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RESEARCH STUDY INTO THE NEAR EARTH APPLICATION OF MILLIMETER RADIO WAVES AS APPLIED TO CERTAIN BATTLEFIELD PROBLEMS

Interim Technical Report

Prepared by

John M. Cotton, Jr.
John W. Dozier

ADVANCED TECHNOLOGY CORPORATION
1830 York Road
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Ballistic Research Laboratories
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Contract No. DA 18-001-AMC-829(x)

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ABSTRACT

This report summarizes the work done in developing three broadband millimeter wave components: a mechanical modulator, a tunable bandpass filter, and a second harmonic mixer. All the components were designed to operate throughout the 75-110 GHz frequency region. The ultimate application of these particular components was a tunable radiometer capable of operation throughout the vicinity of the 94 GHz window. The text describes the techniques used to fabricate the items and gives performance data on the components singly as well as on the completed radiometer assembly.

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1. INTRODUCTION

This report describes work done on the first three assignments of Contract No. DA 18-001-AMC-829(x); subsequent work assignments are still in progress. Strictly speaking, the assignments were to develop and deliver three broad-band components for use in the 75 to 110 GHz region. These components were a mechanical modulator, a tunable bandpass filter and a sealed harmonic mixer. The broader view however was that the components were intended for use as the critical items in the assembly of a tunable radiometer.

The following sections of this report review the development of each item in turn and finally present preliminary data taken while using them in a complete radiometer system. In the development work this ultimate use was always borne in mind to keep the design goals for the individual components in a proper perspective. As a result, the individual components met the specifications set for them and also worked well in the assembled system.

Each of the components described, however, has merits on its own and their ultimate applications are not limited to radiometry. It is easy to conceive of many millimeter wave systems wherein any one of these items could find applications either as they are or with minor modifications. It is hoped that the development of these components will help advance the millimeter wave field to the time when systems in this frequency range are commonplace.

2. BROADBAND DICKE TYPE CHOPPER

For a radiometer application it is not absolutely essential to have a three port switch. It is of course convenient for some purposes for example, if a very cold reference temperature is desired. However when the use of the radiometer will be primarily directed to hot targets such as the sun, the system calibration can be made with a noise tube. Then the necessary device for a Dicke type system is one which will chop the reference and simultaneously furnish a known reference temperature. As long as moderate chopping rates are involved, this can be done with a mechanical device such as the one described below. The initial concept (which proved entirely practical) was to use a rotating card inserted through a slot in the broad wall of the waveguide, which would alternately attenuate and transmit the incoming signal. In its attenuation state it would provide a reference temperature achieved by maintaining the resistor card at a preselected temperature. Details of the development and fabrication follow.

2.1 Electrical Considerations

2.1.1 Attenuation and VSWR

Since it was felt that one of the principal targets for the radiometer would be the sun (which would represent a temperature of 4000° K or more) the attenuation phase of the chopping cycle would have to be sufficient to reduce this to a negligible amount, i.e. at least 30 dB of attenuation. Accordingly the first phase of the development was an experimental one devoted to the determination of attenuation of metallized mica vanes inserted in slotted waveguide.

First, pieces of 0.004" thick mica metallized with nichrome were used to determine the film resistance needed. Using disks 0.300" diameter cut from material with film resistances of 150, 200, 400 and 1000 ohms per square, measurements were made of attenuation and VSWR versus insertion into the waveguide (WR-10) at 90 GHz. It was apparent that 150 to 200 ohms per square was the range required for maximum attenuation. Disks of 0.001" mica were then prepared with evaporated gold films ranging from 70 to 200 ohms per square, with the 150 ohm film giving maximum attenuation. VSWR in all cases was less than 1.2. Measurements at 100 GHz similar to those at 90 GHz verified that bandwidth would not be a problem. Tables 1 and 2 show values measured at 90 GHz.

To increase the attenuation attainable, the disk size was increased first to 0.750" dia., then finally to 1.50" dia. For all sizes, the nichrome film gave higher attenuation than the gold, with no degradation of VSWR.

TABLE I

Attenuation at Maximum Insertion

f = 90 GHz, Film Resistance = 150 ohms/sq

	0.300" dia.	0.750" dia.	1.50" dia.
Nichrome on 0.004" mica	28 db	33 db	-
Gold on 0.001" mica	22 db	26 db	37 db
Nichrome on 0.001" mica	-	-	46 db

TABLE II

VSWR at Maximum Insertion

f = 90 GHz, Film Resistance = 150 ohms/sq

	0.300" dia.	0.750" dia.	1.50" dia.
Nichrome on 0.004" mica	1.18	1.12	-
Gold on 0.001" mica	1.18	1.16	1.14
Nichrome on 0.001" mica	-	-	1.06

It should be noted here that the VSWR is of special significance in the radiometer application. Consider the system here using a harmonic mixer. The local oscillator power applied at the diode for mixing will also result in the generation of harmonic energy which will propagate toward the antenna. Upon reaching the chopper, some of this energy will be reflected to the mixer due to the VSWR. This energy will then have some slight effect on the diode crystal current and hence the noise output of the mixer. Now when the switch changes state, if the VSWR changes in either magnitude, phase or both, this small contribution to the crystal noise will also change. Since these changes are occurring at the chopping frequency, the signal processor will see this modulated noise as an input signal when in reality none exists. Hence the object here was not only to reduce the VSWR, which is desirable in any case, but to maintain that VSWR as constant as possible in alternate states of the switch.

For this reason, eccentric wheels or mechanical vibrational insertion techniques were not investigated. It was decided that the best approach would be to keep the mica disk in the guide at all times and achieve the desired chopping cycle by properly shaping the evaporated nichrome.

The first experiments along this line were made to determine the shape of the coating which would furnish a rapid rise time on the waveform without any points of excessive variation in the VSWR. Ultimately a configuration was experimentally determined by measuring insertion loss and VSWR as a function of angle on a variety of shapes. Starting from the pass or minimum attenuation position, the best configuration tried gave 3 dB attenuation in the first 5° of rotation and 30 dB in 27°. This gave a satisfactory approximation to a square wave for the radiometer modulation.

On the early disks, some difficulties were experienced in maintaining constant attenuation while rotating through the maximum attenuation portion of the cycle. Assuming that the cause was due to uneven evaporation of the nichrome, a jig was assembled which provided rotation of the disk during the vacuum coating process. This procedure served to smooth out the deposition and subsequent disks were quite constant in their "off" state. To insure that this uniformity was maintained and also to furnish protection in handling a coating of magnesium fluoride was vapor deposited over the nichrome.

In order to evaluate the structural properties of the .001" mica at high rotation speeds, one disk was rotated at 6300 RPM for a prolonged period. No protective edge coating was used. There was no evidence of delamination or disintegration. Since the final assembly was to run at 2250 RPM, and be sealed at the edge, no mechanical difficulties were anticipated or experienced.

When the desired shape of the absorber on the disk was determined, an accurate mask was made for the deposition. This has since proved to give excellent repeatability in the fabrication of duplicate disks. After the magnesium fluoride was deposited, this piece of mica was sandwiched with another .001" disk so that the coated side was completely protected. The sandwich was then held in a jig and rotated with its edge in a very low viscosity silicone varnish. The varnish was dried while continuing to spin the assembly at a fairly high speed, thus removing any excess varnish and keeping the total width of the lamination as thin as possible. This technique resulted in a quite uniform sealed edge on the mica and a total width at the edge of just under .003".

With a constant diameter attenuator of this type, the attenuation will, of course, be frequency dependent. Hence when this disk was put in to the final assembly, the insertion into the waveguide was adjusted to provide at least 30 dB of attenuation at 80 GHz. This insertion depth then gave 36 and 38 dB maximum attenuation at 92 and 97 GHz, respectively.

2.1.2 Synchronizing Signal

One of the requirements on this modulator was that it furnish a synch signal to an existing signal processor. This processor had a requirement of a 6 volt square wave input, and also had built in capability for fine control on the delay of this signal. Hence all that was required was a coarse control on the modulator itself.

It was decided that one simple method of obtaining a reference signal from the chopper blade shaft was via a magnetic pick up. A cam made of Armco iron with two square cut lobes was placed on the chopper drive shaft. A 3000 ohm miniature earphone coil was positioned beside the cam as the transducer. This then generated two positive and two negative voltage spikes per revolution. This signal (approximately 2 volts) was then applied to a two transistor bistable multivibrator located in the chopper control panel. This wave shaper, located in the control panel, then produced the desired 6 volt square wave output. The coarse adjustment of phase relationship with the RF signal is readily achieved by positioning of the cam on the shaft. The pick up coil is fixed.

This type of synch signal output should now be compatible with any signal processor which requires an input with two equally spaced zero crossings per cycle.

2.2 Mechanical Considerations

2.2.1 Motor and Gear Drive

Since the parameters of a 75 cycle chopping frequency and two cycles per revolution had been established, a 2250 RPM shaft speed was determined. No standard 2250 RPM synchronous gear head motor could be located. Consequently, an 1800 RPM synchronous motor was used with a 4 to 5 step up gear train fabricated in the model shop. When the motor was received, it was quickly apparent that the vibration it produced necessitated shock mounting.

The resulting mechanical design then had the waveguide and disk assembly rigidly mounted to a base plate and the motor and gear box assembly shock mounted on this same plate. The original intent was to use a flexible rubber coupling between the output shaft of the gear train and the drive shaft for the disk. The requirements in the motor shock mounts, however, permitted considerable change in actual shaft position with orientation of the chopper assembly such as might be experienced if the unit were located on an antenna mount. The rubber coupling was inadequate for this application and was abandoned in favor of a bellows coupling, and this is shown in the photographs included later. The actual model as delivered, however, went one more step to a magnetic coupling with an air gap between the shafts. This permitted complete vibration isolation along the driving shafts and provided good coupling in any orientation of the unit.

2.2.2 Mechanical Alignment

As noted earlier, the finished mica card had a total width of approximately .003". The slotted waveguide section was made with a .006" saw slot in its top broad wall. This left very little margin for error in positioning of the waveguide or run out of the disk. The disk itself was 1.500" in diameter and was clamped between two tapered aluminum hubs 1.320" in diameter. This left a .090" mica rim which was sufficient to permit penetration of the .040" waveguide wall and as much insertion as was desirable into the .050" dimension of the waveguide. The finished model actually ran with almost .010" clearance from the bottom waveguide wall.

The slotted waveguide section was soldered to a brass block and provision was made in the waveguide and disk housing to furnish vertical adjustment of this assembly. The lateral adjustment of the rotating vane relative to the slot was simply accomplished by proper location of the aluminum clamping hub on the shaft. The shaft in turn was located in the housing with extreme accuracy and supported in pre-loaded bearings. The combination of precision machine work and considerable care in assembly resulted in an excellent centering of the disk in the slot with run out of less than .001" in operation.

2.2.3 Temperature Control

In order to furnish a reliable reference temperature for the radiometer, the disk and waveguide assembly was enclosed in an insulated and heated housing. The power to the heater can be adjusted by a transformer on the control panel. The current was set to provide 8 watts of heater power. With this amount, the warm-up time to 40° C in a room temperature environment was 15 minutes.

The heater current is controlled by a thermal regulator on the panel. The regulator uses a thermistor in the chopper housing as a sensing element, and has a sensitivity of ±0.05° C.

A thermocouple was also located in the chopper housing at a point remote from the thermistor location. Monitoring the temperature on the thermocouple with the regulator operating showed temperature stability to be within ±0.2° C inside the housing. The rotating disk and cam apparently provide sufficient air agitation within the housing to maintain a quite uniform temperature, as was expected. Hence a very reliable reference temperature is provided.

2.3 Completed Assembly

The photographs labelled Figures 1, 2 and 3 show the details of the final assembly. The sole exception to this is, as noted earlier, that the bellows coupling was ultimately replaced with magnets.

The originally specified design goals for this component were that it operate over the 80 to 105 GHz region, that it have a VSWR of less than 1.5 with a variation of less than 0.3; that the temperature be maintained at 40° C ± 0.2° and the chopping rate at 75 ± $\frac{1}{2}$ cps; and that the insertion loss be less than 1 dB.

The delivered item met or exceeded these specifications in every detail and are summarized below:

Operating Frequency Range	75-110 GHz
VSWR over this Range	< 1.25
VSWR Variation	< .15
Temperature Control	40° ± 0.2° C
Chopping Frequency	75 ± 0.1 cps
Attenuation	≤ 30 dB
Insertion Loss	< 1 dB

In conclusion, it is felt that development has produced a very useful component for broad band radiometry. The techniques developed are applicable to essentially any standard waveguide and will furnish good operation over the entire waveguide band.

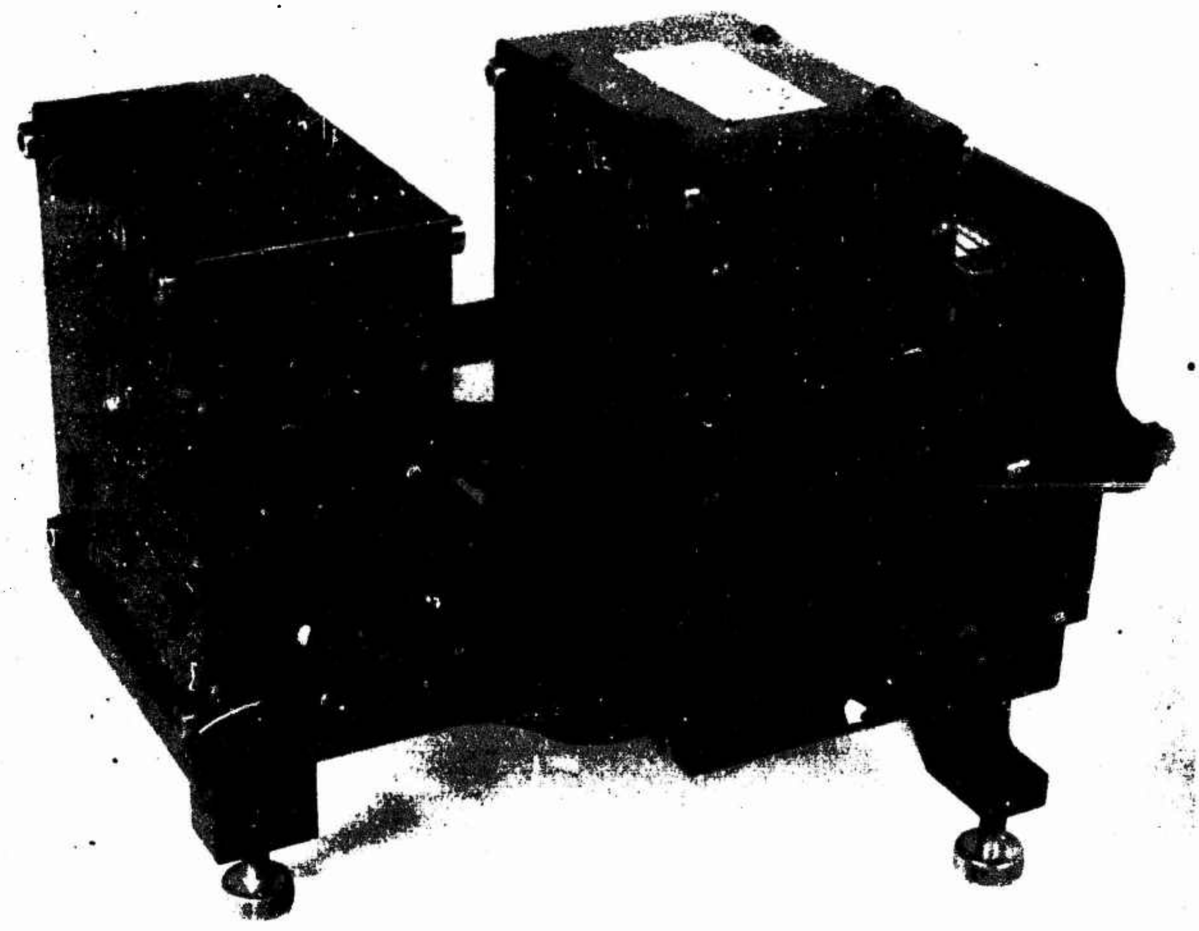

FIG. 1 - Mechanical Modulator Assembly

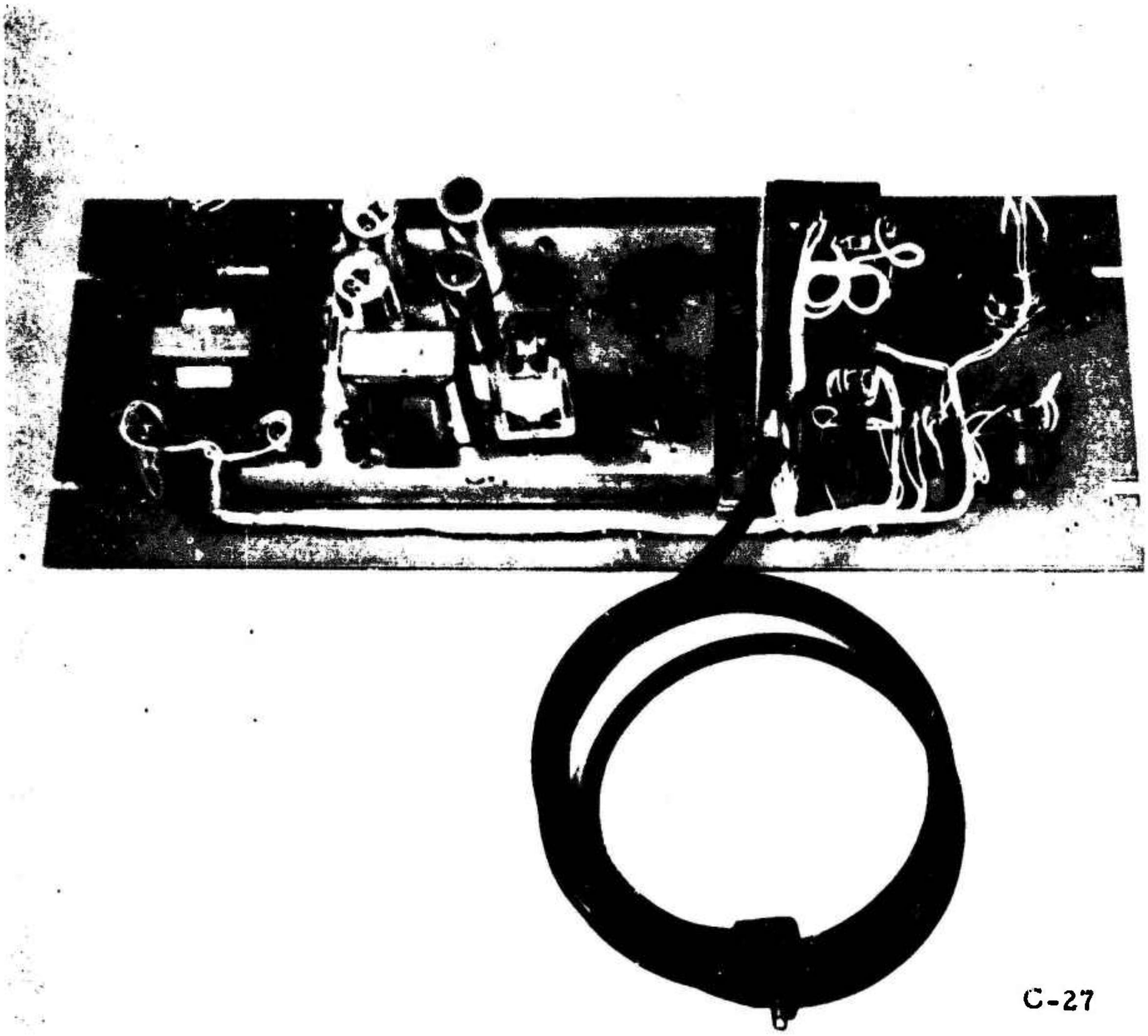

FIG. 2 - Modulator Control Panel Assembly

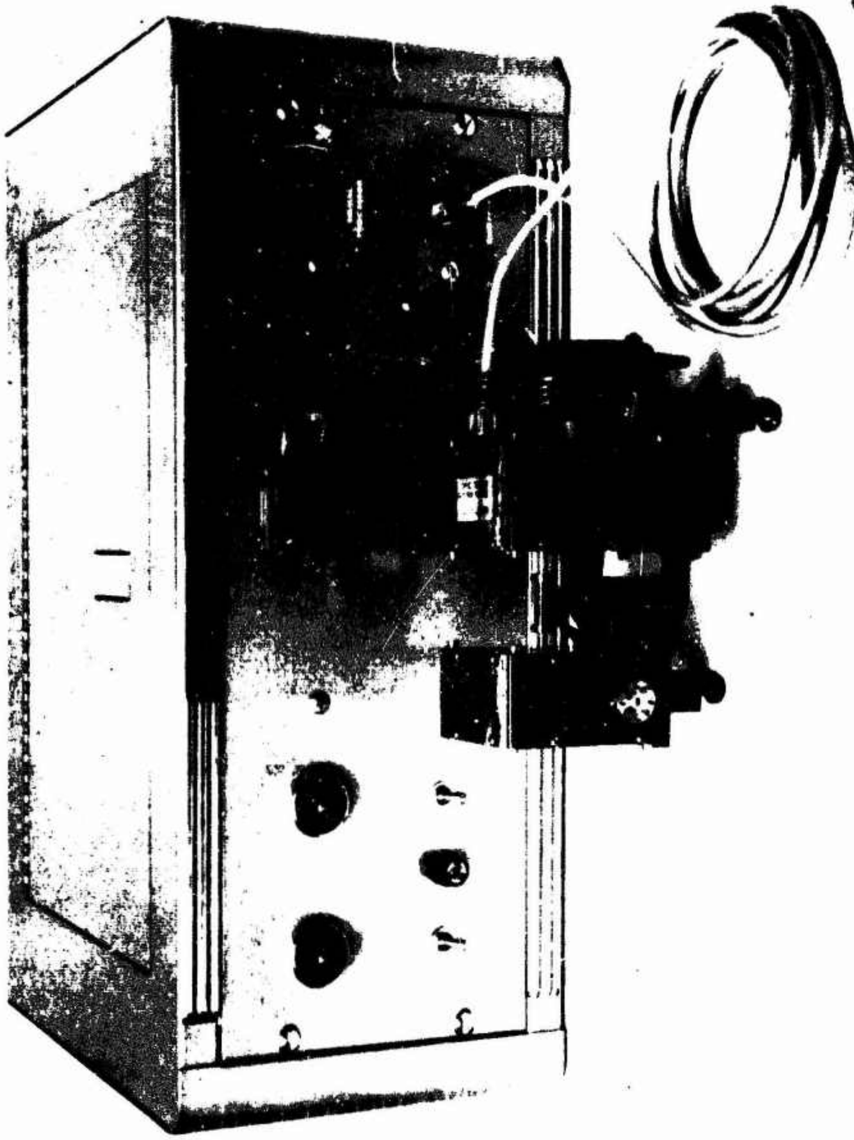

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FIG. 3 - Completed Modulator Unit

3. A BANDPASS FILTER

In the concept of a tunable radiometer, some method of RF preselection is necessary in order to determine which sideband of the mixer is actually being used. Some of this selection is obtained in the mixer itself which will not have a 16 GHz instantaneous bandwidth; tuning of the adjustable short circuits in the mixer will result in a reduced conversion loss of one sideband as compared to the other. This, however, is not adequate for many practical situations, particularly those in which the signal band is on the wings of the atmospheric window and the image band centered in it. In these cases, the rejection afforded by the mixer would be offset by the differences in atmospheric attenuation.

The original concept on this program was the use of a single stage tunable cavity as a bandpass filter. Here, in principle, the ratio of the power transmitted off resonance to that at the center frequency is given by

$$\frac{T(\omega)}{T(\omega_o)} = \frac{1}{1 + Q_L^2 \left(\frac{\omega_o}{\omega} - \frac{\omega}{\omega_o}\right)^2}$$

Hence taking an example of a loaded Q of 20, and a center frequency of 93 GHz, then the power at 92 and 94 GHz would be down approximately 0.3 dB whereas the 83 GHz power is down 13.4 dB and the 79 GHz power 16.4 dB. If one were scanning only the middle 2 GHz of the IF passband, the variation in signal strength would be 0.3 dB whereas the image signal would always be rejected by 14 dB or more.

In practice, however, this filter was very difficult to achieve. Although the desired filter was built, using a directional coupler, it bore little resemblance to the solution proposed.

In the following, some of the difficulties involved in working with the cavity approach will be outlined, and the successful coupler filter described.

3.1 Circular Cylindrical Cavities

3.1.1 Sidewall Coupled Cavity

In the early work on this program, a variety of circular cavities were electroformed. The first of these is shown in Figures 4 and 5. This consisted of a cylindrical cavity with an adjustable short with irises cut in the side walls. These walls had been machined down to a thickness of .003" at their narrowest point. The coupling waveguide sections were made to be removable so that the irises could be accessible between measurements. These preliminary measurements were undertaken in an effort to determine what magnitude of cavity Q was practically achievable in this frequency range.

Measurements were made by mixing the signals from two klystrons, keeping one CW and sweeping the other through its oscillating mode. Following a 1 to 2 GHz travelling wave tube IF amplifier the width of the klystron mode could be quite accurately measured with an IF wavemeter and oscilloscope display, and frequency markers could be established along the scope trace. Then inserting the cavity and tuning its peak transmission to the center of the klystron mode, the bandwidth could be measured.

From the data on bandwidth, insertion loss and VSWR, calculations indicated a loaded Q of 1980 and $Q_o = 2400$. The theoretical Q_o for a gold cavity of these dimensions resonant in the TE_{111} mode is 2800. The reduction from the theoretical value is probably due to losses in the short circuit. Since the first .002" of electroformed material was gold (with copper then plated for wall strength up to the finished thickness) the shorting plunger

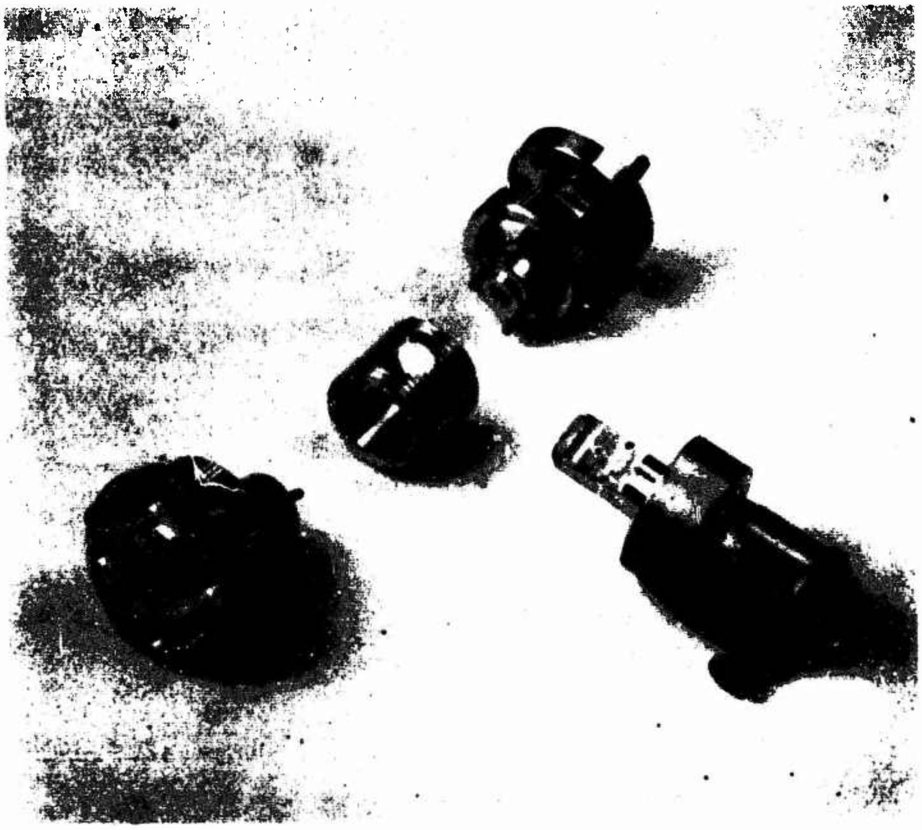

FIG. 4 - Side-wall Fed Cavity, Disassembled View

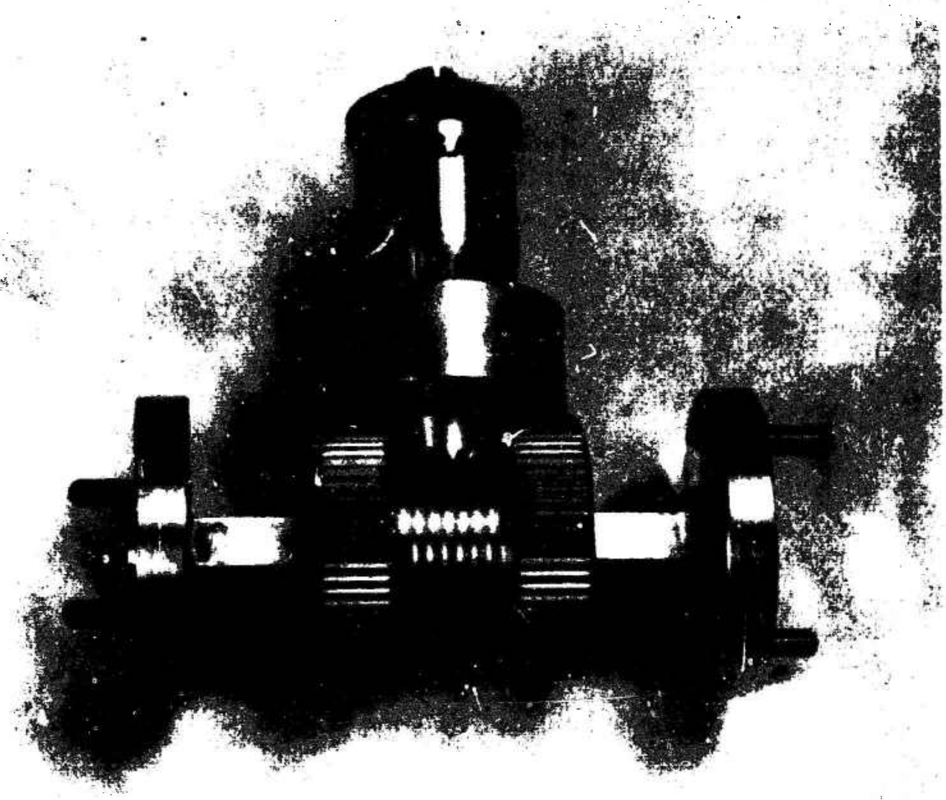

FIG. 5 - Side-wall Fed Cavity, Assembled View

had to be made of very soft material to avoid wear on the cavity walls in use. For this purpose, very soft aluminum was used. The face which would form the moving end of the cavity was first finished and gold plated. Then the outer diameter was turned to the proper size for a fit in the cavity and mounted with a non-rotating micrometer drive assembly. Working with soft materials such as this, a good close fit is difficult to achieve, and it is not unexpected that some losses would be encountered.

After preliminary measurements had been performed the coupling apertures on the sides of the cylinder walls were opened up to progressively larger sizes to lower the Q and lower the insertion loss. It soon became apparent that this was an unsatisfactory experimental approach for a variety of reasons. Most significant among these were the time consumed in the process between measurements and the irreversibility of the technique. Hence it was decided to adopt a configuration which would eliminate these primary objections.

3.1.2 End Fed Circular Cavity

Figures 6 and 7 illustrate the physical configuration of the next unit tested. The circular cavity was again electroformed for TE_{111} operation. The cavity diameter here as in the first was such that the TE_{21} mode was cutoff for frequencies below 105 GHz and it was hoped here, as before, that the TM_{01} mode would be suppressed due to the methods of excitation.

As can be seen in the photographs, both waveguides were set up to feed the same end of the cylinder. It would, of course, be more desirable to have one feed at each end. That approach was rejected, however, because of the problems of maintaining a tunable structure with that configuration.

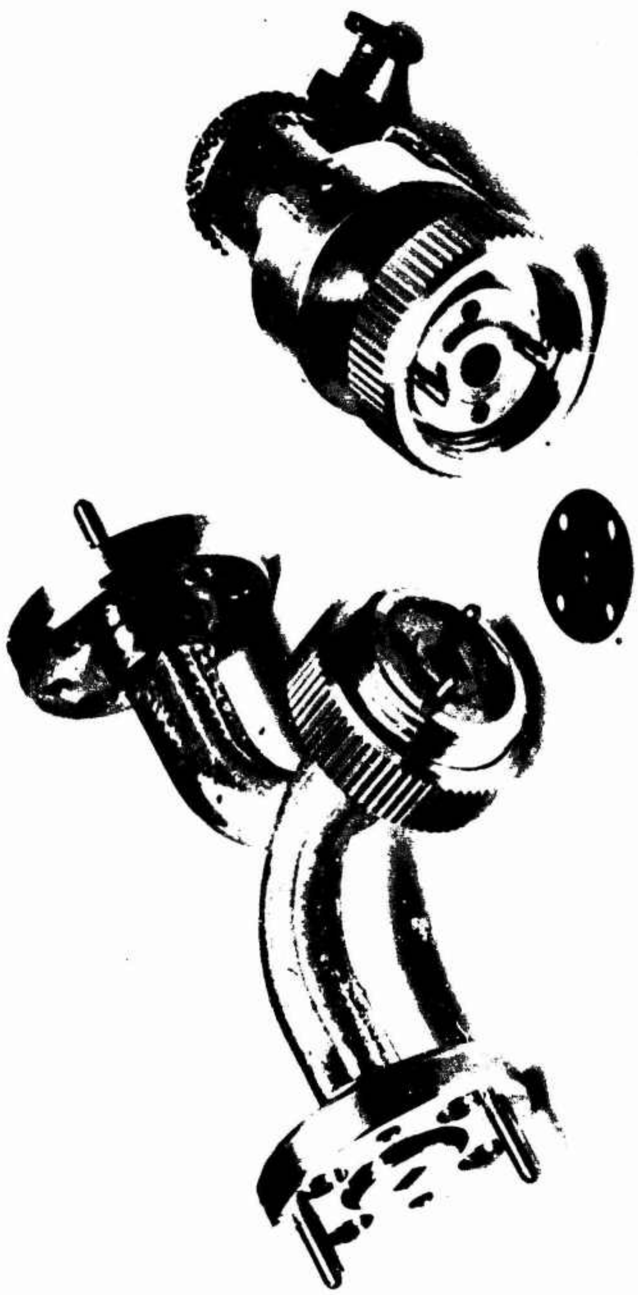

FIG. 6 - End Fed Cavity, Disassembled View

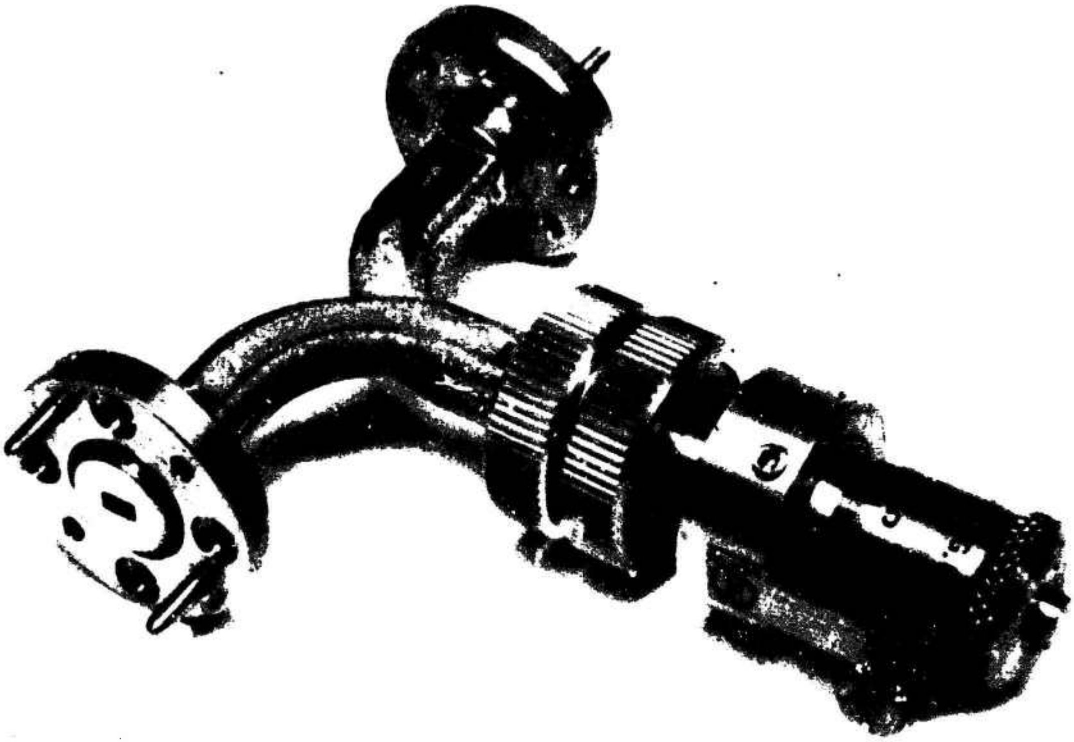

FIG. 7 - End Fed Cavity, Assembled View

The waveguide feed assembly was electroformed in order to achieve a very thin common wall between the waveguide. This was accomplished by first plating .0015" of gold on each waveguide mandrel and following this with a copper flash. At this point the bends were held together with a wooden clamp and the assembly copper plated. Once sufficient copper had been built up to hold the pieces together, the clamp was removed and the plating finished.

The experimental advantage which this offered is indicated by the circular brass shim pictured in Figure 6. Clamping the shim between the flanges provided a .002" material thickness for the irises and a great number of iris shapes, sizes and positions could be evaluated, and indeed were. With this filter the TE_{111} mode of operation was successfully demonstrated with a 1/2 dB insertion loss. This result was obtained with irises which were positioned symmetrically with respect to the vertical and horizontal centerlines of the circular cross section. (The thin common wall of the two waveguides feeding the circular cavity is here considered to be aligned with the horizontal centerline.)

In this configuration, however, the TM_{011} resonance was also found at a frequency approximately 14 GHz higher with an insertion loss of about 1 dB. For the final system application, this was of course intolerable. If the frequency of operation were, for example, to be 80 GHz, with the filter tuned to 80 GHz and the local oscillator at the subharmonic of 86, the image band would be 90 to 94 GHz and the latter frequency would be passed through the cavity with almost no suppression.

The first efforts to provide suppression of the TM_{011} mode involved making the irises antisymmetric with respect to the vertical centerline. This did furnish additional rejection of the undesired mode. However, the position of the irises was no longer optimum for the waveguides in that the irises have been moved to a position of lower electric field. In this position, by the time they were made sufficiently large to achieve insertion losses approaching one dB, their mode suppression capability was diminishing to the point where it was only some 7 dB.

The next approach was to change the orientation of the coupling slots. Heretofore, the slots had always been set up with their long axis parallel to the waveguide broad wall (the horizontal centerline of the cavity.) Now the slots were set up to have their long axes aligned with radials passing through the center of the circular cross section to achieve additional TM_{01} suppression. The suppression achieved was excellent – in excess of 40 dB. However, here again, the slot locations were not optimum for the waveguide and the low value of loaded Q required could not be achieved.

Another attempt was made for mode suppression within the cavity itself. A high density foam material was machined to fit inside the cavity and serve as a holder for a metal shim. This shim, positioned exactly parallel to the horizontal centerline, should have a negligible effect on the TE_{11} mode since the E field is normal to the plane of the shim. It should, however, have a serious effect on the TM_{01} mode which has components of E parallel to the shim. The presence of the foam, and the difficulty in precise positioning detracted from this effort, however. The TM_{01} was affected less and the TE_{11} affected more than was originally hoped.

The next step taken to achieve mode suppression was the fabrication of a semi-circular cylindrical cavity. Suppression of the TM_{01} mode was of course complete since the structure would not support it. With this mode eliminated from consideration it was possible to make the cylinder diameter larger and still have single mode operation throughout the range. This in part offset the decreased volume to surface ratio of the new cavity thereby helping to maintain a reasonable loaded Q. This cavity was fabricated to be compatible with the feed structure already in use.

The disadvantage of the semicircular cross section was the loss of symmetry in the irises when feeding from the end. One iris was coupled to a region of higher field strength than the other, and hence different sizes were required to achieve the same coupling coefficient in each port.

The first dozen irises were machined in a fashion such that one iris was held constant while the other varied. Two series of six types were made in this fashion to permit 24 possible coupling configurations. The preliminary data indicated the proper direction to take and ultimately one iris pair was machined which exhibited the proper Q and an insertion loss of one dB.

It should be noted here the type of measurement which was being made. These preliminary evaluations kept the frequency fixed and adjusted the cavity length. This enabled one to make the inferences: "If the cavity is tuned for 82 GHz transmission, its insertion loss at 94 GHz is so much; if the cavity is tuned for 90 GHz transmission, its insertion loss at 94 GHz is so much. . ." etc. for the entire region of frequency interest. In the absence of swept frequency sources, this provided a rapid means for rejection of some irises.

Hence, after an iris was found which seemed proper, it was thoroughly tested over a broad frequency, and found to be extremely frequency dependent. At the low end of the band the insertion loss at resonance was as high as 8.5 dB. At this juncture the decision was made to find an entirely new approach.

3.2 Directional Coupler Filter

Consider a four port, three 3B coupler which is lossless and has infinite directivity. Because of the 90° phase shift which occurs with the coupling, one can take this device, apply power to one port and position short circuits on the through arm and the normal coupling arm and transmit all the power to the fourth port. If the short circuits are set up at positions equidistant from the coupling holes, this will occur with frequency independence. If, at a given frequency, one of the shorts is moved $1/2\ \lambda_g$ from its initial position, the junction will still transmit at that frequency. This $1/2\ \lambda_g$ now corresponds to some round trip line length differential $\Delta \ell = \lambda_g$. If the frequency is now changed to the point where $\Delta \ell = n/2\lambda'_g$, the phase relationships in the junction become such that the device is totally reflecting; the incident power is all returned out the input port. Hence, we have a periodic frequency filter. The Q of this is adjustable in a digital fashion by making the initial round trip displacement $\Delta \ell = n\lambda_g$ where the Q increases with the integer n.

As a first step in the investigation of the coupler approach, calculations were made for the transmission coefficient assuming a perfect coupler in WR-10 waveguide. These calculations are summarized in the following figures. The double hatched shading indicates the pass-band and the image band if one uses a 5-7 GHz IF, and the single hatching extends this to a 4-8 GHz bandwidth.

These calculations demonstrated that a 2 or 3 λ_g short displacement could provide good image band rejection over the entire waveguide range. The crossover from 2 to 3 λ_g occurs around 100 GHz where neither setting gives an optimum rejection for 88 GHz, but either one is adequate, i.e., of the order of 14 dB. (See Figure 11.)

With this theoretically established encouragement, some experiments were performed using an available RG-99 hybrid junction.

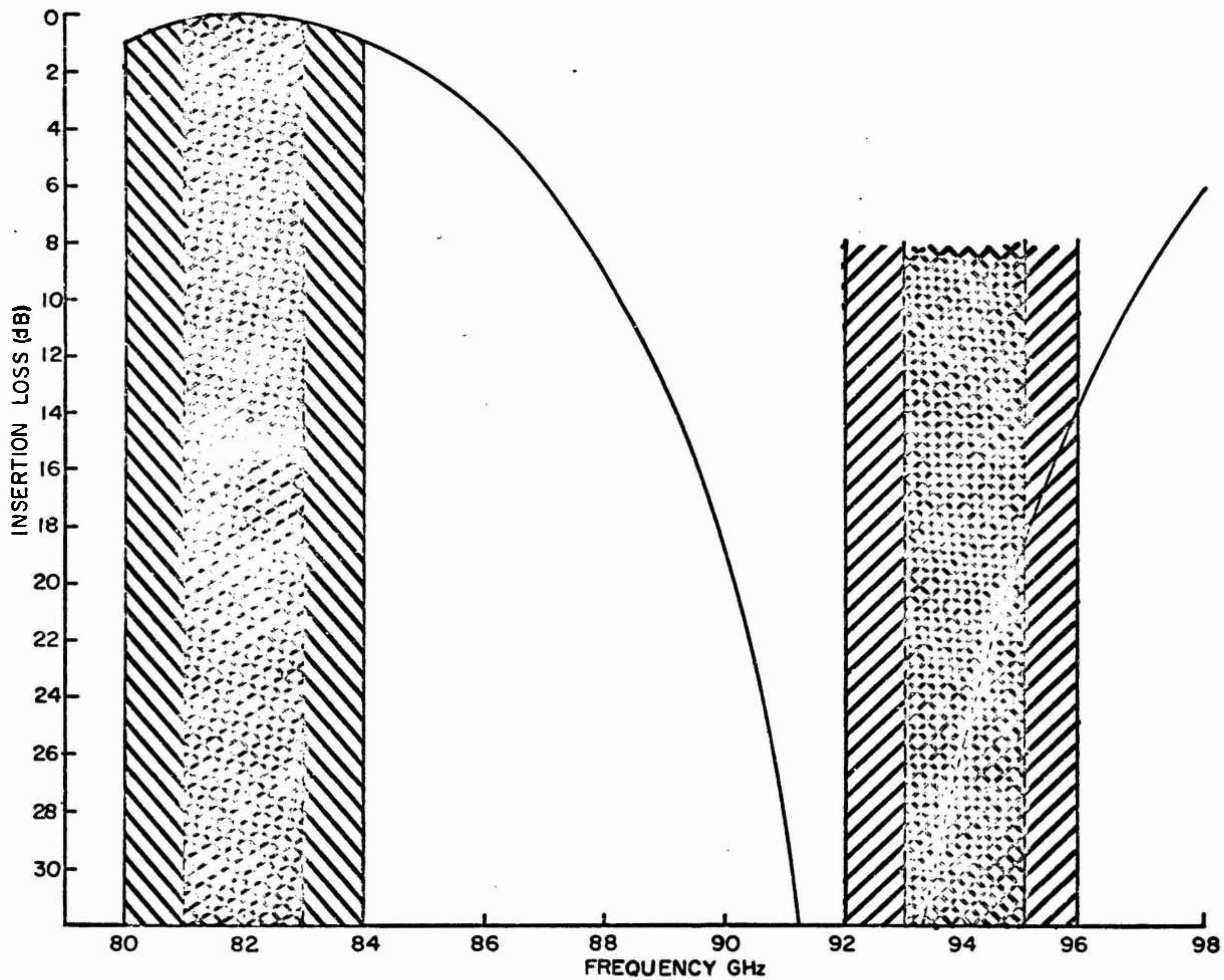

FIG. 8 - COUPLER FILTER TUNED FOR 82GHz $\Delta L = 2\lambda$.

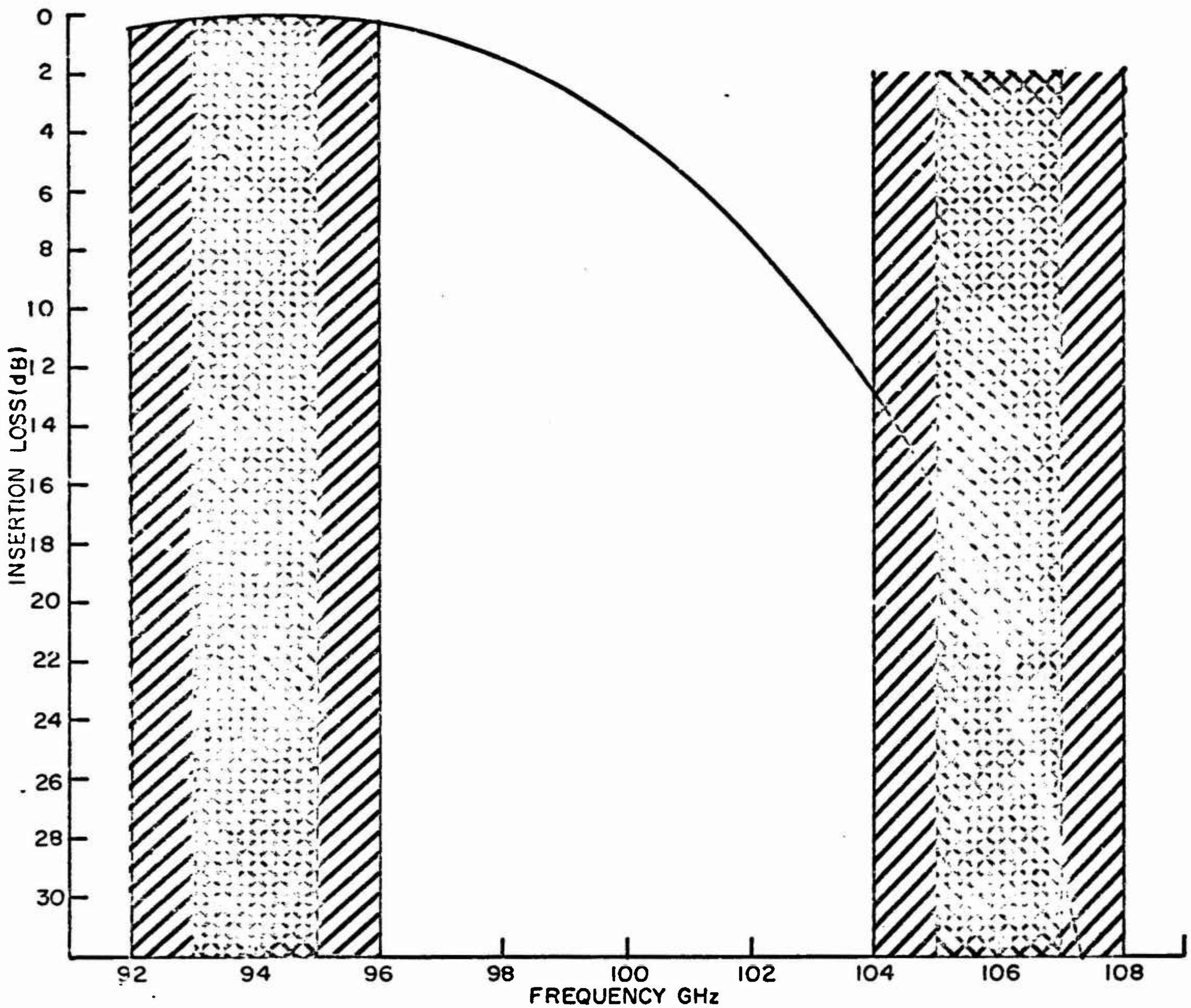

FIG. 9-COUPLER FILTER TUNED FOR 94GHz $\Delta L = 2\lambda$ LOWER SIDEBAND MIXING

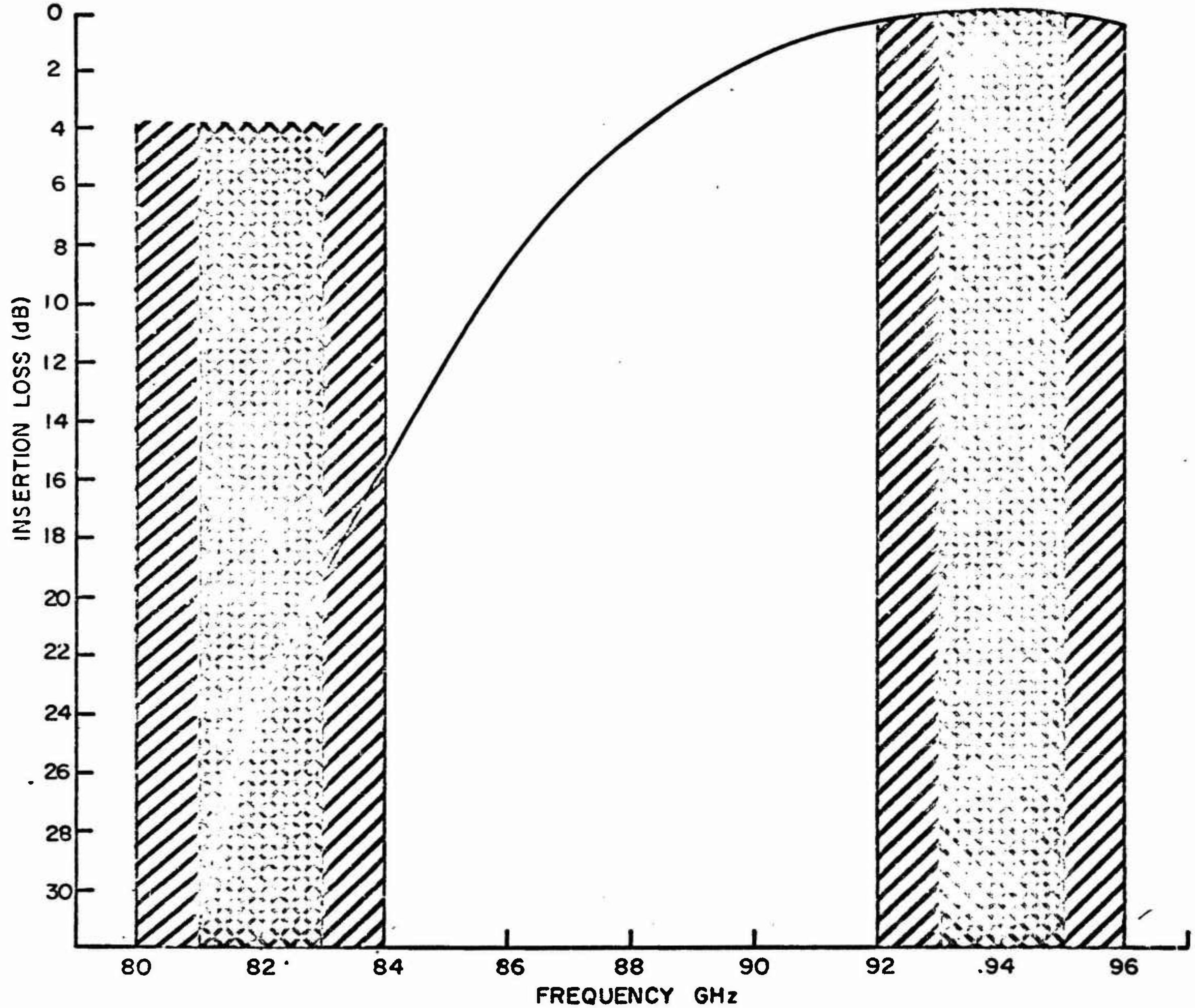

FIG.10–FILTER COUPLER TUNED FOR 94 GHz $\Delta L = 2\lambda$; UPPER SIDEBAND MIXING.

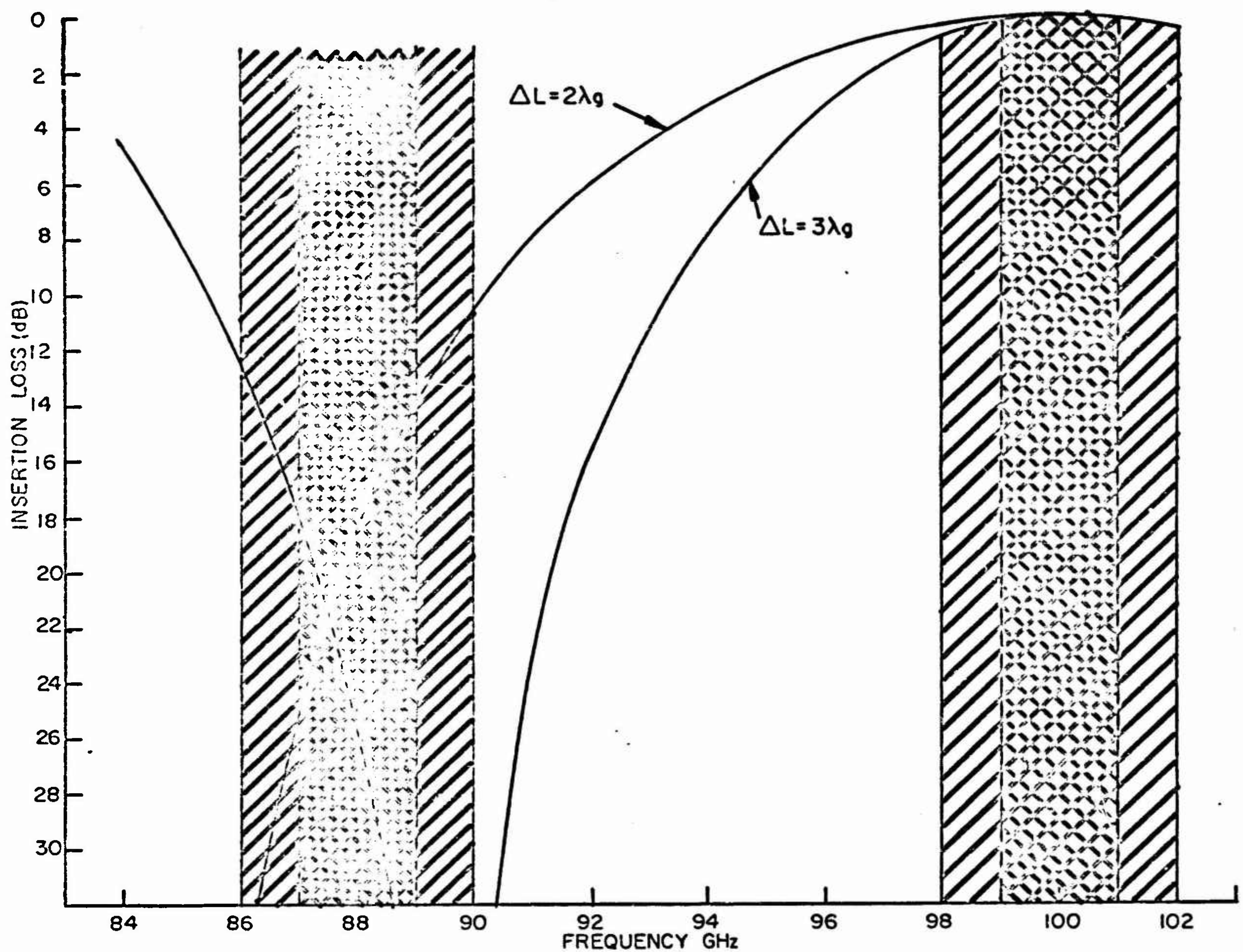

FIG.11-COUPLER FILTER TUNED FOR 100 GHz SHOWING CROSSOVER OF 2λ AND 3λ POINTS

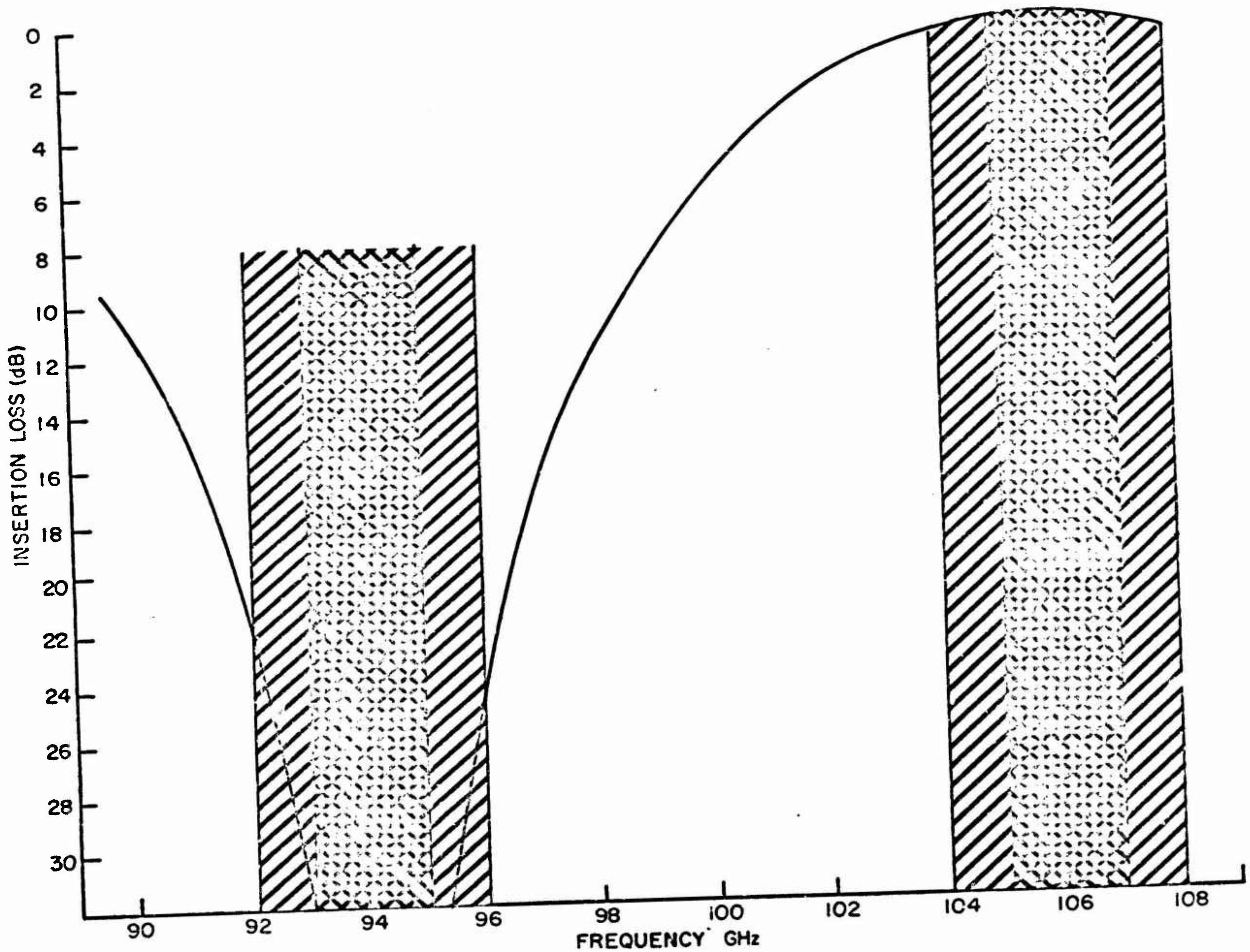

FIG.12- COUPLER FILTER TUNED FOR 106GHz $\Delta L = 3\lambda$.

The hybrid junction differs from the coupler in that it provides 0° phase shift on the H arm and 180° on the E arm and hence there is no position of shorts on the through arms which provide frequency independent operation. The lowest Q position will be when one path from the junction to the short is $1/2\ \lambda_g$, and the Q's then increase as $\frac{(2n+1)}{2}\ \lambda_g$. Another noticeable difference is that the straight hybrid junction is a much more difficult device to match out over a waveguide band than a directional coupler. However, it was decided to get a first check on the theory using available components and about in the middle of their frequency range.

The results achieved are shown in the accompanying figures. The line in the middle of the shaded regions represents a theoretical curve based on perfect splitting and no losses. The limits of the shaded area represent the theoretical limits of the deviation from this curve if one assumes 26 dB directivity and permits this misdirected energy to have a random phase with respect to everything else.

In both cases, the signal was fed into the H-arm, shorts were located on the through arms and signal level measured at the E arm after the shorts were fixed for the initial transmission. In Figure 13, a $5/2\ \lambda_g$ path difference was used with 0.6 dB insertion loss at 73.5 GHz and the rejection coming in good agreement with the 82 GHz calculated. In Figure 14, a $7/2\ \lambda_g$ difference was established at 82.5 GHz with one dB insertion loss, and although the Q seems slightly sharper than predicted, the agreement is quite reasonable. Since the curve is derived from a function which has the form $\log\left(\frac{1+\cos\theta}{2}\right)$, we have a slowly varying function where θ is small, i.e., near resonance which is the desired behavior.

This approach to a filter had the definite advantage of being entirely calculable, and on the basis of the preliminary data warranted further investigation. Hence, a 3 dB coupler was designed and fabricated.

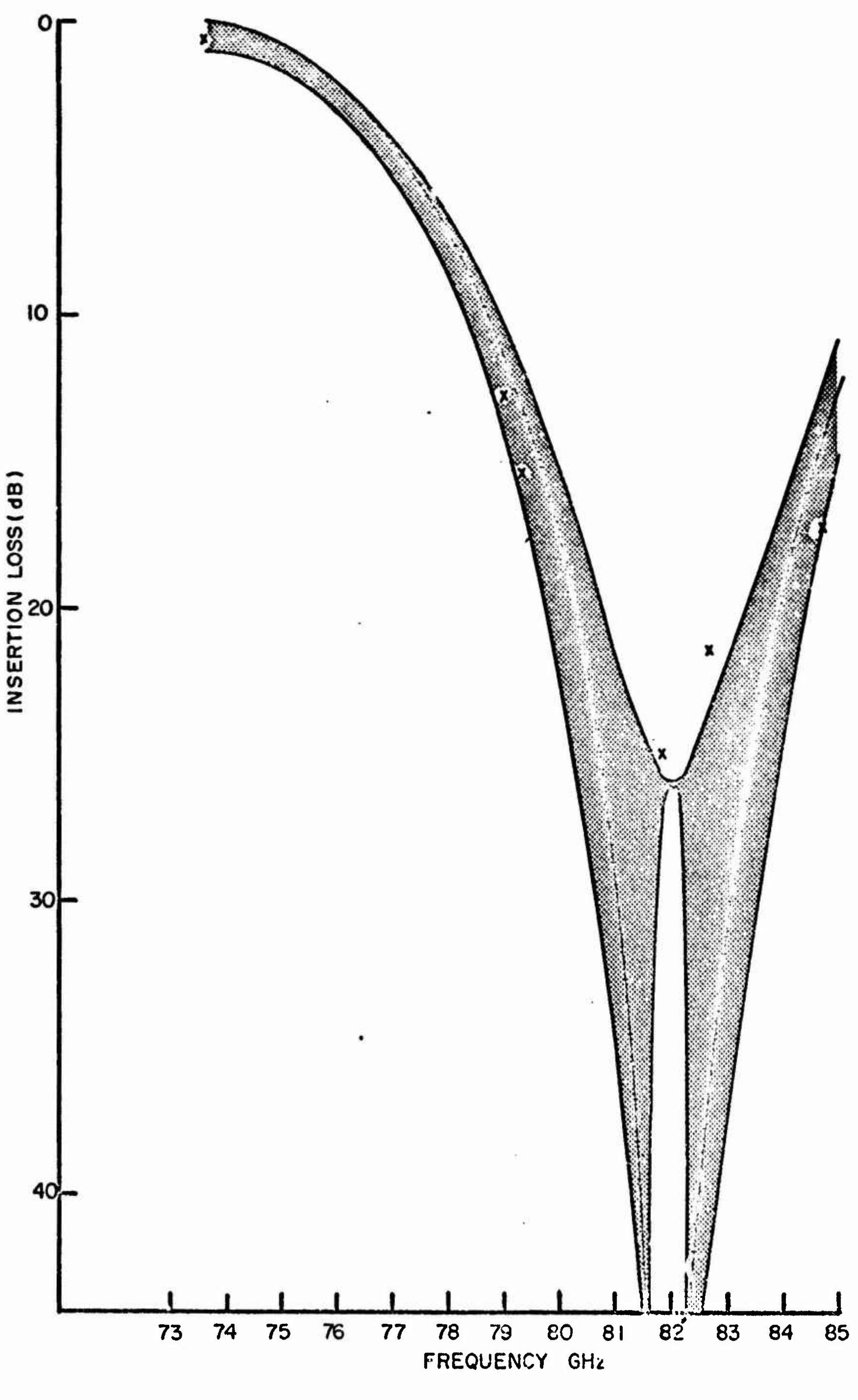

FIG.13-CALCULATED INSERTION LOSS OF A HYBRID JUNCTION FILTER
(MEASURED POINTS X)

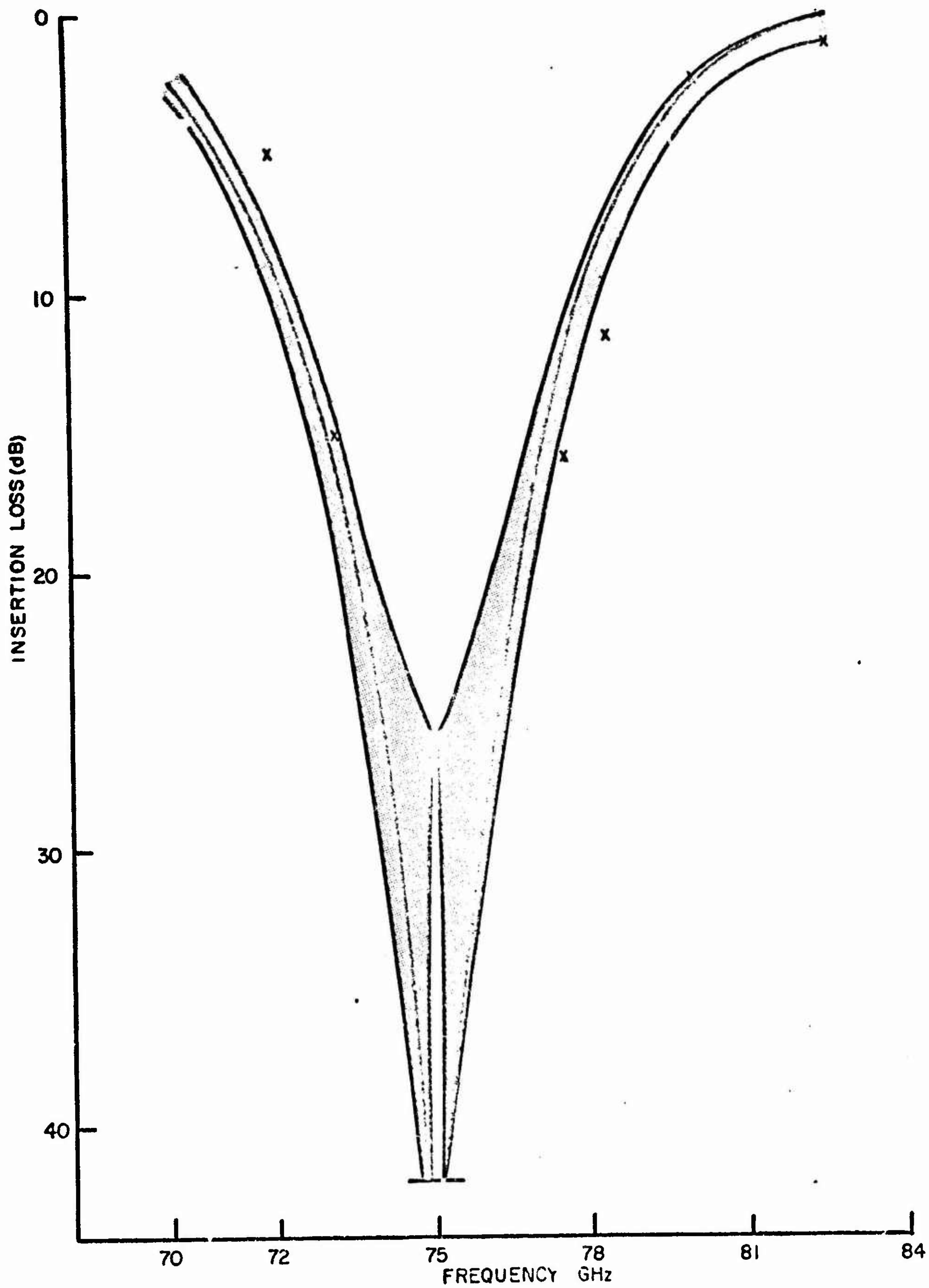

FIG.14-CALCULATED INSERTION LOSS OF A HYBRID JUNCTION FILTER.
(MEASURED POINTS X)

Consideration was given to the fact that ultimately one would deal here with incoherent noise and rely on phase cancellation effects for the filter. However, it has been determined that as long as the length of the delay in one arm of the filter is small compared with the wavelength of the process, the effect can be neglected. At the worst case, the delay will be about 1/2" (at 75 GHz) and a full 4 GHz process has a wavelength of about 3", which fairly well meets the requirement. In the narrow band (YIG-filtered) radiometer mode, the approximation is excellent.

The 3 dB coupler was again an electroformed structure. The coupling holes were drilled in a brass shim which was then gold plated and precisely glued between two polystyrene mandrels. This assembly was then electroformed as a unit. One of the first such units, with the top wall cut away, is shown in Figure 15. The final design of the coupler resulted in a power split giving signals down 3.3 and 3.5 dB out the coupling and through arms, respectively. This represented about 0.4 dB of dissipative loss through the device, or 0.8 dB in the two passes which are necessary in the filter application. Within the limits of measurement error at least, the imbalance of power and the directivity add negligibly to this quantity in mid range. At 94 GHz, the insertion loss as a filter was 0.8 dB. At the lower end of the range this increased to as much as 1.2 dB at 75 GHz due to the lack of symmetry in the power split.

The coupler was then converted to its permanent filter configuration. One port was terminated in a fixed short soldered to the waveguide. A second port was terminated in a moveable short with micrometer drive. The adjustable short circuit was made in a non-contacting dumbell configuration. This short was first checked over the waveguide band and indicated a VSWR in excess of 30 to 1 at all frequencies in the range. Since all interior finishes are gold

FIG. 15 - Cut-away View of 4 Port, 3 dB Coupler

and there are no contacting moving parts, the electrical characteristics should remain unchanged indefinitely.

The final version of the tunable filter then is shown in Figure 16, and its filter characteristics are indicated in summary form in Figure 17. This latter figure deserves a word of explanation.

The solid line is the theoretically calculated position of the shorting plunger for minimum insertion versus frequency and indicates the change from $2\lambda_g$ to $3\lambda_g$ at 100 GHz. The X's on the diagram indicate the experimentally measured points corresponding to this curve. The cross hatched areas are the regions corresponding to a 4-8 GHz image band for the solid curve, and the small circles represent measured points of maximum insertion loss vs frequency (all > 30 dB).

Thus, for example, the micrometer setting which was measured to give optimum transmission at 82 GHz is the same setting which was measured to have maximum rejection at 93 GHz which provides the image band rejection desired. Similarly, the setting which gives optimum transmission at 84 and 109 GHz has its maximum rejection for 96 GHz, thus demonstrating the image band rejection for either upper or lower sideband mixing.

In all cases, rejection at the maximum point was in excess of 30 dB and transmission at the minimum was less than 1 dB loss (with the exception of the very low end of the band where this rose to 1.2 dB). This approach to a filter is considered very practical where a low value of Q_L is desired and the periodic effects of the filter characteristic can be accepted or, as in this case, even used to advantage.

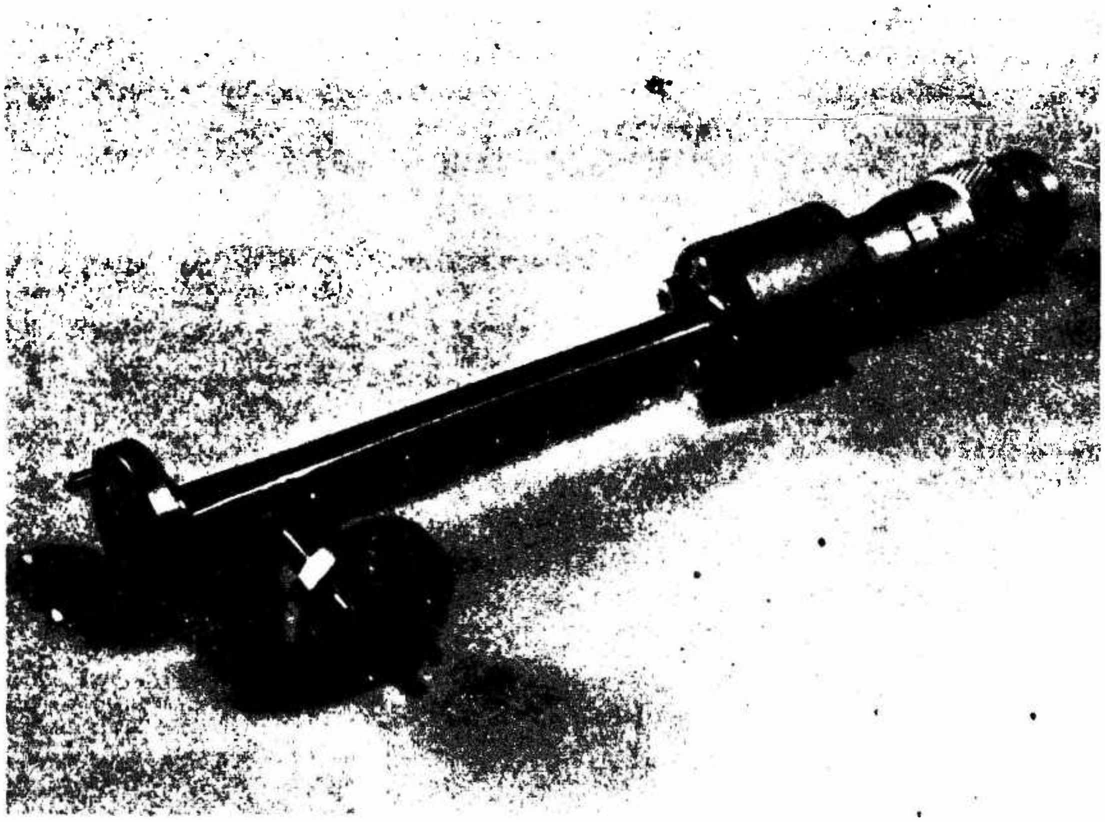

FIG. 16 - Tunable Coupler Filter

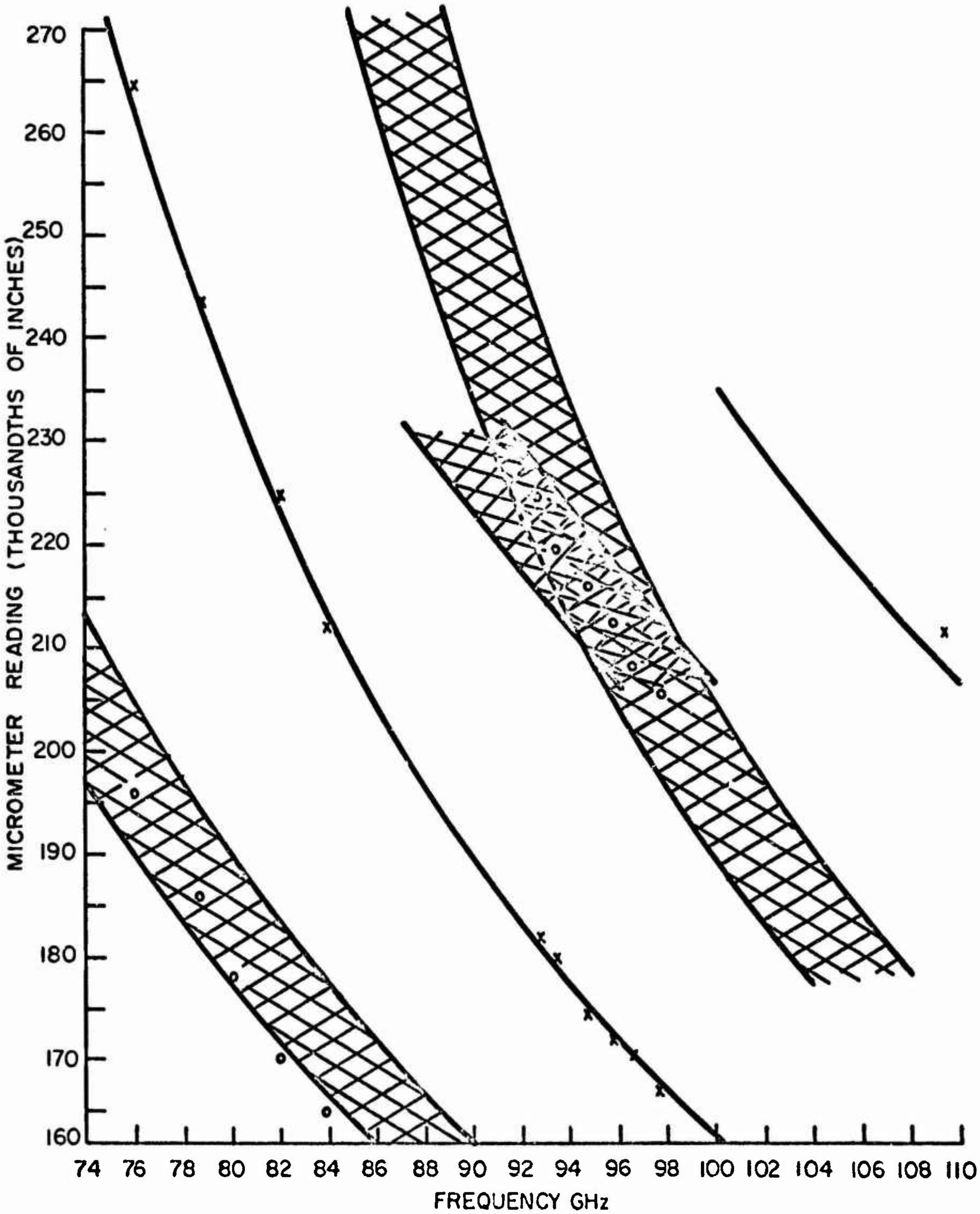

FIG.17-CALIBRATION CURVE FOR TUNABLE FILTER (X,O MEASURED POINTS)

4. A SEALED BROADBAND HARMONIC MIXER

Of the three major components requisite for a tunable radiometer system, the mixer, which at the outset seemed the least troublesome item, proved in fact to be the most difficult. Although mixer technology in this frequency range has been well developed at ADTEC the requirement of full waveguide bandwidth operation on a crossed waveguide structure had not previously been thoroughly explored. Unfortunately, the diode junction in this structure is not a problem which is amenable to anything like a rigorous theoretical analysis. The boundary value problem is much too complex to be treated in this fashion. Hence, after some reasonable estimates of the parameters involved, one uses all the variables at hand to effect an empirical solution to the problem.

4.1 Mechanical Structure

The mechanical problems of fabricating an hermetically sealed structure were met fairly readily. Figures 18 and 19 show the basic block as it was machined with the crossed RG-96/U the WR-10 waveguides. The circular grooves around the waveguides in Figure 18 provide a well for the epoxy used to hold the teflon windows in place. (Although the original design called for .001" mica, the teflon was found to have a better VSWR in the short millimeter wavelength region.) The hole immediately below the WR-10 flange provides a means of achieving a mechanical lock on the post carrying the whisker wire.

In Figure 19 the circular rings are O-ring grooves, and the circular holes are drilled to accomodate the bellows which provide the internal seal and permit the moveable short circuit to be located within the sealed area.

Figure 20 is a cut-away view of the assembly. In this

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FIG. 18 - Harmonic Mixer, Front View

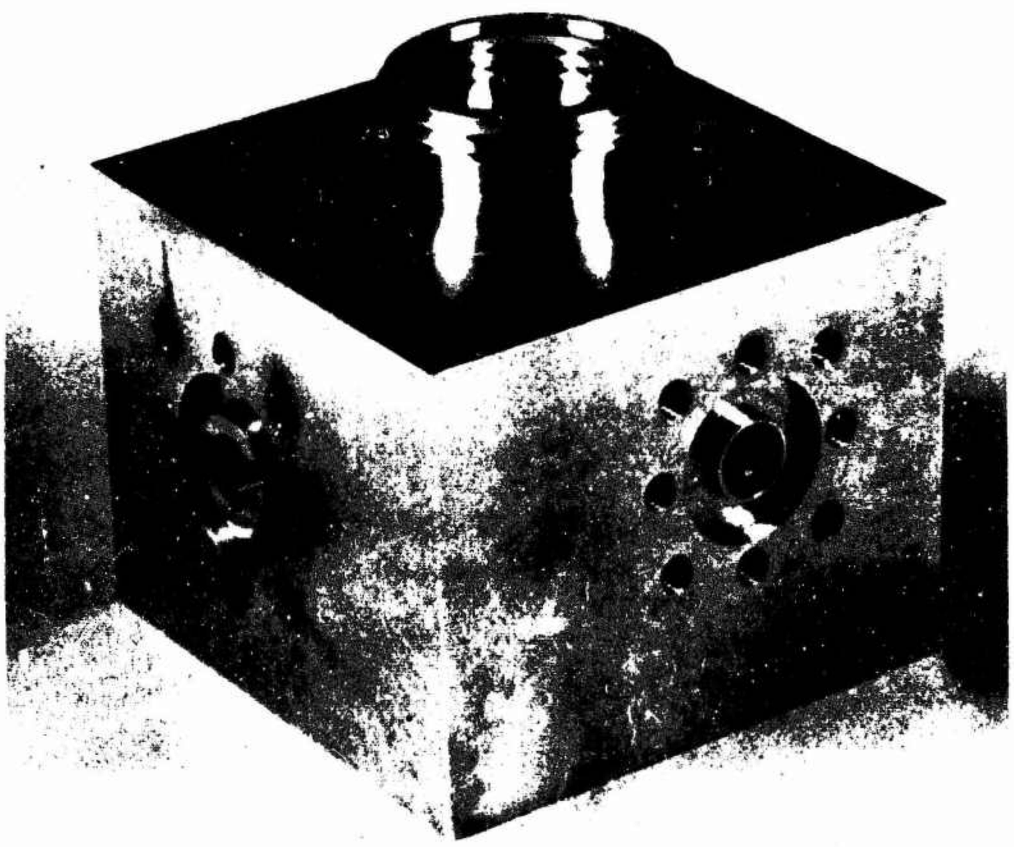

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FIG. 19 - Harmonic Mixer Body, Rear View

it can be seen that positive stops are provided in both directions such that the bellows cannot be overextended or compressed. The bellows themselves are .001" nickel and permit greater than 1/2 λ_g travel of the short circuit in either waveguide.

Before the diode is formed in the mixer, the windows are put on the waveguides Mounting of the shorting plungers then seals the opposite ends of the waveguides. After the mixer is made, the whisker and crystal post holes are sealed. One purging hole is left in the open area between the bellows and one shorting plunger. The mixer is then placed in a bell jar which is alternately evacuated and replenished with dry nitrogen through several cycles. After the final cycle, the last purging hole is sealed while the mixer is still in an inert atmosphere. Thus any deterioration which would occur at the junction due to the presence of water vapor and/or oxygen is essentially eliminated.

4.2 Electrical Performanc

There are several requirements which must be met simultaneously in order to achieve good mixing performance through the range. These are:

a. The combination of the whisker post, whisker wire, adjustable short and coupling hole must present a good match to the local oscillator and efficiently couple the energy into the smaller waveguide and into the junction.

b. The crystal post, diode junction, whisker wire adjustable short and coupling hole must also present a match in the signal waveguide which will couple the energy to the diode junction and not into the local oscillator waveguide.

c. The capacitive bypass at the crystal post must be sufficient to provide a shunt keeping the L.O. and signal power

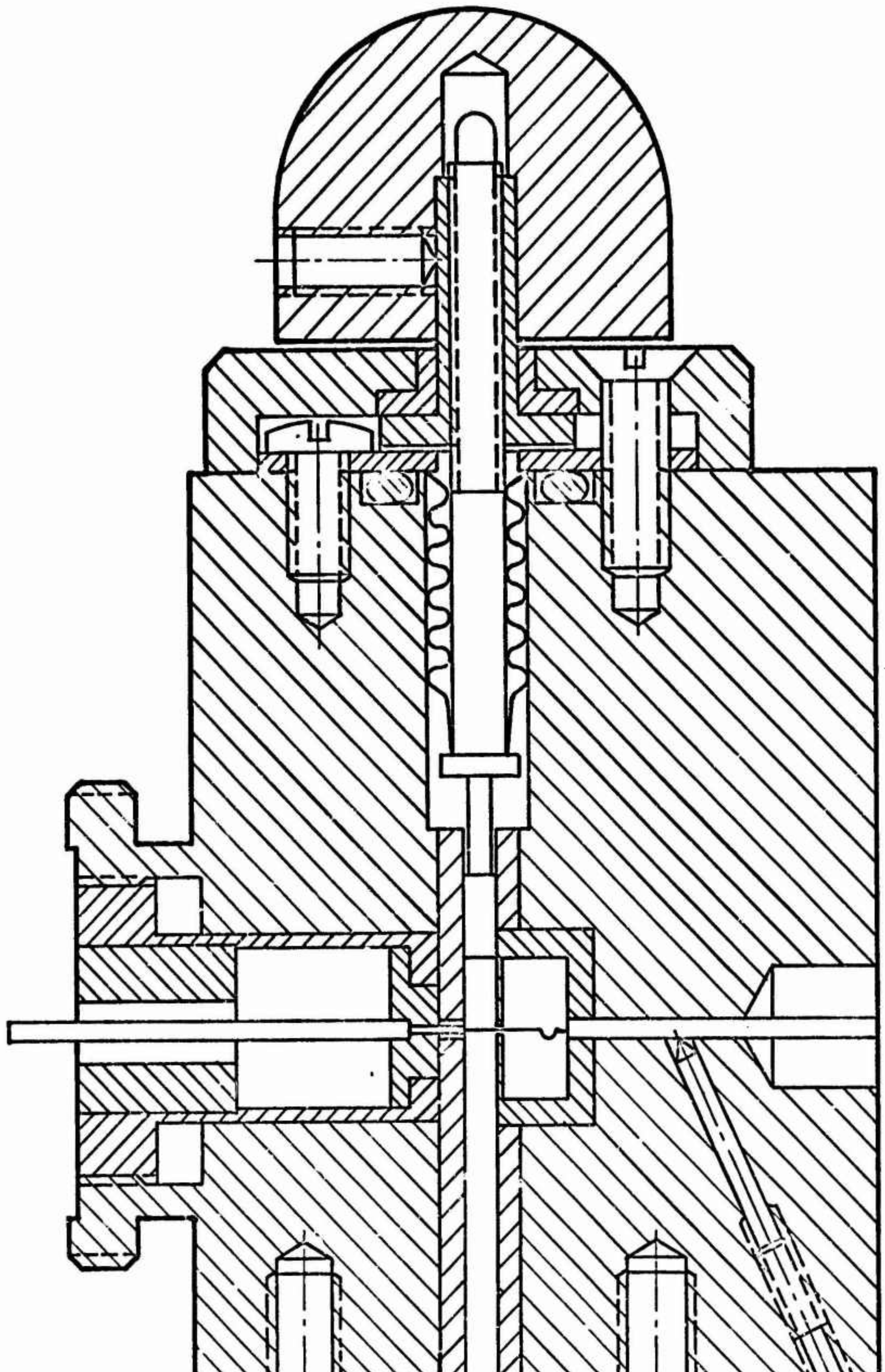

FIG. 20 - CROSS SECTIONAL VIEW OF MIXER ASSEMBLY

from leaking out the coax input; but this capacity must be small enough that a resistive match can be achieved at the IF terminal.

One of the main difficulties in attempting any sort of analysis as to how best to meet these conditions is the lack of information concerning the mode of propagation of the energy along the whisker wire. One could assume in the region of the common wall a TEM coaxial type mode would be present. But at the local oscillator frequency, this section is only an eighth of a wavelength and the effects of the abrupt transitions are probably at least as significant as the length of line itself. In the small wave-guide which is beyond cut-off for the L.O. frequency, one can still determine that the fields extend some distance from the wire since moving the short circuit in the signal waveguide has an effect on the crystal current produced by the local oscilllator.

In an effort to eliminate signal leakage out the L.O. waveguide, the common wall was made to be $\lambda/4$ at the middle of the signal range, assuming a TEM mode. The theory here is that the high impedance (approximately 400Ω) of the L.O. waveguide would transform to a low impedance at the signal guide wall. This approach is limited by the physical aspects of the problem, however. Although larger wires were tried - up to .003" dia. gold - the only good mixer performances were achieved with .001" phosphor bronze. To accept some flexing of this wire and some error in concentricity, the coupling hole was .010" dia. This results in a coaxial impedance on the order of 140 Ω which only permits a transformation of roughly 400 to 30 Ω on the quarter wave section, even if the problem were that simple. Even in the final mixer assembly, some of the loss quoted as transducer loss is actually due to non-recoverable signal leakage into the L.O. waveguide.

This was a very significant item in early stages of the mixer development. A diode would be formed using a signal frequency of 94 GHz and an L.O. frequency of 44 GHz and conversion loss measured. The signal frequency would then be changed to 82 GHz thus introducing no change in IF or local oscillator frequency. The conversion loss after adjusting the shorts for optimum performance might be as much as 10 dB worse with no measurable change in the RF input VSWR.

The only adjustments available for matching purposes within the structure were the crystal and whisker posts so a rather prolonged series of tests were begun. With the crystal post at a fixed insertion depth into the signal guide (where we construe this depth to be negative when the post is not into the guide) a series of contacts would be made and measured at at least two widely separated frequencies with successively shorter whisker lengths. Sometimes the data was difficult to interpret. When one makes a series of contacts trying to maintain identical conditions there is a certain spread in the data which at times is considerable. This random spread was sufficient at times to conceal any trend which might have been apparent using identical diodes.

Ultimately a configuration was reached which apparently matched the structure at all frequencies checked. This configuration had the crystal post inserted some .008" into the signal guide. The whisker carrier post crossed the local oscillator waveguide to a point about .010" from the common wall. The C bend in the whisker wire (as depicted in Figure 20) was eliminated. Contact pressure was maintained by a bow in the wire itself. The diode junction itself, of course, was welded by the forming current.

There is no claim made here that this is the optimum arrangement for this type of mixer. However at all frequencies checked from 80 to 110 GHz it was possible, by adjusting the shorting plungers and L.O. drive, to achieve a total loss through the mixer of 14 dB or less, which was the design goal. This figure includes the losses due to RF and IF mismatches as well as an unknown amount of signal leakage. This measure of conversion loss on a coherent signal was later verified using the complete radiometer system.

5. ASSEMBLED RADIOMETER SYSTEM

The last phase of the work performed on these assignments was to check them in concert and evaluate the system as a whole in both the narrow band and broad band modes of operation. As a subtask of the mixer development work, a YIG filter had been purchased and a control panel assembled. These are pictured in Figures 21 and 22. The driver panel was designed such that a 10 turn precision potentiometer controlled the center frequency of the filter in a linear fashion from 4-8 GHz.

The YIG filter itself was a 4 sphere, 2 section filter which could be used as either a single or two stage device. Tests with a swept frequency source indicated that the spurious responses were always greater than 50 dB below the pass band response for any setting of the center frequency. These tests also showed that each section of the filter must be well isolated from any other component in order to achieve a good bandpass characteristic. Measurements were made on the gain of the IF through the cascaded pair of TWT amplifiers, and then through the pair with the filter between them. Lacking isolators at this frequency, the isolation of the filter was achieved through the use of attenuator. Hence the insertion loss of the filter seems high. These data are presented in Figures 23, 24, and 25. Most of the fluctuations vs frequency are introduced by the TWT pair. With the exception of the data taken at 4.5 and 6.1 GHz, the YIG introduced only an additional ± 1 dB variation.

Assembly of the complete radiometer system revealed some of the problems which arise when integrating several components. Foremost among these was a problem of 60 cycle pick up in the 75 cycle synch signal circuitry. Re-wiring of the chopper control panels and the addition of shielding in the cables has apparently eliminated this difficulty.

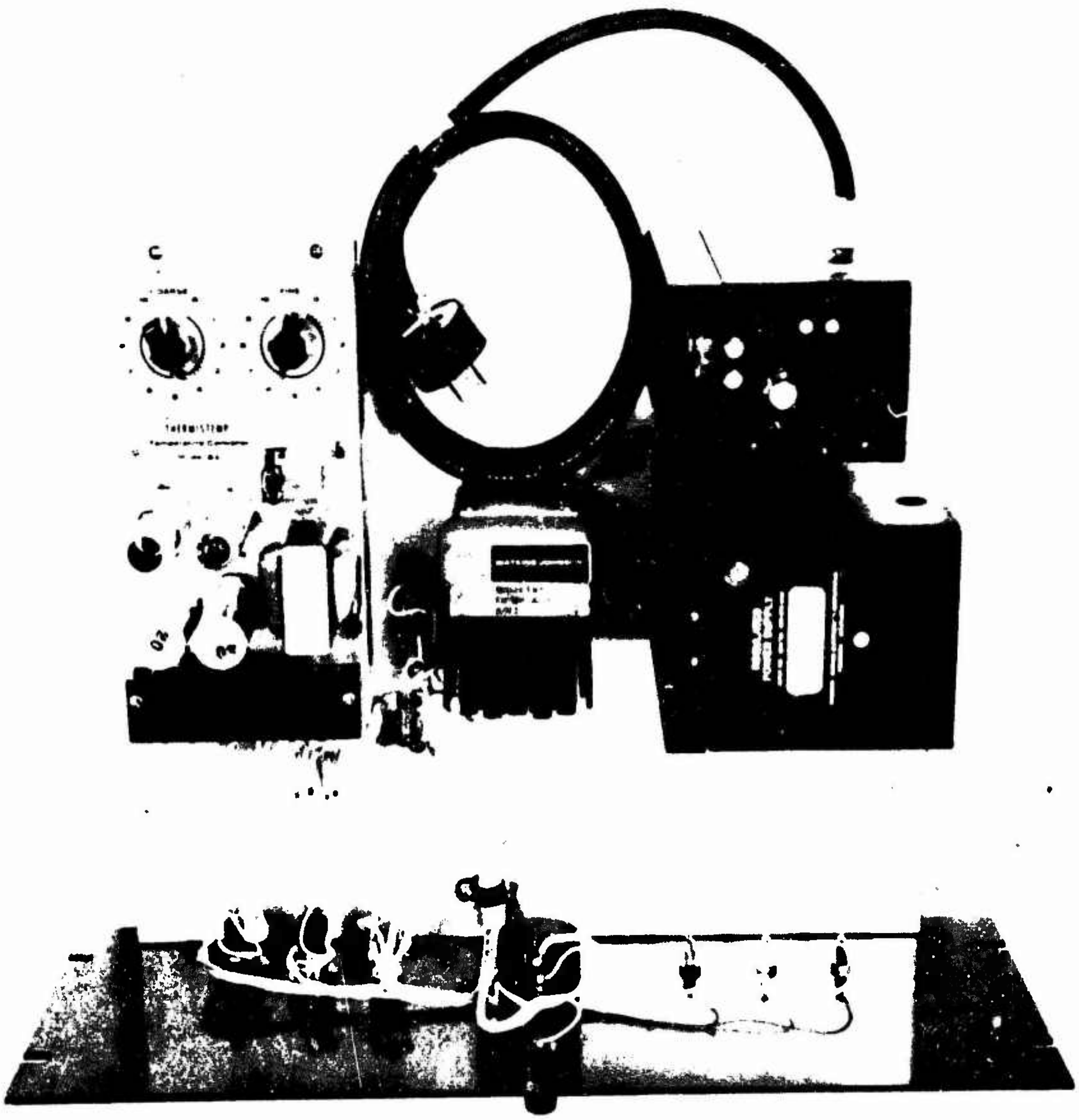

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FIG. 21 - YIG Filter Control Panel

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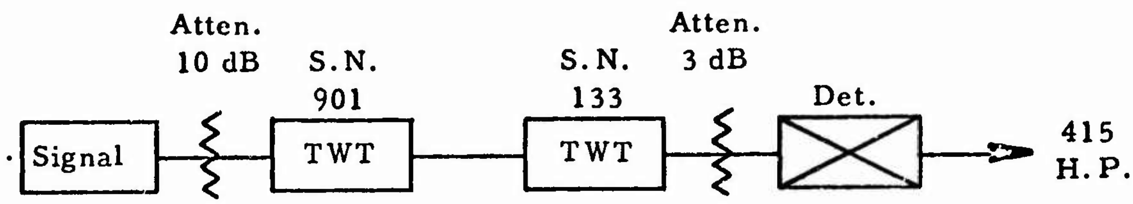
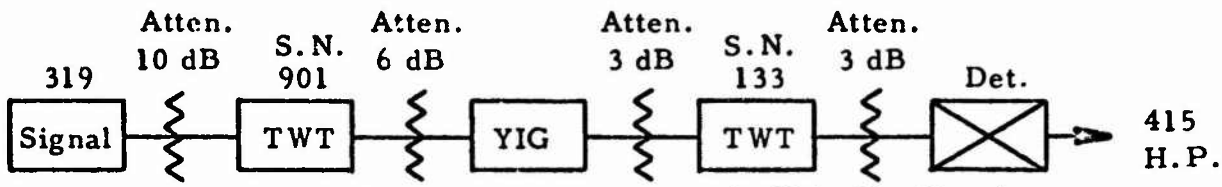
FIG. 22 - YIG Filter



Frequency (GHz)	Gain (+ dB)	Frequency (GHz)	Gain (+ dB)
3.8	71.6	5.8	72.3
3.9	72.5	5.9	72.0
4.0	73.0	6.0	72.0
4.1	73.3	6.1	72.0
4.2	74.5	6.2	72.0
4.3	75.0	6.3	71.8
4.4	75.0	6.4	71.7
4.5	75.0	6.5	71.5
4.6	74.3	6.6	71.0
4.7	74.0	6.7	70.5
4.8	74.1	6.8	69.0
4.9	74.2	6.9	67.0
5.0	73.6	7.0	68.0
5.1	73.8	7.1	69.0
5.2	73.0	7.2	68.0
5.3	73.0	7.3	68.5
5.4	73.0	7.4	69.0
5.5	72.8	7.5	68.5
5.6	72.0	7.6	68.0
5.7	72.2		

FIG. 23 - Measured IF Gain Characteristics

FIG. 24 - Measured IF Gain Characteristics;
Narrow Band



Frequency (GHz)	Gain (dB)	Pot. Position for YIG Filter
3.8	54.5	0.050
3.9	55.0	0.240
4.0	56.2	0.450
4.1	57.0	0.665
4.2	58.0	0.890
4.3	59.0	1.120
4.4	59.0	1.350
4.5	57.0	1.600
4.6	57.0	1.850
4.7	58.0	2.100
4.8	57.0	2.370
4.9	58.0	2.640
5.0	57.0	2.920
5.1	58.0	3.180
5.2	58.0	3.450
5.3	56.5	3.730
5.4	57.0	4.010
5.5	57.0	4.290
5.6	56.0	4.570
5.7	56.5	4.370
5.8	56.0	5.120
5.9	55.5	5.890
6.0	55.0	5.680
6.1	54.0	5.950
6.2	55.0	6.210
6.3	55.0	6.490
6.4	54.6	6.750
6.5	55.0	7.010
6.6	54.5	7.260
6.7	53.0	7.500
6.8	51.0	7.790
6.9	51.0	7.990
7.0	52.0	8.200
7.1	52.0	8.410
7.2	51.0	8.620
7.3	51.0	8.820
7.4	50.0	9.040
7.5	51.0	9.220
7.6	52.0	9.410

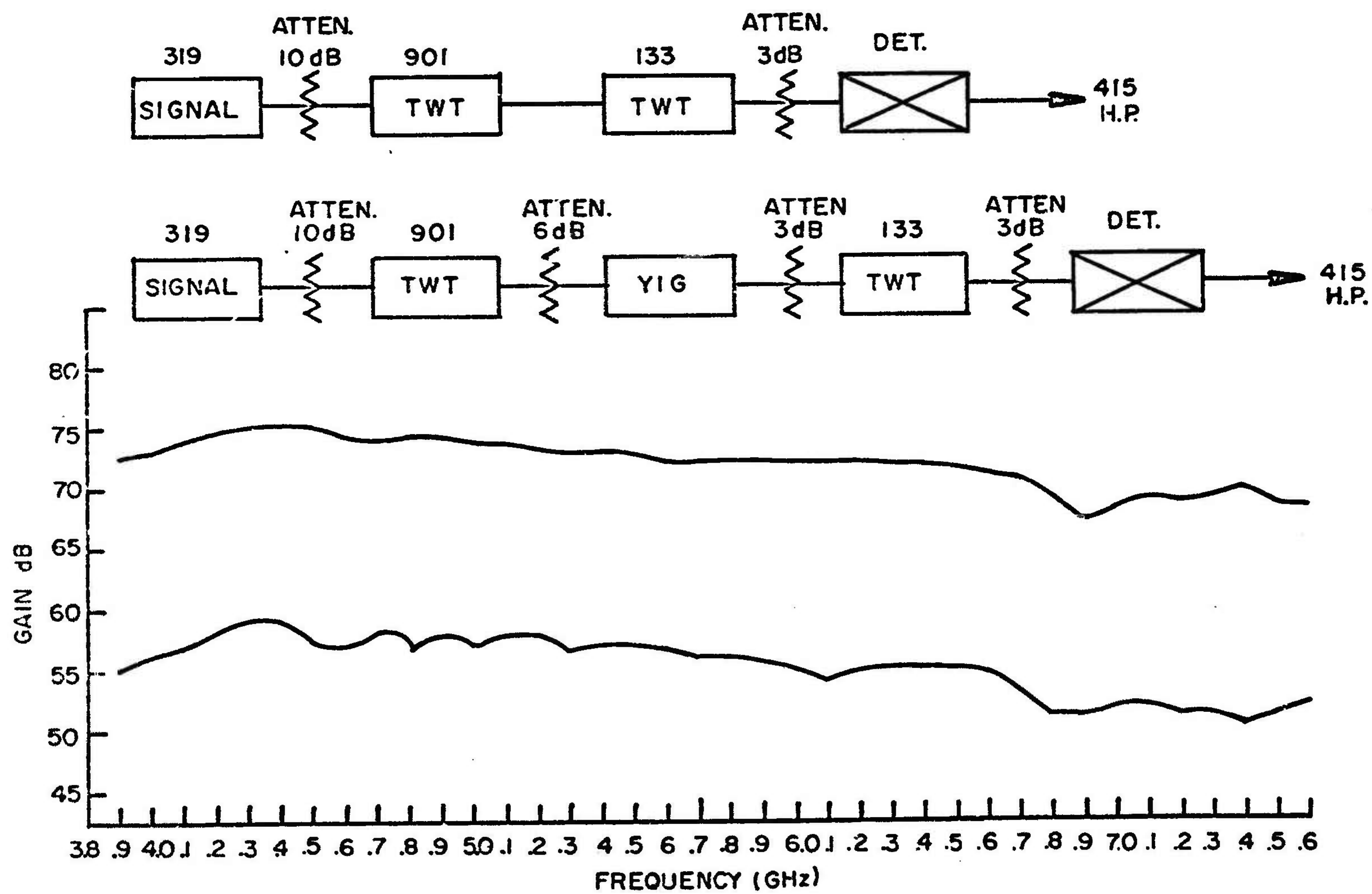


FIG.25-CASCADE GAIN CHARACTERISTIC OF W.J. TWT'S SERIAL NUMBER 901 & 133 4-8(GHz) AND THROUGH I STAGE YIG FILTER.

The assembled radiometer apparently worked as expected. The overall loss of the mixer using a coherent source had been measured between 13 and 14 dB. Assuming a crystal noise ratio of two or less and accepting the manufacturer's measurements of 5.9 dB noise figure of the TWT, the single sideband radiometer noise figure would be between 20 and 21 dB. Then adding to this the one dB insertion loss of the single sideband filter, using an ambient temperature of 298°, with a bandwidth of 4 GHz and integration time of 0.1 seconds, $(\Delta T)_{min}$ calculates to be between 1.87° and 2.36°. Depending on the criteria used, the tangential signal level would then be somewhere in the range of 7.5° to 14°.

Tests of the assembled radiometer system using ice water and dry ice in alcohol as reference temperatures indicated that the peak to peak noise fluctuations on the output were actually of the order of 6°, hence the tangential signal level would fall in the proper range to give agreement between the system sensitivity as measured with a coherent source and when used with broadband radiation.

The completed items as shown in Figure 26 were delivered to the Ballistic Research Laboratories at Aberdeen Proving Grounds in late October. It is felt that these components have been completed in accordance with the original design goals and should furnish good operation in a radiometer system. Further testing of the overall system and data acquisition will be carried out by the Ballistic Research Laboratories.

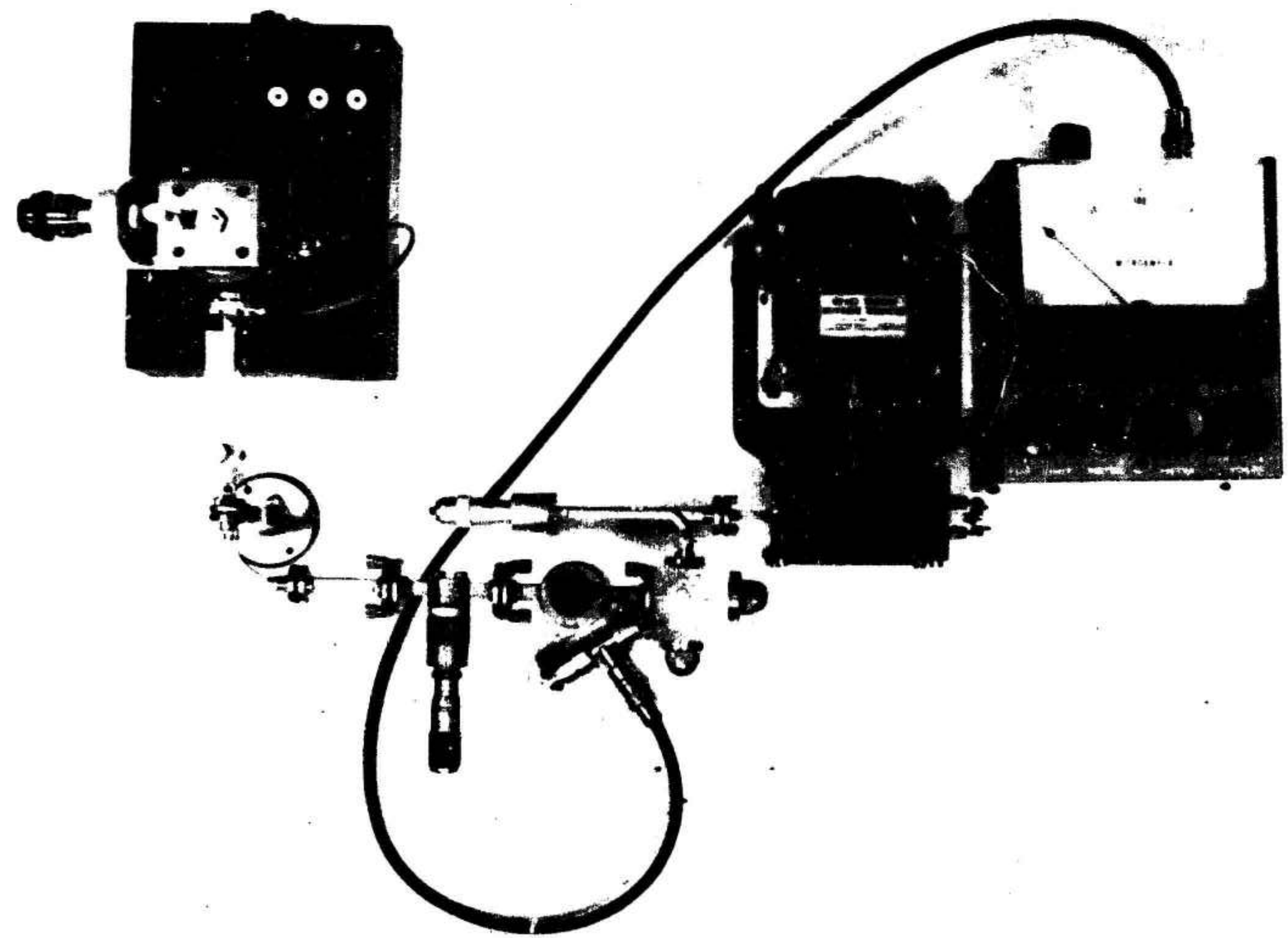


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FIG. 26 - Radiometer System Components

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3. REPORT TITLE

Research Study into the Near Earth Application of Millimeter Radio Waves as Applied to Certain Battlefield Problems

4. DESCRIPTIVE NOTES (Type of report and inclusive dates)

Interim Engineering Report 19 July 1965 - 31 October 1966

5. AUTHOR(S) (Last name, first name, initial)

Cotton, John M., Jr.
Dozier, John W.

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
November 30, 1966	52	--

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13. ABSTRACT

This report summarizes the work done in developing three broadband millimeter wave components: a mechanical modulator, a tunable bandpass filter, and a second harmonic mixer. All the components were designed to operate throughout the 75-110 GHz frequency region. The ultimate application of these particular components was a tunable radiometer capable of operation through the vicinity of the 94 GHz window. The text describes the techniques used to fabricate the items and gives performance data on the components singly as well as on the completed radiometer assembly.

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Security Classification

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14. KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
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Mixer, Harmonic						
Filter, Tunable						
Modulator, Mechanical						

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