allows the secondary leakage reactance of T1 to be series resonated with the average capacitance of the bridge circuit at the fundamental frequency. Assume for the sake of argument that equal sinusoidal currents of the fundamental frequency are flowing downward through the right and left hand branches of the bridge from A to B. Any second harmonic voltage developed across CR2 will be opposite in phase to a second harmonic voltage developed across CR1 because the diodes are reversed. However, in terms of the second harmonic potential difference between C and D the two are series additive. If the varactors are identical, twice the second harmonic voltage developed across one varactor will appear at the output. C3 and L1 form a series resonant filter tuned to the second harmonic output frequency. C1 and C2 permit output impedance matching. R1 provides a DC return path for the self developed bias voltage stored on C4 and C5. C4 and C5 permit compensation for dissimilarities in varactors CR1 and CR2.

Performance:	Power Input at 100 Mc	36 mw
	Power Output at 200 Mc	16 mw

3.4.10 Lower Sideband Up Converter - DC-16-3005-4

The Lower Sideband Up Converter is similar in operation to the USUC. In this application the lower sideband output frequency was desired from the translator. This up converter translates a 200 Mc input signal frequency

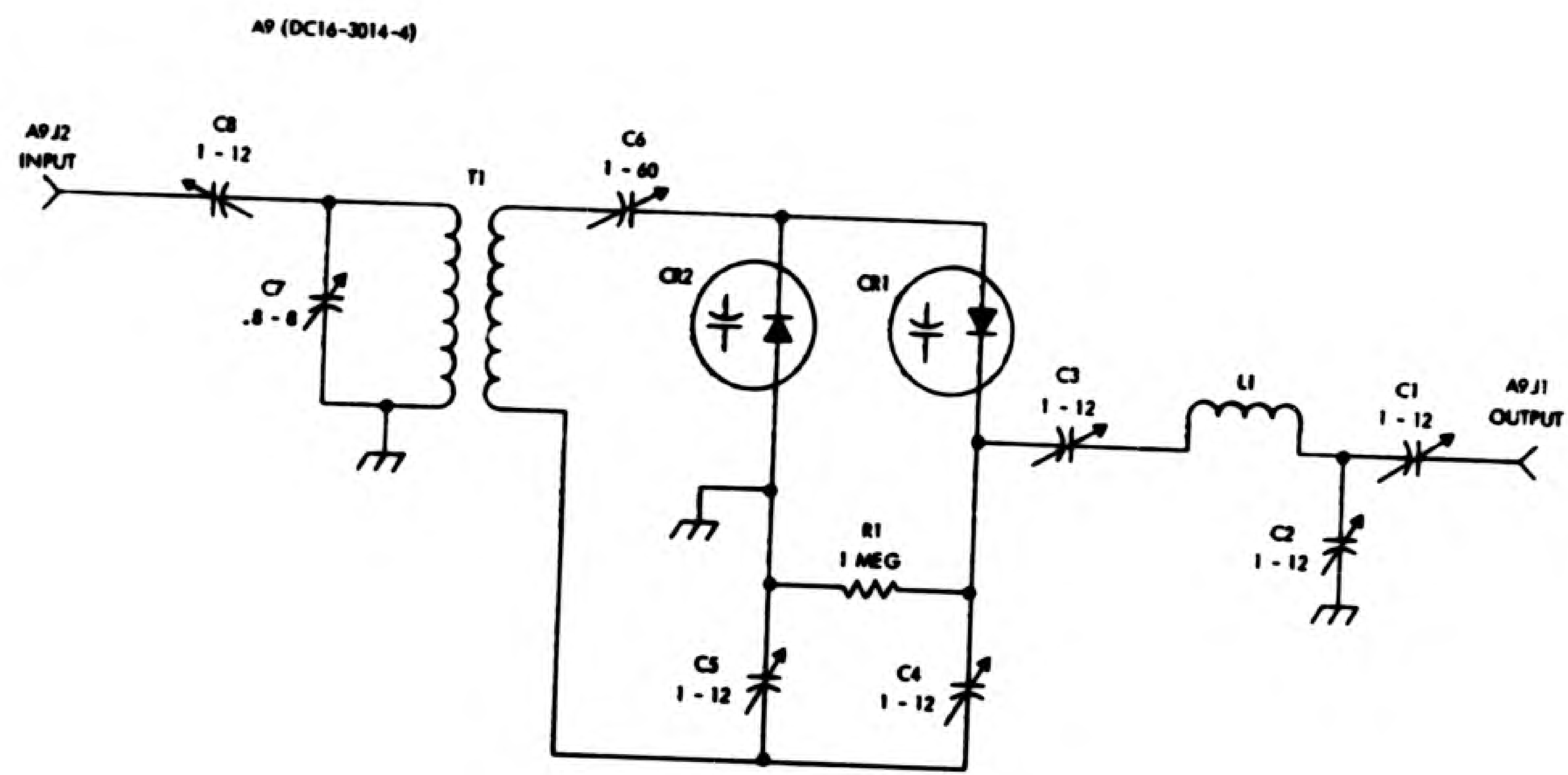


Figure 3-16. X-2 Varactor Multiplier (100/200 Mc)
Schematic Diagram

to an output frequency of 7800 Mc. A schematic diagram is shown in Figure 3-17.

The Lower Sideband Up Converter consists of an input filter, a translator, and an output filter. A diagram of the waveguide components is also shown in Figure 3-17. The input filter is a two stage inductive iris coupled band-pass filter. This filter passes the pump frequency with minimum loss and rejects the translated frequencies.

The translator, as in the Upper Sideband Up Converter, incorporates a varactor diode positioned across the center of the waveguide and functions as a mixer. The 200 Mc input frequency is coupled to the diode through a coaxial bias tee. The diode is operated self-bias and a terminating resistor of 3600 ohms is located on the bias tee.

A quarter wave coaxial RF choke and mica bypass capacitor present a low impedance to the pump and translated frequencies, resulting in isolation of the high frequencies from the 200 Mc input. The input 200 Mc signal is not propagated by the waveguide circuitry, and is, therefore, filtered out. A pair of capacitive tuning screws, one located on each side of the varactor, is used to match the diode to the waveguide circuitry. The distance from the diode to the input and output flange is made unsymmetrical to obtain an impedance match to the filters at their respective frequencies. The only difference between the translator used in the Upper Sideband Up-Converter and the Lower Sideband Up-Converter is the distance to the output flange. This length is longer in the Lower Sideband Up-Converter because of the lower output frequency.

The output filter is a narrowband five stage inductive iris coupled filter that is tuned to 7800 Mc. Its function is to pass the output frequency with minimum loss and to reject all other frequencies. Capacitive buttons,

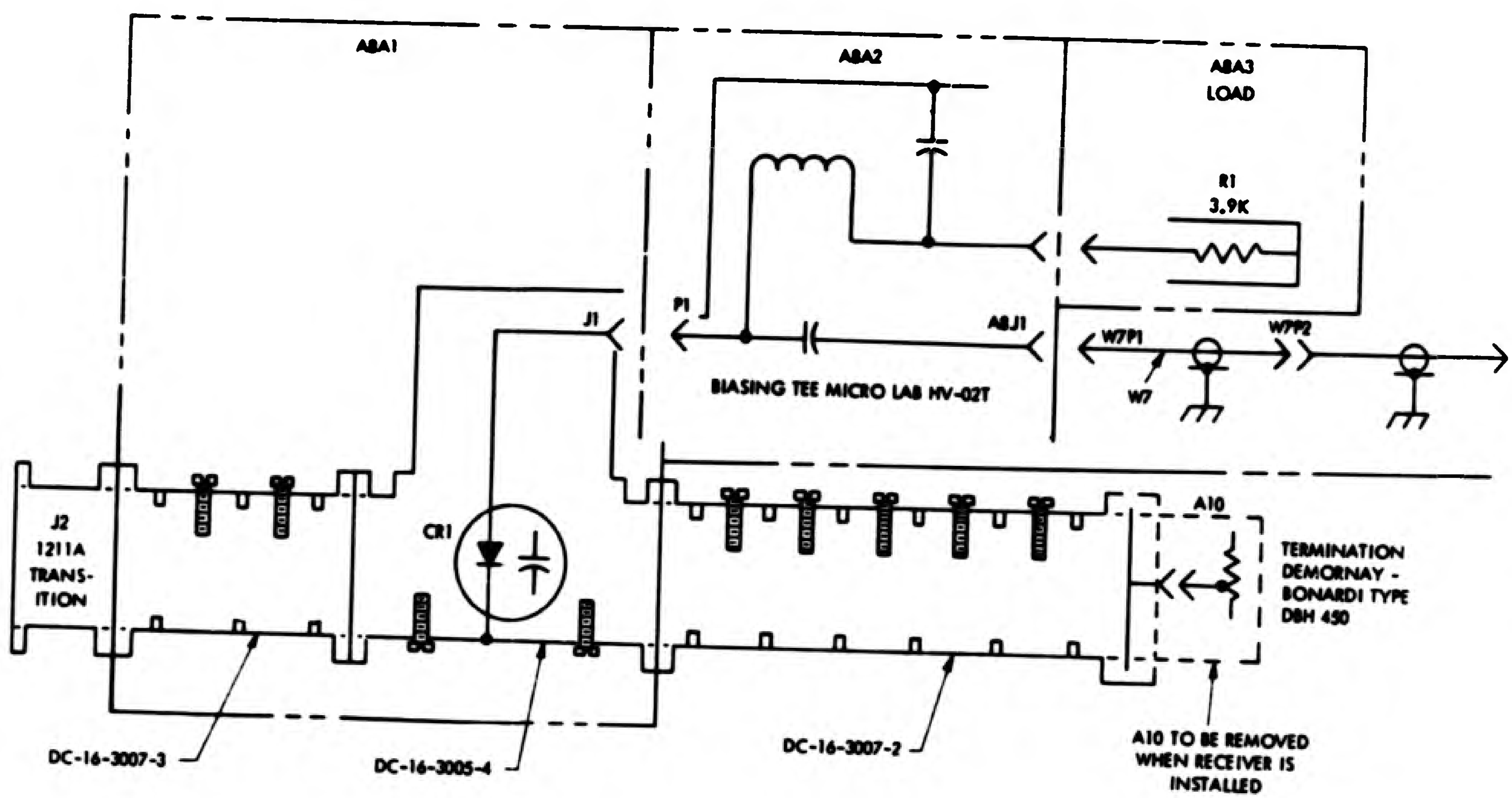


Figure 3-17. Lower Sideband Upconverter (8000/7800 Mc)
Functional Diagram

located on the narrow waveguide wall, were employed to obtain the required bandpass characteristics. The electrical characteristics of the output bandpass filter are as follows:

Insertion loss at 7800 Mc	-	4.5 db
3 db bandwidth	-	20 Mc
Attenuation at ±20 Mc	-	38 db minimum

Typical operation utilizing a D4211K diode is as follows:

fp - 8000 Mc	Pp - 50 mw
fs - 200 Mc	Ps - 10 Mc
fo - 7800 Mc	Po - 8.6 Mc

Electrical characteristics of D4211K are:

C_{y-6} - .46 pf

F_{c-6} - 138 gc

Bv - 18^{v}

The Lower Sideband Up Converter was tested at various temperatures to determine operational limits. With a few changes made in the output filter, the Up Converter operated from -25°F to a +115°F with a maximum power output deviation of 0.78 db, and 1.05 db over a -25°F to +150°F.

3.5 Klystron Phase Servo Reference Frequency Generator

The servo system which maintains an exact additive phase relationship between the two VA823 klystrons employs a local oscillator and two mixers to heterodyne the X-band signal samples to a 50 Mc intermediate frequency for amplification and phase comparison. When the transmission frequency is changed from 7.630 Gc to 8.330 Gc, the local oscillator frequency in the phase servo must be changed also. The phase servo reference frequency (LO) is derived from the stable clock when the transmitter is operated at

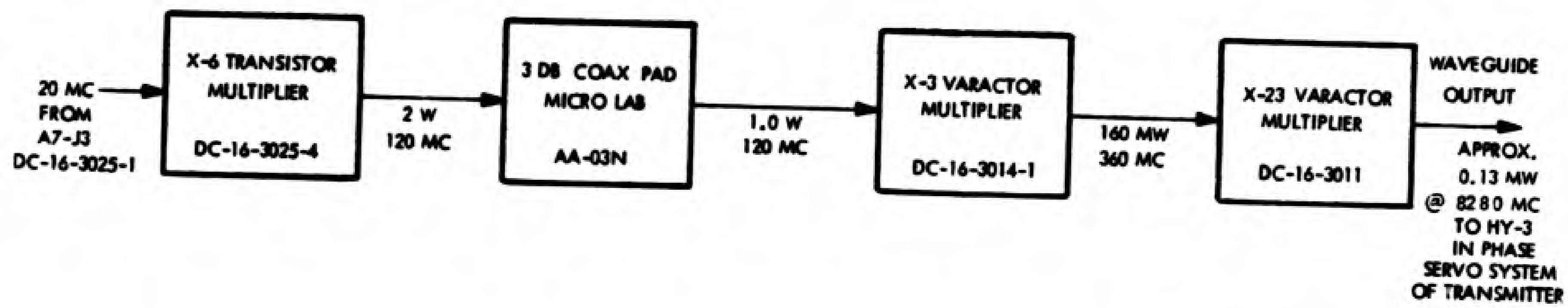


Figure 3-18. Klystron Phase Servo Reference Frequency Generator (8280 Mc) Block Diagram

8.330 Gc. Figure 3-18 is a functional diagram of the equipment involved. The X4 and X5 multiplier (DC-16-3025-1) is described in Section 3.4.1. It provides four milliwatts at 20 Mc to the X6 Transistor Multiplier - Amplifier (DC-16-3025-4).

3.5.1 X-6 Transistor Multiplier/Amplifier - DC-16-3025-4

3.5.1.1 Description

The X-6 Transistor Frequency Multiplier accomplishes multiplication in the first stage to provide the 120 Mc output from a 20 Mc input. The 120 Mc signal is then amplified to an 2.0 watt level by eight stages of transistor amplification. The block diagram and schematic diagram for this module are shown in Figure 3-19 and 3-20.

3.4.1.2 Specifications

The design specifications are listed below:

Temperature Range:	15°C to 44°C
Output Power Variation:	±.2 db
120 Mc Power Output:	2.0 watts nominal
20 Mc Input Power:	4 mw
Spurious Response: (In band)	90 db down

3.5.1.3 Design Considerations

The method of multiplication and circuit design of the first five stages employed is discussed in Section 3.1.3. The sixth and seventh amplifiers use 2N2369 transistor in a ground emitter configuration and provide the required drive to the final amplifier which employs an RCA 2N2876 operating Class C to obtain the required output of 2.0 watts. Class C operation is used in this stage to realize high efficiency operation.

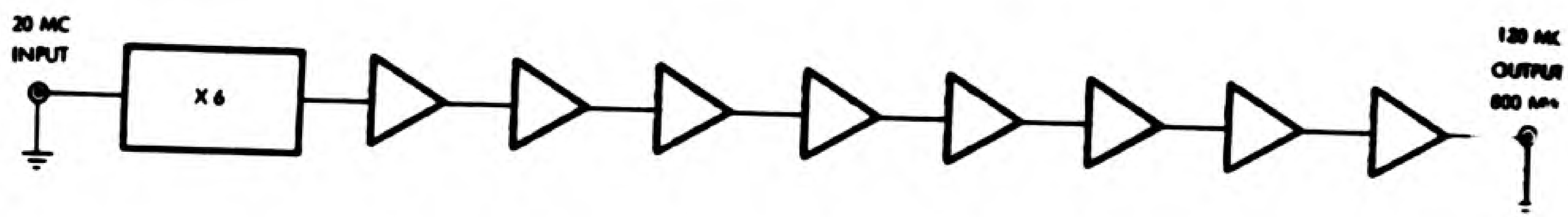
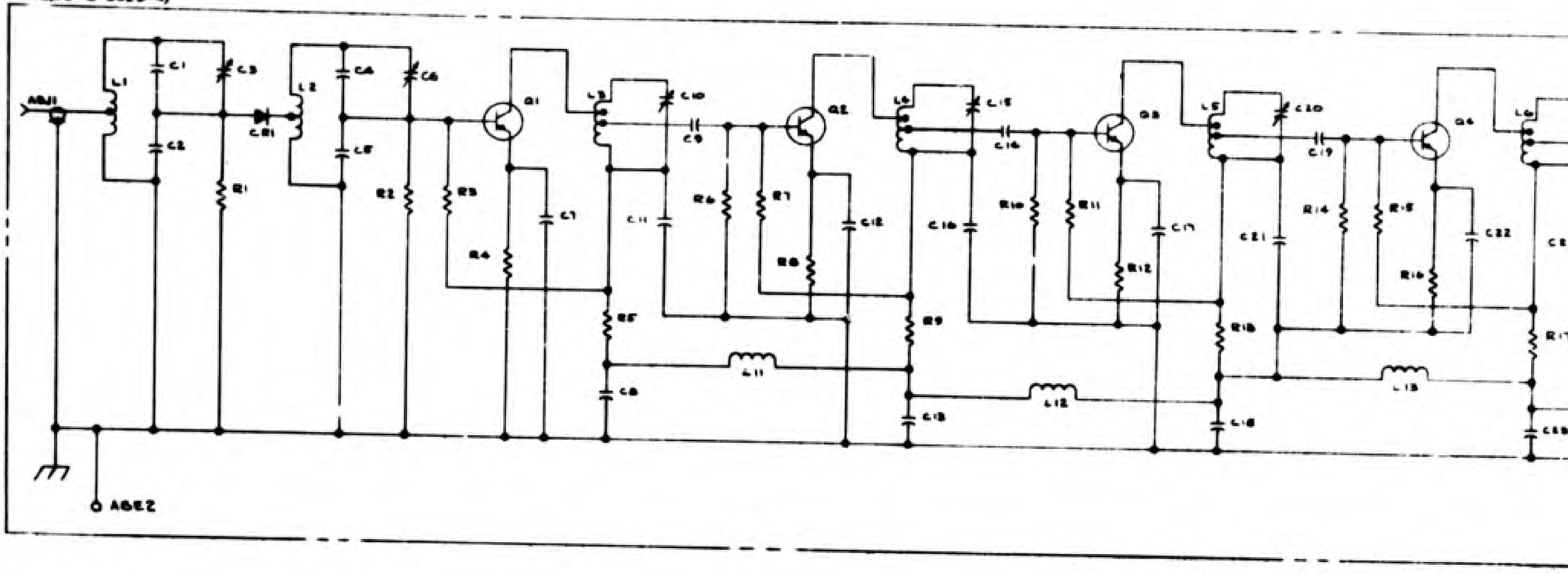


Figure 3-19. X-6 Transistor Multiplier (20/120 Mc) Block Diagram

A6(DC-16-3025-4)



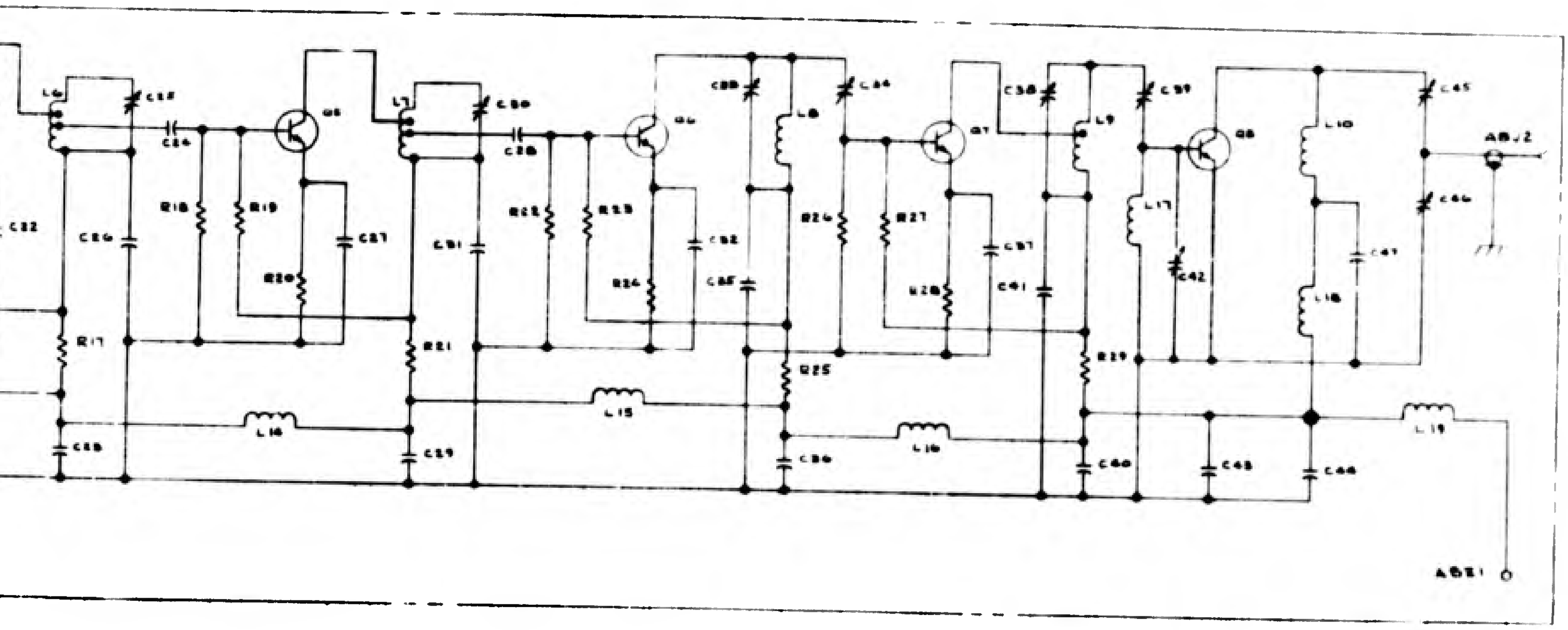


Figure 3-20. X-6 Transistor Multiplier (20/120 Mc.) Schematic Diagram

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The bandpass characteristics of each amplifier are designed to provide a response 10 db down at B'm = 4 from that obtained at the center frequency of 120 Mc. Therefore, each tuned tank provides approximately 10 db rejection at the frequencies of 100 and 140 Mc. The rejection of the 100 and 140 Mc spurious frequencies realized in the 120 Mc amplifier was 70 db. The spurious frequencies within ±5 Mc of the center frequency were suppressed more than 90 db.

3.5.2 X-3 Varactor Frequency Multiplier (120/360 Mc) DC-16-3014-1

Figure 3-21 is a circuit schematic of this multiplier. Capacitors C1 and C2 provide a means of matching the multiplier input impedance to the source impedance. Similarly, C7 and C8 permit matching the multiplier output to the load impedance. The elements L1, C3 and the bridge network form a series resonant filter at the 120 Mc fundamental frequency. The elements L2, L3 and C6 form a bandpass filter which passes the 360 Mc third harmonic and rejects the fundamental frequency.

The left and right hand branches of the bridge configuration are identical except for the reversal in varactor polarity. As a result of this, any second harmonic voltage developed across one branch will be of reverse phase from that across the other. This means that the second and all even harmonic voltages taken in sequence in passing around the bridge quadrangle will add in series. This implies that even harmonic currents will circulate freely around the bridge and through the varactors. Expressed in other words, the second harmonic is "idled" in the bridge composed of CR1, CR2, C4 and C5. This circulation of second harmonic through the varactor results in a mixing of the second harmonic with the fundamental and augmentation of the third harmonic output.

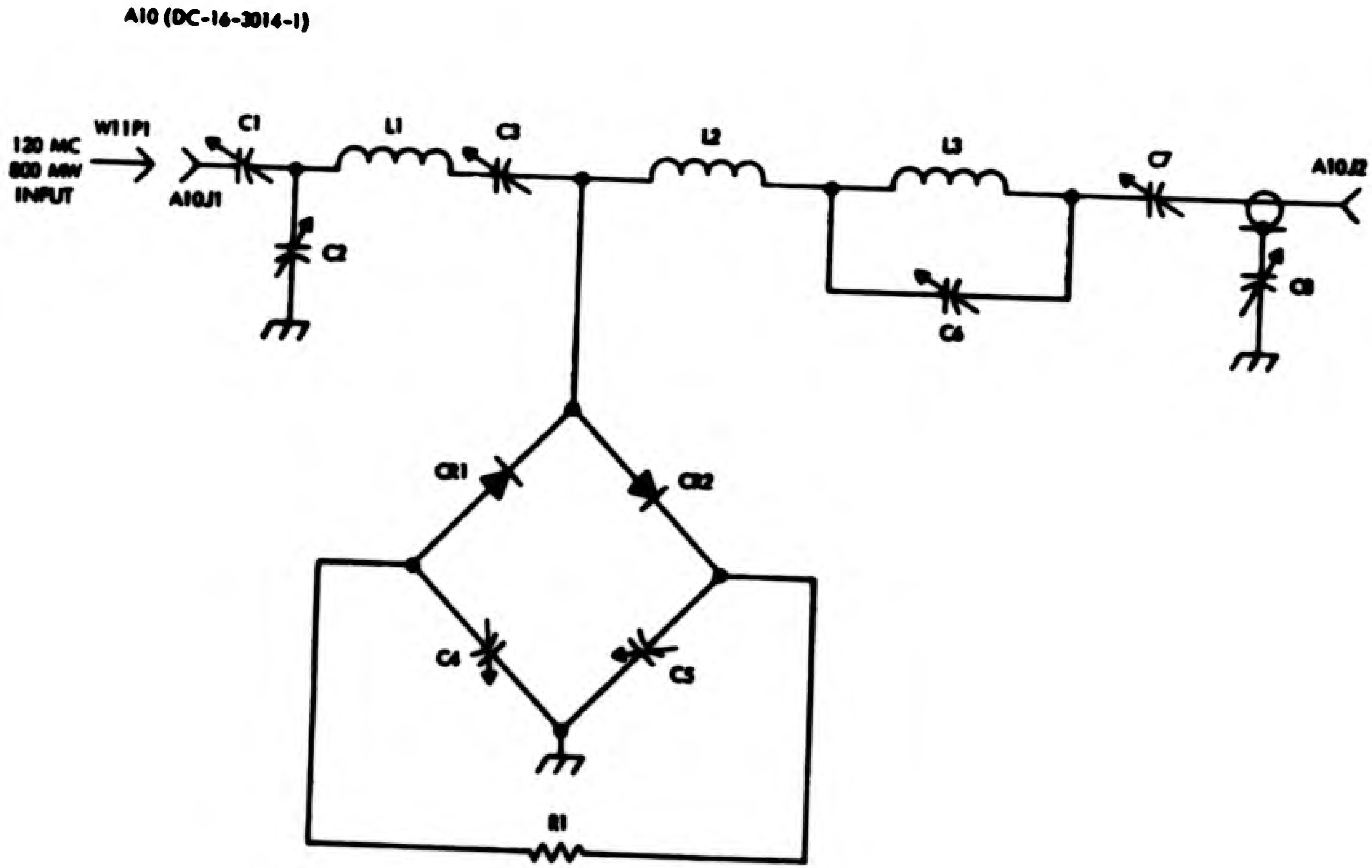

Figure 3-21. X-3 Varactor Multiplier (120/360 Mc) Schematic Diagram

3.5.2.1 Performance

Input Power at 120 Mc:	500 mw
Output Power at 360 Mc:	230 mw
Spurious Outputs:	120 Mc < 0.1 mw
	240 Mc 0.4 mw
	480 Mc 0.1 mw

3.5.3 X-23 Varactor Frequency Multiplier (360/8280 Mc) DC-16-3011

This module is substituted for the 7580 Mc multiplier XCR8, isolator E-18, and filter FL10 in the GFE transmitter when the transmission frequency is 8330 Mc. (Symbol designations refer to G.E. Drawing C48C600188.) In the original transmitter design, the drive for the 7580 Mc multiplier was obtained from a separate oscillator-amplifier chain. This system, which supplies a local oscillator reference signal to the klystron phase servo, is not altered when the transmitter is operating at 7630 Mc. When the transmitter is operated at 8330 Mc, the phase servo L.O. signal is obtained from the frequency standard via the transistor multipliers DC-16-3025-1, DC-16-3025-4, the X3 varactor multiplier DC-16-3014-1 and this module (DC-16-3011).

The X-23 multiplier was designed in a hybrid lumped element-coaxial-waveguide circuit. This configuration is a natural consequence of the choice of input frequency and multiplication ratio. The input and output frequencies are respectively 360 Mc and 8280 Mc. A varactor diode is employed as the non-linear element.

A diagram of the multiplier circuit is shown in Figure 3-22. The lumped element circuitry performs the function of matching the input power source to the multiplier input. The coaxial structure incorporating the varactor diode is positioned across the center of the waveguide and serves to support

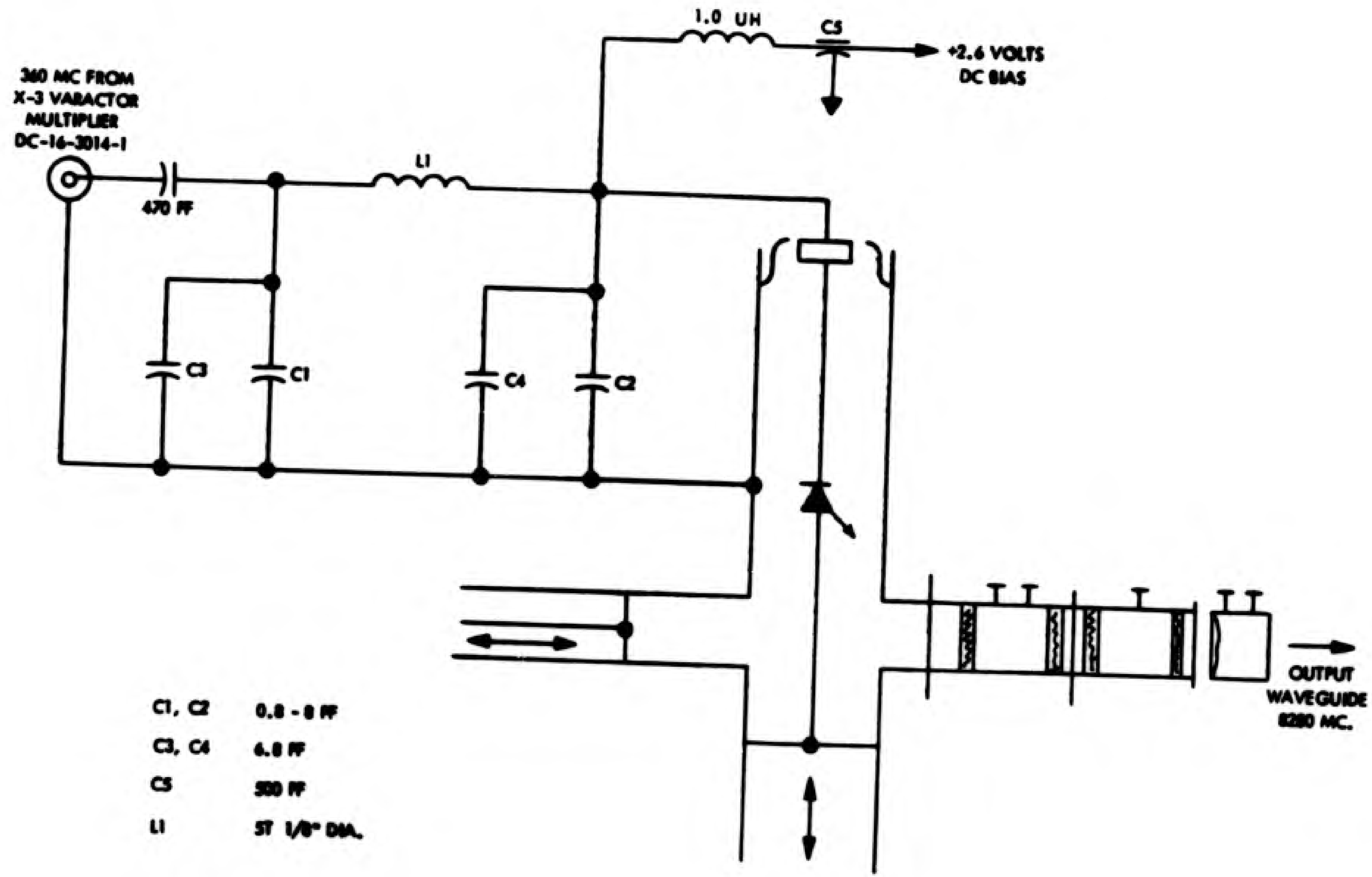

Figure 3-22. X-23 Varactor Multiplier (360/8280)
Schematic Diagram

the lower frequency components not supported in the waveguide and of transitioning to the waveguide. The waveguide portions of the circuit perform the functions of output frequency selection, output matching and high frequency idler tuning. Output frequency to input circuit isolation is achieved by an annular choke and mica bypass capacitor. The input frequency and lower harmonics are rejected by the waveguide circuit. Selection of the 23rd harmonic is accomplished by a two post capacity tuned filter. Output matching is provided by a two post capacitive screw tuner. Idler adjustment is afforded by a two post tuner preceding the filter. The diode case capacity and the average junction capacity at the output frequency is parallel resonated by the adjustable short circuit. The coaxial adjustment extension is used to tune out the effects of the waveguide-coaxial bottom gap. The tuning of these various elements are interdependent so that the functions delineated are to be considered as first order effects.

3.6 700 Mc Offset Frequency Generator

One of the requirements of the Modification Program was to provide means by which the Terminal transmission frequency could be shifted from 7.630 Gc to 8.330 Gc. The ability to operate at 7.630 Gc was not to be impaired by the changes. The requirement was met by supplying a kit of interchangeable components; one set for each frequency. The exciter frequency shift is accomplished by adding an upper sideband up converter (mixer) in the signal path from the original exciter. A 700 Mc offset frequency derived from the stable clock is mixed with the 7630 Mc output of the original exciter to obtain 8330 Mc having the stability of the station clock.

In this way FM modulation of the 7630 Mc waveform is preserved in the 8330 Mc output. When the 7630 Mc carrier is unmodulated, the 8330 Mc output becomes the carrier input to the single sideband modulator. Figure 3-23 shows the equipment used to shift the exciter frequency.

3.6.1 100 Mc Amplifier - DC-16-3025-3

This unit supplies 100 milliwatts at a frequency of 100 Mc to the X-7 varactor multiplier. It is described in Section 3.4.2.

3.6.2 X-7 Varactor Frequency Multiplier - DC-16-3014-3

Figure 3-24 is a schematic diagram of this unit. It resembles the X-5 varactor multiplier DC-16-3014-2 and the X-3 varactor multiplier DC-16-3014-1. In the X-7 multiplier, idler circuits are provided to permit free circulation of 3rd and 5th harmonic currents. The tapped coil in the input network acts as an impedance transformer to match to the relatively low input impedance of the varactor bridge. This circuit is superior to the series resonant input network in that variations in the average varactor capacitance do not affect it so seriously.

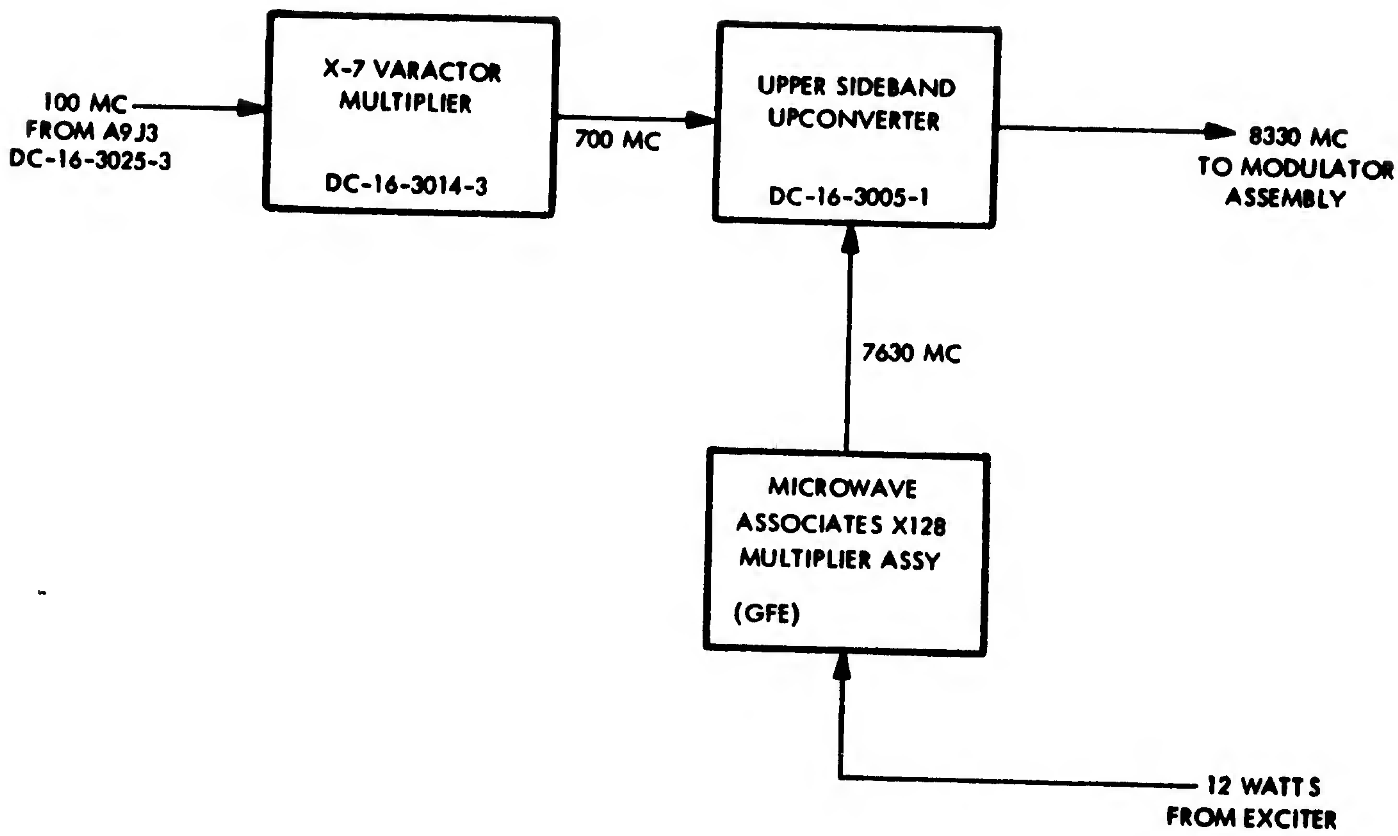

Figure 3-23. 700 Mc Offset Frequency Generator Block Diagram

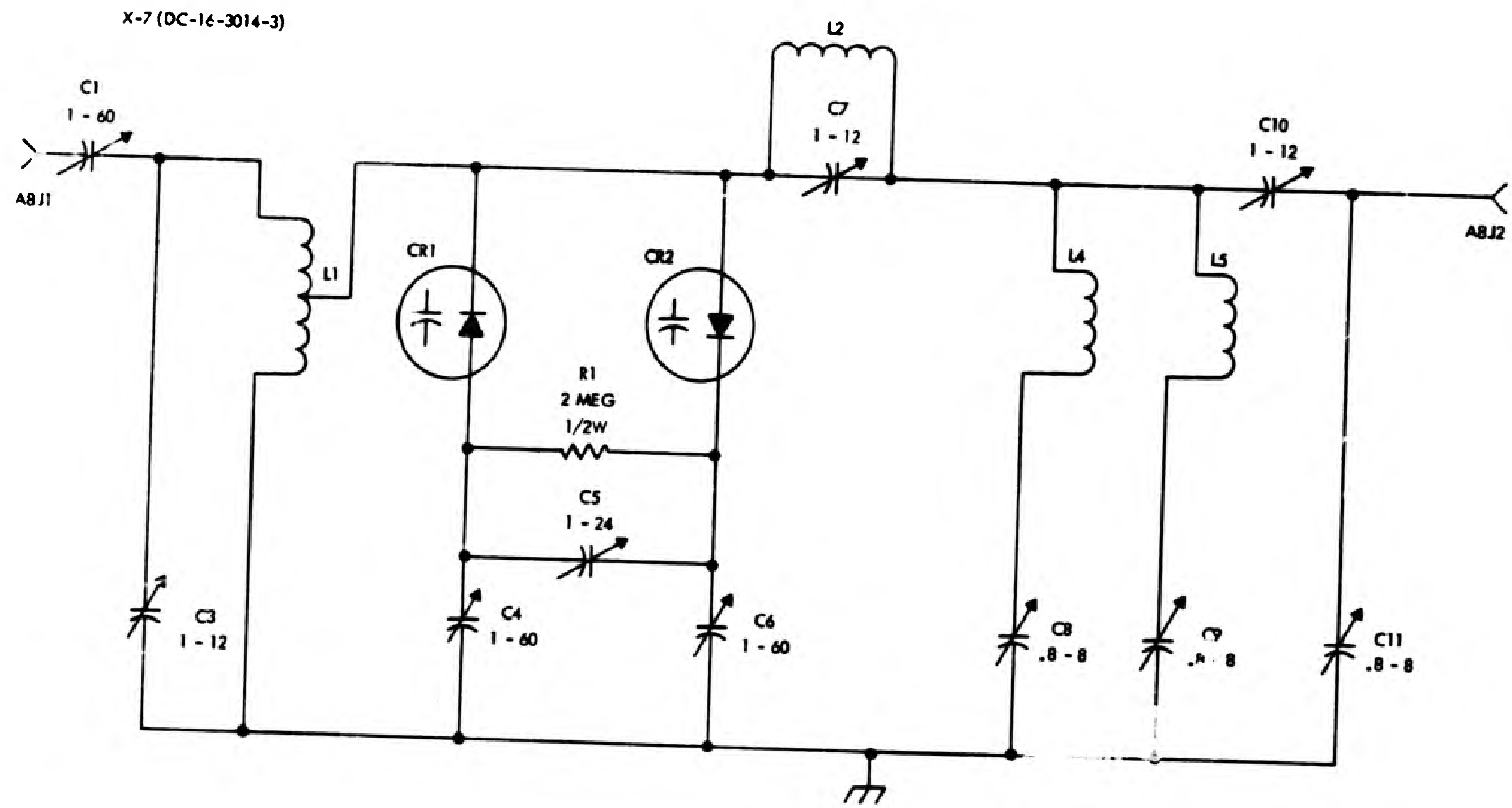

Figure 3-24. X-7 Varactor Multiplier (100/700 Mc)
Schematic Diagram

3.6.3 Upper Sideband Up Converter - DC-16-3005

Upper Sideband Up Converter is implemented in waveguide and translates an input frequency of 700 Mc to an output frequency of 8330 Mc. This requires a translator frequency of 7630 Mc. When the terminal is operated on FM, this is the modulated carrier. A functional diagram is shown in Figure 3-25.

The Upper Sideband Up Converter consists of an input filter, a translator, and an output filter. A diagram of the waveguide components is shown in Figure 3-26. The input filter is a three stage inductive iris coupled bandpass filter used to pass the pump signal frequency with minimum loss and to reject the translated frequencies. Electrical characteristics of the input filter are as follows:

Insertion loss at 7630	-	0.55 db
3 db bandwidth	-	88 Mc
Attenuation at ±100 Mc	-	24 db minimum

The translator incorporates a varactor diode that functions as a mixer and is positioned across the center of the waveguide. The 700 Mc signal frequency is coupled to the diode through a coaxial bias tee with the bias arm terminated in 10 ohms. An RF quarter wave choke and mica bypass capacitor combination presents a low impedance to the pump and translated signals resulting in isolation of the high frequencies from the 700 Mc input. The input frequency is not propagated by the waveguide circuitry and is, therefore, filtered out. A pair of capacitive tuning screws, one located on each side of the varactor, is used to match the diode to the waveguide circuitry. The distance from the diode to the input and output flange is made unsymmetrical to obtain the proper impedance match to the filters at their respective frequencies.

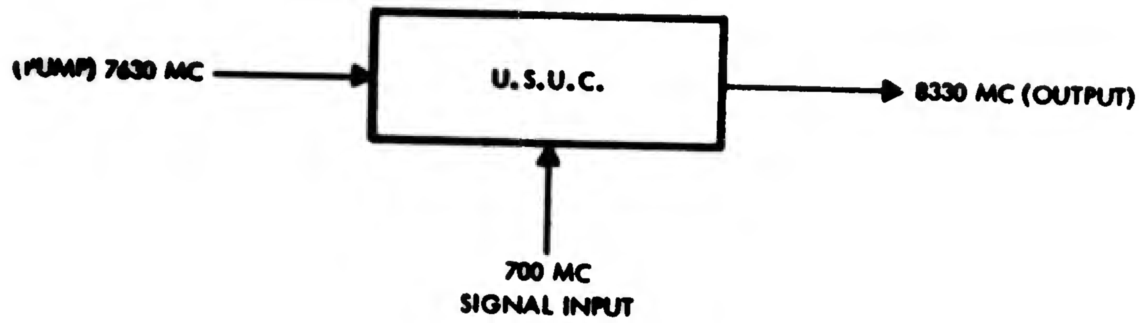

Figure 3-25. Upper Sideband Upconverter (7630/8330 Mc) Functional Diagram

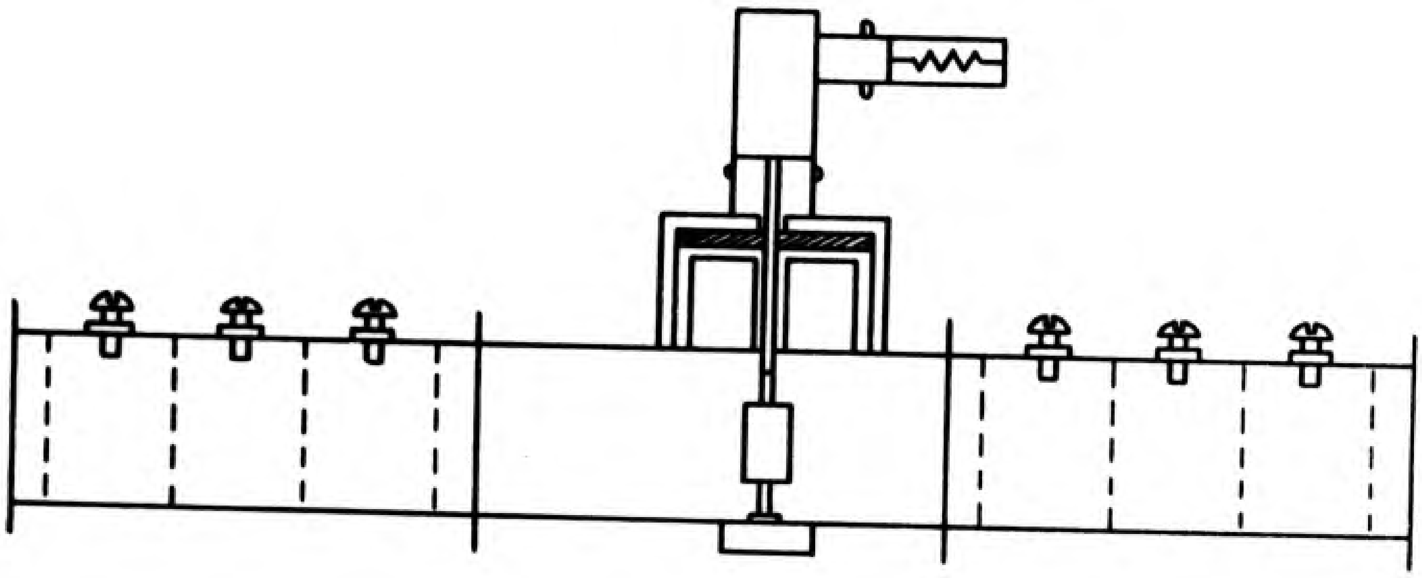

Figure 3-26. Upper Sideband Upconverter (7630/8330 Mc) Structural Diagram

The output filter is a three stage inductive iris coupled filter that is tuned to the output frequency of 8330 Mc. Its function is to pass the output frequency with minimum loss and reject the pump and lower sideband. The electricalcharacteristics of the bandpass filter are as follows:

Insertion loss at 8330 Mc	- 0.45 db
3 db bandwidth	- 85 Mc
Attenuation at ±100 Mc	- 27 db minimum

Typical operation utilizing a D4211K diode is as follows:

Fp - 7630 Mc	Pp - 50 mw
Fs - 700 Mc	Ps - 10 mw
Fo - 8330 Mc	Po - 3.7 mw

Diode characteristics of D4211K:

C_{-6} - .80 pf

F_{c-6} - 141 gc

Bv - 18^{v}

4.0 OPERATOR'S CHECK LIST

4.1 Operation of Transmitter at 7630 Mc

A. The upper sideband up converter (DC-16-3005-1) must be removed and the waveguide piece (DC-16-3003-12) inserted in its place.

B. The 7630 Mc antenna feed assembly (DC-16-3015-7) must be installed. The rod support wires (struts) must be at 45° to the vertical.

C. The diplexer waveguide coupler (DC-16-3003-4) and the waveguide section (DC-16-3003-8) must be removed and stored. The ends of the unused guides should be covered with plastic caps or taped, to keep out dirt and insects.

D. Transmission line waveguide section (DC-16-3003-14) must be installed to bypass the diplexer.

E. The X-band receiver cannot be used when the transmitter is operated at 7630 Mc. If it is present, the parametric amplifier should be disconnected from the waveguide system.

F. The traveling wave tube amplifier must be turned on. After about three minutes, adjust the cathode current to a level between 10 and 16 ma. The final setting will depend upon the drive required by the VA823 klystrons.

G. The 0.903172 Mc synthesizer must be turned on. If the clock power has been interrupted, it may be necessary to press the guarded red reset button on the clock panel to restart the internal frequency divider.

4.2 Operation of Transmitter at 8330 Mc

A. The waveguide section (DC-16-3003-12) must be removed and the upper sideband up converter (DC-16-3005-1/DC-16-8002-4) substituted for it.

B. The 8330 Mc antenna feed assembly (DC-16-3015-8) must be installed. The rod support struts should be oriented at a 45° angle to the vertical.

C. The waveguide section (DC-16-3003-14) must be removed from the antenna equipment box and stored.

D. The diplexer coupler (DC-16-3003-4) must be put in place between (DC-16-3003-5) and the diplexer.

E. Waveguide section (DC-16-3003-8) must be installed.

4.3 Changing the Transmitter Polarization

Waveguide sections DC-16-3003-6 and DC-16-3003-5, each have two alternate positions. To operate with horizontal polarization, insert DC-16-3003-5 in the upper waveguide branch from the dual mode transducer between sections DC-16-3003-2 and DC-16-3003-4. Insert DC-16-3003-6 in the lower waveguide branch between DC-16-3003-3 and DC-16-3003-7 (switch coupler). To operate with vertical polarization, rotate each section 180° about an axis parallel to its length so that DC-16-3003-5 now connects DC-16-3003-4 and DC-16-3003-3, and DC-16-3003-6 now connects DC-16-3003-7 and DC-16-3003-2.

APPENDIX I

Antenna Feed System Development

Two basic adaptations were required of the X-band antenna feed. These were to provide for both vertical and horizontal polarizations simultaneously and to permit transmission at the new transmitter frequency of 8330 Mc. Because the new feed would be used for both transmission and reception it would have to be well matched and produce a satisfactory radiation pattern at both the transmit and receive frequencies.

The change in transmitter frequency from 7630 Mc to 8330 Mc would, when effected, be permanent. Thefeed design problem was simplified by the decision to provide two separate feed elements, one for each transmission frequency.

Figure I-1 is a photograph of the original X-band feed assembly. It consists of a disc-on-rod or "cigar" type of slow wave structure supported in the mouth of a short rectangular horn by means of a transverse metal vane. The geometry of this horn and the presence of the transverse vane limit this feed assembly to a single linear polarization even though the disc-on-rod slow wave structure will support waves polarized at any angle about the axis of the rod. Two perpendicular polarizations can be launched from horns having axial symmetry such as square or circular horns. A square section wave launcher was considered but was discarded as being more difficult to fabricate and as a result of constructional difficulties also more likely to produce cross coupling between the two orthogonal modes than would a horn of circular cross-section. Directly related to the feed horn problem is the problem of launching two separate orthogonally polarized waves into the horn from separate drive ports.

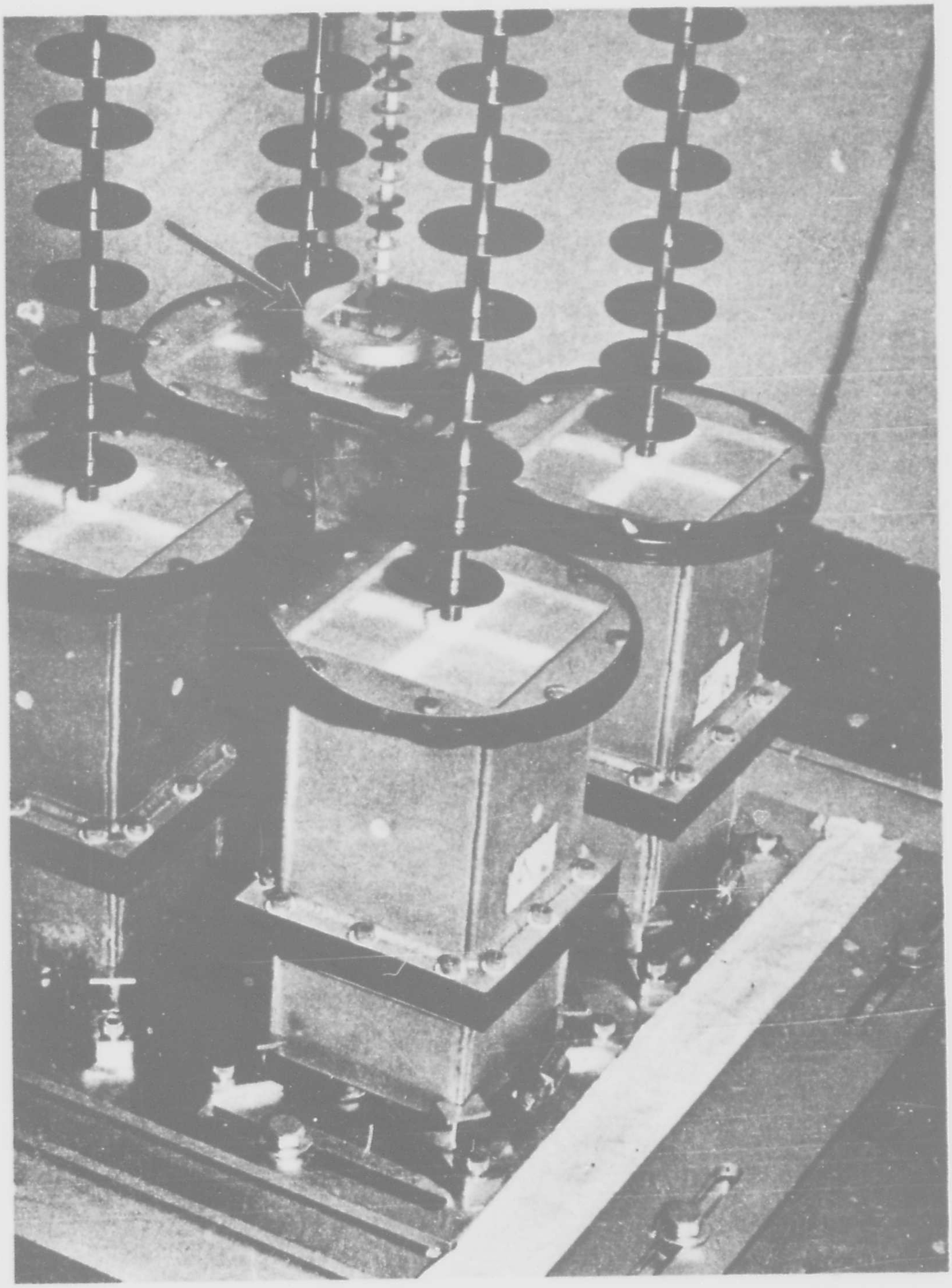

Figure I-1. Original X-Band Feed Assembly

This requires a three port junction or "dual mode transducer". Because the perpendicular modes in the common arm are not coupled, the junction behaves much the same as a four port device. The design of a junction of this form which would remain matched over the required frequency band is a problem in itself. Fortunately devices of this kind find application in Faraday rotation circulators and several designs have been produced by specialists in this field for their own use. A Ferrotec Inc. model DMT-120 was selected as being suitable for the TPSC Terminal feed. This unit has two inputs in WR112 waveguide and a circular output guide having an inside diameter of 1.047 inch. Isolation between the input parts when the output is properly terminated is greater than 40 db. The vendor's inspection test data for the mode transducer used in the TPSC terminal antenna feed are tabulated below:

<u>Dual Mode Transducer</u>

Ferrotec Inc., Model DMT 120, Serial No. 2. Data from vendor's tag.

Frequency	Isolation	VSWR Port 1	VSWR Port 3
7500	44.5	1.06	1.04
7600	44 +	1.10	1.01
7800	44 +	1.08	1.02
8100	44 +	1.02	1.06
8300	44 +	1.10	1.07
8500	47 +	1.11	1.09

Selection of the Ferrotic DMT-120 dual mode transducer defined the diameter and shape of the guide connecting it to the feed horn.

At this point the feed horn design problem was defined to the extent that it was known that it was to be driven from a 1.047 inch I.D. circular waveguide propagating the TE_{11} mode, was to have a circular cross-section and should produce the same radiation pattern as the original feed system. One way of looking at the operation of surface wave antennas of this type is to regard the overall radiation pattern as resulting from the superposition of the radiation directly from the feed horn operature and the radiation from the slow wave structure. In the TPSC terminal feed modification it was desired to preserve the original radiation pattern, so the area of the circular horn operature was set equal to the area of the original rectangular operature, (without the transverse vane.) The length of the tapered section was made about three guide wavelengths. This horn together with the dual mode transducer described above but without the disc-on-rod slow wave structure produced the following standing wave ratios:

Freq.	7630	7730	7750	8330	8350
VSWR	1.18	1.06	1.035	1.17	1.11

To check the effect of the disc-on-rod element upon the match, the element was supported in the appropriate position with a piece of low density polystyrene foam. The effective dielectric constant of this material is so close to one that it produces virtually no reflections when placed in or in-front-of the horn. A VSWR of 1.05:1 or less could be achieved with this horn and a replica of the original disc-on-rod element supported in a polystyrene foam block. The experiment just described was encouraging but there still remained the question of how to obtain a solid support for the slow wave structure in the horn without creating a mismatch.

A change in the dielectric constant of the medium filling a waveguide will not produce any net reflected components if the dielectric loaded portion of the guide is some integer multiple of one half of the guide wavelength in the loading medium. This approach was employed to provide a support for the rod. The rod was threaded into the center of a teflon cylinder one guide wavelength long machined to provide an interference fit in the mouth of the horn. By positioning the rod axially it was possible to obtain a VSWR of less than 1.02:1 at the design center frequency. Radiation pattern measurements were performed on this assembly. The patterns obtained were almost identical to those for the original feed assembly. As this design appeared satisfactory on a low power basis, it was decided to test it at full power. The feed assembly and mode transducer were connected to the transmitter at the TPSC terminal and were allowed to radiate 13 Kw Cw vertically (without any reflectors) for about 20 minutes. At the end of this time the teflon support had become seriously deformed through the combined effects of thermal expansion and the constraint of the horn. Not being able to expand radially, the teflon expanded axially instead. Upon cooling it shrank both radially and axially to the extent that what had previously been a tight press fit became a very loose fit with perhaps 0.030" clearance. As the thermo-mechanical behavior of teflon was unsatisfactory a fused quartz support plate one guide wavelength long was made. Because of the difficulty in machining such a hard material it was not feasible to form a threaded center hole. In lieu of this a thin-walled metal sleeve was inserted and retained by press fitted collars. The quartz cylinder was held in the horn by shrinking the horn onto the quartz. The "cold" interference on diameters was approximately 0.002". This assembly matched out satisfactorily and suffered no ill effects under the high Cw power level. However, when it was thermally cycled the quartz cracked at -30°F. This was due to the very large compressive stress developed because of the differing coefficients of thermal expansion of the quartz and the brass horn. In the assembly tested, the horn wall surrounding the quartz cylinder was 0.062" thick. It is believed that were the wall made thinner or if a metal having a smaller thermal expansion coefficient were used for the horn, this fracture problem would be overcome. The fused quartz supports were not pursued further because of the time and expense required to obtain new sample pieces and because a simpler support was developed. Metal supporting members may be used to hold the slow wave structure in the launching horn if they do not disturb the axial symmetry of the horn and if they do not create large reflections. Thin metal struts or vanes perpendicular to the electric field will not seriously perturb the guide, however, where crossed fields are to be transmitted simultaneously such supports would always be parallel to one of the fields. Four radial struts 90° apart and each inclined at 45° to the direction of the electric field will appear the same to each of two perpendicular fields and will not produce any unbalanced orthogonal reflections which could couple one mode to the other. The final design of the feed assembly employs two in-line-sets of four radial wires each 0.030" in diameter to support a thin-walled internally threaded bushing into which the disc-on-rod element is screwed. The axial spacing between the two sets of wires is chosen so that the shunt susceptance introduced by one set is largely cancelled by the second set; i.e. by its susceptance transformed through a suitable length of line. Trimming adjustment to obtain VSWR's under 1.05:1 is achieved by axial positioning of the disc-on-rod element and where needed by altering the length of the rod locking nuts. The assembly of the support structure requires careful attention to spacing and centering of the parts. In order to facilitate assembly the supporting structure is built into a separate collar which is later brazed to the tapered horn. Initial tests and adjustments were conducted on both the 8330 Mc and the 7630 Mc assemblies before the collars were brazed in place. The brazing had little effect on the 7630 Mc unit but did alter the 8330 unit somewhat. Pattern plots for the two final assemblies are shown in Figures 1-8 and 1-9. Figures 1-10 and 1-11 are the corresponding VSWR plots. Figure 1-7 is a photograph of the completed feed assembly.

APPENDIX II

SSB-DSB Modulator Theory of Operation

The exciter portion of the TPSC Terminal transmitter employs a chain of varactor frequency multipliers to derive the X-band carrier from a crystal controlled low frequency standard. These multipliers will not pass amplitude modulated waves without introducing a large amount of distortion. Because of this it was decided to perform the double sideband suppressed carrier and single sideband modulation functions by operating on the X-band carrier directly. A phase discrimination type of modulator was selected. Figure II-1 is a functional representation of this configuration. If the two balanced modulators are identical,

$$V_1 = f \cdot F \cos\alpha \cos\beta$$

$$V_2 = f \cdot F \sin\alpha \sin\beta$$

using trigonometric identities

$$V_1 = \frac{f \cdot F}{2} \quad \cos(\alpha - \beta) + \cos(\alpha + \beta)$$

$$V_2 = \frac{f \cdot F}{2} \quad \cos(\alpha - \beta) - \cos(\alpha + \beta)$$

If V_1 and V_2 are summed

$$V_{out} = V_1 + V_2 = f \cdot F \left[\cos(\alpha - \beta)\right]$$

Or the difference can be taken as

$$V_{out} = V_1 - V_2 = f \cdot F \left[\cos(\alpha + \beta)\right]$$

This phase discrimination system is implemented in the TPSC terminal modification in the manner shown in Figure 2-5.

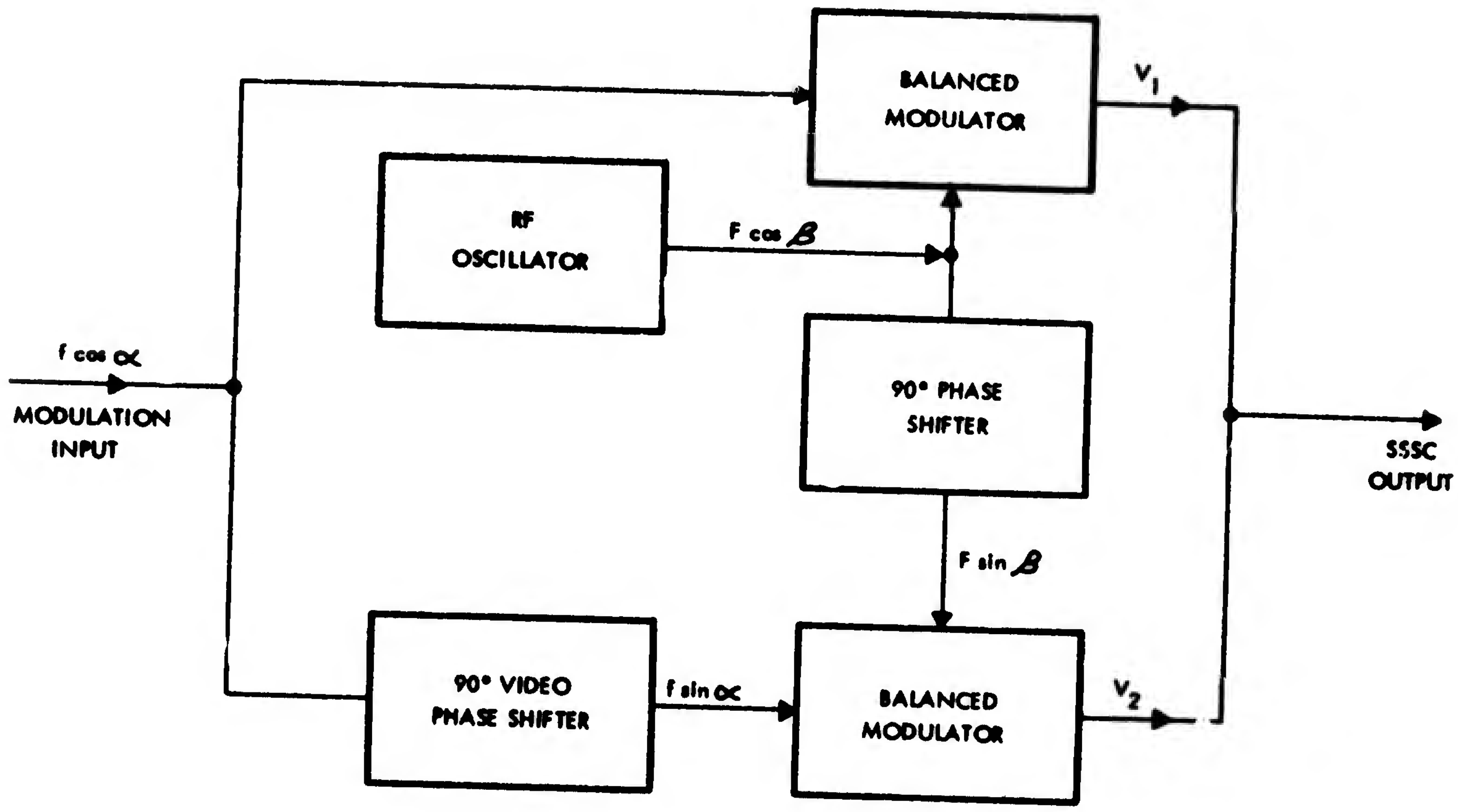

Figure II-1. Phase Discriminator SSB Modulator

Here the 3 db sidewall hybrid provides the necessary carrier power split and the 90 phase separation between outputs. Summing of the balanced modulator outputs occurs in the magic tee.

If the audio phase shift is removed, the balanced modulator output voltages become:

$$V_1 = f \cdot F \cos \alpha \cos \beta$$

$$V_2 = f \cdot F \cos \alpha \sin \beta$$

$$V_1 = \frac{f \cdot F}{2} \left[\cos (\alpha - \beta) + \cos (\alpha + \beta) \right]$$

$$V_2 = \frac{f \cdot F}{2} \left[\sin (\alpha + \beta) - \sin (\alpha - \beta) \right]$$

$$V_1 + V_2 = \frac{f \cdot F}{2} \cos (\alpha - \beta) + \frac{f \cdot F}{2} \cos (\alpha + \beta) + \frac{f \cdot F}{2} \sin (\alpha + \beta) - \frac{f \cdot F}{2} \sin (\alpha - \beta)$$

$$= \frac{f \cdot F}{2} \left[\cos (\alpha - \beta) - \sin (\alpha - \beta) + \cos (\alpha + \beta) + \sin (\alpha + \beta) \right]$$

$$= \frac{f \cdot F}{2} \left[\cos (\alpha - \beta) + \theta \right] + \frac{f \cdot F}{2} \left[\cos (\alpha + \beta) + \phi \right]$$

Both sidebands are present, and the carrier is missing. Double sideband suppressed carrier operation is achieved by removing the quadrature phase relationship between the two audio drive signals to the balanced modulators.

The degree to which the carrier and unwanted sidebands can be suppressed in a modulator of this kind depends upon the balance of the crystals in the individual balanced modulators and the similarity of the two balanced modulators. The modulator assembly used in the TPSC Terminal equipment is a commercial unit. Tuning screws have been added to the crystal mounts to permit balancing of stray reactive components of diode impedance.

APPENDIX III

Operation and Maintenance Notes - Test Data

Spectral Composition of Microwave Associates
X-128 Frequency Multiplier (Original Equipment)

Conditions:

Input signal to exciter was supplied from the FE1000A frequency standard and the 0.903172348 Mc synthesizer. Equipment was set up as shown in Figure III-1.

Test Data:

Frequency (gc)	Power Level (dbm)
7.540	-38
7.575	-37
7.600	-27
7.635	+6
7.640	+1
7.660	+6
7.720	-16
7.780	-18.5
7.840	-28
7.875	-31.5
7.900	-11.0
7.960	-30
8.040	-30
8.400	-23
8.648	-20.5

Total RF power output, primarily at 7.630 gc was 60 mw.

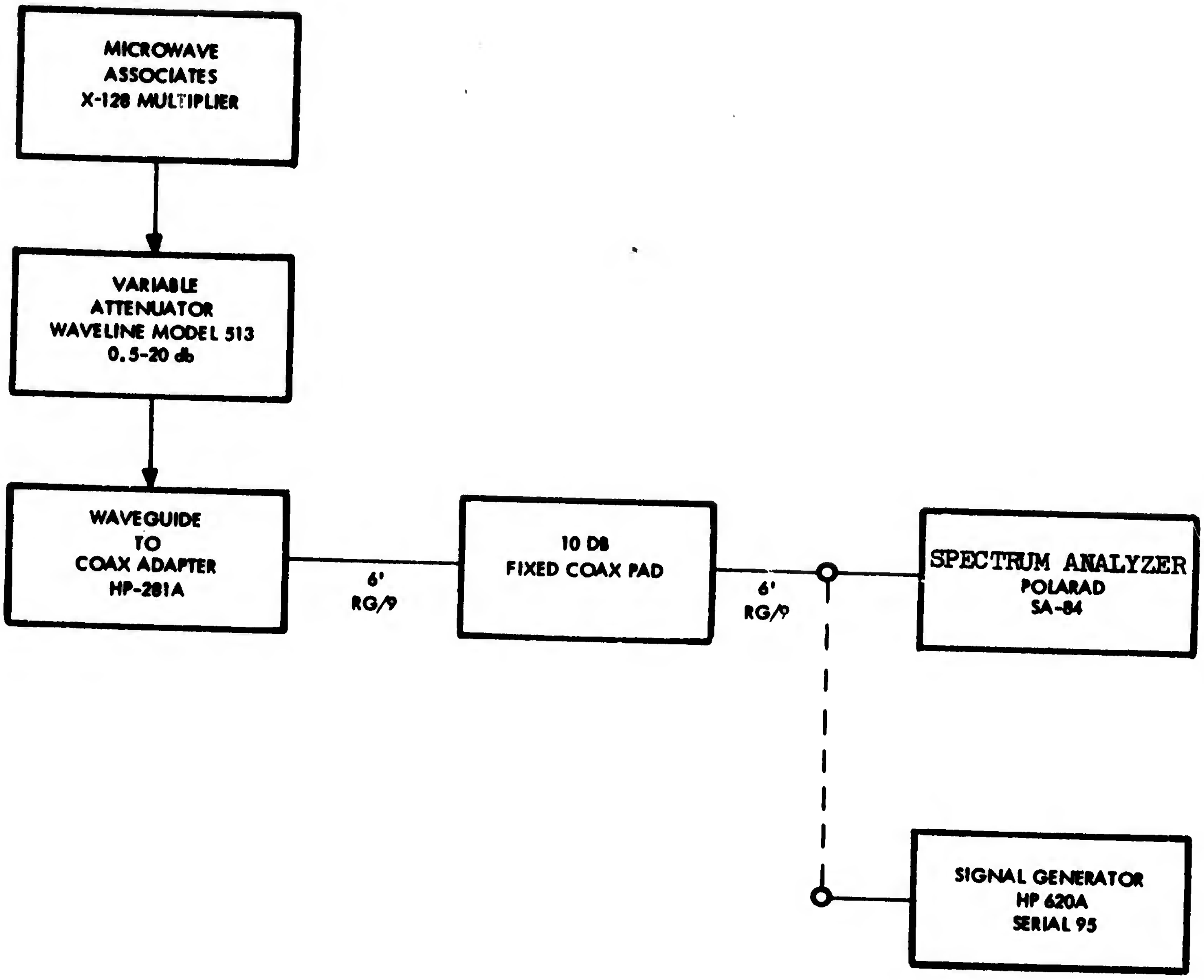

Figure III-1 Equipment Set-Up for Measuring Spectral Composition of Exciter Output

Test of Klystron Phase Servo Reference Frequency Generator

Initial attempts at adjustment of the multiplier chain following installation in the equipment van disclosed a considerable interaction between the tuning adjustments of the three multipliers. To circumvent this difficulty the transistors in the 6th and 7th stages of the X-6 transistor multiplier (DC-16-3025-4), were changed from type 2N914's to 2N2369's. This change increased the power output to 2.0 watts permitting the insertion of a 3 db pad (Microlab AA-03N) between the X-6 transistor multiplier (DC-16-3025-4) and the X-3 varactor multiplier (DC-16-3014-1) for isolation of these two units. With the addition of the 3 db pad, the adjustments of these units became independent of one another. In the original installation the X-3 varactor multiplier was located in the auxilliary rack #18 whereas the X-23 varactor multiplier which is driven by it, is located in the right hand bay of the transmitter cabinet. The X-3 multiplier was removed from rack 18 and mounted on a bracket to the X-23 multiplier. This change allows for ease of tuning of both multipliers and removes the effects of the long connecting cable.

The location of the 1 x 4 x 5 transistor multiplier (DC-16-3025-1) was changed to facilitate the replacement or checking of a fuse located on the front panel of the +28v DC power supply (DC-16-3019-1). In the original installation, the 1 x 4 x 5 transistor multiplier was located in front of the +28v power supply in rack 18, drawer 4. This 1 x 4 x 5 transistor multiplier was moved to the left side of the +28v power supply in the same drawer.

The schematic representation of the X-23 varactor multiplier is incorrect as shown on sheet 1 of DC-16-9001. The corrected schematic is shown in Figure 3-22.

The X-23 multiplier requires an external biasing voltage replacing the self biasing network. This voltage is derived from the +28v regulated DC power supply (DC-16-3019-1) located in the Time Base and Control Rack via an adjustable voltage divider. The voltage divider shown schematically in Figure III-2 is enclosed in a small metal "mini" box attached just to the right of the rear door of the right hand bay of the transmitter cabinet. In order to accommodate the additional wires needed for the varactor bias, J6(W105P1) shown on DC-16-9003 and DC-16-9004 have been changed from a Winchester PM6SLRN and PM6PLS, to an Amphenol H53102A-18-1S and H53106A-18-1P respectively. The new pin connections and wiring changes are shown in Figure III-3.

Following the changes just described and after retuning the multiplier chain, an output of 250 μw was obtained from the X-23 varactor multiplier. The spectral composition of this signal is shown in Figure III-4a.

The power output and frequency of each unit is listed as follows:

Unit	Power Output	Freq.
1 x 4 x 5 Mult. (DC-16-3025-1)	4 mw	20 Mc
X-6 Mult. (DC-16-3025-4)	1.8 w	120 Mc
3 db Pad	0.9 w	120 Mc
X-3 Mult. (DC-16-3014-1)	340 mw	360 Mc
X-23 Mult. (DC-16-3011)	250 μw	8.280 gc

Test of 700 Mc Offset Frequency Generator and Upper Sideband Up Converter

After installation of this multiplier chain and up converter was completed, it was found that insufficient power output from the 100 Mc amplifier (DC-16-3025-3) was available to drive the X-7 varactor multiplier (DC-16-3014-3), and that a high level of spurious signals was present. These

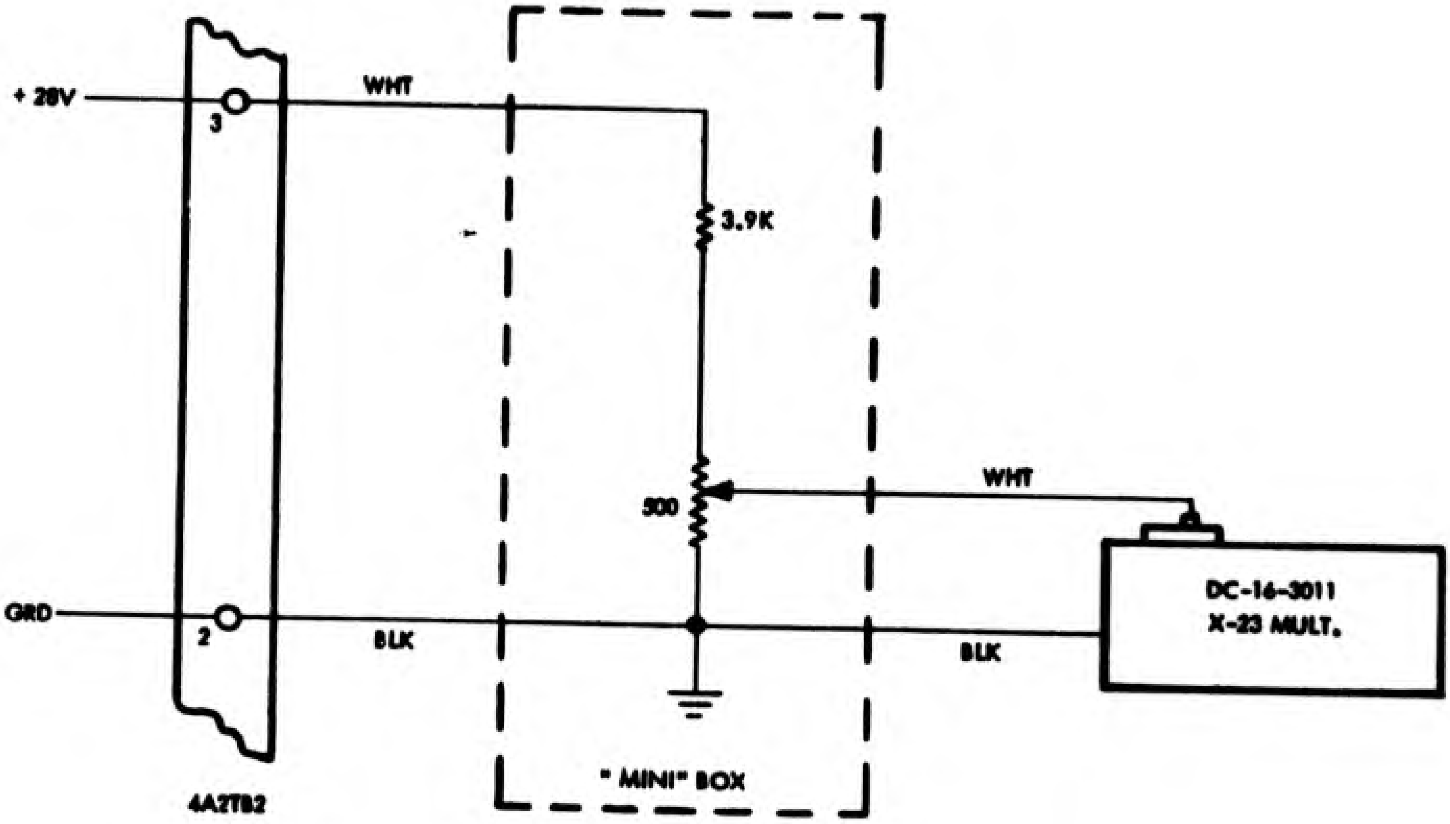

Figure III-2. Bias Circuit for X-23 Multiplier

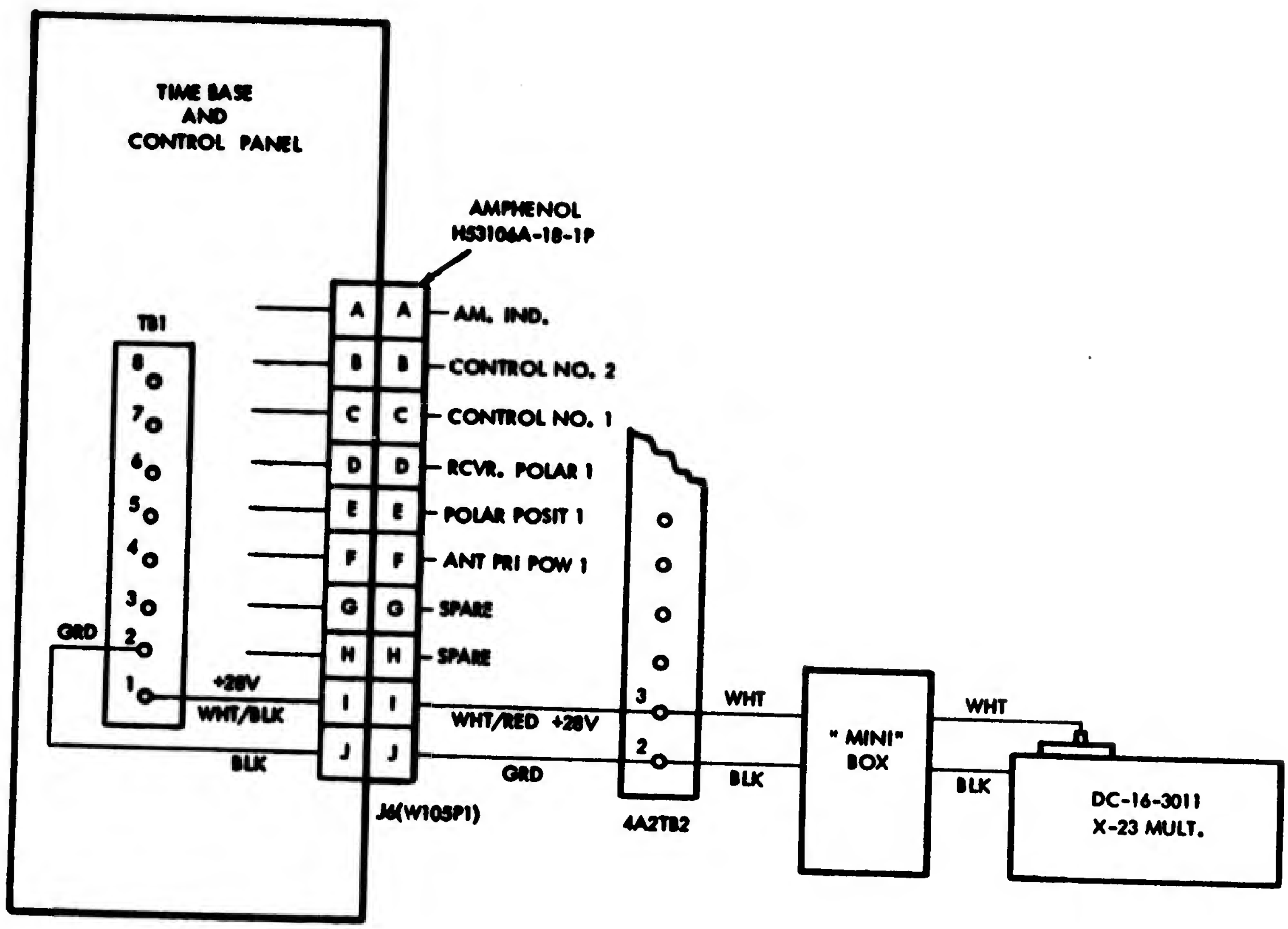

Figure III-3. Modification to Control Chassis Wiring difficulties were resolved by replacing a defective by-pass capacitor; providing an additional B+ RF bypass capacitor and retuning of the circuitry.

Difficulty in obtaining the necessary power output of the X-7 varactor multiplier at 700 Mc resulted in the following changes to the circuitry. The two PS1 diodes, CR1 and CR2 (PC-124's) were replaced with PC-141's. The series resonant circuit was replaced with a step down transformer to provide a better match into the diodes; and the elimination of two variable capacitors. The revised schematic of the X-7 varactor multiplier is shown in Figure 3-24. With minor adjustments made in the Upper Sideband Up Converter (DC-16-3005) a power output of 4 mw was obtained at 8.330 gc. The spectral composition of this signal is shown in Figure III-4b and c. Figure III-4c shows the signal with 18 db of attenuation removed.

The power output and frequency of each unit is listed as follows:

Unit	Power Output	Frequency
1 x 4 x 5 Mult. (DC-16-3025-1)	12 μw	100 Mc
100 Mc Ampl. (DC-16-3025-3)	94 mw	100 Mc
X-7 Mult. (DC-16-3014-3)	6 mw	700 Mc
U.S.U.C. (DC-16-3005)	4 mw	8.330 gc

Test of X-Band Receiver Local Oscillator Synthesizer

Initial attempts at obtaining the necessary power output of the multiplier string tested in the equipment van, disclosed that the low power output of the Lower Sideband Up Converter (DC-16-3005-4) was a result of difficulty in the S-band multiplier (DC-16-3017). Investigation of this unit resulted in the removal of the original biasing resistor and replacing with a variable potentiometer. The S-band multiplier was retuned and operated satisfactory with approximately 3K ohms. A fixed 2.7K resistor was installed with no degradation in performance. The multiplier string was completed and

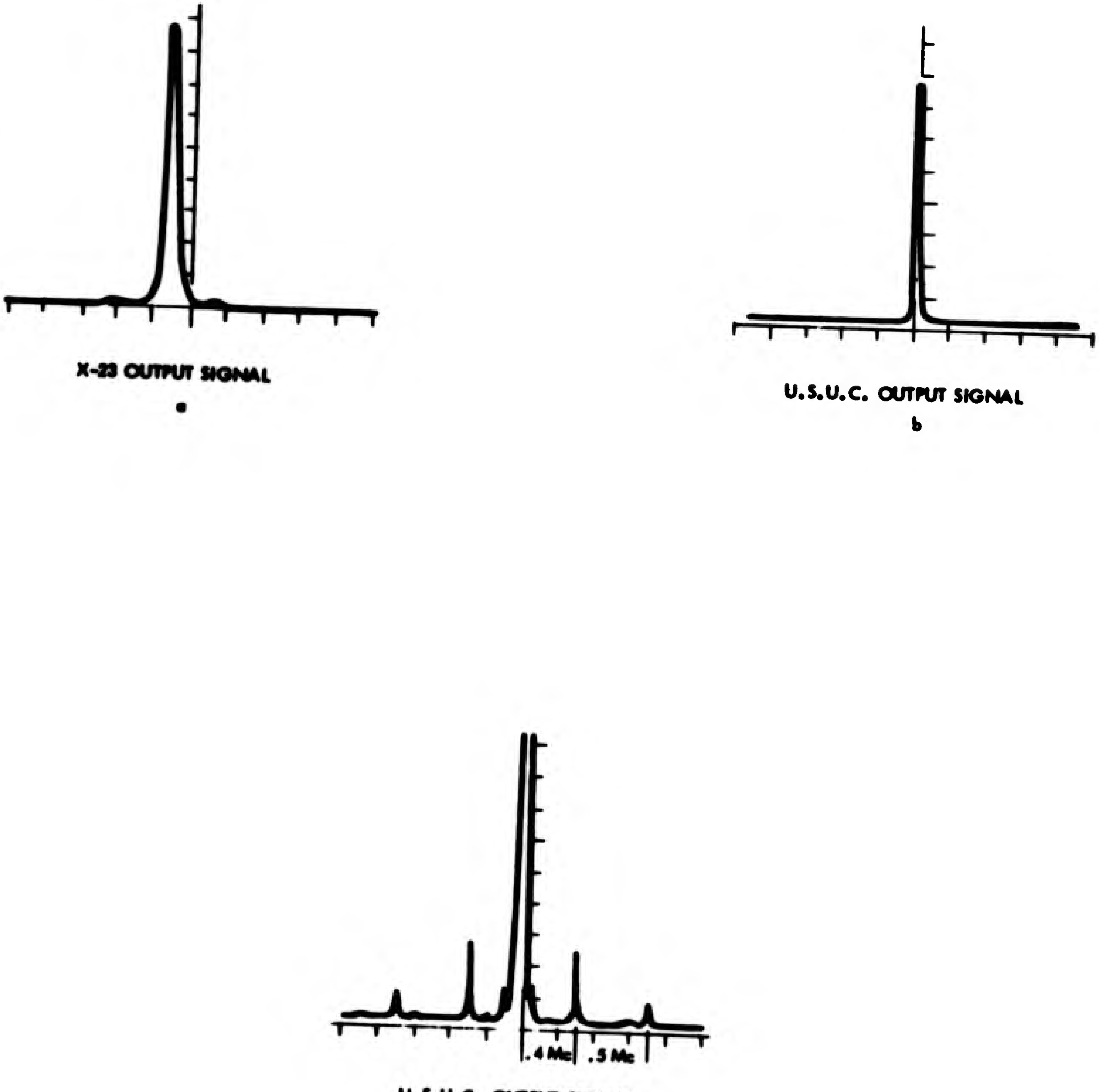

Figure III-4. Output Spectra - X-23 Varactor Multiplier and U. S. U. C.

with minor adjustments made to the C- and X-band multipliers; the necessary power output of 2 mw was obtained at the output of the Lower Sideband Up-Converter.

The power output and frequency of each unit is listed as follows:

Unit	Power Output	Frequency
X-5 Mult. (DC-16-3014-2)	3.6 watts	500 Mc
X-2 Mult. (DC-16-3010)	960 mw	1.000 gc
X-2 Mult. (DC-16-3017)	360 mw	2.000 gc
X-2 Mult. (DC-16-3012)	120 mw	4.000 gc
X-2 Mult. (DC-16-3013)	40 mw	8.000 gc
L.W.U.C. (DC-16-3005-4)	2 mw	7.800 gc

Mechanical changes were made in the modulator waveguide to accommodate the addition of power-set attenuators. These changes are shown in Figure III-5 and represent corrections to Drawing DC-16-3003.

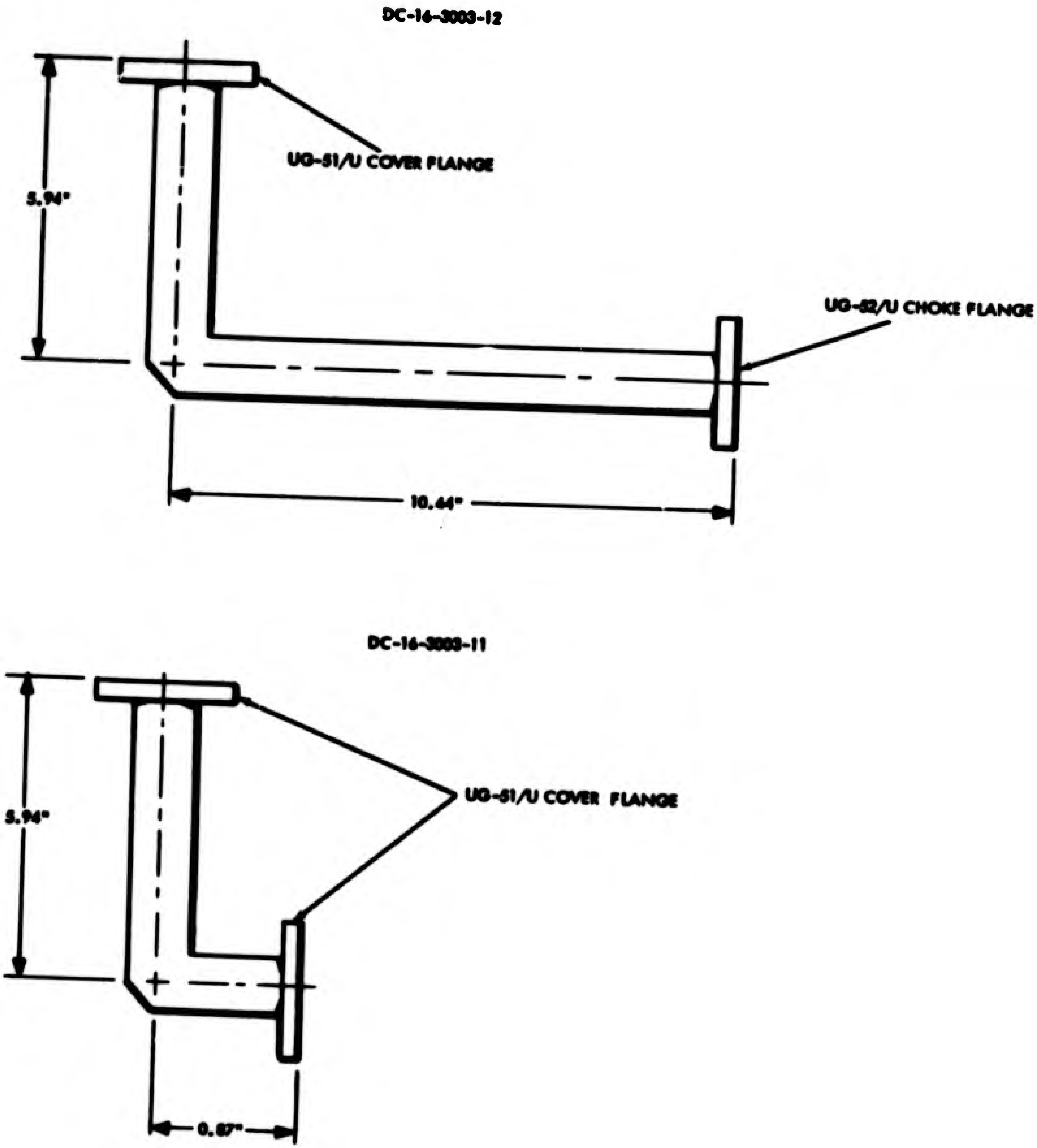

Figure III-5. Modulator Waveguide Change

APPENDIX IV

DESIGN OF A TE_{10} TO TE_{01} MODE TRANSDUCER

This appendix describes the design of a replacement mode transducer for the Transportable Passive Satellite Communications Terminal. The replacement transducer was to be designed to operate at 8.330 Gc with a bandwidth between 6 and 10 percent, and a maximum insertion loss of 0.025 db. It must mate with 3.000 inch diameter circular waveguide and WR137 rectangular waveguide.

The TE_{10} mode is the dominant mode in rectangular waveguide. The problems of design when using this mode are, therefore, limited to achieving a usable impedance match. The TE_{01} mode, on the other hand, is the third H mode in circular waveguide. Further, the TE_{01} and the TM_{11} modes have the same phase velocity, allowing coupling between them. Further yet, the TM_{11} mode is the second E mode in circular waveguide. Therefore, there are a minimum of five modes (TE_{11} , TM_{01} , TE_{21} , TE_{01} , and TM_{11}) that can propagate in any waveguide which will support the TE_{01} mode. The problems of design when using the TE_{01} mode are twofold; to obtain low VSWR and good mode purity. In the usual single mode system, a discontinuity will produce a reflected wave and/or higher order mode patterns in the vicinity of the discontinuity, the latter being necessary to satisfy the boundary conditions. These higher order modes cannot propagate in the guide and hence exist only near the disturbance in symmetry. In a multimode system other modes thus created may or may not be beyond cut off.

There are several methods of transitioning from the TE_{10} mode to the TE_{01} mode. Most of these are modifications or combinations of two basic approaches, coupling and gradual transition from rectangular to circular waveguide. The former approach did not seem to offer the bandwidth which the design required. The latter approach seemed to offer more than sufficient bandwidth.

The simplest design employing the gradual transition approach is that which gradually transforms a single rectangular waveguide into a circular sector waveguide and then into a circular waveguide. However, this is a rather long transducer. A modification which yields a much more compact design employs a series folded tee. This is the approach which was used.

The series folded tee is used as a two port device, the E arm being one port and the folded coplanar arms being the other. The coplanar arms were made up of half height waveguide in order to achieve a good impedance match without employing an appreciable amount of reactive matching, the presence of which would have limited the bandwidth of the tee. Except for the presence of the middle wall, the output of the folded coplanar arms, taken together, is the TE_{20} mode. For this reason, the output of this port is labeled the pseudo-TE_{20} mode. It must be remembered when considering this "mode" that each half is a separate TE_{10} wave which may be attenuated or given an arbitrary phase shift relative to the other half. Such separate action would partially destroy its mode properties and be a source of reflection and/or mode impurity for the transducer. In theory, from the start of the coplanar arms until nearly the end of the transducer, the two waves are acted upon identically, but separately.

The next section of the transducer is a transition from double rectangular to double sector waveguide. Each sector waveguide is removed from the other by 180° and is joined to the other at its apex (Figure IV-1). In designing this section of the transducer an attempt was made to maintain a constant impedance. This attempt resulted from an intuitive feeling that the shift from the sinusoidal E field distribution of rectangular waveguide ($\sin \frac{x}{a}$) (Figure IV-2) to the Bessel distribution of sector waveguide $J_1 (3.83 \frac{r}{a})$ (Figure IV-3), if accompanied by an appreciable impedance change, would give rise to excessive coupling to unwanted modes. It was felt that since the sinusoidal distribution and the Bessel distribution were so close in shape, that a reasonable first approximation to constant impedance might be obtained by maintaining the cross sectional area of this section of the transition as a constant. This was tried with good results although the impedance equations calculated much later show the assumption to be only partially valid (see Table 9). The cross-section of this part of the transducer (Figure IV-4) was designed as a constant area, truncated sector of a circle of decreasing radius and linearly increasing apex angle. The truncated sectors are located by the arc which passes through the outer corners of the originating rectangle projected parallel to the waveguide longitudinal axis. The sector is truncated by the longitudinal continuation of the center wall separating the two rectangular waveguides. The center wall was also linearly reduced in thickness and removed in this section.

The final section of the transducer is an impedance transforming section. The sector waveguides are linearly opened to semi-circles which together form the circular waveguide. Also, at the beginning of this

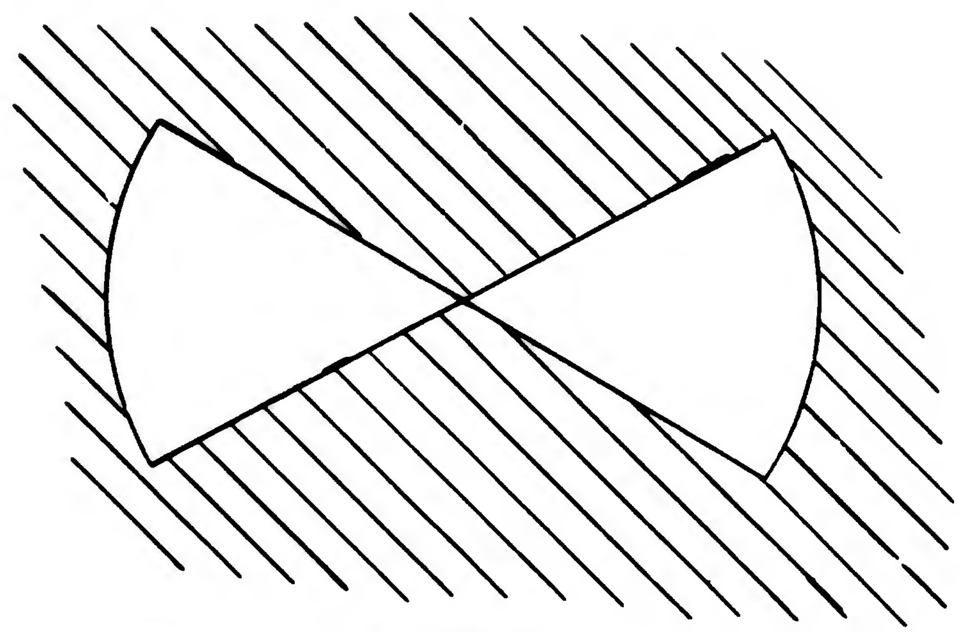

Figure IV.1

Typical Dual Sector Waveguide Cross Section

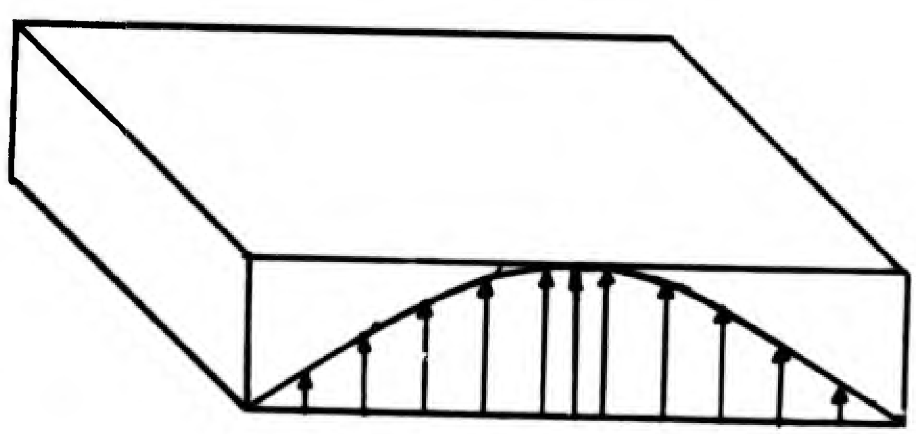

Figure IV-2

Rectangular Waveguide "E" Field Distribution

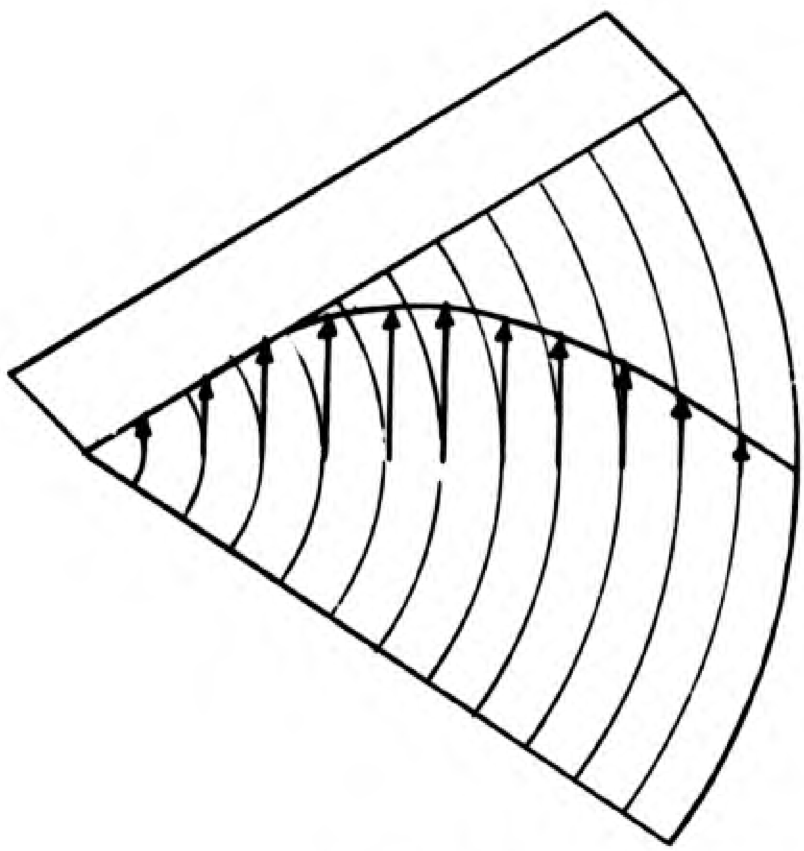

Figure IV-3

Sector Waveguide "E" Field Distribution

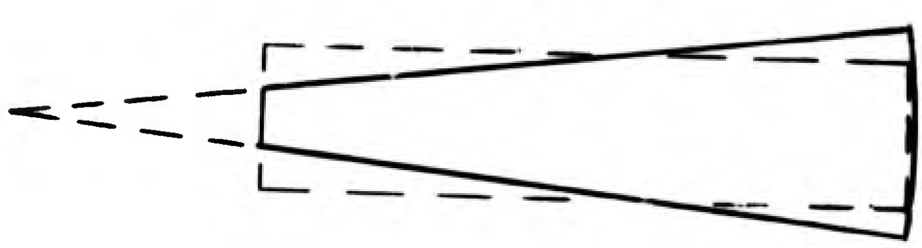

Figure IV-4

Typical Transitional Cross Section
Between Rectangular and Sector Waveguide

section the radius of the sectors is linearly increased to the radius of the circular waveguide. The last two sections of the transducer were produced as a casting.

The VSWR of the transducer described is shown in Table 1a. Since the tee is reasonably well matched over this range (Tables 1b and 2), the cast transition VSWR (Table 1c) is rather marginal. This would seem to indicate that the casting was designed too short. Fortunately it was possible to compensate for the mismatch with an inductive iris (Table 3) placed between the tee and the casting. However, it must be realized that the mismatch, if in the sector portion (final section) of the casting, probably contributed to the mode impurity.

Once the VSWR conditions for the transducer had been met, it was necessary to determine the TE_{01} mode purity. A qualitative technique used was to probe the radial electric fields to determine change with rotational position. This measurement unmistakably displayed the presence of the TE_{21} mode. Probe measurements of the maximum circumferential E field strength and the maximum radial E field strength indicated that the maximum radial field strength lay 16-18 db below the maximum circumferential field strength. Still, the question of tolerable limits existed. Two transducers were placed on either end of a short section of circular waveguide (Figure 5) and rotated, one relative to the other. The VSWR versus relative rotational position was unsatisfactory (Table 4). It therefore became necessary to achieve a higher degree of mode purity.

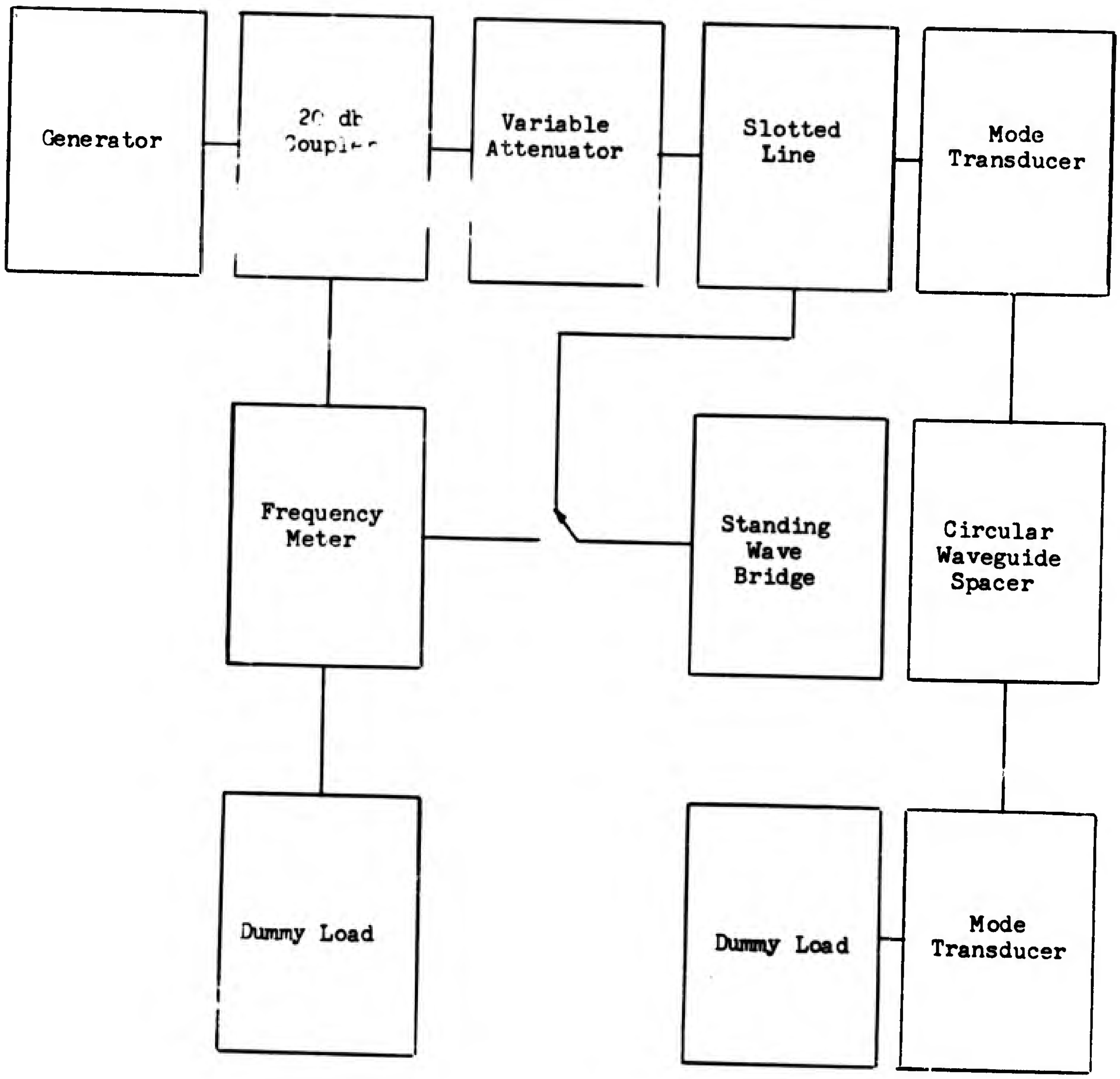

Figure IV-5

Test Set-Up for Measuring "WOW"

Reactive mode filtering using eight radial metal plates was tried to filter out and suppress the unwanted TE_{21} mode. The plates did not have a noticeable effect on the transmission of the TE_{21} mode (or the TE_{01}). It is theorized that the waveguide diameter may have been too large to prevent the passage of this unwanted mode with the number of plates employed. Fabrication problems with more plates caused the idea to be abandoned.

TE_{21} mode inhibition by the use of circumferentially placed wires along the arc of the sector cross section proved very promising. The wires, which are placed one next to another and which are DC insulated from one another (coil wire) inhibit the flow of longitudinal wall currents. These currents are necessary to support any mode except for the circularly symmetrical TE_{on} modes. Low power tests with transducers modified in this fashion were quite an improvement (Table 5). However, under high power, arcs developed across large numbers of the wires in the longitudinal direction and circumferential TE_{01} burns were evident on bunches of wires causing individual wires to be freed from the adhesive holding them in place and causing arcs. This technique also had to be abandoned.

Selective mode attenuation using slots cut circumferentially through the sector arc wall and filled with a lossy epoxy mixture also looked quite promising (Table 6). However, the power thus absorbed heated the castings to rather high temperatures. It was apparent that this power loss would make the system line loss excessive. These slots, being in the transducer, may have caused regeneration of the TE_{21} mode. Some tests were made on slots placed in circular waveguide. Initial tests on circumferential slots one third of the circumference long failed to show either the previously experienced heating or effective mode attenuation. Tests on full circumference slots cut into a circular waveguide wall were impossible due to lack of time.

Because the attempts to prohibit unwanted modes were unsuccessful it was decided to try another transition design between the tee and the circular waveguide. The most hopeful approach seemed to be reduction of the unwanted mode by lengthening and somewhat modifying the original shape. Due to the cost of trial pieces having a smoothly changing inner surface, it was decided to try a quarter wave stepped inner contour accepting the risk of poor mode performance due to the presence of small finite discontinuities. It was also decided to make the characteristic impedances of the steps follow a binomial theorem distribution so as to yield a good bandwidth. The first trial transition contained 13 steps (Table 7a). As previously, probe measurements of maximum field strengths were made. The field strengths were separated by a very discouraging 6 db. Field strength separation was improved to 20-22 db by the addition of 21 additional steps having a characteristic impedance equal to the geometric mean of the steps on either side. The additions were made in three separate modifications (Tables 7b, c, and d) until the largest single impedance jump was 11.2% of the average value of the two steps involved. This final version of the stepped transition with the tee has a VSWR of 1.07:1.0 at the design frequency of 8.330 Gc. The mode purity of this design while better than that employing the casting is still not good enough for use with a rotary joint unless other restrictions are placed upon it.

The final arrangement used in the TPSCT contract was a transducer employing a stepped transition on one side of the rotary joint and a transducer employing a cast transition on the other side. This combination displayed very nearly the same results as when two stepped transducers were used on either side of the rotary joint. In order to get satisfactory rotary joint performance it was necessary to carefully adjust the phase length between transducers in order to obtain proper phasing of the unwanted mode. Measurements of "WOW" were made with a set of spacers which allowed the separation to be varied by 0.05 λg (TE_{01} at 8.33 Gc) steps (Table 8). The widest areas of satisfactory performance ($VSWR_{max.} < 1.30:1$) versus transducer spacing allowed a variation of approximately 0.15 λg in spacing. Satisfactory performance versus frequency with the spacing set at the middle of a 0.10 λg wide region was found to be 80 Mc wide. These figures when compared with the bandwidth of the transducers immediately show the necessity of a very high degree of mode purity for wideband use with a rotary joint. Of course, transducers such as are discussed here can be used over a wide bandwidth by readjusting the transducer separation, it being the lack of proper phasing of the mode impurity which degrades the operation of the transducer pair.

Recently, some further information has become available on the design of this type of transducer. Shin-ichi Iiguchi (1) has shown that a transducer consisting of a single smooth transition from rectangular to sector to circular waveguide will contain mode impurities at or above a calculated minimal value, no matter how well it is designed. The TE_{11} mode, for which Iiguchi predicts a power suppression of 15 db in his transducer, was not detected at all in the dual waveguide system discussed above. It may be theorized that the high suppression of the TE_{11} mode resulted either from the lack of necessity to open the dual sectors to more than 180 each; or from presenting two component parts of the TE_{01} mode, one of which does not exist in the TE_{11} mode, which inhibit the TE_{11} mode from forming in any strength. The latter theory may be further supported by noting that both of the component parts presented, exist in and support the TE_{21} mode. Either theory, but particularly the latter, argue for the use of four quadrant inputs, thus possibly inhibiting the TE_{21} mode. Also, if the TE_{21} mode can be successfully suppressed, the next troublesome mode is the TE_{41} for which the waveguide may be designed to be beyond cut off. Another possible improvement would be the use of the cruciform (2) rather than sector waveguide. One immediate advantage of the cruciform, even if no other exists, is greatly reduced cost of fabrication.

Thus far it has not been possible to arrive at any objective figure for mode purity due to a lack of a consistent measurement scheme. The technique discussed by Klinger (3) bears consideration for this purpose.

One transducer employing a stepped transition and one transducer employing a cast transition were installed on the TPSCT elevation rotary joint. The first optimum spacing of the transducers was 2.30 λg (TE_{01} at 8.33 Gc) in excess of the rotary joint length, which displayed a maximum VSWR of 1.21:1 through 360°.

TABLE 1

TRANSDUCER USING CAST TRANSITION AND NO INDUCTIVE IRIS

Frequency	1a Transducer VSWR	1b Tee VSWR	1c Casting VSWR
7.9 Gc	1.33:1	1.15:1	1.17:1
8.00	1.33	1.11	1.19
8.10	1.29	1.08	1.19
8.20	1.25	1.04	1.21
8.30	1.23	1.04	1.18
8.40	1.25	1.07	1.18
8.50	1.38	1.14	1.22
8.60	1.95	1.24	1.54

TABLE 2

REPRESENTATIVE VSWR OF SERIES FOLDED TEE

Frequency	VSWR
7.63 Gc	1.23:1
7.90	1.16
8.00	1.13
8.10	1.10
8.20	1.07
8.30	1.04
8.40	1.03
8.50	1.08
8.60	1.19

TABLE 3

TRANSDUCER USING CAST TRANSITION AND INDUCTIVE IRIS

Frequency	VSWR
7.63 Gc	1.07:1
7.90	1.15
8.00	1.12
8.10	1.09
8.20	1.06
8.30	1.03
8.40	1.07
8.50	1.13
8.60	1.25

TABLE 4

TWO TRANSDUCERS SEPARATED BY 6" CIRCULAR WAVEGUIDE SPACER AT 8.33 Gc

Relative Rotational Position*	VSWR
0°	1:26
50°	1.12
55°	1.48
65°	1.06
75°	1.70
90°	1.06
105°	1.70
115°	1.06
125°	1.50
140°	1.13
180°	1.27
215°	1.11
230°	1.47
240°	1.06
255°	1.57
270°	1.06
280°	1.90
285°	1.07
300°	1.53
310°	1.14
360°	1.26

*Degree readings are aporoximate.

TABLE 5

TWO TRANSDUCERS WITH CIRCUMFERENTIAL WIRE LINING SEPARATED BY 6" CIRCULAR WAVEGUIDE SPACER AT 8.33 Gc

Relative Rotational Position*	VSWR
0°	1.22
45°	1.37
90°	1.04
135°	1.37
180°	1.21
225°	1.37
270°	1.04
315°	1.41

TABLE 6

TWO TRANSDUCERS, ONE WITH LOSSY SLOTS, SEPARATED BY 6" CIRCULAR WAVEGUIDE SPACER AT 8.33 Gc

Relative Rotational Position*	VSWR
0°	1.15
45°	1.06
80°	1.15
90°	1.14
105°	1.03
120°	1.03
135°	1.07
180°	1.16

*Degree readings are approximate.

TABLE 7

IMPEDANCE OF STEPS IN QUARTER WAVE STEPPED TRANSITION

(a)

Step No.	Impedance
1	200 OHMS
2	200
3	201
4	208
5	228
6	280
7	394
8	614
9	957
10	1349
11	1660
12	1824
13	1882

(b)

Step No.	Impedance
1	200 OHMS
2	200
3	201
4	208
5	218
6	228
7	253
8	280
9	332
10	394
11	492
12	614
13	767
14	957
15	1136
16	1349
17	1496
18	1660
19	1740
20	1824
21	1853
22	1882

TABLE 7 - (Cont'd.)

IMPEDANCE OF STEPS IN QUARTER WAVE STEPPED TRANSITION

(c) Step No.	Impedance	(c) Step No.	Impedance
1	200 OHMS	16	614
2	200	17	686
3	201	18	767
4	208	19	857
5	218	20	957
6	228	21	1043
7	253	22	1136
8	280	23	1238
9	305	24	1349
10	332	25	1496
11	362	26	1660
12	394	27	1740
13	440	28	1824
14	492	29	1853
15	550	30	1882

TABLE 7 - (Cont'd.)

IMPEDANCE OF STEPS IN QUARTER WAVE STEPPED TRANSITION

(d) Step No.	Impedance	(d) Step No.	Impedance
1	200 OHMS	18	614 OHMS
2	200	19	686
3	201	20	767
4	208	21	857
5	218	22	957
6	228	23	1043
7	240	24	1136
8	253	25	1238
9	266	26	1349
10	280	27	1421
11	305	28	1496
12	332	29	1576
13	362	30	1660
14	394	31	1740
15	440	32	1824
16	492	33	1853
17	550	34	1882

TABLE 8

MAXIMUM VSWR WITH ROTATION OF TRANSDUCER

WITH STEPPED TRANSITION AND TRANSDUCER

WITH CAST TRANSITION VERSUS SEPARATION AT 8.33 Gc

Separation	VSWR	Separation	VSWR	Separation	VSWR
None	1.64	2.60	2.1	3.85	1.15
0.60	1.57	2.65	5.9	3.90	1.80
0.70	2.1	2.70	3.6	3.95	1.47
0.90	2.2	2.75	1.44	4.00	4.2
1.00	1.34	2.80	1.30	4.05	4.5
1.05	1.35	2.90	1.28	4.10	1.92
1.30	8.3	2.95	1.20	4.15	1.58
1.50	1.28	3.00	1.24	4.20	1.37
1.60	1.92	3.05	2.2	4.25	1.29
1.65	2.5	3.10	8.3	4.30	1.27
1.70	2.3	3.15	4.6	4.35	1.31
1.75	6.8	3.20	1.46	4.40	1.54
1.90	1.27	3.30	1.33	4.45	1.50
1.95	1.33	3.35	1.35	4.50	5.4
2.00	1.62	3.40	1.45	4.55	3.1
2.05	2.7	3.45	1.24	4.60	1.71
2.10	2.7	3.50	1.60	4.65	1.53
2.20	4.0	3.55	3.4	4.70	1.35
2.30	1.24	3.60	2.6	4.75	1.32
2.35	1.19	3.65	8.5	4.80	1.34
2.40	1.30	3.70	1.76	4.85	1.78
2.50	1.25	3.75	1.39	4.95	4.8
2.55	4.1	3.80	1.27	5.00	6.8

TABLE 9

RECTANGULAR WAVEGUIDE EQUATIONS FOR THE TE_{10} MODE

1. $E_x = 0$

$E_y = E \sin\left(\frac{\pi}{a}x\right)e^{j\omega t - \gamma z}$

$E_z = 0$

$H_x = -E \frac{\gamma}{j\omega\mu_o} \sin\left(\frac{\pi}{a}x\right)e^{j\omega t - \gamma z}$

$H_y = 0$

$H_z = -E \frac{K}{j\omega\mu_o} \cos\left(\frac{\pi}{a}x\right)e^{j\omega t - \gamma z}$

$$\gamma = j\frac{2\pi}{\lambda}\sqrt{1-\nu^2}$$

$$\nu = \frac{\lambda}{\lambda_c}$$

$$\lambda_c = 2a$$

$$\omega\mu_o = \frac{2\pi\eta_o}{\lambda}$$

$$\eta_o = \sqrt{\frac{\mu_o}{\epsilon_o}} = 377 \text{ ohms}$$

$$K = \sqrt{k^2 + \gamma^2} = \frac{\pi}{a}$$

$$k^2 = \omega^2\mu_o\epsilon_o$$

2. $W_t = \frac{ab}{4} E^2 \frac{\sqrt{1-\nu^2}}{\eta_o}$

3. $Z = \frac{V^2}{W} = \frac{2b\eta_o}{a\sqrt{1-\nu^2}}$

CIRCULAR WAVEGUIDE EQUATIONS FOR THE TE_{01} MODE

4. $E_r = 0$

$E_\theta = E\, J_1\left(\frac{u_1}{a} r\right)e^{j\omega t - \gamma z}$

$E_z = 0$

$H_r = -E \frac{\gamma}{j\omega\mu_o} J_1\left(\frac{u_1}{a} r\right)e^{j\omega t - \gamma z}$

$H_\theta = 0$

$H_z = -E \frac{K}{j\omega\mu_o} J_o\left(\frac{u_1}{a} r\right)e^{j\omega t - \gamma z}$

$u_1 = 3.832$

$\gamma = j\,\frac{2\pi}{\lambda}\sqrt{1 - \nu^2}$

$\nu = \frac{\lambda}{\lambda_c}$

$\lambda_c = \frac{2\pi a}{u_1}$

$\omega\mu_o = \frac{2\pi\,\eta_o}{\lambda}$

$\eta_o = \sqrt{\frac{\mu_o}{\epsilon_o}} = 377$ ohms

$K = \sqrt{k^2 + \gamma^2} = \frac{u_1}{a}$

$k^2 = \omega^2\mu_o\epsilon_o$

5. $W_t = \frac{\pi a^2}{2}\; J_o^{\,2}(u_1)\; E^{\,2}\frac{\sqrt{1 - \nu^2}}{\eta_o}$

6. $Z = \frac{V^2}{W^2} = \frac{3100}{\sqrt{1 - \nu^2}}$

SECTOR WAVEGUIDE EQUATIONS FOR THE TE_{01} MODE

7. Same as 4 above

8. $W_t = \Theta\,\frac{a^2}{4}\; J_o^{\,2}(u_1)\; E^2\,\frac{\sqrt{1 - \nu^2}}{\eta_o}$

9. $Z = 1.309\,\frac{\Theta\,\eta_o}{\sqrt{1 - \nu^2}}$

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