

ADTEC

50,101-21

GPO PRICE \$ _____

CFSTI PRICE(S) \$ _____

Hard copy (HC) 4.00

Microfiche (MF) 1.00

Final Engineering Repc

July 29, 1966

ff 653 July 66

SUBMILLIMETER SATELLITE RADIOMETER

by

R.F. Packard, M. Cohn, R.J. Bauer,
J.M. Cotton, Jr., L.E. Dickens, J.W. Dozier, J.D. Rodgers

Prepared under Contract NASw-1000

for

Office of Space Science and Applications
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
Washington, D.C.

ADVANCED TECHNOLOGY CORPORATION

1830 YORK ROAD

TIMONIUM, MARYLAND

N67 12068

(ACCESSION NUMBER)

(THRU)

(PAGES)

(CODE)

(NASA CR OR TMX OR AD NUMBER)

(CATEGORY)

FACILITY FORM 502

50,101-21

July 29, 1966

Final Engineering Report

SUBMILLIMETER SATELLITE RADIOMETER

Project Manager: R.F. Packard

Technical Supervisor: M. Cohn

Other Contributors: R.J. Bauer
J.M. Cotton, Jr.
L.E. Dickens
J.W. Dozier
J.D. Rodgers

Prepared under Contract NASw-1000

by

ADVANCED TECHNOLOGY CORPORATION
Timonium, Maryland

for

Office of Space Science and Applications
National Aeronautics and Space Administration
Washington, D.C.

ABSTRACT

An experimental program to determine the feasibility of developing an all solid state superheterodyne Dicke radiometer operating at a wavelength of about 0.5 mm has been successfully completed. The critical components of such a radiometer are the harmonic mixer and the final multiplier stage of a solid state local oscillator source. A 560 GHz fourth harmonic mixer having a conversion loss of 29.5 dB with an LO drive level of 10 milliwatts was developed. An experimental epitaxial GaAs varactor having a zero bias cutoff frequency of about 1000 GHz was developed. This varactor was incorporated in a 70 to 140 GHz frequency doubler which yielded an output of 10 mW for an input of 40 mW (25% efficiency).

With devices having these characteristics a sensitive submillimeter radiometer receiver could be built. It would have a calculated minimum detectable temperature, $(\Delta T)_{\min}$, of 7°K for a signal-to-noise ratio of unity. This calculated sensitivity is based upon: RF losses of 3 dB, mixer diode noise ratio of 2, IF noise figure of 4 dB, IF bandwidth of 1.6 GHz, and output integration time of 10 seconds. These receiver component characteristics are realistic values which have been achieved.

TABLE OF CONTENTS

	<u>Page</u>
I. SUBMILLIMETER SATELLITE RADIOMETER	1
A. General Considerations	1
1. <u>Objectives</u>	1
2. <u>Submillimeter Wave Propagation in the Atmosphere</u>	2
3. <u>Requirements for Space-Borne Radiometer</u>	4
B. Superheterodyne Radiometer Receiver	4
C. Key Components Selected for Development	7
D. Estimated Radiometer Sensitivity	8
II. SUBMILLIMETER HARMONIC MIXER	10
A. Design and Construction of the Diode Mount	10
B. The Mixer Diode	16
C. Test Procedures	18
D. Test Data	24
1. <u>Mixer Conversion Loss Measurements on 2-28-66 at 560 GHz</u>	25
2. <u>Mixer Conversion Loss Measurements on 3-7-66 at 560 GHz</u>	29
3. <u>Verification of Operating Frequency</u>	32
4. <u>Mixer Loss Measurements with Dual IF Conversion at 420, 560, and 700 GHz</u>	35
5. <u>Mixer Conversion Loss Measurements on 7-24-65 at 525 GHz</u>	38
6. <u>Mixer Diode Noise Ratio</u>	38
7. <u>Mixer IF VSWR</u>	39
E. Conclusions	39
III. SOLID STATE LOCAL OSCILLATOR	43
A. Design Considerations	43
1. <u>Requirements</u>	43

TABLE OF CONTENTS (Continued)

	<u>Page</u>
2. <u>Previous Work with Harmonic Generators</u>	44
3. <u>Characterization of Millimeter Wave Diodes</u>	48
B. Development of Epitaxial Varactor Harmonic Generator at 140 GHz	54
C. Test Procedures and Data	55
D. Data Analysis	58
E. Conclusions	60
IV. SUBMILLIMETER SIGNAL SOURCE	62
A. Design and Construction	62
B. Test Procedures and Data	64
C. Conclusions	67
V. DIODE DESIGN FOR SPACE ENVIRONMENT	69
A. General Requirements	69
B. Specific Techniques	69
1. <u>Hermetically Sealed Diode Mounts</u>	69
2. <u>Pin Diode Structure</u>	73
3. <u>Epoxied Diodes</u>	76
C. Conclusions	76
VI. CONCLUSIONS AND RECOMMENDATIONS	77
A. Feasibility of Submillimeter Radiometer	77
B. The Next Phase of Development	79
C. Later Phases	80
APPENDICES	
A. Comparison of Submillimeter Radiation Detectors	A-1 - A-6
B. Propagation in Waveguides Below Cutoff	B-1 - B-3
C. Crossed Waveguide Coupling Studies	C-1 - C-2
D. Analysis of Diode Loss Factors	D-1 - D-9

TABLE OF CONTENTS (Continued)

	<u>Page</u>
E. Tunnel and Backward Diodes	E-1 - E-4
F. 140 GHz Second Harmonic Mixer	F-1 - F-2
G. Analysis of Single Source Mixer Test System	G-1 - G-7
H. Strontium Titanate Harmonic Generator	H-1 - H-16
I. 140 GHz Tripler Using Commercial Varactor	I-1 - I-2

REFERENCES

LIST OF ILLUSTRATIONS

<u>Figure</u>	<u>Title</u>	<u>Page</u>
1-1	Satellite Submillimeter Radiometer (Possible Block Diagram)	5
2-1	Submillimeter Harmonic Mixer, Cross Section Drawing	11
2-2	Submillimeter Harmonic Mixer Photograph	12
2-3	Waveguide Mandrel 0.014" x 0.007" in Jig	13
2-4	Partial Electroform and Taper Mandrel in Jig	13
2-5	Waveguide and Taper after Electroforming	14
2-6	Waveguide and Taper Machined for Mixer	14
2-7	Single Conversion Harmonic Mixer Test Setup, 525 to 1.4 GHz, Block Diagram	19
2-8	Dual Conversion Harmonic Mixer Test Setup, 420, 560, or 700 GHz to 115 MHz, Block Diagram	20
2-9	Single Conversion Harmonic Mixer Test Setup, 560.7 to 1.3 GHz, Block Diagram	22
2-10	Measured Fourth Harmonic Mixing Conversion Loss vs Local Oscillator Power, Diode of 2-28-66	26
2-11	Direct Current in Fourth Harmonic Mixer, Diode of 2-28-66	27
2-12	Measured Fourth Harmonic Mixing Conversion Loss vs Diode Current, Diode of 2-28-66	28
2-13	DC Characteristics of Mixer Diode of 2-28-66	30
2-14	Measured Fourth Harmonic Mixing Conversion Loss vs Local Oscillator Power, Diode of 3-7-66	31
2-15	Direct Current in Fourth Harmonic Mixer, Diode of 3-7-66	33

LIST OF ILLUSTRATIONS (Continued)

<u>Figure</u>	<u>Title</u>	<u>Page</u>
2-16	Measured Fourth Harmonic Mixing Conversion Loss vs Diode Current, Diode of 3-7-66	34
2-17	Measured Total Fourth Harmonic Generation and Mixing Loss, Diode of 12-23-65	36
2-18	Measured Third Harmonic Mixing Conversion Loss, Diode of 12-17-65	37
2-19	IF Output Impedance of Submillimeter Mixer, Diode of 2-28-66	40
3-1	Harmonic Generation Conversion Loss vs Frequency and Harmonic Number	46
3-2	Measured Second Harmonic Generation Conversion Loss, 68 to 136 GHz	47
3-3	Current-Voltage Characteristics of Two Different, but Typical, Point Contact n-GaAs ($\rho = 0.015$ ohm-cm), Phosphor-Bronze Diodes	50
3-4	Current-Voltage Characteristics of Two Different, but Typical, Point Contact n-GaAs ($\rho = 0.015$ ohm-cm), Copper Diodes	51
3-5	Current-Voltage Characteristics of a Typical Point-Contact, Varactor, n-GaAs ($\rho = 0.0422$ ohm-cm), Zinc Diode	56
3-6	Crossed Waveguide Mount, Model M-9838, Photograph	57
3-7	Power Output and Conversion Efficiency of a 140 GHz Second Harmonic Generator	59
4-1	Submillimeter Harmonic Generator, Photograph	63
5-1	Hermetically Sealed Fundamental Mixers (26 to 170 GHz), Photograph	70
5-2	Vibration Testing of 94 GHz Mixer Diode	72
5-3	Millimeter Wave Diode (Pin Construction), Cross Section Drawing	74

LIST OF ILLUSTRATIONS (Continued)

<u>Figure</u>	<u>Title</u>	<u>Page</u>
5-4	Millimeter Wave Back Diode (Pin Construction), Photograph	75
C-1	Fourth Harmonic Generator for Crossed Waveguide Coupling Studies, Photograph	C-2
C-2	Epoxied Diode, Microphotograph	C-3
C-3	Data for Fourth Harmonic Crossed Waveguide Device	C-3
D-1	Diode Equivalent Circuit	D-2
D-2	Geometry of Point-Contact Junction	D-4
E-1	Current-Voltage Characteristics of the Point-Contact, Tunnel, and Backward Diodes	E-2
E-2	Typical Measured I-V Curve for Tunnel Diode	E-3
F-1	Measured Second Harmonic Mixing Conversion Loss vs Local Oscillator Power, 141.65 to 1.65 GHz	F-2
G-1	Single Source Test System, Block Diagram	G-2
H-1	Dielectric Constant and Loss Tangent vs Temperature for a Ferroelectric Ceramic	H-2
H-2	Dielectric Constant and Loss Tangent vs DC Voltage for a Ferroelectric Ceramic	H-3
H-3	Typical Dielectric Constant and Loss Tangent Variations vs Frequency for Ferroelectrics	H-5
H-4	Experimental Strontium Titanate Harmonic Generator	H-10
H-5	Strontium Titanate Harmonic Generator, Photograph	H-13
I-1	Commercial Varactor and Harmonic Generator Mount, Photograph	I-2

I. SUBMILLIMETER SATELLITE RADIOMETER

A. General Considerations

1. Objectives

The submillimeter wavelength portion of the electromagnetic spectrum has been used for few experiments and fewer systems applications because of a lack of suitable components. Radio astronomers have investigated radiation from planets, the sun, and other extra-terrestrial sources at both the longer millimeter wavelengths and the shorter infrared wavelengths by means of ground-based and space-borne radiometers. The primary goal of the work described herein has been to develop components that will permit practical experiments in astronomical radiometry in the unexplored region of submillimeter wavelengths. Only after such experiments have been made can the data be interpreted and the contribution of such experiments to a better understanding of the universe be evaluated. At present, the lack of submillimeter radiometer experimental programs results in a gap in information for a sizable portion of the electromagnetic spectrum.

Another objective of the development of submillimeter wave receiver components has been to provide devices for space communications systems feasibility studies. The tradeoffs in future space communications systems are speculative now. The antenna directivity achievable with structures of moderate size may well make submillimeter wave systems attractive in years to come. In addition to the use of the developed components for communications receivers, the submillimeter wave harmonic generators may ultimately evolve to the stage of being used as communications transmitters. Information could be transmitted with a variety of modulation techniques. Thus, this development has contributed to the variety of design choices available to the systems engineer.

An important by-product of this work is the capability established for laboratory measurements of materials and other components at submillimeter wavelengths. This results, in part, from the necessity of setting up a measurement system in order to evaluate the components being developed. Further improvement of the devices and the design of auxiliary components will be aided by the availability of data on properties of materials at submillimeter wavelengths. At present, information such as the dielectric constant and loss tangent of insulating plastics must be extrapolated from measurements at other wavelengths.

Experimental physicists have made some measurements at submillimeter wavelengths by using 1) thermal radiators as sources, 2) optical-type filters arranged to have relatively wide pass bands, and 3) direct detectors of the thermal or quantum type. The work reported here is Phase I of a program to develop components for superheterodyne receivers in order to achieve good frequency resolution and high sensitivity with large dynamic range. The components will be designed to withstand the environmental conditions anticipated for radio astronomical and communications systems in space.

2. Submillimeter Wave Propagation in the Atmosphere

The attenuation of the atmosphere is a major consideration in the application of submillimeter wave systems. Recently Chang and Lester reported⁽¹⁻¹⁾ an average of measured values at 590 GHz of 34 dB per kilometer per gram of water vapor content per cubic meter of air. They used a superheterodyne radiometer⁽¹⁻²⁾ designed and constructed for them at this laboratory and made measurements outdoors for ranges from 3 to 67 meters. The following table illustrates the very high propagation losses for various terrestrial humidity conditions:

Air Temperature °C	Relative Humidity %	Water Vapor Content g/m ³	Propagation Loss Based upon 34 dB/km/g H ₂ O/m ³ dB/km
0	10	0.5	17
0	100	5.	170
20	10	1.7	58
20	60	10.	340
20	100	17.	580

Table 1-1 Propagation Loss in the Atmosphere at 590 GHz

Most of the absorption in the atmosphere at submillimeter wavelengths is believed to be due to water vapor although other gases may contribute. The interpretation of the experimentally measured attenuation data in terms of the conditions present in a given radio-astronomical experiment is difficult because of variation in water vapor content over the propagation path through the atmosphere. However, Rosenblum, in a 1961 review⁽¹⁻³⁾ of published data, showed that total vertical attenuation for one-way transmission through the atmosphere is greater than attenuation per kilometer for horizontal propagation at

all frequencies from 10 to 400 GHz. Attenuation at angles appreciably off vertical is much greater. The data in Table 1-1 are based upon propagation loss measurements at 590 GHz. Strong and Vanasse⁽¹⁻⁴⁾ have shown with relative measurements that the loss increases to a peak at about 550 GHz, decreases to a minimum of undetermined value at 660 GHz, and then increases to a peak at 750 GHz. From these facts and the data in the preceding table one may deduce that the earth-bound radio astronomer or the designer of earth-to-space communications systems must contend with tens of decibels of attenuation at sub-millimeter wavelengths.

As will be discussed in Subsection D, the anticipated receiver sensitivity at 0.5 mm wavelengths expressed as a temperature contrast is 7°K for a signal-to-noise ratio of unity and an integration time of 10 seconds. The extra-terrestrial source producing the strongest 600 GHz signal at the earth's surface is the sun. If it is assumed that it has the same brightness temperature at 600 GHz as Coates reported⁽¹⁻⁵⁾ measuring at 70 GHz, namely $7000^{\circ} \pm 700^{\circ}\text{K}$, it is apparent that a state-of-the-art radiometer (with antenna beam completely subtended by the sun) could not see the sun if the total vertical attenuation exceeded 30 dB. The preceding table shows that the propagation loss far exceeds this value except during unusually favorable weather conditions. Detection of much weaker radiation at submillimeter wavelengths from planets or other heavenly bodies, some of which must be observed at angles near the horizon, is out of the question with earthbound instruments.

Although improved receivers will ultimately be built, it will be shown in Appendix A that the theoretical minimum detectable temperature contrast for an ideal receiver (with $B_{\text{IF}} = 1.6 \text{ GHz}$, $B_{\text{O}} = 0.1 \text{ Hz}$) is 0.002°K which is just 34.7 dB better than the 7°K quoted above. Further improvements could be obtained by increasing the input bandwidth and by decreasing the output bandwidth. At present the biggest gain in submillimeter radiometer detection capability which appears feasible is to eliminate the atmospheric attenuation by, for example, placing the radiometer on a space platform. A submillimeter radiometer in space above the earth's atmosphere will permit radio astronomical observations which are unattainable otherwise for this portion of the spectrum.

Similarly, it appears that submillimeter wave communications systems will be restricted to situations where the propagation path does not pass through a region of appreciable water vapor content. Communications systems which would utilize a coherent oscillator are further restricted because present submillimeter wave signal

sources have quite limited output power. Furthermore, it may not be desirable to improve receiver sensitivity by decreasing output bandwidth as was suggested for the radiometer because this would reduce the information handling capacity.

3. Requirements for Space-Borne Radiometer

In order to achieve the primary objective of submillimeter radio astronomy by utilizing a space platform for a radiometer receiver, certain requirements are placed on the system and components. The following are the general requirements to be met:

- 1) Over-all system weight must be minimized;
- 2) Over-all system volume must be minimized;
- 3) Total system primary power consumption must be minimized;
- 4) The system must meet standards of reliability;
- 5) The system must have a useful operating lifetime;
- 6) The system must withstand the conditions of a vehicle launch environment;
- 7) The system must withstand the conditions of the space environment;
- 8) System performance must be sufficient to permit a useful experiment.

These requirements have served as guides in the development of the radiometer components described in this report.

B. Superheterodyne Radiometer Receiver

Various methods may be used to detect submillimeter radiation. Direct detection may be accomplished with a video crystal diode detector, with a photoconductive detector, or with a bolometric thermal detector. Predetection amplification is impossible because no submillimeter wavelength amplifiers are available. Superheterodyne detection has been found to give high sensitivity when used in millimeter wave⁽¹⁻⁶⁾ and submillimeter wave⁽¹⁻²⁾ Dicke radiometer receivers and is the method chosen for development on this program. A comparison among submillimeter detectors is presented in Appendix A.

Figure 1-1 is a block diagram of a possible submillimeter Dicke radiometer designed to meet the requirements of space operation. This radiometer differs from the submillimeter radiometer previously built

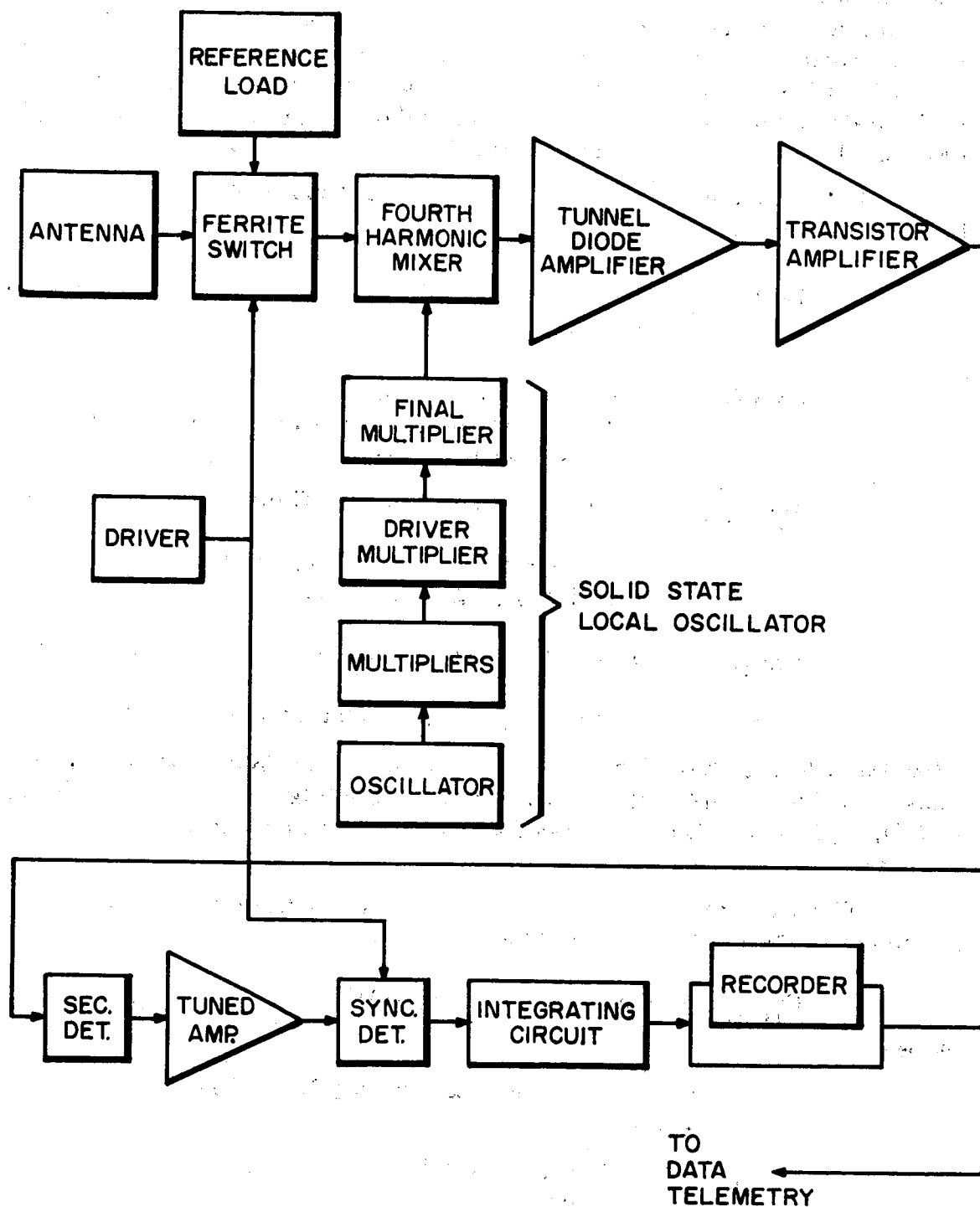


FIG. 1-1 - Satellite Submillimeter Radiometer
(Possible Block Diagram)

and operated at 590 GHz in a number of significant respects. Weight, volume, and primary power consumption are minimized by using a ferrite switch instead of a mechanical chopper, by using solid state IF amplifiers instead of traveling-wave tube amplifiers, and by using a solid state multiplier chain as local oscillator instead of an electron tube oscillator and associated high voltage power supply.

The sensitivity of a superheterodyne Dicke radiometer is given by:

$$(\Delta T)_{\min} = \frac{F_{\text{RAD}} \cdot T_o}{\sqrt{B_{\text{IF}} \cdot \tau}} \quad (1-1)$$

where

$(\Delta T)_{\min}$ is the rms temperature change in degrees Kelvin which produces an output signal-to-noise ratio of unity;

F_{RAD} is the radiometer noise figure (two sidebands);

T_o is 290°K and is the reference temperature at which F_{RAD} is defined;

B_{IF} is the IF bandwidth in Hz;

τ is the output time constant in seconds.

Equation 1-1 may be further refined by the inclusion of various constants whose values depend upon circuit details but whose combined effect on $(\Delta T)_{\min}$ is negligible compared to that of the other factors for typical values.

The radiometer noise figure, F_{RAD} , depends upon receiver component properties as follows:

$$F_{\text{RAD}} = 0.5 L_{\text{RF}} L_c (t + F_{\text{IF}} - 1) \quad (1-2)$$

where

0.5 is the factor used when both RF sidebands are passed;

L_{RF} is the loss ratio of all RF components ahead of the mixer;

L_c is the mixer conversion loss ratio;

t is the mixer diode noise ratio;

F_{IF} is the IF amplifier noise figure.

Equations 1-1 and 1-2 are used to evaluate receiver components in terms of system performance and to indicate where development efforts must be concentrated.

C. Key Components Selected for Development

It was recognized at the beginning of this program that the feasibility of a radiometer such as is shown in Figure 1-1 depended upon the development of a solid state local oscillator (SSLO) and a mixer which could operate together to provide useful radiometer sensitivity. The performance required of these two devices was beyond that which had been demonstrated up to that time. With the exception of the ferrite switch, the other components were realizable within the existing state-of-the-art. The development of a ferrite switch was deferred to a later phase, however, because: 1) a mechanical chopper might be used in its place, 2) it would not be needed unless the mixer and SSLO could be developed and 3) it would require a submillimeter setup for its development.

The choices of design goals for the mixer and for the solid state local oscillator were interdependent. The output power which a SSLO can be designed to generate decreases as the output frequency is increased. Moreover, the SSLO was expected to generate less than the optimum drive power for a mixer. The SSLO frequency determines the order of harmonic mixing necessary to receive submillimeter wavelengths. The mixer conversion loss will increase as the SSLO power decreases but will be least for the lowest order of mixing (highest SSLO frequency). Although fundamental mixing would have the lowest conversion loss, it was clearly unrealistic to expect to develop a SSLO at submillimeter wavelengths. Second harmonic mixing was known to be feasible but, again, it was believed doubtful that a 1 mm SSLO source could be developed. Therefore, it was decided to try to extend the harmonic mixing technique to fourth harmonic mixing and to extend harmonic generation techniques to produce a usable SSLO at 2 mm wavelengths. The required advances in the state-of-the-art were thus shared between the two components.

The development of the fourth harmonic mixer and the development of the final multiplier stage in the solid state local oscillator are described in Sections II and III which follow. Experimental models of these devices were successful in reaching performance levels which prove the feasibility of an all solid-state 0.5 mm radiometer receiver.

It may be that other possible tradeoffs between SSLO output power and frequency and corresponding mixer driving power and harmonic order would result in better performance, but it was beyond the scope of this program to investigate all possible combinations.

D. Estimated Radiometer Sensitivity

Equation 1-1 has been used to calculate a minimum detectable temperature of 7°K for a possible submillimeter radiometer. A measured value of 29.5 dB for mixer conversion loss as reported in Section II was the basis of the calculation. The following assumptions were made: RF losses of 3 dB, mixer diode noise ratio of 2, IF noise figure of 4 dB, IF bandwidth of 1.6 GHz, and output integration time of 10 seconds. The 3 dB of RF losses assumed here is a nominal value to allow for the insertion loss of a lens antenna, feed horn sidelobe spillover, and ferrite circulator insertion loss. The input waveguide attenuation losses are included in the measured value of conversion loss by reason of the method of measurement. The mixer conversion loss was measured at 560 GHz with an LO drive power of 10 mW. It will be shown in Section III that this level of output was generated with a multiplier which would be the output stage of a SSLO.

The value of $(\Delta T)_{\min}$ is based upon a chain of calculations and the best measured value of conversion loss to date. The evidence is that minimum detectable temperatures under 10°K can be achieved with fourth harmonic mixing at wavelengths of about 0.5 millimeter. This value may be compared to measurements⁽¹⁻²⁾ made at 600 GHz with a superheterodyne radiometer having a second harmonic mixer and a backward wave oscillator tube as LO. The minimum temperature contrast between a thermal source of extended area and the ambient atmospheric background which that radiometer could detect was 26°K . The signal-to-noise ratio during the measurement could not be determined precisely but it was in the range 2.5 to 8.3. Therefore, the $(\Delta T)_{\min}$ for $S/N = 1$ was between 3.1° and 10.4°K . Thus the value of 7°K calculated from a fourth harmonic mixer conversion loss of 29.5 dB as measured on this project is a credible value. Measured values of mixer conversion losses cannot be compared because it was not possible to measure these during the earlier project.

Let us consider the performance of a submillimeter radiometer in space having a sensitivity, $(\Delta T)_{\min}$, of 7°K . If the sun is assumed to have a brightness temperature of 7000°K at 0.5 mm as has been measured at 4.3 mm⁽¹⁻⁵⁾ and is assumed to fill the antenna beam of a radiometer in space, then signal-to-noise ratios of 30 dB should be attainable. If

Mars or the Earth's moon is the object of interest, a brightness temperature of the order of magnitude of 200°K may be assumed (in agreement with Mayer's discussion⁽¹⁻⁷⁾ of recent radiometric measurements at other frequencies). Signal-to-noise ratios of 14 dB might then be achieved at 0.5 mm wavelengths with a radiometer in a space probe close enough to the body to fill the radiometer antenna beam.

The sensitivities quoted will not be achieved in practice because no practical antenna has a main beam efficiency of 100%, i.e., some energy is received through sidelobes. Also, in some cases, the antenna may not be large enough or the space platform close enough to the body for the body to subtend the entire solid angle of the main beam. However, a more probable condition is exemplified by the antenna of the 600 GHz radiometer mentioned earlier.⁽¹⁻²⁾ It is a lens antenna which has a 10-inch aperture and a calculated half-power beamwidth of one-seventh degree or 2.5 milliradians and a tenth-power beamwidth of 3.5 milliradians. The latter angle would be subtended by Mars at a range of 1,200,000 miles. In conclusion, one would expect that at submillimeter wavelengths the object of interest would fill the greater part of the radiometer receiving antenna beam pattern.

II. SUBMILLIMETER HARMONIC MIXER

A. Design and Construction of the Diode Mount

A crossed waveguide device was constructed for use as a 0.5 millimeter wavelength harmonic mixer. Figure 2-1 is a drawing which shows the principal features of the mixer structure, and Figure 2-2 is a photograph of the completed mixer. The signal waveguide is $0.014'' \times 0.007''$ in cross section, which corresponds to a calculated cutoff frequency of 420 GHz for the TE_{10} mode and 840 GHz for the TE_{20} mode. The local oscillator waveguide is $0.080'' \times 0.040''$, the equivalent of RG-138/U, and has a calculated cutoff frequency of 74 GHz for the TE_{10} mode or 148 GHz for the TE_{20} mode. Therefore, the device can be operated as a fourth harmonic mixer at about 420 to 592 GHz with first order mode propagation of LO and signal or as a third harmonic mixer at about 420 to 444 GHz. Fifth harmonic mixing at 420 to 740 GHz is also possible. Mixing at higher frequencies for each harmonic is possible if the LO input is operated above the TE_{20} mode cutoff.

A new method was developed for making waveguides of very small size such as was required. The finished waveguide has smooth surfaces, square corners, and uniform cross section. The method consists of molding a polystyrene mandrel, electroforming a waveguide about the mandrel, and then dissolving the mandrel. The inside surfaces of the completed waveguide are as smooth as the original mold. The mold is made of four metal pieces whose molding surfaces have been highly polished. These are clamped together and heated; then polystyrene which has been heated to a plastic state is forced into the cavity under pressure. Upon cooling, the finished mandrel is removed from the mold. A conductive coating of gold is applied by vacuum evaporation. Then the waveguide is built up by electroforming a one or two mil thickness of gold followed by copper to any desired thickness.

This mixer has a tapered section connecting to the signal waveguide which serves either as a receiving antenna horn or as a transition from a larger connecting waveguide. The tapered section and the $0.014'' \times 0.007''$ signal waveguide were made as one piece by mounting a tapered mandrel and the partially formed signal waveguide together in a jig and electroforming over both. The mandrels and electroform in various stages of manufacture are shown in Figures 2-3 to 2-6.

The above procedure was developed for making small waveguides in order to minimize waveguide attenuation due to surface roughness. Gold was used for the inner surfaces in order to prevent corrosion which would roughen the surfaces. Special attention to minimizing

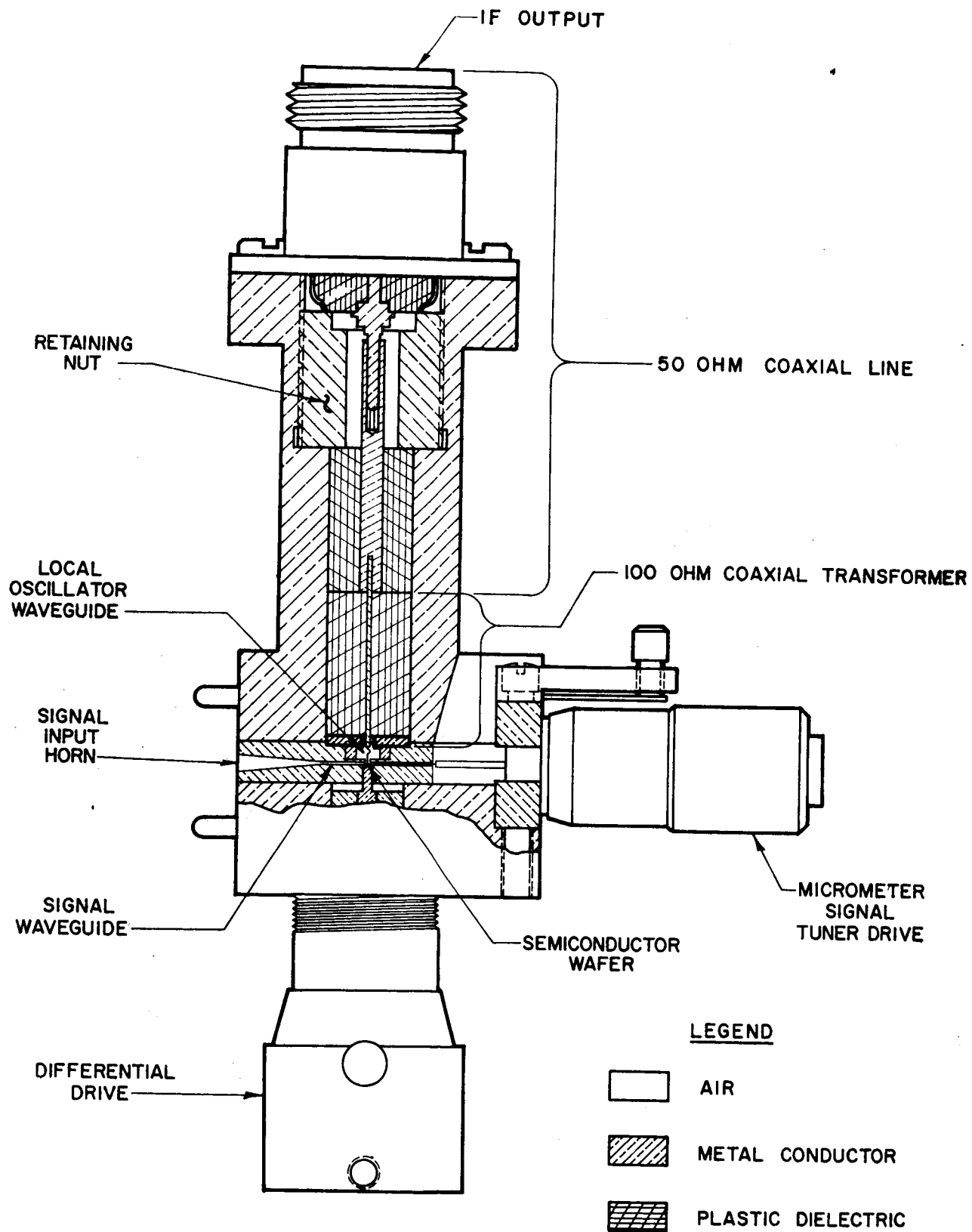


FIG. 2-1 - Submillimeter Harmonic Mixer

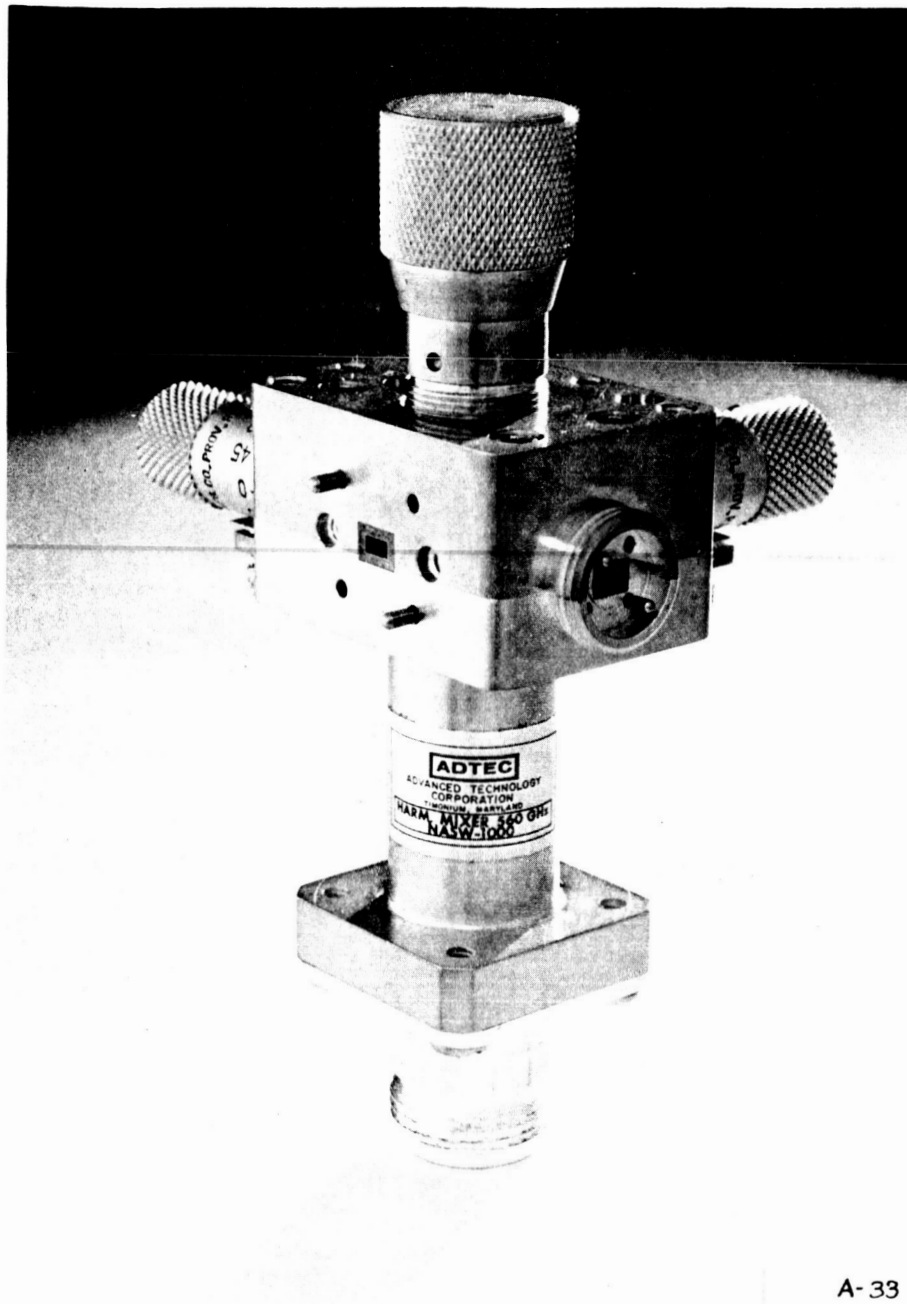


FIG. 2-2 - Submillimeter Harmonic Mixer

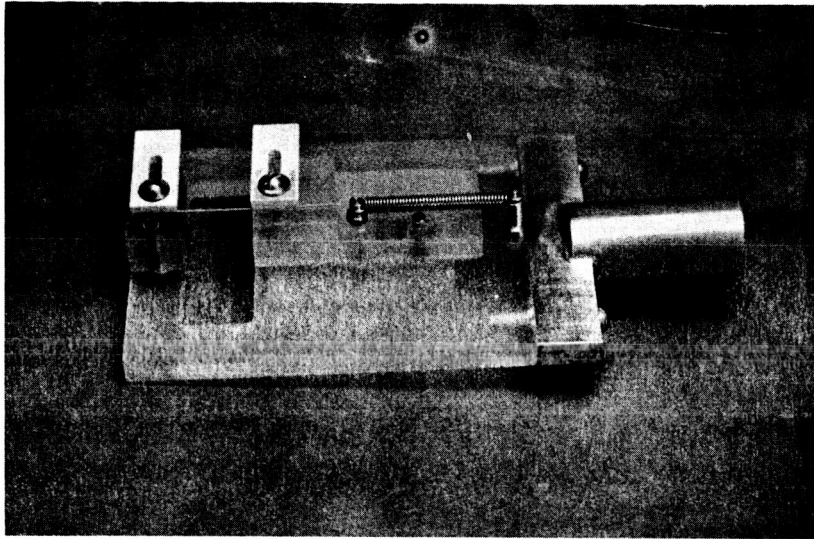


FIG. 2-3 - Waveguide Mandrel 0.014" x 0.007" in Jig

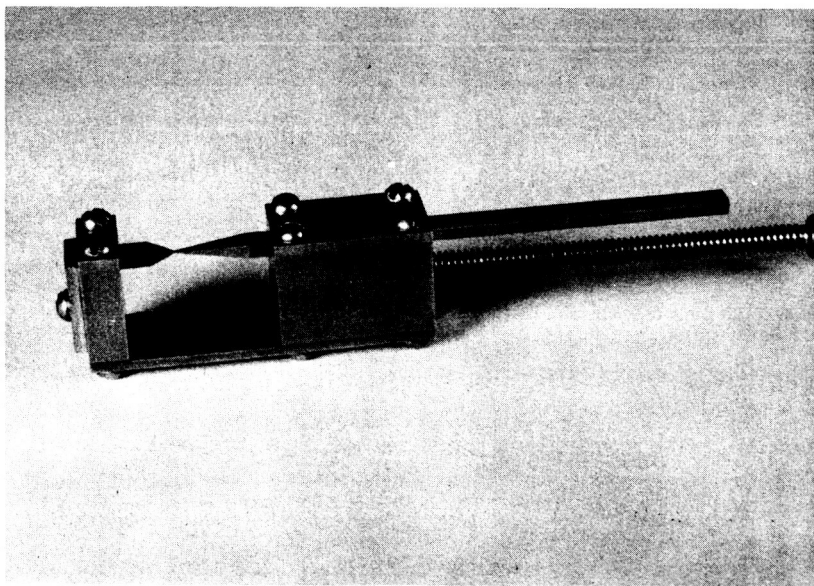


FIG. 2-4 - Partial Electroform and Taper Mandrel in Jig



FIG. 2-5 - Waveguide and Taper After Electroforming



FIG. 2-6 - Waveguide and Taper Machined for Mixer

attenuation is necessary at submillimeter wavelengths because the attenuation is relatively high even for ideal structures. The attenuation due to the finite conductivity of a perfectly smooth gold waveguide is 1.8 dB per inch at 560 GHz. Harvey⁽²⁻¹⁾ states that a surface roughness of 50% of the skin depth will increase the attenuation about 20%. The skin depth in gold at 560 GHz is about 4 micro-inches.

In addition to achieving smooth surfaces, it is possible to minimize attenuation by designing the shortest possible waveguide. In this mixer the distance from the throat of the horn to the edge of the crystal wafer is 0.148". This is about five guide wavelengths at 560 GHz; the waveguide should have less than 0.5 dB attenuation at the signal frequency. A minimum length of signal waveguide is required as a high pass filter because the input to a harmonic mixer should not pass energy at frequencies near the fundamental LO frequency or near harmonics lower than the desired harmonic, in this case the fourth. The choice of a 0.148" length was arbitrary since design criteria are not clear cut. This problem is discussed in Appendix B.

The larger LO waveguide was electroformed about a polystyrene mandrel which had been made in a machined mold. Surface roughness is not as critical at the LO frequencies, but low waveguide losses are still highly desirable. Furthermore, the use of a waveguide electroformed to precise dimensions allowed a close-fitting tuning short to be installed. This minimized leakage and dissipative losses in the tuner.

The LO energy is coupled into the semiconductor junction in the signal waveguide by the whisker wire through a hole 0.008" in diameter in the wall common to the two waveguides. The wall is 0.008" thick. In effect, the two waveguides are coupled by a coaxial line whose length is 0.065 wavelength at 140 GHz and 0.26 wavelength at 560 GHz. The wall thickness does not appear to be critical although it may affect the conversion loss by a few decibels. As will be described in the next subsection, the mixer was operated successfully at several frequencies. An experimental study was made at a lower frequency of the performance of a fourth harmonic crossed waveguide device as a function of frequency. The study is described in Appendix C. Some anomalous data were obtained so that no positive guide to the choice of common wall thickness can be given. A dimension of 0.008" was chosen for the submillimeter mixer because it is about the minimum wall thickness that is practical to machine and there is some evidence, albeit not unequivocal, that a quarter wavelength at the signal frequency is desirable.

A special feature of this mixer mount is the method of supporting the whisker carrier and the construction of the coaxial IF output. The mount has been designed to prevent whisker motion due to attachment or flexing of the connecting cable. This has been a frequent source of diode instability in past mounts. The key features of the special construction may be seen in Figure 2-1. The 50-ohm coaxial line center conductor was made a snug fit in the dielectric which was held in the mixer body by the retaining unit. The connection to the output connector (a type N panel-mounted fitting) was a sliding fit. The IF output was built with a 50-ohm characteristic impedance and a 100-ohm quarter-wave transformer to improve the impedance match between the diode and the IF port. Measured data on the IF impedance at 1 to 4 GHz are included in the next subsection.

The rest of the construction is typical of the diode mounts that have been made for several years at this laboratory. The waveguides are mounted in a split brass block. Two ports are fitted to mate with waveguide flanges. The block also holds the tuner drives which are micrometer drives for convenience in measuring tuning short positions and hence guide wavelengths. A differential tuner in the signal waveguide would permit easier tuning in this experimental mixer. The ultimate operational mixer for a satellite radiometer may well be fixed tuned. A desirable modification to the crystal differential drive would be a different method of locking its position once a diode junction is made. The present set screw usually cannot be set without disturbing the diode junction.

B. The Mixer Diode

The submillimeter mixer was designed to accommodate a semiconductor wafer 0.008" in diameter mounted on a metal post with an ohmic contact and positioned in the broad wall of the signal waveguide by a differential drive. The semiconductor wafer was polished to a thickness of about 0.004". A phosphor bronze whisker wire of 0.001" diameter which had been electrolytically pointed was brought into contact with the semiconductor wafer.

Since monocrystalline n-gallium arsenide was used previously⁽¹⁻²⁾ with success in a 600 GHz second harmonic mixer, the same type of semiconductor was used again in the fourth harmonic mixer for this project. Wafers were prepared from a different ingot (Code No. K-3) for this project since the supply (Code No. M-4) used previously was exhausted. The properties of the two different ingots are as follows:

<u>Code No.</u>	<u>Dopant</u>	<u>Carrier Concentration (atoms/cm³)</u>	<u>Mobility (cm²/V-s)</u>	<u>Resistivity (ohm-cm)</u>
M-4, prior use	Te	18×10^{17}	2800	1.3×10^{-3}
K-3, present use	None	1.14×10^{17}	4100	13.4×10^{-3}

The K-3 material was chosen because it has been used at this laboratory in the construction of a variety of millimeter wave mixers and harmonic generators which have had good performance characteristics. For example, it was used in the best 140 GHz second harmonic generator tested on Contract NASw-662.⁽²⁻²⁾

Prior to the construction and test of the submillimeter mixer there was no certain knowledge of the LO requirements and the mixer conversion loss that might be achieved. The diode properties contributing to mixer conversion loss were studied and are discussed in Appendix D. That analysis gave some insights into what might be tried to minimize the conversion losses. The general goal was to reduce resistance to RF current flow in the paths to the active region of the diode which is the junction of whisker tip in a small area of one surface of the crystal wafer. A variety of schemes for reducing the path length for RF currents in or over the semiconductor wafer to the ohmic contact were considered. Other semiconductor materials, including epitaxial types which might have better rectifying properties, were looked for as well as materials forming distinctly different diode types such as backward diodes and tunnel diodes.

Some experiments were performed at millimeter wavelengths to seek better diodes. It was possible to do this with existing diode mounts while the submillimeter mixer was being constructed. It was also desirable to experiment with materials at lower frequencies as much as possible where the operation of the RF source was less expensive and the size of the parts permitted more rapid modifications. The fabrication and test of tunnel diodes in fundamental mixer mounts and backward diodes in second harmonic mixer mounts is reported in Appendix E. In addition to experiments on this project, work on other projects was reviewed for possible techniques of improving submillimeter mixer performance. So far, the best mixer performance on this project and others at this laboratory has been with diodes formed with n-GaAs and phosphor bronze whiskers. Test data measured with diodes of this type in a M-9838 mount operated as a 140 GHz second harmonic mixer are presented in Appendix F. There remains considerable experimenting

to be done, however, before the possibilities of further mixer improvement are exhausted. It is expected that future work will be directed to developing rugged diode structures with the same or lower conversion loss.

C. Test Procedures

Several different examples of mixing have been demonstrated with this mixer. These include fourth harmonic mixing at 525 and 560 GHz, third harmonic mixing at 420 GHz, and fifth harmonic mixing at 700 GHz. The submillimeter mixer was first successfully operated in the single conversion test setup of Figure 2-7. A signal at 525 GHz was detected with the local oscillator set at 130.9 GHz; this implies that fourth harmonic mixing took place. The signal was the seventh harmonic from a harmonic generator driven at 75 GHz. The harmonic generator is described in Section IV. The test setup of Figure 2-7 was later modified to a dual conversion system as shown in Figure 2-8 with which signals at 420, 560, and 700 GHz were detected. Finally, a combination of the features of both test setups was used.

These test setups required two sources of millimeter wave energy. A 2 mm wavelength backward wave oscillator was used for one, and a 2 or 4 mm wavelength klystron was used for the other. These tubes were expensive to operate because of a high initial cost and a limited lifetime. For reasons of economy a potential system using one oscillator was studied in some detail. It is described in Appendix G. It was not used because it would have required the development of a 2 mm wavelength single sideband modulator, a difficult and expensive task in itself.

In the test setup of Figure 2-7 the desired pair of signal and local oscillator harmonic frequencies were mixed and detected because the fundamental frequencies have common harmonics only at 525 GHz or at multiples of 525 GHz. RF losses at 1050 GHz or higher multiples would be so high that it was assumed the mixer output was due to 525 GHz. The beat frequencies between other signal and LO harmonics were 17 GHz or higher and would not pass the IF amplifiers. The signal and LO fundamental frequencies were set to an accuracy better than ± 0.5 GHz so that harmonics other than the seventh and fourth, respectively, did not create a beat in the 1 to 2 GHz IF passband.

The dual conversion system of Figure 2-8 was assembled in order that a higher drive frequency for the harmonic generator might be used. With this setup the same order harmonics of both signal oscillator and local oscillator were detected. The different orders were distinguished by the second IF amplifier which had a narrow passband (after Cotton⁽²⁻³⁾).

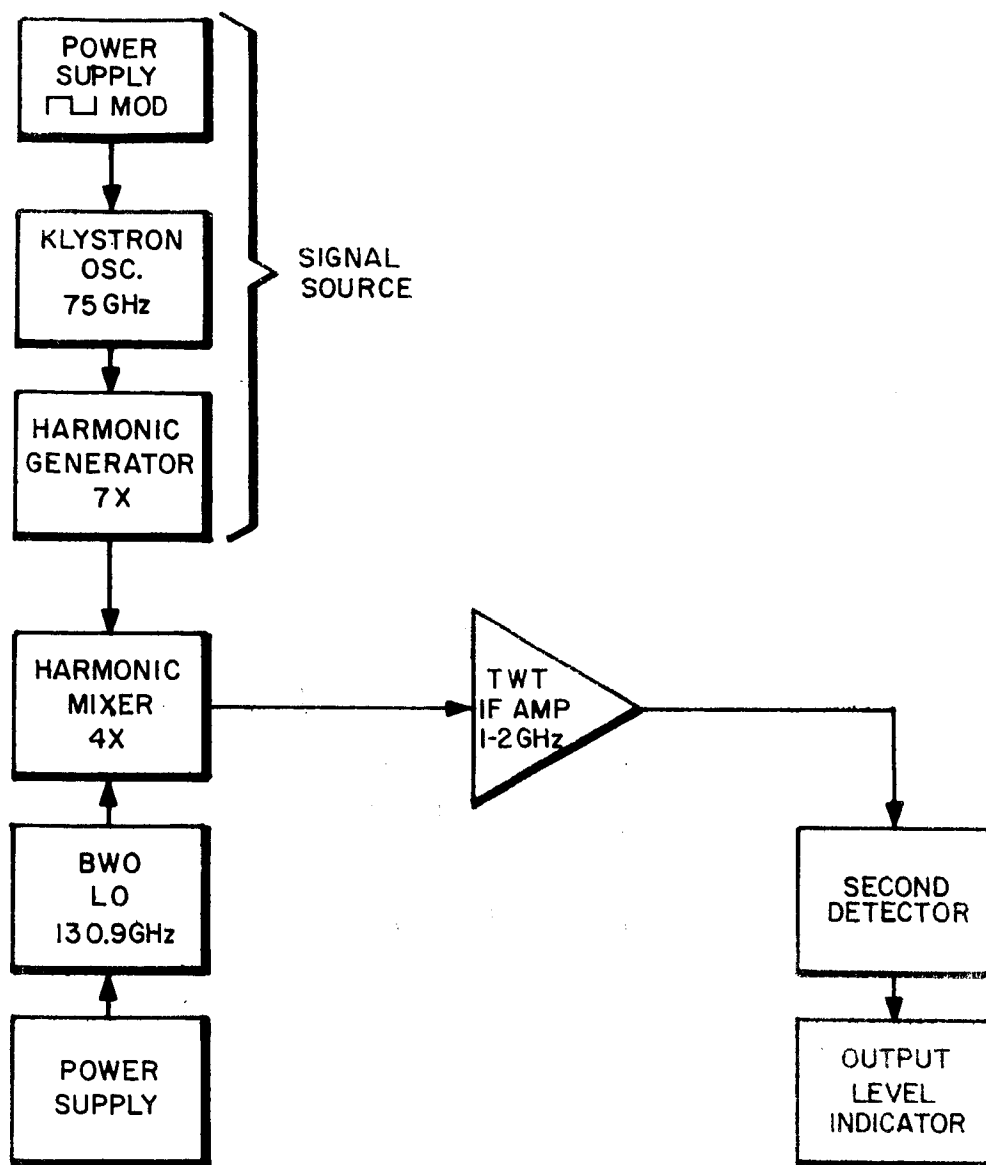


FIG. 2-7 - Single Conversion Harmonic Mixer Test Setup,
525 to 1.4 GHz

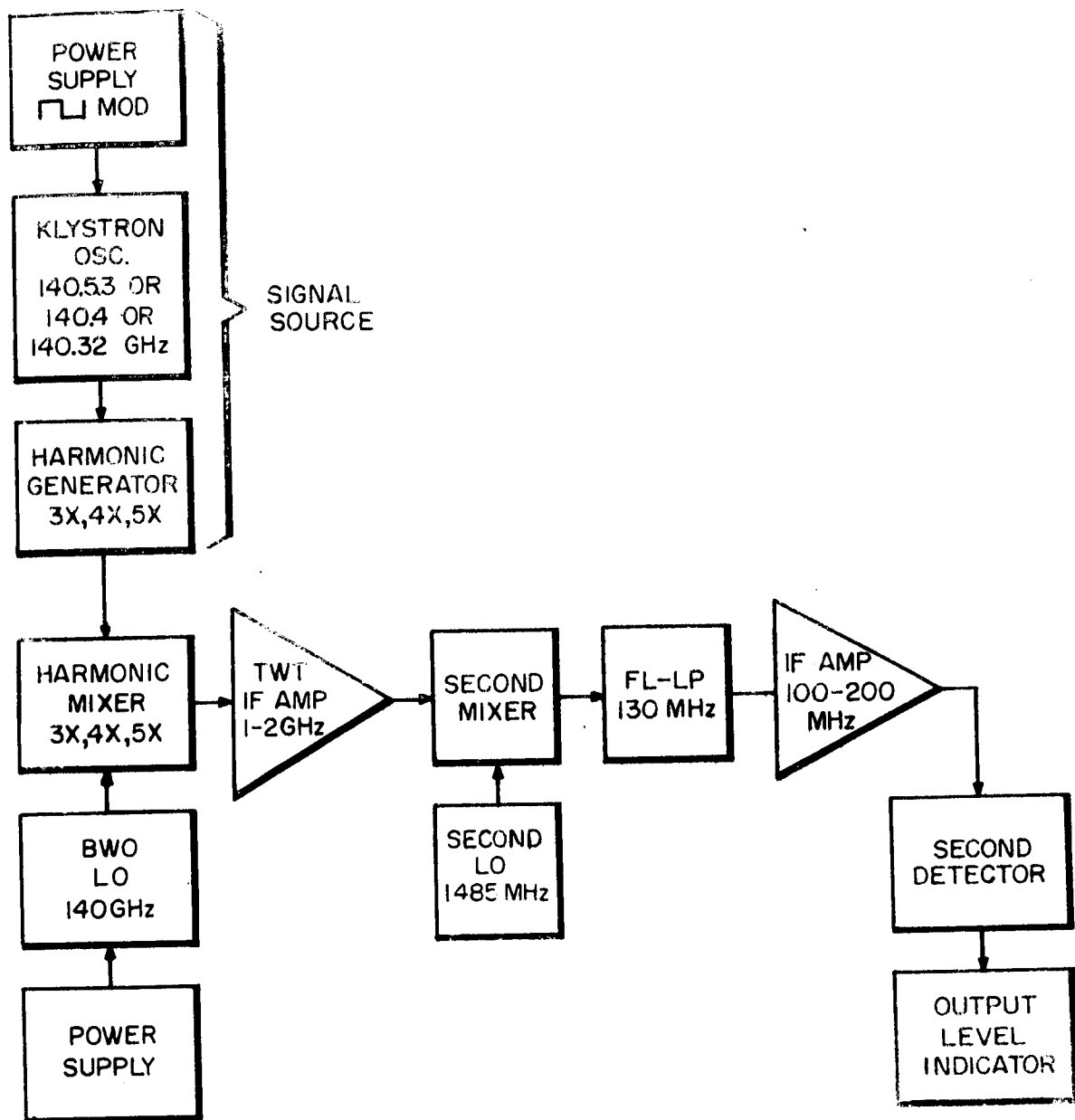


FIG. 2-8 - Dual Conversion Harmonic Mixer Test Setup
420, 560, or 700 GHz to 115 MHz

For example, the submillimeter mixer LO was set at 140 GHz, and the source driving the harmonic generator was set at 140.4 GHz. The IF signal due to the beat of the fourth harmonics of these signals was 1600 MHz and was detected by a second mixer whose local oscillator was set at 1485 MHz. The second IF bandpass was 115 ± 15 MHz. The beat of the third harmonics was 1200 MHz which produced 285 MHz in the second mixer and was rejected by the IF amplifier. The beat of the fifth harmonics was 2000 MHz which gave 515 MHz in the second mixer and was also rejected by the IF amplifier. Either by adjusting the difference between submillimeter LO and harmonic generator driving source or by tuning the second LO frequency, the beat of some other desired order was detected to the exclusion of the remaining beats.

The harmonic generator (see Section IV) that was used as a signal source in these tests of the submillimeter mixer was designed to have an output waveguide as large as possible yet propagate 560 GHz in only one mode. This was done to achieve minimum waveguide attenuation at 560 GHz. The output waveguide is 0.020" x 0.010" which has a calculated cutoff frequency of 295 GHz. Therefore, lower order harmonics down to 295 GHz propagated also. As a result, measurements of the power at the harmonic generator output port included the contribution of all harmonics above 295 GHz of which, in practice, the lowest harmonic was dominant. Therefore, the signal level at the desired higher order harmonic had to be estimated. In the case of the single conversion test setup (Figure 2-7) the conversion loss at the fifth harmonic was measured, and the increased loss at the seventh harmonic was estimated. In the case of the dual conversion test setup (Figure 2-8) the conversion loss at the third harmonic was measured, and the increased loss at the fourth harmonic was estimated.

Measurements of signal power at the desired harmonic were accomplished in the final months of the project by using a high pass filter especially constructed for the purpose. The filter consisted of a waveguide squeeze section having inside cross-sectional dimensions of 0.012" x 0.006" for a nominal cutoff frequency of 500 GHz. The filter was used in the test setup of Figure 2-9 which is similar to that of Figure 2-7 but shows the oscillator frequencies adjusted for mixing at 560.7 GHz. The RF portion of the setup of Figure 2-9 was also used for some measurements with the dual IF system as shown in Figure 2-8. This latter arrangement with its narrow second IF band was particularly convenient for identifying the intermediate frequency with a wavemeter. It also had about 2 dB better sensitivity. However, the single conversion setup was simpler to operate.

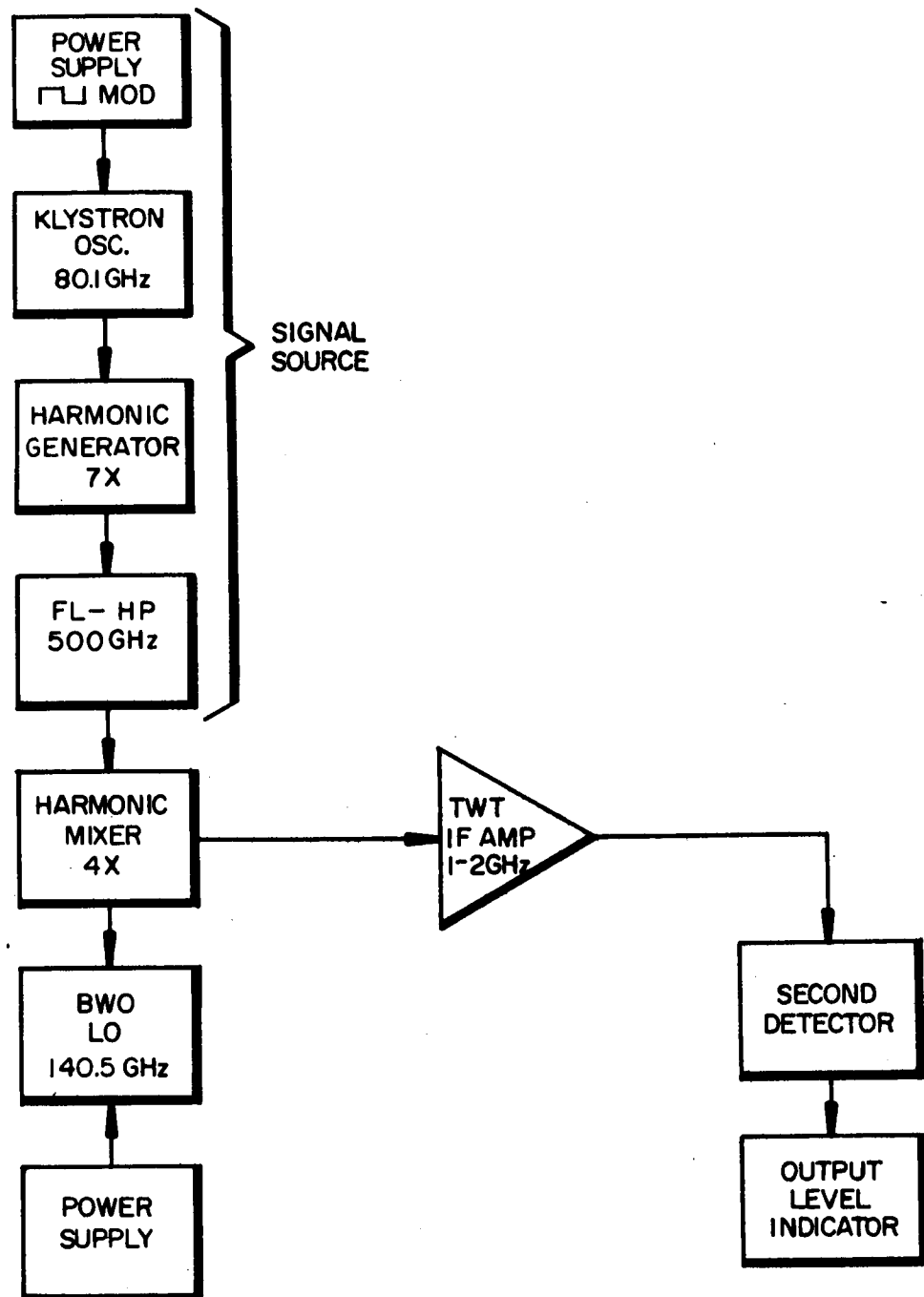


FIG. 2-9 - Single Conversion Harmonic Mixer Test Setup,
560.7 to 1.3 GHz

In all of these test setups the frequency of the signal was verified by measuring the signal guide wavelength with the mixer signal micrometer tuner. There was especially good distinction between 420 and 560 GHz mixing in the dual conversion test setup because 420 GHz is the mixer signal waveguide cutoff frequency and, consequently, has a disproportionately longer wavelength.

The power output of the signal sources in Figure 2-9 was measured with a Wollaston wire bolometer in a mount having the internal dimensions of RG-138/U waveguide (0.080" x 0.040"). A typical power level was -58 dBm. Because the power had to be measured at such a low level, the bolometer was operated as a video detector rather than in a bridge circuit with dc power substitution. The output indicator for the bolometer was a tuned amplifier with a bandpass of 4 Hz. Even so, it was barely possible to make this measurement since the signal-to-noise ratio was only 4 dB.

A precision rotary vane attenuator was used with the bolometer and output indicator to obtain a calibration in terms of absolute power level by measuring the output of a 140 GHz source with a water calorimeter. Two possible sources of error in the calibration and measurement procedures existed which would have caused the measured values of the signal source output to tend to be lower than the true values. During calibration the water calorimeter may have indicated a value lower than the true value because of failure to absorb all of the incident energy. Similarly, during the actual measurement the bolometer may have reflected a portion of the incident energy. A compensating error was introduced by the fact that there was harmonic energy present at the filter output at all harmonics above the filter cutoff frequency. The bolometer responded to the total power whereas the mixer detected only one harmonic, whose power must have been less than the measured value. For example, if the power at the $(n + 1)^{\text{th}}$ harmonic was 5 dB less than the power at the n^{th} harmonic, then the power at the n^{th} harmonic was 1.7 dB less than the total power seen by the bolometer at both harmonics. Additional filters which would have permitted the measurement of power levels of different groupings of harmonics were not available. However, some idea of the power level at different harmonics may be obtained from the fact that the signal source output within the filter was -24 dBm. Since the harmonic generator was being driven at 80.1 GHz and its output waveguide has a cutoff frequency of 295 GHz, the -24 dBm power must have contained fourth and higher order harmonics. With the filter in place, only the seventh and higher harmonics were passed and the power level dropped to -58 dBm or a decrease of 34 dB by eliminating fourth, fifth, and sixth harmonics. Of course some of the 34 dB decrease was due to the filter insertion loss above the cutoff frequency. This was found with

the mixer test setup to be about 2 dB. Therefore, on the average, the power in each successively higher harmonic from fifth to seventh was 10.7 dB less. This is only an approximation to the true value because the level of each successively higher harmonic does not decrease as much as the preceding. In any case, the magnitudes of the errors in calibration and measurement due to reflections are unknown also and it is impossible to determine the net error.

A reflection error was also introduced by the mismatch looking into the mixer signal port. A further error was introduced by the insertion loss of the signal waveguide. In effect, the conversion loss values quoted are for the mixer mount, not just the mixer diode. Hence, they are conservative. However, it would be desirable to investigate their magnitude to determine if design changes to reduce their value would be worthwhile. No means of measuring the input VSWR at the signal frequency is available.

Let us return to the consideration of the procedure used for measuring the mixer conversion loss. The method of measuring the input signal power has just been described. The converted output power at the intermediate frequency was measured by a substitution method. A signal generator was set at the IF and connected to the input to the first IF amplifier in place of the mixer. The output power of the signal generator was calibrated and its output attenuator was set at the level which was required to give the same indication on the mixer test setup output meter as was produced with the mixer, signal source, and LO connected. In this procedure an error was introduced by the impedance mismatch between mixer and first IF amplifier. The mixer VSWR was measured with a slotted line and the mixer conversion loss data were corrected for this mismatch loss.

D. Test Data

The measurement of the submillimeter harmonic mixer conversion loss and associated properties was made with progressively more refined methods during the term of the project. Improved performance was obtained as experience was gained in positioning the diode junction in the mixer mount. Many diode junctions were formed and tested; detailed data were recorded in selected cases only. The data which are presented in this subsection are representative of the various methods of measurement.

1. Mixer Conversion Loss Measurement on 2-28-66 at 560 GHz

Of the diodes measured, one was chosen as best suited for use as the mixer in a satellite submillimeter radiometer. The data plotted in Figures 2-10 to 2-12 were measured with this diode. The conversion loss of 29.5 dB was the lowest value measured for any diode in the submillimeter mixer with a local oscillator drive power of 10 mW. As will be shown in the next section this was the level of power which was generated with a 140 GHz doubler and therefore is the probable level of power available from a solid state local oscillator. The lowest mixer conversion loss (L_{HM}) for 10 mW drive was obtained with an external dc bias of 0.2 V. It may be seen in Figure 2-10 that for 10 mW drive, zero bias resulted in 2 dB more loss and 0.4 V bias also gave more loss. As is typical of these diode mixers, zero bias gives the lowest conversion loss if the optimum level of LO drive is available. For zero bias this diode required 12.6 mW of drive to attain a L_{HM} of 29.5 dB and 20 mW to attain 26.5 dB. The latter value was the lowest obtained for any bias with this diode. Additional LO drive might have resulted in even lower L_{HM} , but the test was not extended because at 20 mW the rectified diode current was 2.8 mA. In the judgment of the engineer testing the mixer, 2.8 mA was about the maximum value which could be allowed without danger of burning out the diode junction.

Figure 2-11 is a plot of the rectified diode current as a function of LO power for zero bias and three positive values of forward bias. The limiting resistance (R_{LIM}) of the external dc circuit was 25 ohms. The curve of current for each bias level increases gradually at the lower values of applied power but rises steeply at the higher drive levels. The ability of the diode junction to dissipate the heat generated by the current in the junction resistance determines the operating current limit.

The curves in Figure 2-10 cross each other in a complicated manner. There are two inflection points each in the curves for 0 V and 0.2 V bias. The data from which Figures 2-10 and 2-11 were plotted was rearranged and plotted in Figure 2-12 to show mixing conversion loss as a function of total diode current. This procedure resulted in curves for different bias levels which do not cross. It is probable that the point on the 0.4 V curve for 1.3 mA current is an erroneous reading and that the 0.4 V and 0.2 V curves do not meet. It is evident from Figure 2-10 that at some particular LO drive levels bias voltage was necessary to obtain minimum conversion loss. However, Figure 2-12 shows that at any value of diode current, zero bias gives the best conversion loss. Therefore, the decision to use external bias depends upon the amount of available LO drive power and upon the total current that the diode can withstand.

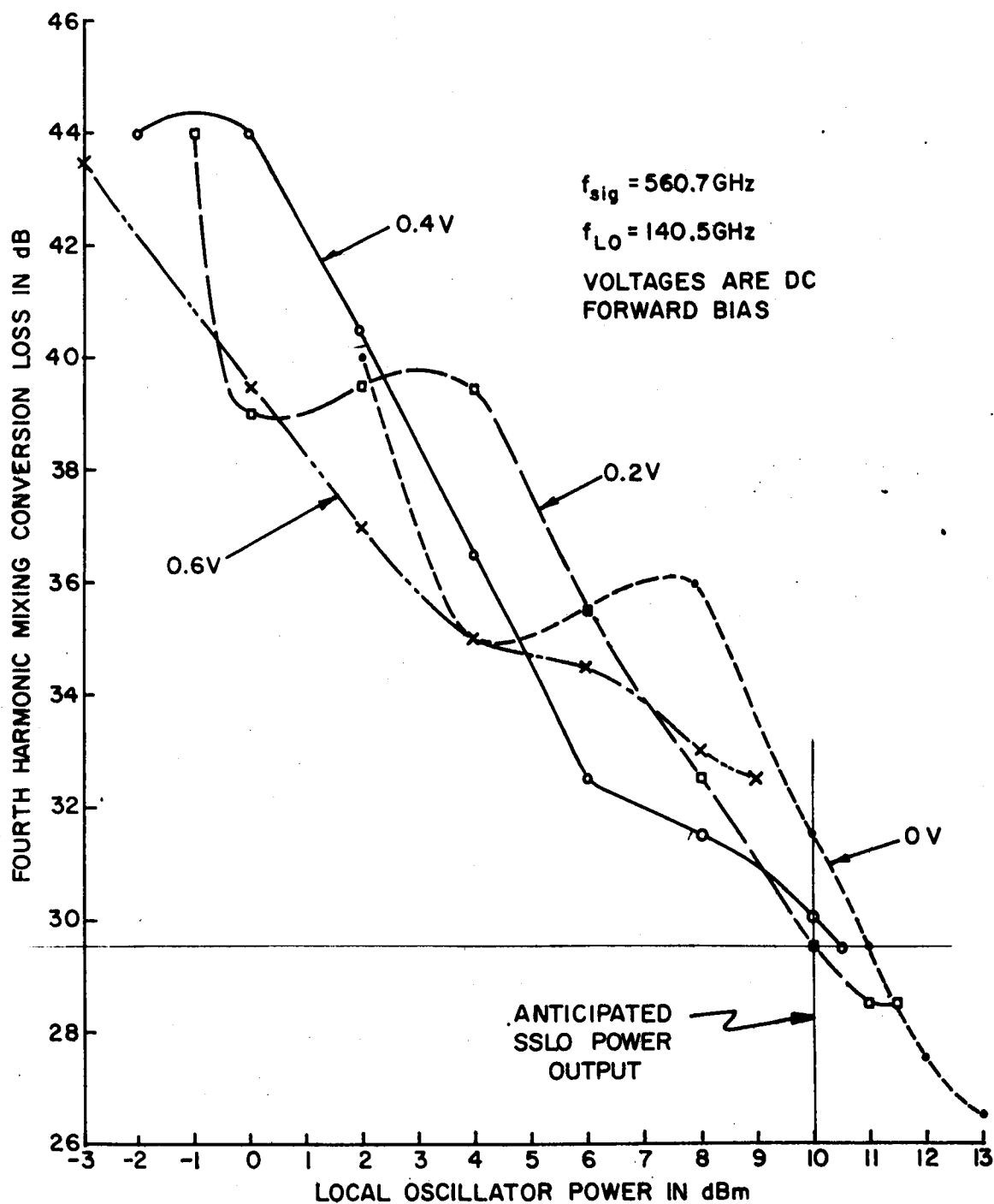


FIG. 2-10 - Measured Fourth Harmonic Mixing Conversion Loss
 vs Local Oscillator Power, Diode of 2-28-66

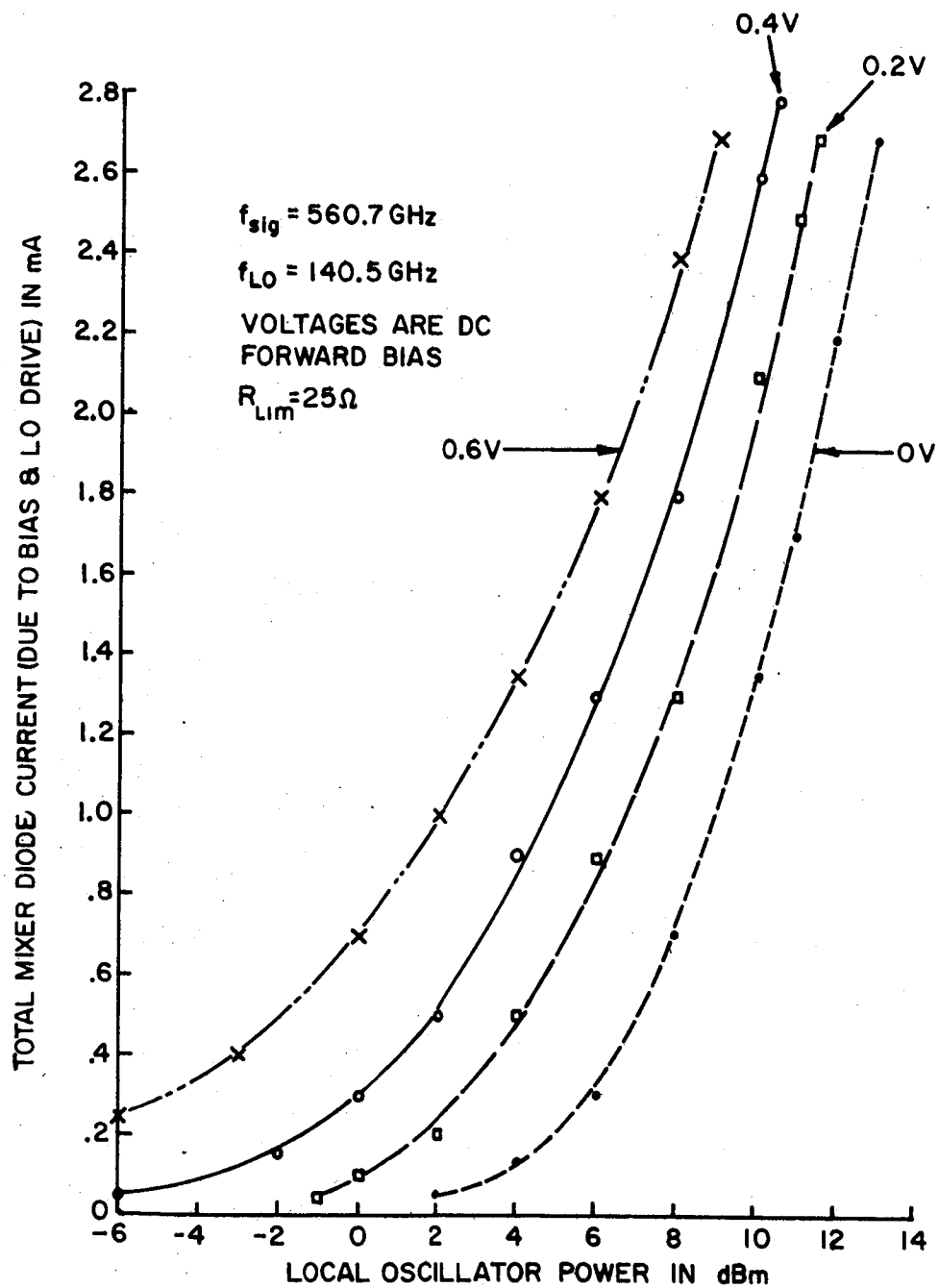


FIG. 2-11 - Direct Current in Fourth Harmonic Mixer,
Diode of 2-28-66

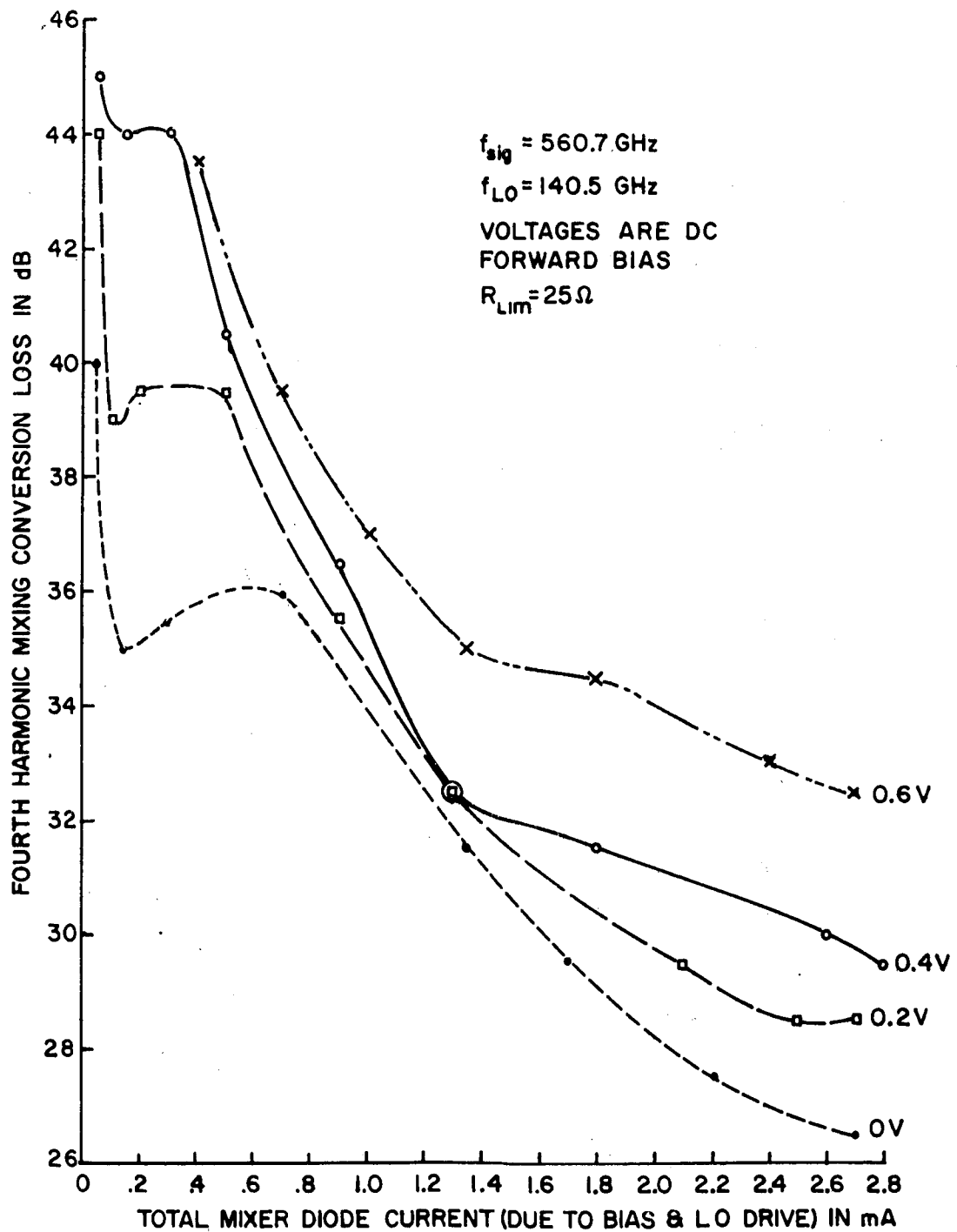


FIG. 2-12 - Measured Fourth Harmonic Mixing Conversion Loss vs Diode Current, Diode of 2-28-66

It is clear from Figure 2-12 that the dip in conversion loss at each bias level occurs for about the same value of total mixer current. These dips are probably caused by substantial variation of the diode impedance at signal and LO frequencies as the diode current changes. It is characteristic of mixer diodes that their impedance varies rapidly as a function of current at low current values. Since there is no impedance measuring equipment available at submillimeter wavelengths, this point cannot be verified experimentally. It is, however, a common occurrence, and it has been verified experimentally at lower frequencies where impedance measuring instrumentation is available.

The bias voltages given on the graphs in this section are external voltages applied from a battery bias box having an internal resistance of approximately 25 ohms. The actual operating point on the diode I-V curve is different for it is affected also by the self bias produced by the rectified LO power. The self bias current flows in the easy direction through the diode as does the current produced by the dc forward bias, but it develops a voltage across the resistance of the external source which opposes the external voltage. This effect is small since the bias source resistance is only 25 ohms.

The dc characteristic curve of the diode used for the mixer tests of 2-28-66 is shown in Figure 2-13, which is a composite tracing of photographs of the reverse and forward conduction regions as displayed on an oscilloscope. The "break" or maximum change of slope in the forward direction occurs at about 0.6 V as is typical for gallium arsenide diodes. The diode dc resistance is about 12,000 ohms at that point. In the reverse direction, 10 μ A of conduction occurs at -3.5 V which corresponds to a back resistance of 350 K ohms.

2. Mixer Conversion Loss Measurements on 3-7-66 at 560 GHz

On 3-7-66 a diode was formed which would withstand greater LO drive power and diode current than the diode of 2-28-66. Figure 2-14 is a plot of mixer conversion loss as a function of LO drive. The curves drawn for the individual bias voltages do not appear to reach a minimum at 40 mW of LO drive, which was the highest LO drive available at the mixer. The lowest conversion loss at 560 GHz measured for any diode was obtained with this diode with 40 mW of drive and 0.1 V bias and was 24 dB. Conversion loss for 0.1 V bias for other LO drive levels were measured but are not plotted because the other curves are shown for 0.2 V bias increments. By comparison with the data of Figure 2-10 one would expect even lower L_{TM} for zero bias and greater LO drive. The system signal-to-noise ratio on 3-7-66 was typically 30 to 34 dB for 40 mW LO drive.

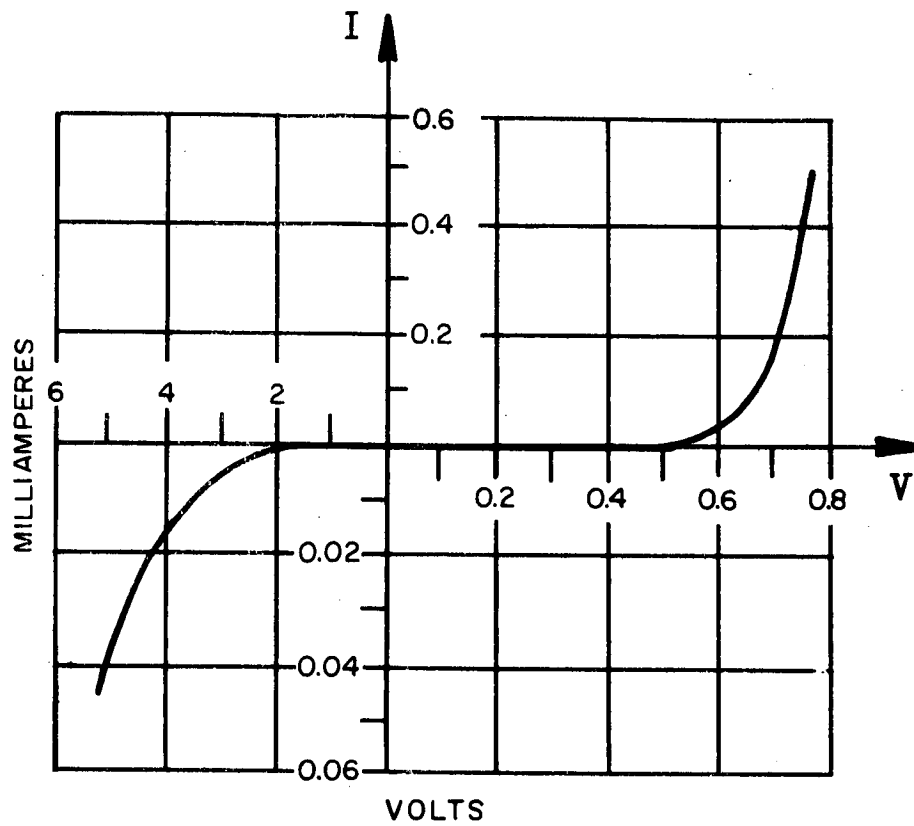


FIG. 2-13 - DC Characteristic of Mixer Diode of 2-28-66
(n-GaAs, Phosphor Bronze)

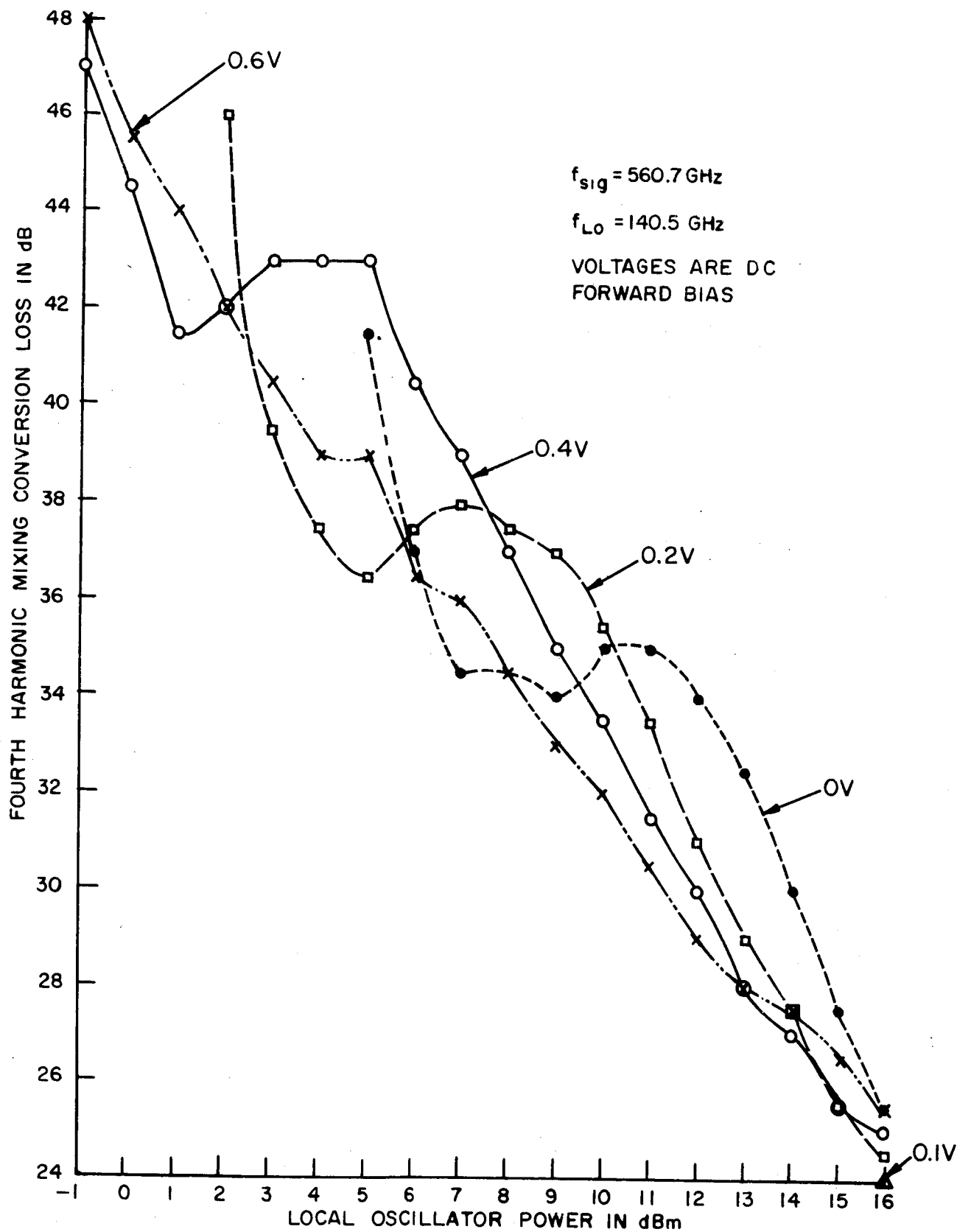


FIG. 2-14 - Measured Fourth Harmonic Mixing Conversion Loss vs Local Oscillator Power, Diode of 3-7-66

The diode current is plotted as a function of LO power and dc bias in Figure 2-15. Unlike the diode of 2-28-66 whose current is plotted in Figure 2-11, the diode of 3-7-66 was limited by available LO drive power. It withstood 5.4 mA of current which was reached only with 0.6 V external bias in addition to 40 mW of LO drive. Note that the scales of Figure 2-15 are compressed in comparison with those of Figure 2-11.

Again, the data from which Figures 2-14 and 2-15 were plotted have been regrouped and plotted in Figure 2-16, conversion loss as a function of total diode current. Here it is also evident that a lower mixer conversion loss could have been achieved with more LO drive. Comparing these curves with the data in Figure 2-12 one would expect minimum conversion loss for zero external bias and sufficient LO drive to reach 5.4 mA. By extrapolation of the zero bias curve of Figure 2-15 to 5.4 mA it appears that 70 mW of LO drive would be required. Then by extrapolation in Figure 2-14 or 2-16 the conversion loss that would be obtained for 70 mW drive is found to be on the order of 20 to 22 dB. While 70 mW is 8 dB above the present capability of a solid state local oscillator, it is interesting to know that there is the potential of achieving 20 dB fourth harmonic mixer conversion loss at 560 GHz. Future development may show that the conversion loss property of the diode of 3-7-66 can be achieved with the LO drive characteristic of the diode of 2-28-66.

3. Verification of Operating Frequency

The measurements on 2-28-66 and 3-7-66 were made with the test setup of Figure 2-9. The signal source power was measured with a Wollaston wire bolometer as described in the preceding subsection. Evidence that the power measured by the bolometer was power at 560 GHz and not a lower harmonic of the 80 GHz signal source klystron was obtained by measuring the guide wavelength with the harmonic generator output micrometer tuner. Five successive output maxima, representing two guide wavelengths, were observed and found to be spaced over 1.24 mm or an average measured guide wavelength of 0.62 mm. The calculated guide wavelength for 560 GHz in 0.020" x 0.010" waveguide is 0.63 mm; for the next lower harmonic, 480 GHz, it is 0.78 mm; for the next higher harmonic, 640 GHz, it is 0.525 mm. Clearly, the power measured was predominantly 560 GHz. The mixing frequency was verified by measuring the signal guide wavelength with the mixer signal micrometer tuner. Seven successive output maxima which represented three guide wavelengths were observed and found to be separated by 2.37 mm or an average measured guide wavelength of 0.79 mm. The calculated guide wavelength for 560 GHz in 0.014" x 0.007" waveguide is 0.816 mm.

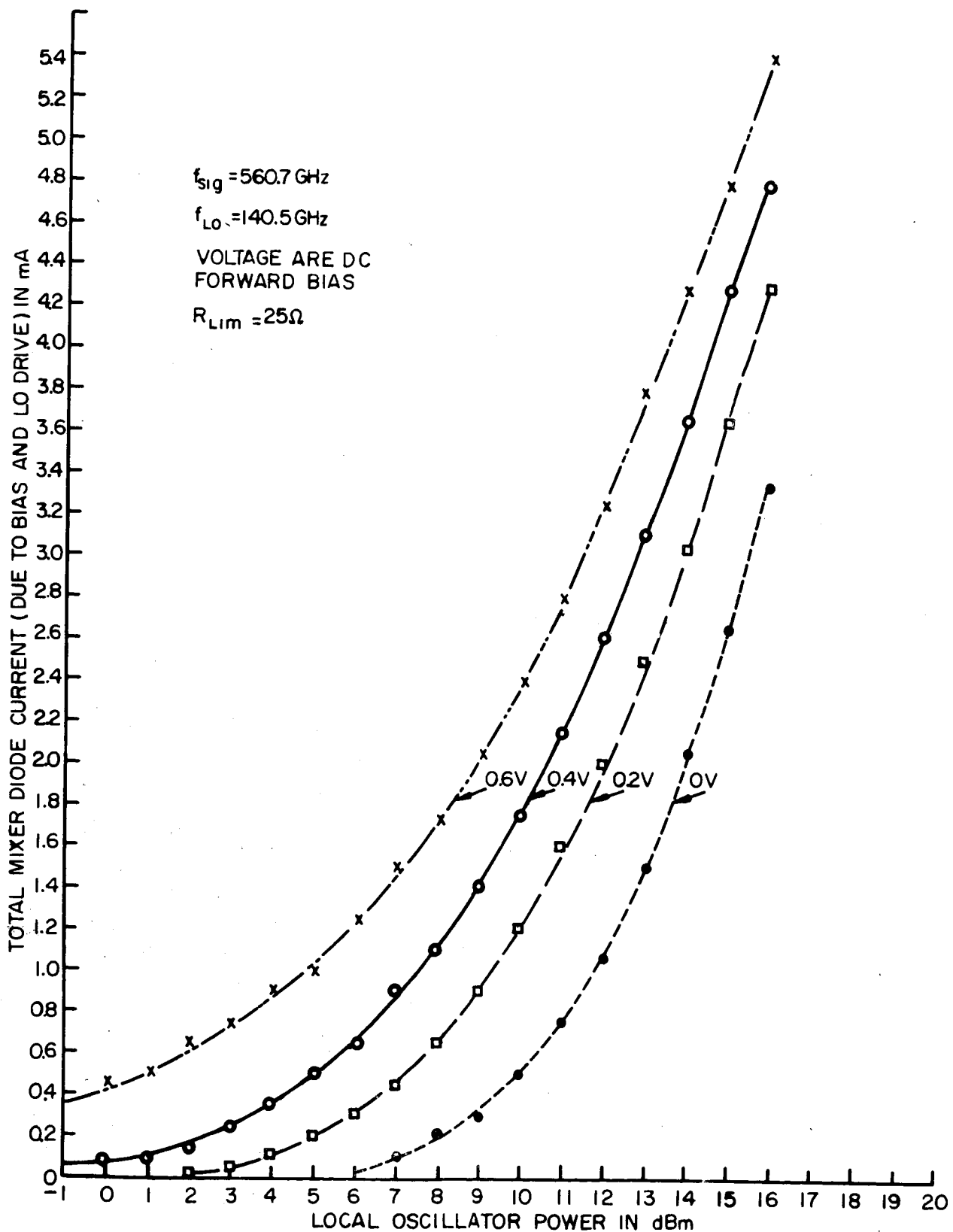


FIG. 2-15 - Direct Current in Fourth Harmonic Mixer, Diode of 3-7-66

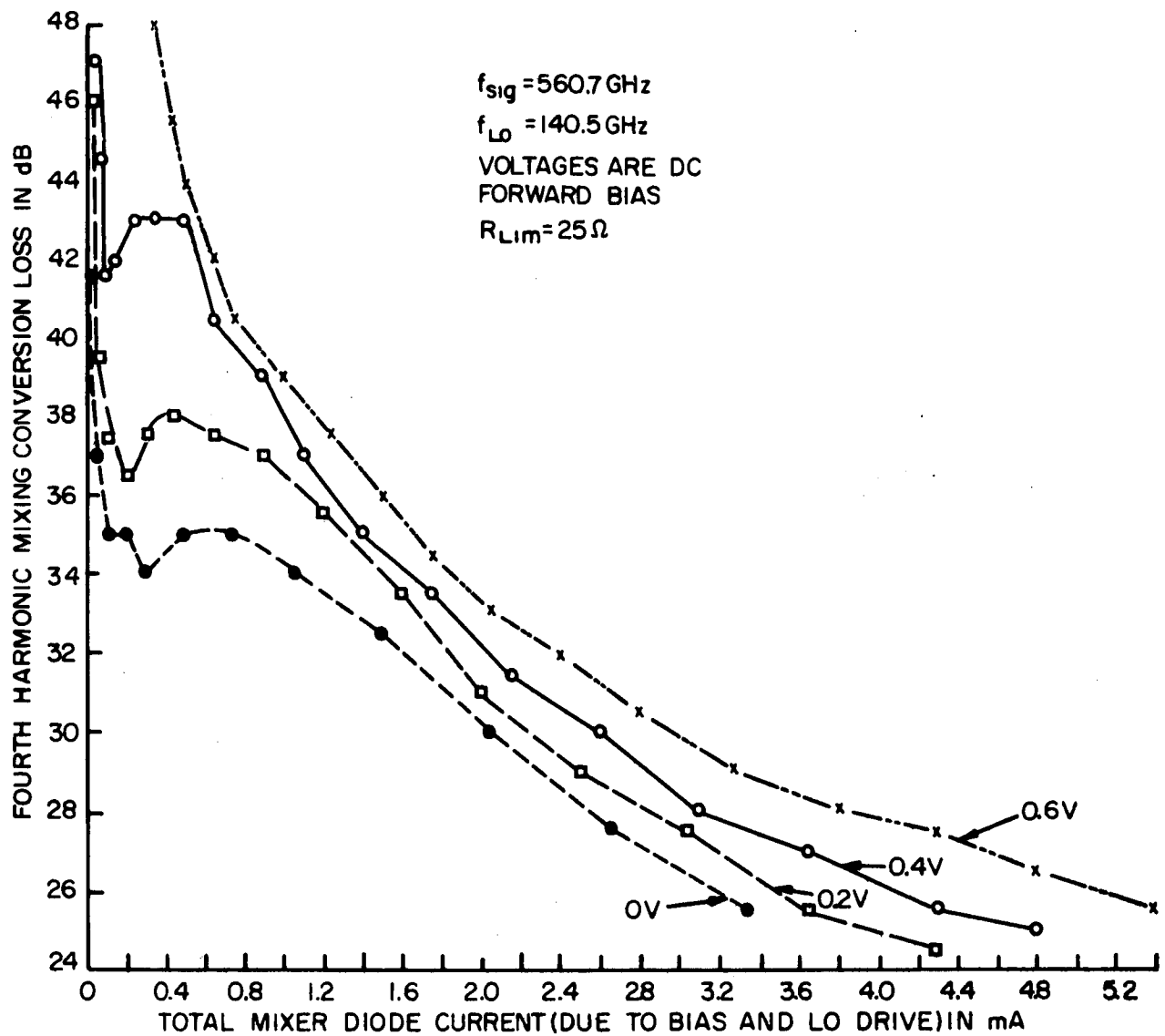


FIG. 2-16 - Measured Fourth Harmonic Mixing Conversion Loss vs Diode Current, Diode of 3-7-66

4. Mixer Loss Measurements with Dual IF Conversion at 420, 560 and 700 GHz

Harmonic mixing was also observed earlier with the test setup of Figure 2-8. Since both the local oscillator and the klystron in the signal source were operating at nearly the same frequency, 140 GHz, the distinction between third, fourth, and fifth harmonics was made with a narrow band second IF conversion as explained in the preceding section. The frequency being mixed was verified by guide wavelength measurements. At that time the high pass filter was not available to use at the output of the signal source, and signal power measurements at 560 and 700 GHz could not be made. The combined harmonic generator and harmonic mixer conversion loss was recorded instead. The increased harmonic generator conversion loss at fourth and fifth harmonics was estimated from measurements at the third harmonic and the mixer conversion loss at 560 and 700 GHz deduced from that.

Figure 2-17 is a plot of total fourth harmonic generation and mixing loss as a function of local oscillator drive power measured on 12-23-65. It is estimated that the harmonic generation portion of the conversion loss was between 46.5 and 52 dB. The lowest conversion loss shown in Figure 2-17 occurred for zero bias and 8 to 10 mW LO drive. The mixer portion is deduced to be between 31.5 and 37 dB. The conversion loss curves in Figure 2-17 for different bias voltages are generally similar to those already presented in Figures 2-10 and 2-14. The two inflection points in the 0 and 0.2 volt bias curves occur at higher levels of LO drive in Figure 2-17. This may be due to a different diode characteristic on 12-23-65, or it may be that the LO power reaching the diode was less than indicated. If the curves of Figure 2-17 could have been extended by applying higher LO drive, lower conversion losses might have been achieved as were, in fact, measured later in February and March.

Since the harmonic generator output waveguide had a cutoff frequency which fell between the second and third harmonic of 140 GHz, its output power was predominantly third harmonic and the third harmonic mixing conversion loss could be separated from the total conversion loss. Figure 2-18 is a plot of 420 GHz mixer conversion loss for different bias voltages. The 0.5 V value was used instead of 0.6 V. These curves differ in character from those for fourth harmonic mixing in that the lower bias values do not have two inflection points. The minimum conversion loss was 32.5 dB and occurred for zero bias and 20 mW LO drive. Measurements with other diodes on three prior occasions gave third harmonic mixer conversion losses at 420 GHz of 32 to 34 dB. However, it is believed that substantially

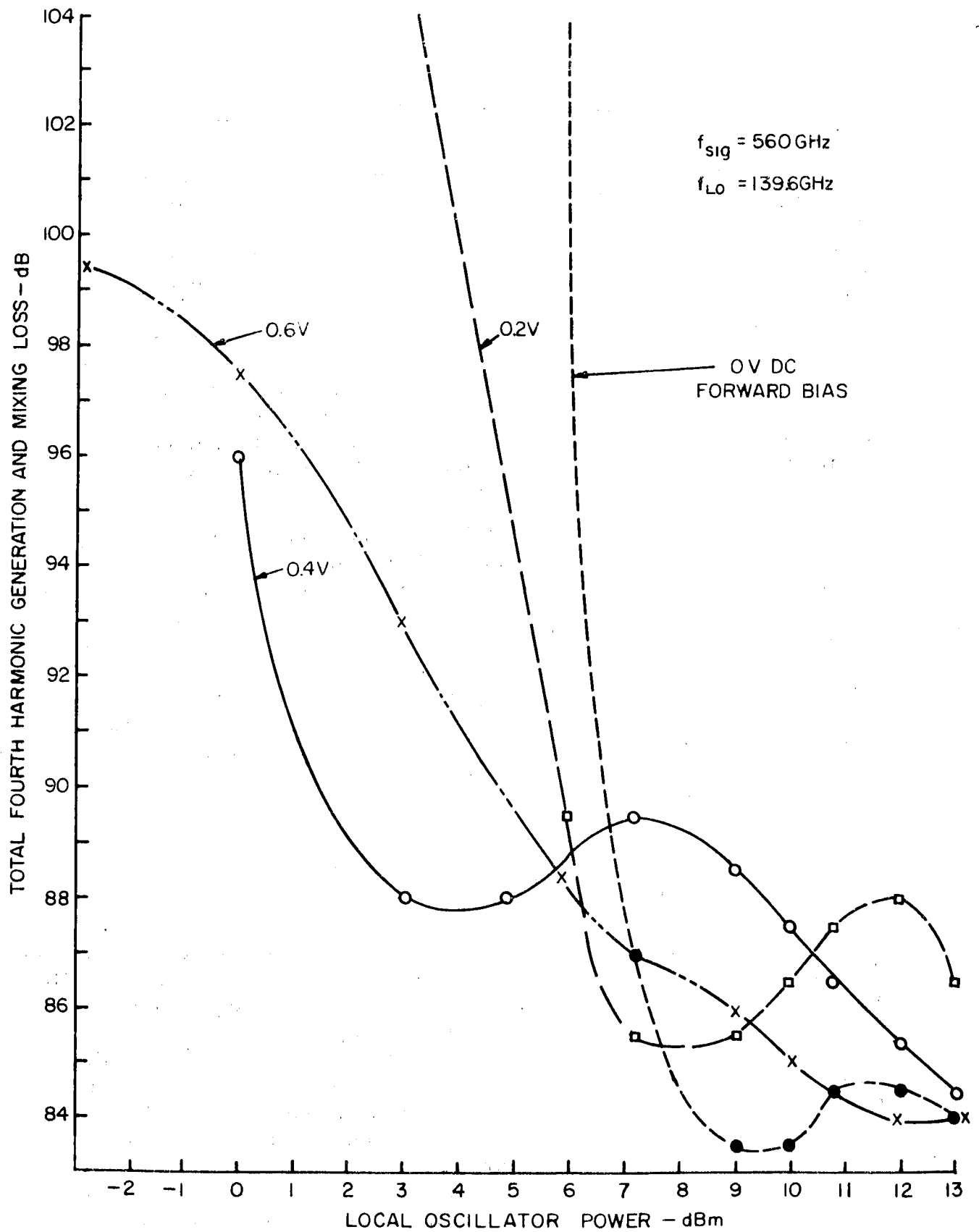


FIG. 2-17 - Measured Total Fourth Harmonic Generation and Mixing Loss, Diode of 12-23-65

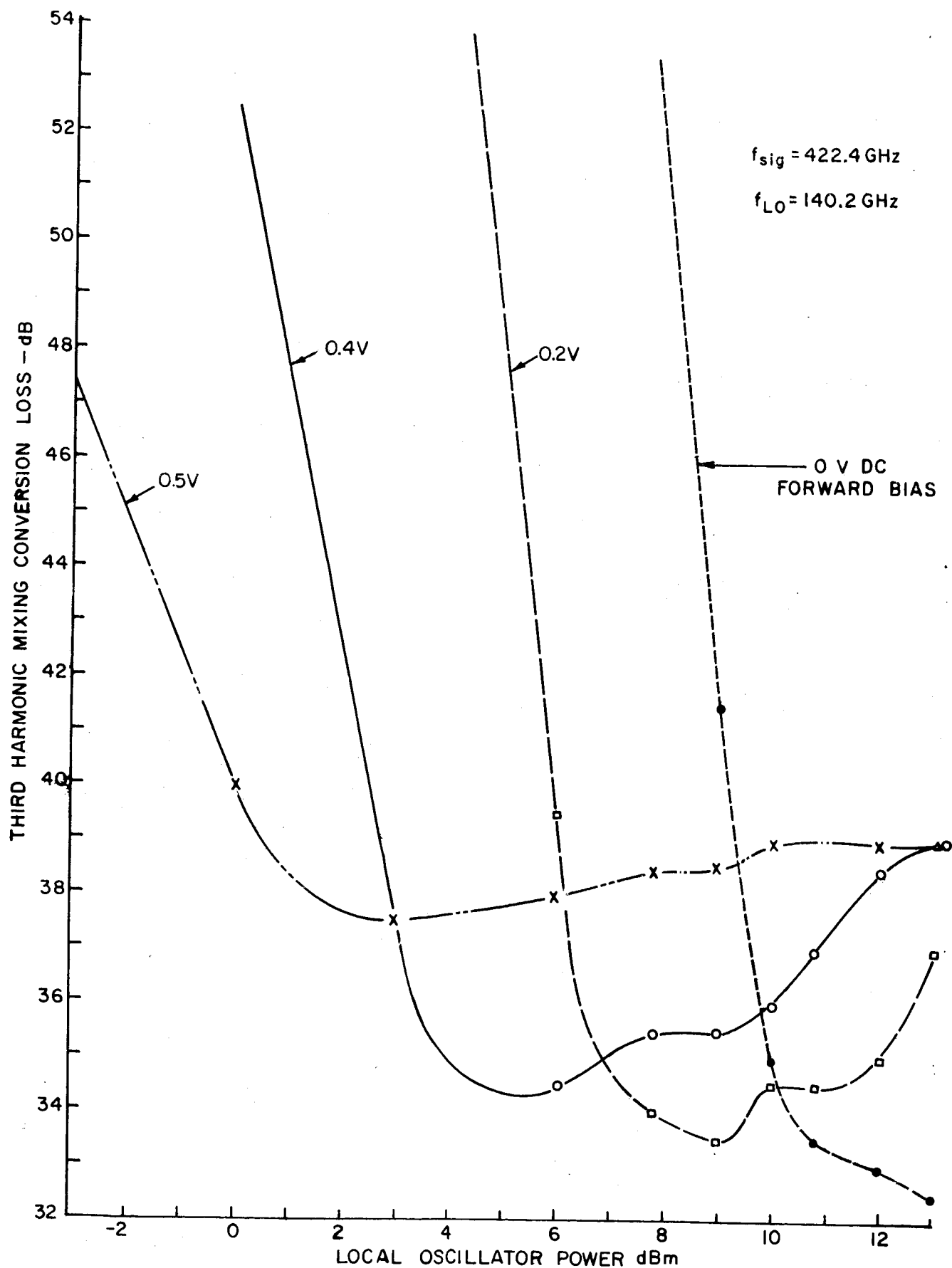


FIG. 2-18 - Measured Third Harmonic Mixing Conversion Loss,
Diode of 12-17-65

lower losses could be achieved with further development in view of the value of 26.5 dB measured later at 560 GHz for 20 mW LO drive. A contributing factor to the measured conversion loss at 420 GHz is the loss due to operation at the nominal cutoff frequency of the signal waveguide which is 420 GHz. Since third harmonic mixing at 420 GHz was not a project goal, further development was not attempted.

By careful tuning of the test setup of Figure 2-8, mixing at the fifth harmonic, i.e., 700 GHz, was achieved with a signal-to-noise ratio of 15 dB. The combined harmonic generation and mixing loss was 99 dB; LO drive was 10 mW. Harmonic generator conversion loss for 420 GHz was measured to be 45 dB; L_{HG} at 700 GHz may have been 60 dB. Data were not measured for other levels of LO drive. The frequency being mixed was verified by guide wavelength measurements with the micrometer tuner in the mixer signal waveguide. Six successive output maxima, representing two and one-half guide wavelengths, were observed and found to be spaced over 1.32 mm or an average measured guide wavelength of 0.53 mm. The calculated guide wavelength for 700 GHz in 0.014" x 0.007" waveguide is 0.54 mm.

5. Mixer Conversion Loss Measurements on 7-24-65 at 525 GHz

The earliest operation of the submillimeter harmonic mixer was in July, 1965 using the test setup shown in Figure 2-7. The signal source produced 525 GHz from a seventh harmonic generator. The harmonic power which was predominant in this case was the fifth at 375 GHz and was measured with a bolometer to be -31 dBm for an L_{HG} of 40 dB. The seventh harmonic conversion loss was estimated to be 50 dB. Since the combined losses of generator and mixer measured 102 dB, the mixer loss must have been on the order of 52 dB. Only a limited number of diodes were formed and tested in the mixer at this time. A new signal waveguide was installed in the mixer before the tests which began in December.

6. Mixer Diode Noise Ratio

The noise ratio of the mixer diode of 3-7-66 was measured by the Y-factor method⁽²⁻⁴⁾ for 10 mW LO drive, 0 V external bias, and an IF band of 1 to 2 GHz. The resultant value of noise ratio was 3. An IF amplifier was unavailable for a measurement for 2 to 4 GHz, but it is estimated that the noise ratio would be 2 for that band. The measured value depends upon the noise output of the local oscillator and will have to be measured ultimately with the SSLO.

7. Mixer IF VSWR

The IF output VSWR of the submillimeter mixer was measured during the conversion loss tests in order to correct for mismatch loss. In general, the VSWR decreased markedly as LO drive was increased and decreased to a lesser extent as external dc bias was increased. The measurements of VSWR for these corrections were made at the intermediate frequency of the test setup only. Since a radiometer will use a broad IF band, impedance data was measured and plotted in Figure 2-19 for one diode over the band 1 to 4 GHz. LO drive power was set at 20 mW; there was no external bias. The VSWR is less than 3:1 from 1.6 to 3.5 GHz, a frequency band greater than an octave. The mixer as measured had an arbitrary 100 ohm matching transformer in the IF output line. The impedance referenced to the input to the transformer (which is very close to the diode junction, see Figure 2-1) is also plotted. The impedance plot of Figure 2-19 is presented as evidence that a good IF match can be achieved with an IF output design optimized for the IF band ultimately selected for a radiometer. Similar data on a number of diodes would be collected upon which to base an optimum matching network design.

E. Conclusions

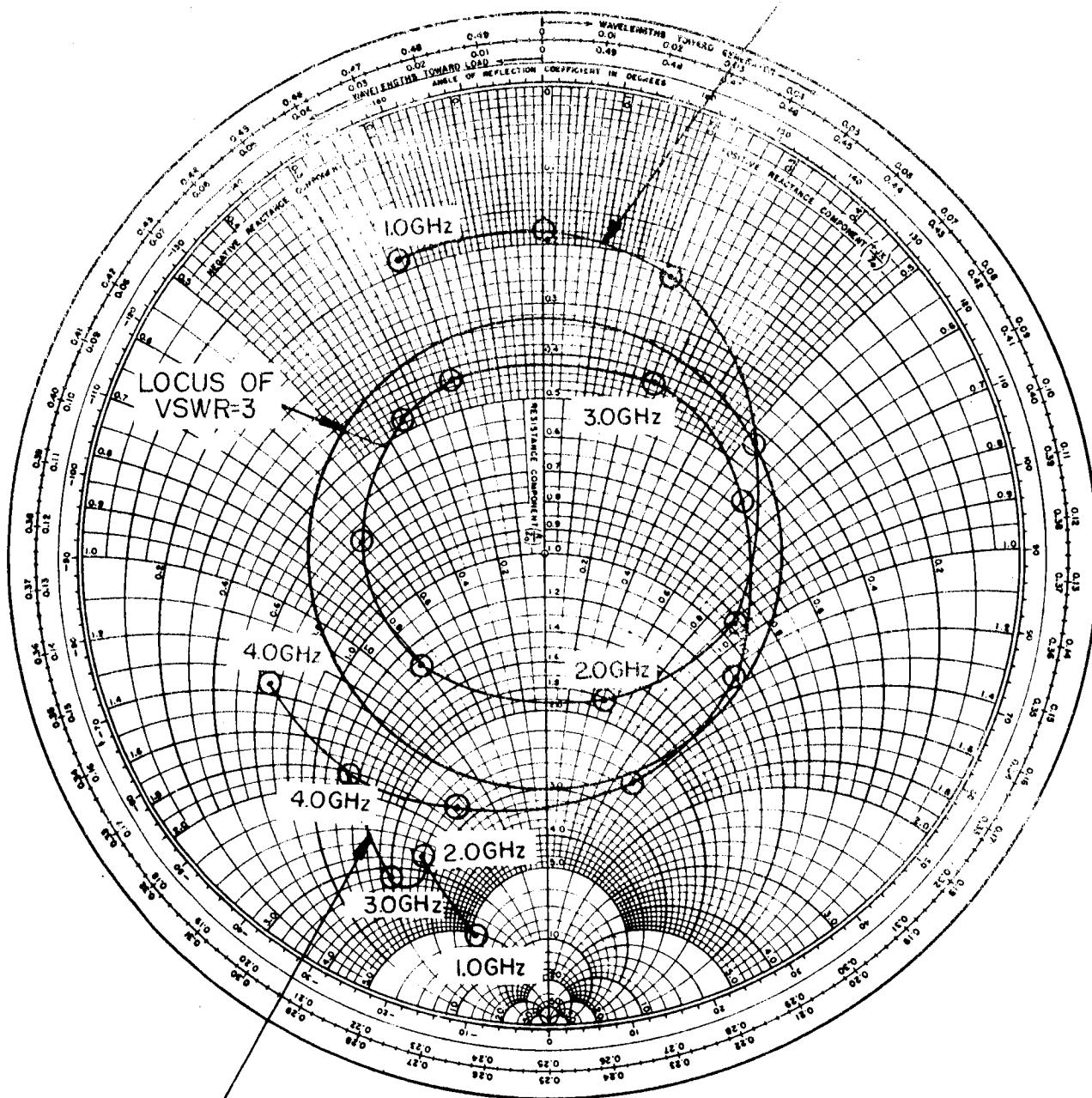
The data on conversion loss of the submillimeter mixer have been summarized in Table 2-1. In general, lower conversion loss measurements were obtained as the testing progressed. The earlier fourth and fifth harmonic measurements are enclosed in parentheses to indicate that they are estimates. The development of the submillimeter mixer has been far from exhaustive. The greater portion of the effort expended on the mixer has been to construct the device and to develop means of testing it. The measured results have demonstrated the feasibility of a submillimeter satellite radiometer utilizing a harmonic mixer. Continued developmental testing of the submillimeter mixer would undoubtedly yield better performance. Furthermore, harmonic mixing superheterodyne receivers could be constructed at frequencies at least as high as 700 GHz.

Table 2-1 includes data measured on the signal-to-noise ratio at the output of the test setup. This represents the dynamic range available for testing at submillimeter wavelengths with the test setup. For example, the insertion loss of waveguide components such as ferrite switches, the transmission loss of lenses, and the major feature of antenna radiation patterns could be measured. In a laboratory setup, greater LO power could be used with probable reduction in conversion losses and an increase in dynamic range for testing.

$P_{LO} = 20\text{mW}$

0V EXTERNAL BIAS

REFERENCED TO IF OUTPUT
CONNECTOR



REFERENCED TO BEGINNING OF
100 OHM TRANSFORMER

FIG. 2-19 - IF Output Impedance of Submillimeter Mixer,
Diode of 2-28-66

Date of Measurement	Signal Frequency	Order of Harmonic Mixing	Mixer Conversion Loss for Local Oscillator Power of			Signal Level	Signal-to-Noise Ratio for Local Oscillator Power of		
			10 mW	20 mW	40 mW		10 mW	20 mW	40 mW
	GHz		dB	dB	dB	dBm	dB	dB	dB
7-24-65	525	4		(52)		(-41)		15-20	
12-15-65	700	5					15		
12-17-65	420	3	33.5	32.5		-39		42	
12-23-65	560	4	(31.5-37)	(32-37.5)		(-42 to -36.5)	23		
2-28-66	560	4	29.5	26.5		-48			
3-7-66	560	4	32	28	24	-51	22-26		30-34

Note: All numbers in parentheses are estimates.

TABLE 2-1
Measured Data for Submillimeter Harmonic Mixers

Factors contributing possible errors to the data of Table 2-1 have been discussed in Subsection C on test procedures. These include possible errors in the calibration of the bolometer which are compensated to some degree by the inability of the bolometer to distinguish between the desired harmonic and higher harmonics. Losses due to reflections at the signal port of the mixer are included in the values measured for mixer conversion loss. To the device designer this represents an error in the true value of diode conversion loss, but for the system designer signal mismatch losses need not be separated from diode conversion loss. The achievement of a conversion loss of 29.5 dB for only 10 mW of LO power at 140 GHz makes it possible to predict that a satellite radiometer can be built having a $(\Delta T)_{\min}$ for $S/N = 1$ of 7°K . Improvements in the mixer such as mechanical changes, IF matching, and diode development have been described which, if carried out, will lead to even better performance. The design of the diode and mixer mount to meet environmental requirements will be described in Section V.

III. SOLID STATE LOCAL OSCILLATOR

A. Design Considerations

1. Requirements

The solid state local oscillator was selected as one of two key components to be developed to demonstrate the feasibility of a satellite submillimeter radiometer. The requirements for the SSLO were discussed in Subsection C of Section I with respect to the requirements for the other key component, the submillimeter harmonic mixer. The SSLO is envisioned as an RF oscillator followed by a chain of multiplier stages. The term "multiplier" is used interchangeably for "harmonic generator" which may be a second, third or higher order harmonic generator.

The estimated requirements on the SSLO output were several milliwatts at 2 mm wavelengths with reasonable efficiency. If a multiplier could be developed to meet these requirements, then it could serve as the output stage of a chain of multipliers. Similar techniques could be applied to the preceding multipliers in the chain which, operating at lower frequencies, would not be as difficult to design. However, it is very important to achieve good efficiency in the output stage so that the power output required from the preceding stage will be a realistic requirement.

The development of the SSLO was concentrated on the design of a harmonic generator to use as the output stage. A nominal output frequency of 140 GHz was chosen in view of the design frequency selected for the submillimeter mixer and the frequencies of available klystron sources for test purposes. Two main approaches to the design of a 140 GHz harmonic generator were investigated.

One approach was to develop a new type of device which would operate by virtue of the nonlinear permittivity as a function of the voltage field of certain ferroelectric materials at low temperature. This nonlinear effect is a bulk effect rather than a junction effect and offers promise of good power handling capability. Capacitors of the ferroelectric, strontium titanate, were assembled for test in a waveguide mount capable of being cooled to liquid nitrogen temperature. A description of the operating principles and the experimental work on the strontium titanate harmonic generator is given in Appendix H.

The other approach to the development of a 140 GHz multiplier was to extend and improve the techniques for harmonic generation using semiconductor diode junctions. A discussion of the principles

of these devices and the application of these principles to the design of a successful 140 GHz second harmonic multiplier is given in the following sections. Most of the work on diode junctions was devoted to second harmonic generation rather than third or higher because the theory and the practical device construction of doublers are simpler and promised earlier success. The added complication of tuning at idler frequencies in higher order multipliers was left as a possible future refinement.

2. Previous Work with Harmonic Generators

For many years varactor diodes, which are nothing more than reverse-biased p-n junction diodes, ⁽³⁻¹⁾ have been used to obtain microwave RF energy by harmonic multiplication from some lower radio frequency power source. Theoretically, ⁽³⁻²⁾ a varactor pumped at some fundamental frequency can produce an output at any desired harmonic with 100% efficiency. In practice, however, the conversion efficiency decreases very quickly with increasing fundamental frequency and desired harmonic number.

The results which have been obtained to date are actually quite impressive, that is, for the frequency range up to and including X-band. One supplier reports ⁽³⁻³⁾ such performance as a 1 to 4 GHz quadrupler with 2 watts input power and a conversion efficiency of 45% minimum or a 4 to 12 GHz tripler with one watt input power and a conversion efficiency of 30%. Another company has reported ⁽³⁻⁴⁾ a device which triples from 11.6 to 34.8 GHz with 0.172 watts input power and has an output power of 0.100 watts for a conversion efficiency of 58%, or 2.3 dB loss. Recently, an output of 2 mW at 60 GHz has been reported ⁽³⁻⁵⁾ for a tripler from 20 GHz having an efficiency of 2.5%. At 3 mm wavelengths (104 GHz) DeLoach ⁽³⁻⁶⁾ has generated 1 mW output with 8.3% efficiency for a doubler using a diode assembled in waveguide rather than in a diode package. At 2 mm (136 GHz) workers at this laboratory ⁽³⁻⁷⁾ have obtained 1 mW with 5.6% efficiency using a variable resistance diode also assembled directly in a waveguide mount. Very little other information has been published on varactors applied as harmonic generators for output frequencies in excess of 35 GHz. An attempt was made at the beginning of this project to use a commercial varactor in a tripler from 47.6 to 140 GHz. The device is described in Appendix I. The third harmonic conversion loss was 49 dB. The frequency cutoff of the commercially available varactor diodes is just too low for efficient operation any further into the millimeter wave region.

The original development of harmonic generators with millimeter wave output frequencies used the point contact variable resistance diode as the nonlinear element rather than a variable reactance. Therefore, the devices were limited to a maximum conversion efficiency⁽³⁻⁸⁾ of $1/n^2$ for generating the n^{th} harmonic. This number was reduced further by all the circuit losses and the diode parasitic losses. Figure 3-1 presents a general summary of typical conversion losses achieved at this laboratory with variable resistance diode harmonic generation techniques. For all cases these were gallium arsenide devices with electrically "formed" junctions. The crossed waveguide structure was common to all, and all but one were open, unsealed devices which thereby were susceptible to having unpredictable lifetimes. The conversion losses indicated are typical and do not reflect the superlative performance which is only occasionally encountered. In doubling to 140 GHz, for example, conversion losses as low as 10 dB have been observed, but these cannot be readily repeated. As will be described in Subsection C following, this performance has been surpassed with variable reactance diodes formed and tested on this project. A varactor doubler to 140 GHz was measured to have less than 6 dB conversion loss.

The harmonic generators, like mixers, are quite sensitive to variations in input power at the fundamental frequency. Unlike the mixers, however, attempts to compensate for low drive power with external dc bias are rarely beneficial. For almost all cases the lowest conversion loss is achieved with no additional dc voltage supplied. Some dc path must be provided for the flow of rectified current so that the diode will set its own operating point. A rough rule of thumb has been employed for some time that the dependence on input power carries the exponent of the harmonic number involved.⁽³⁻⁹⁾ That is, in doubling, the output power would be proportional to the square of the input power, tripling would be proportional to the cube, etc. Under this rule, the conversion loss, being the ratio of the powers involved, would carry this exponent reduced by one. Recent experience with gallium arsenide has shown that this rule is valid only when the input power is considerably below that required for optimum performance. As one approaches the proper drive level, this dependence is less pronounced and, in fact, ultimately reverses. This is shown in Figure 3-2 for a doubler to 136 GHz, built and tested under Contract NASw-662.^{(3-7)*} For this particular unit the optimum conversion

* Figure 3-2 is based upon the same measurements as Figure 39 of Reference (3-7) but has been corrected for input mismatch loss.

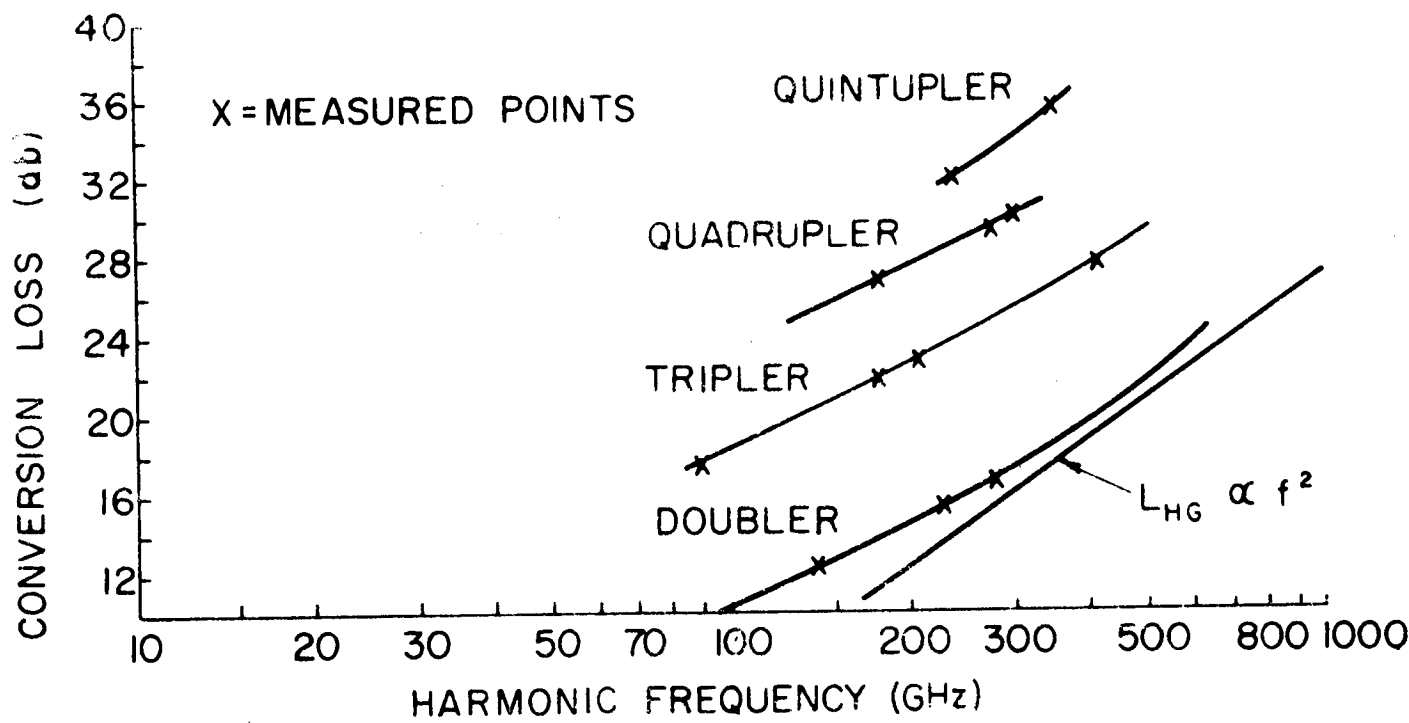


FIG. 3-1 - Harmonic Generation Conversion Loss
vs Frequency and Harmonic Number

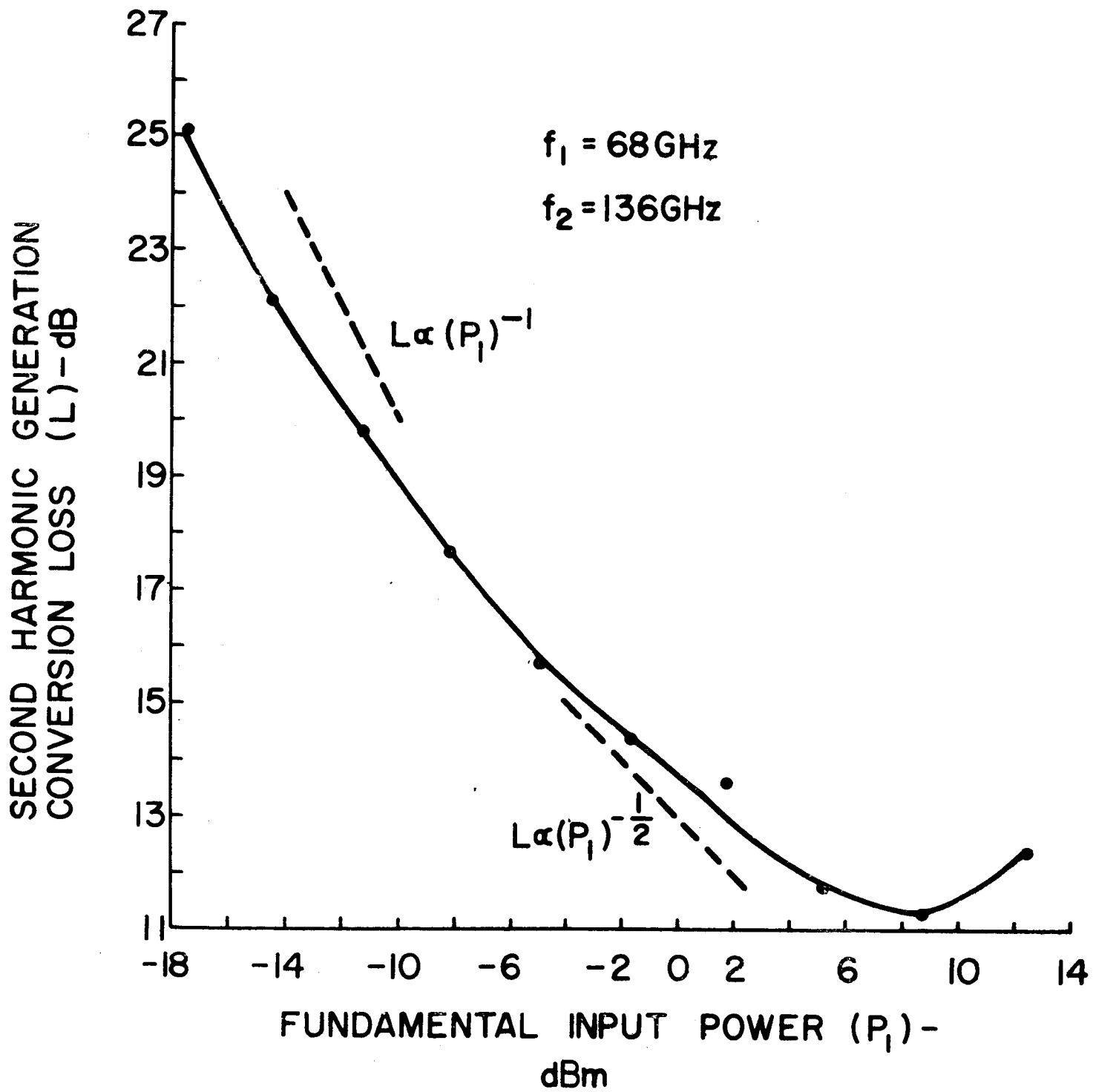


FIG. 3-2 - Measured Second Harmonic Generation Conversion Loss

loss was achieved with about 7 mW of drive power. As the input power was increased toward 18 mW the harmonic output power continued to increase to the one milliwatt level; the efficiency of harmonic generation, however, had begun to decrease. As one continues to increase power to the point of burnout, the situation is usually encountered where additional drive power results in decreased harmonic power. Most of the devices measured have shown an optimum operating point, in terms of conversion loss, in the two to ten milliwatt range of input power.

3. Characterization of Millimeter Wave Diodes

It has been known for some time that point contact rectifiers, be they p-n junctions, Schottky barrier junctions, or some type in between, have a sharp voltage-dependent junction capacitance.⁽³⁻¹⁰⁾ When these junctions are reverse biased they can be characterized in the same manner as any varactor diode. A program was initiated at this laboratory to determine more precisely the nature of the junction being used so successfully throughout the millimeter wave range as a mixer and as a variable resistance harmonic generator and to determine, if possible, what modifications in the selection of materials or in the junction forming technique might allow the maximizing of the efficiency and/or power handling capability of the junction as they may apply to the variable reactance form of harmonic power generation.

All of the sharply breaking I-V curves which have been formed by electrically pulsing the junction with the 60 Hz sweep voltage have followed closely the equation⁽³⁻¹⁰⁾

$$I_f = I_s \left\{ \exp \left[\frac{q}{\eta k T} (V_f - I_f R_s) \right] - 1 \right\} \quad (3-1)$$

where

I_f is the forward current,

I_s is the saturation current,

q is the electronic charge,

k is Boltzmann's constant,

T is the absolute junction temperature,

V_f is the forward (applied) voltage,

R_s is the junction spreading resistance.

The quantity η is a slope parameter which, for the diodes measured, had values between 1.2 and 1.4. If the junctions were primarily p-n junctions, then the expected value of η would be about 2.0.⁽³⁻¹¹⁾ If the junctions were ideal Schottky barrier junctions, then η would be equal to 1.0.⁽³⁻¹²⁾

The first combination of materials which was examined was n-type gallium arsenide as a semiconductor and a phosphor-bronze pointed contact. This combination has received the most publicity⁽³⁻¹³⁾⁽³⁻¹⁴⁾⁽³⁻¹⁵⁾⁽³⁻¹⁶⁾ and is the one most commonly employed in millimeter wave mixers. Figure 3-3 shows two typical dc current-voltage characteristic curves obtained for GaAs of resistivity $\rho = 0.015$ ohm-cm. The junctions were formed by electrically overloading the junctions with the 60 Hz sweep signal of the I-V curve tracer. The results of Figure 3-1 were obtained with diodes typified by the I-V characteristics of Figure 3-3. As the slope parameters were on the order of 1.3 to 1.4 for the n-GaAs/phosphor-bronze rectifying contacts, it would appear that these rectifying contacts were predominantly majority carrier rectifiers of the Schottky type. However, the static characteristics of these diodes were not as close to the ideal Schottky characteristic as those reported by Kahng⁽³⁻¹⁷⁾ Kahng reported an η of the order 1.01 to 1.025 for large area diodes of n-type GaAs and gold contacts.

In an effort to decrease the slope parameter and thereby improve the millimeter wave operation of the devices, pure copper was substituted for the phosphor-bronze. It is believed that, during the forming operation, the contact surfaces of the phosphor-bronze and GaAs are raised to such a temperature that an intimate, metal-semiconductor interface is obtained. This may or may not be attendant with a diffusion of copper⁽³⁻¹⁴⁾ from the wire point into the GaAs. The diffusion of copper into the GaAs is not necessary for the formation of the Schottky barrier; only the true metal-semiconductor interface is needed. In fact, a diffusion of the copper may tend to grade the junction thereby giving less than the extremely abrupt junction desired. An additional reason for the change from phosphor-bronze to copper is the fact that phosphor-bronze is made up of 95% copper and 5% tin. Tin is a very active donor in GaAs, and a small amount of tin diffusion at the time of forming may also grade the junction and cause an increase in the measured slope parameter. Figure 3-4 shows two typical current-voltage characteristic curves obtained for GaAs ($\rho = 0.015$ ohm-cm) with pure copper wire point. Again the junction was formed by electrical pulsing while the I-V curve was displayed. While the slope parameter for the phosphor-bronze units typically stayed in the range 1.3 to 1.4, η for the copper-bonded units deviated little from 1.2 ($1.19 \leq \eta \leq 1.21$).

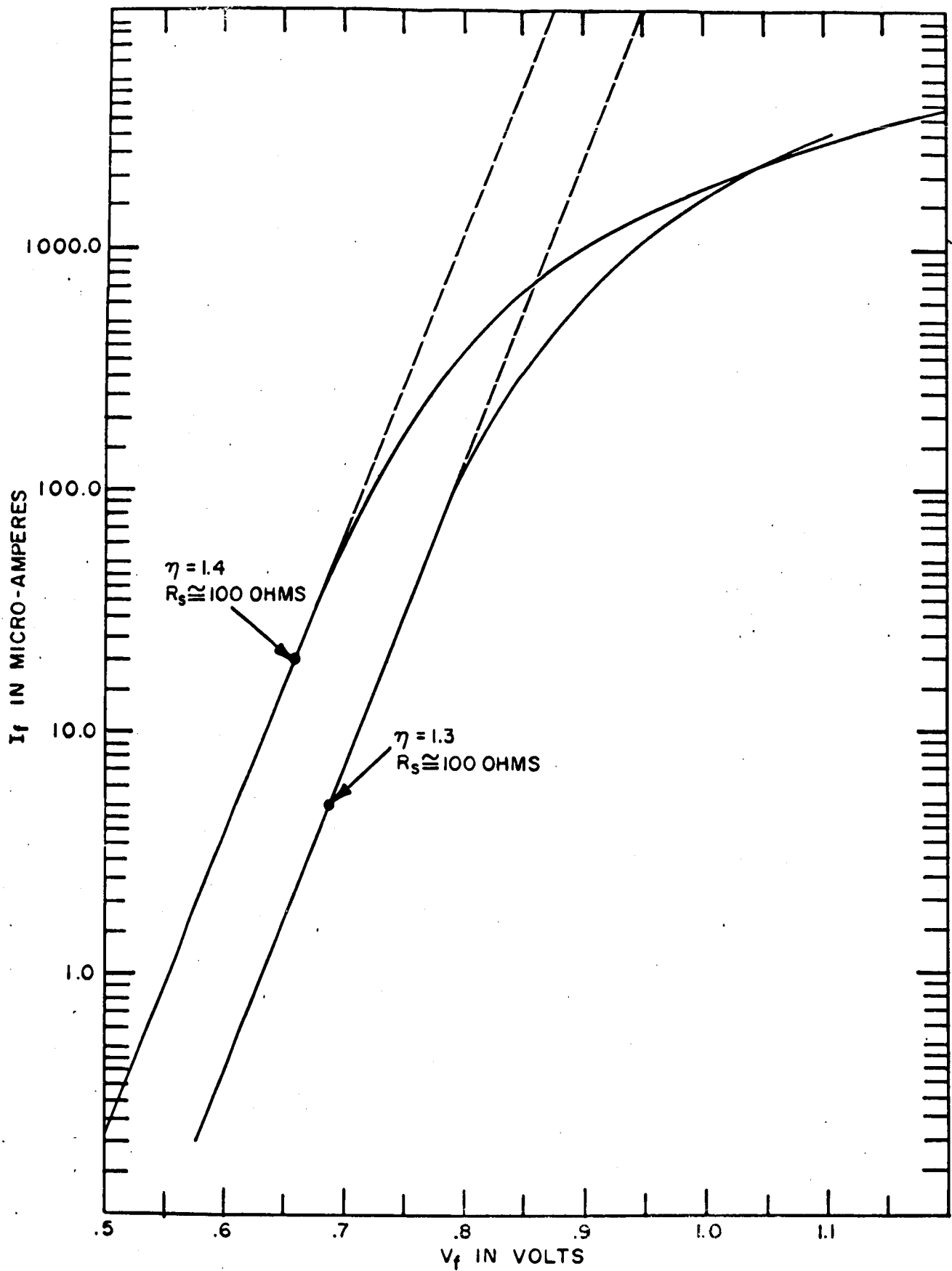


FIG. 3-3 - Current-Voltage Characteristics of Two Different, But Typical, Point Contact n-GaAs ($\rho = .015 \text{ ohm-cm}$), Phosphor-Bronze Diodes

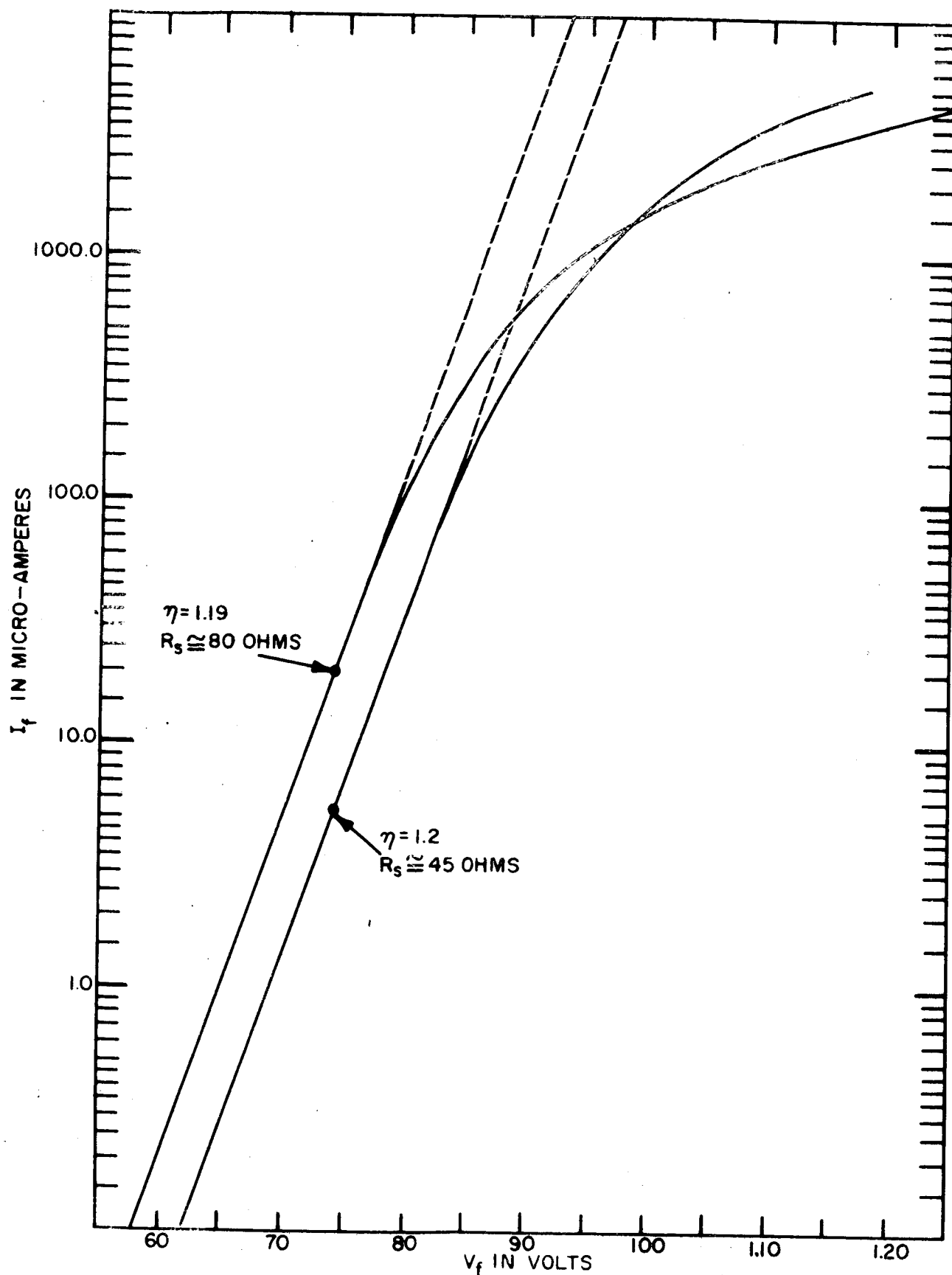


FIG. 3-4 - Current-Voltage Characteristics of Two Different, But Typical, Point-Contact n-GaAs ($\rho = .015 \text{ ohm-cm}$), Copper Diodes

In operation as a harmonic generator, the improvement with copper-bonded diodes was readily seen. Specifically, in the 70 GHz to 140 GHz doubler, an improvement in conversion efficiency of 2 to 4 dB was obtained over that data shown in Figures 3-1 and 3-2. The best conversion efficiency measured for the copper-bonded diodes was 8.8 dB. The power handling capability was quite similar to that shown in Figure 3-2 in that the junction would not yield appreciably more than 1 mW of 140 GHz power.

Some analytical justification of the above-noted results may be made which then can lead to further improvement. The following calculations are applicable to either the n-GaAs/phosphor-bronze devices or the n-GaAs/Cu devices as both had the same basic crystal material. The resistivity of the bulk material used was 0.015 ohm-cm. The spreading resistances, as interpreted from the measured I-V curves, ranged from 40 ohms to 180 ohms with an average figure of about 70 ohms. The reverse breakdown voltage (measured at the 10 micro-ampere conduction point) was 6 to 8 volts. For a uniformly doped semiconductor, the barrier capacitance of an abrupt junction (ideal step junction) depends on the reverse voltage in accordance with the well known equation⁽³⁻¹²⁾

$$\frac{C}{A} = \left(\frac{\epsilon q N}{2V_T} \right)^{\frac{1}{2}} \quad (3-2)$$

where

C is the capacitance,

A is the junction area,

ϵ is the permittivity,

N is the donor concentration,

V_T is the total voltage across the junction including the diffusion voltage, V_D , (i.e., $V_T = V_D - V$).

The spreading resistance R_s can be given as

$$R_s = \frac{\rho}{4a}, \quad (3-3)$$

where

ρ is the semiconductor resistivity,

a is the effective radius of the point contact.

If one takes a resistivity value of 0.015 ohm-cm and an R_s of 70 ohms, then Equations 3-2 and 3-3 allow an estimate for the junction capacitance of approximately 0.002 pF. The cutoff frequency, $f_{co} = (2\pi R_s C)^{-1}$, is calculated to be 1140 GHz at zero bias. This is truly a very high cutoff diode. According to presently accepted varactor theory, (3-18)(3-19) this diode as a doubler to 140 GHz should yield a conversion efficiency of about 80% (a conversion loss of 1.0 dB). However, this estimate does not seem to fit the data as shown in Figure 3-1. Of course, circuit losses and reflection (mismatch) losses may contribute in a large way to the total loss. Also consider the fact that the phosphor-bronze whisker can exert a rather large stress on the very small point contact. The effect of a large stress applied to a small area junction is to cause the "ideal" current through the small area to increase. (3-20) For high stress levels the current through the stressed area can become much larger than that through the unstressed area; hence the total diode current is essentially that flowing in the stressed area. It is very possible that the stressed area (the effective area of the subject diode) is only 20% of the total area. This could mean that the effective shunting capacitance is 0.008 pF rather than 0.002 pF. The effective cutoff frequency drops to 220 GHz and the calculated conversion loss jumps to a figure of the same order of magnitude as has been measured, about 6 dB. The theory now fits the measurements solely because of an allowance for additional parasitic capacitance due to stress on the diode.

The figure of 1.0 dB conversion loss is idealistic only in the sense that it represents the minimum possible losses of the model junction in an ideal (lossless) circuit.

A further point that should be made is that the active junction variable capacitance of 0.002 pF represents in itself a limit in power handling capability. It can be shown (3-21) that a good approximation of the maximum input power, P_{in} , which can be efficiently transformed to harmonic power by an abrupt junction, is given by the equation

$$P_{in} = 0.2 f_1 C_b (V_D - V_b)^2 \quad (3-4)$$

where

f_1 is the input frequency,

C_b is the active junction capacitance at the breakdown point,

V_b is the breakdown voltage.

This can also be given in terms of the zero bias capacitance, C_o , and the breakdown and diffusion voltages by

$$P_{in} = 0.2 f_1 C_o (V_D - V_b)^{3/2} (V_D)^{1/2} . \quad (3-5)$$

If one uses either Equation 3-4 or 3-5 and a junction capacitance of 0.002 pF, then the value 0.58 mW is obtained for the maximum input power. Even if the ideal conversion loss were realized, the output power would be insufficient for our application in a solid state local oscillator. When the diode is overdriven, this power level may be increased by taking advantage of the abrupt capacitance change between zero and the diffusion voltage (V_D).

Now if the junction area can be increased but at the same time the parameters again be achieved as have been reported, (3-14)(3-22)(3-23) then more reasonable power levels may be attained. If $R_s = 18$ ohms and $C_o = 0.01$ pF, then $P_{in} = 2.2$ mW and $f_{co} = 870$ GHz which gives a doubler conversion efficiency of about 74% or a loss of about 1.3 dB. If a junction capacity of 0.1 pF can be attained and R_s reduced to 3 ohms, then even with the f_{co} at zero bias dropping to 500 GHz, an efficiency of about 59% and 12.7 mW of output power at 140 GHz can be calculated.

B. Development of Epitaxial Varactor Harmonic Generator at 140 GHz

Reduced R_s , increased C_o and high f_{co} is attainable through the use of pulse formed p-n junctions. The junction and the forming technique is that proposed by Burrus⁽³⁻²³⁾ and consists of pulse forming junctions between an electrolytically pointed zinc whisker and epitaxial n-GaAs. The best results to date, for both power handling and efficiency, were obtained with GaAs that had a 4.0 μ m thick epitaxial layer and a carrier concentration of 5.9×10^{16} atoms/cm³. The resistivity of the epitaxial layer was 0.0422 ohm-cm and the substrate resistivity was 0.49×10^{-3} ohm-cm. The diodes made on this material had very sharp reverse breakdown voltages in the range 16 to 19 volts. Junctions could consistently be formed which could be made to yield on the order of 7 dB conversion efficiency (20%) with output power at 140 GHz of better than 10 mW.

The diodes were initially characterized at low frequencies. The capacitance measurements were made at a frequency of 1 MHz on a Boonton, Model 65B, capacitance bridge. The majority of the zero bias capacitance values fell within the range of 0.01 to 0.02 pF. The

series resistance, R_s , was obtained in two ways. First, the deviation from the straight line plot of the $\log_e I_f$ vs V_f was used to estimate R_s and, secondly, the slope of the I-V curve measured in the vicinity of the 10 mA region was used as an estimate of R_s . Relatively good agreement was obtained between the two methods. The values of R_s fell within the range of 10 to 15 ohms. If the extremities of the grouping for R_s and C_o are used, the cutoff frequencies are expected to range from 500 GHz to 1500 GHz at zero bias.

Figure 3-5 presents the I-V characteristic of a typical varactor diode of the n-GaAs/Zn type. The slope parameter of 1.85 corresponds to that value expected for a true p-n junction. The slope parameter was measured for about ten separate junction samples. For all samples, η fell within the range 1.8 to 2.0.

C. Test Procedures and Data

The varactor diodes were all formed in open, crossed guide structures of the type previously used for variable resistance multipliers.⁽³⁻²⁴⁾ A photograph of the ADTEC M-9838 structure that was used is shown in Figure 3-6. The 0.005" thick semiconductor wafer was mounted on a 0.020" diameter pin which was inserted and adjusted in the RG-138/U waveguide by use of the differential screw control. The zinc whisker was mounted on a 0.020" diameter whisker carrier which was inserted through the RG-98/U waveguide and was held in place by the N-connector seen in Figure 3-6. The contact between the zinc wire and the GaAs wafer was made by adjustment of the differential screw, and this junction was pulse formed by pulsing through the N-connector.

The test procedures for measuring the harmonic generator conversion loss were straightforward. An E-H tuner was used to match the 70 GHz to the impedance of the RG-98/U input to the diode. A bolometer was used to detect the 140 GHz output. The bolometer was calibrated against a water calorimeter at a frequency of 140 GHz. The 70 GHz available power was also measured by use of a water calorimeter. The power measurements may have been in error due to mismatch between the bolometer and the fundamental power source and between the bolometer and the harmonic generator output. The latter case probably introduced the greater error resulting in a low value of output power and a high value of conversion loss. However, this conservative effect was counterbalanced to some degree by another type of error, which was that the harmonic generator output contained harmonics of order higher than second. This caused a somewhat higher power reading than the second harmonic alone would have.

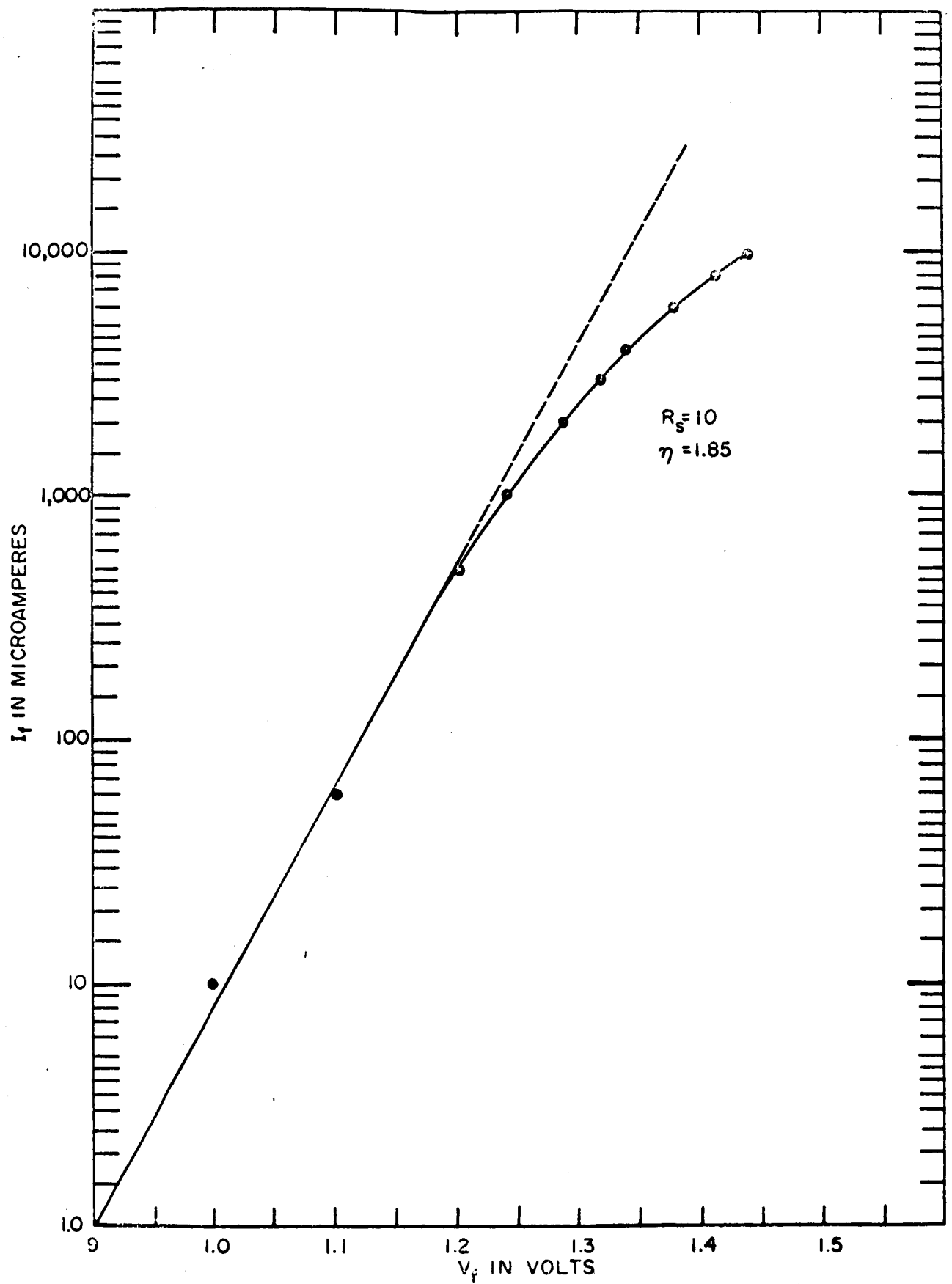


FIG. 3-5 - Current-Voltage Characteristics of a Typical Point-Contact Varactor n-GaAs ($\rho = 0.0422$ ohm-cm) Zinc Diode

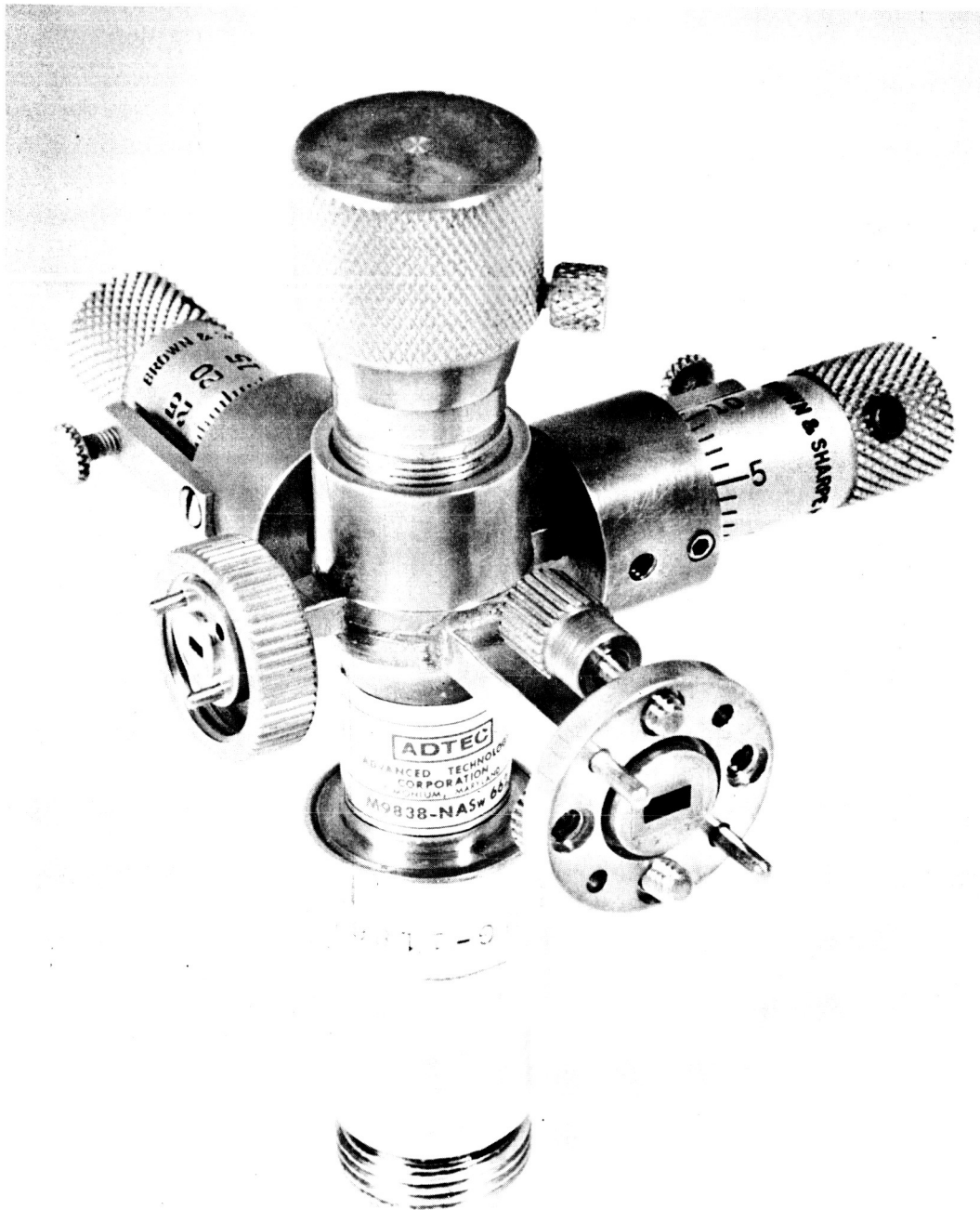


FIG. 3-6 - Crossed Waveguide Mount, Model M-9838

Figure 3-7 presents the measured data for five different varactor diodes and one variable resistance diode. The points representing diodes A and B show the variation of conversion efficiency with input peak power. Diode A is close to the average for the number of diodes formed. Diode B has a similar efficiency (about 20%) but a maximum 140 GHz output of 12 milliwatts of power. Data for three other diodes are shown which represent the variations that were obtained. The best efficiency (about 30%) occurred with an output power of 11.5 milliwatts. The highest power of 16 milliwatts was obtained with diode D at an efficiency of 20%. Diode E represents the performance needed for the solid state local oscillator, namely, 10 mW output with 25% efficiency. For a small fractional increase in input power above the greatest value plotted for each varactor diode complete dynamic detuning appeared to result with a sharp loss in output power. Diode R is plotted for comparison; it was the best variable resistance diode.

D. Data Analysis

The calculations in Subsection A2 can be modified slightly to see just how the measured data fit the predictions. With a breakdown voltage, V_B , of 16 volts and a zero bias junction capacitance, C_0 , of 0.015 pF, an estimated maximum input power of 28 milliwatts is obtained. It has been estimated by Morrison⁽³⁻²¹⁾ that by driving slightly into the forward region a maximum input power 3 to 5 times this input power may be permitted. Therefore, it is expected that this small capacitance of 0.015 pF can accept the level of power as shown in Figure 3-7. Using an R_s of 10 ohms, and neglecting the fact that this R_s may vary with reverse bias, one may estimate a zero bias f_{co} to be about 1000 GHz. The doubler efficiency that might be expected (assuming no other losses) would be 1.1 dB or 78%. In the crossed waveguide structure used, no reduction in height of either waveguide existed, so a high VSWR existed between the diode and the E-H tuner on the input side and most probably between the diode and the load on the output side. The VSWR between the E-H tuner and the diode on the input side was measured to be between 10:1 and 16:1. Therefore, it would not be unreasonable to assume 1.5 dB loss for the RG-138/U components (sliding short tuners, flange connections, and section of waveguide) and on the order of 3 dB loss for the E-H tuner, the two inches of resonant RG-98/U waveguide and the sliding short. This may account for the losses shown in Figure 3-7. If this is correct then by use of a suitable design in reduced height waveguide, the varactors herein discussed should give even more output power at 140 GHz or operate efficiently at even higher frequencies.

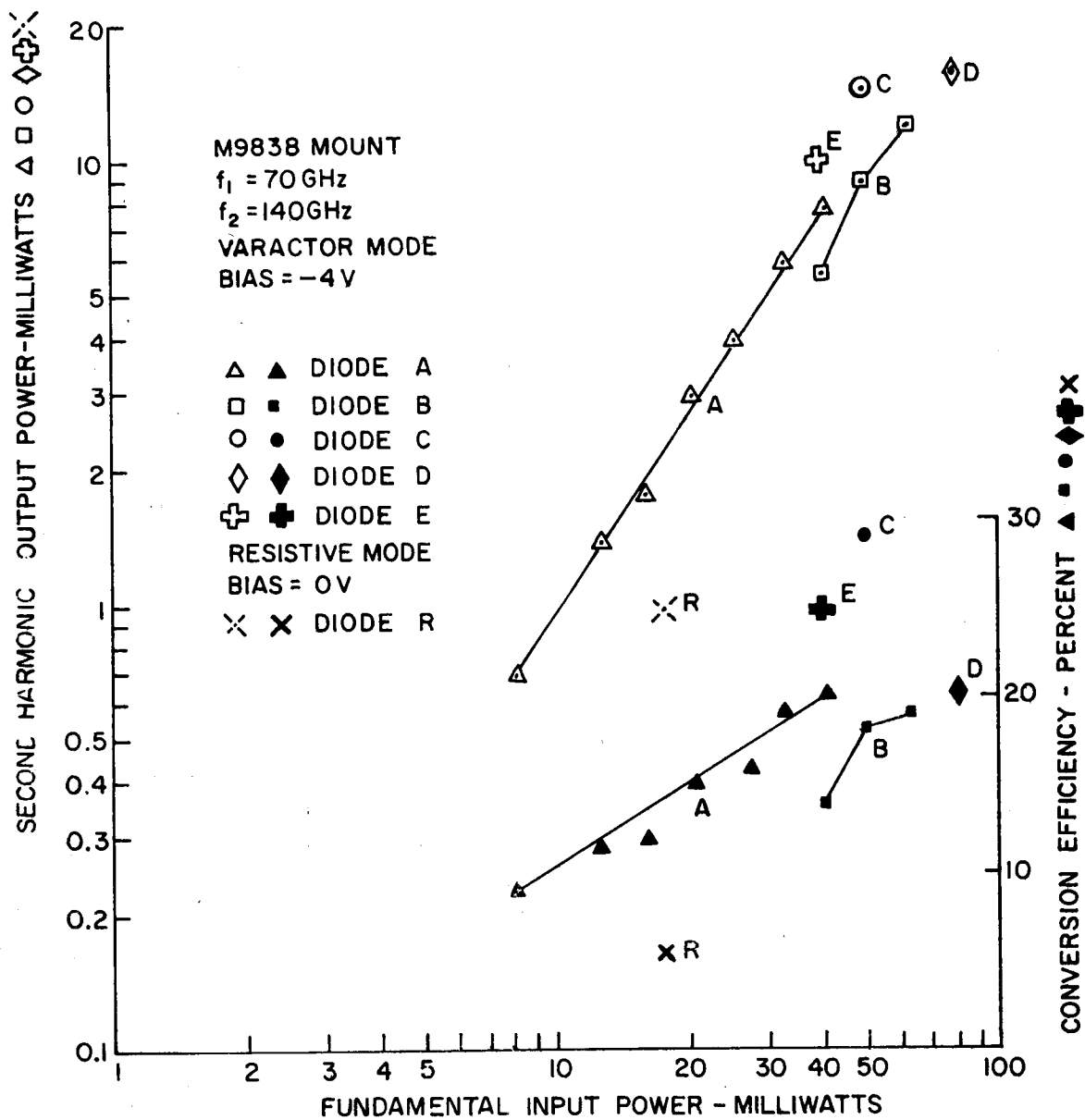


FIG. 3-7 - Power Output and Conversion Efficiency of a 140 GHz Second Harmonic Generator

It is of interest to calculate the series equivalent input and output resistances into which the guide impedance must transform if one is to begin to reach the quoted efficiencies. The input resistance, R_{in} , and output resistance, R_{out} , for the doubler are⁽³⁻¹⁹⁾

$$R_{in} = R_{out} = R_s \left[1 + (K_1 Q_1)^2 \right]^{\frac{1}{2}}, \quad (3-6)$$

where Q_1 is the quality factor at the dc bias point for the input frequency, K_1 is the modulation factor; values for K_1 are given in Reference (3-19). For the abrupt junction $K_1 = 0.286$. For $C_o = 0.015$ pF, $R_s = 10$ ohms, and an input frequency of 70 GHz, $Q_1 = 30$. Using this data and Equation 3-6 one obtains $R_{in} = R_{out} = 86$ ohms. This is a very low value of resistance and may present quite a matching problem. To match the relatively high waveguide impedance to such a low value may require the use of a matching transformer which (as noted above) could easily be more lossy than the diode multiplier itself.

E. Conclusions

On the basis of the above discussion and test data, it is believed that the idea of a solid state local oscillator at 2 mm wavelengths for a submillimeter mixer has been proven feasible. In particular, the construction of a 70 GHz to 140 GHz varactor doubler with an output of 10 mW for an input of 40 mW has been demonstrated. This was accomplished with an n-GaAs/Zn epitaxial diode fabricated in a crossed waveguide mount. The construction of a mount of electroformed reduced height waveguides should permit better impedance matching to the varactor diodes and thereby reduce the overall conversion loss. The analysis of the data showed that the diodes themselves have a potential efficiency of 78%. It is important that the potential exists for greater efficiency because lower conversion loss in the output stage reduces the requirement on power output from the preceeding stages.

The output level of 10 mW is adequate to serve as local oscillator power for the submillimeter mixer as has been reported in Section II. With 10 mW of LO drive a mixer conversion loss of 29.5 dB was measured. It is calculated that a radiometer built around these key components would have a minimum detectable temperature contrast of 7°K for a signal-to-noise ratio of one and an output integration time of 10 seconds.

The varactor diodes formed in the M-9838 mounts were open structures. The diode junction area was very small. The soft zinc whiskers did not bear on the semiconductor with the pressure that phosphor bronze whiskers do as used in variable resistance diodes. Therefore, these diodes did not have as great mechanical stability. There are three promising approaches to making these diodes rugged enough for a space satellite environment. One is to bond the zinc whisker to the semiconductor wafer with a drop of epoxy resin or other adhesive. A second approach is to seal the diode in an inert atmosphere, such as dry nitrogen, by using waveguide windows. The third possibility is to construct a diode as an integral unit on the end of a metal pin using a hollow ceramic cylinder as a housing. This has been called the "pin" diode structure. These methods of ruggedizing diodes are discussed in more detail in Section V of this report.

IV. SUBMILLIMETER SIGNAL SOURCE

A. Design and Construction

The crossed waveguide device used as a signal source in tests of the submillimeter harmonic mixer was the same device used for both the 7 X 75 GHz multiplier in Figures 2-7 and 2-9 and the 3X, 4X or 5X 140 GHz multiplier in Figure 2-8. A photograph of this device is reproduced in Figure 4-1. Although a submillimeter harmonic generator is not a part of a submillimeter radiometer, this device has been useful in testing the harmonic mixer being developed for the radiometer and in developing design information on diodes and waveguide structures which may be applied to radiometer components.

The submillimeter signal source was designed to have several specific features. The two crossed waveguides were inserted in a block and the waveguide flanges machined in the body of the block. The fundamental input waveguide has the inside dimensions of RG-99/U, which is recommended for 60 to 90 GHz. The flange was machined to mate with the UG-387/U type. The harmonic waveguide is smaller than any standard waveguide and has inside dimensions of 0.020" x 0.010". This size has a first order mode cutoff at 295 GHz and a second order mode cutoff of 590 GHz. It was chosen as about the largest size which would guide a 560 GHz wave in a single mode. This was desirable to minimize attenuation. Its flange is the compact FXR type which was chosen for all waveguides smaller than RG-99/U used on this project. The harmonic waveguide has been made as short as possible, also to minimize attenuation. A taper transition was built to connect it to RG-138/U (0.080" x 0.040"). The transition may be used as a test horn if desired. The harmonic generator was built with the diode in the smaller waveguide in order that the harmonic be established in the first order TE_{10} mode.

The submillimeter signal source was constructed so that the diode could be sealed in dry nitrogen. In general, a crossed waveguide structure has six openings: four waveguide ports, a coaxial port, and a mounting hole for a differential drive. Unlike a harmonic mixer, there are no microwave impedance matching requirements on the coaxial port of this device. A simple miniature connector was used which could be sealed in place. The whisker carrier was designed to be sealed in place also, after being adjusted to achieve a good diode by means of an external differential drive. The harmonic generator was mounted in a jig which also held the differential drive while sealing the whisker carrier. The waveguide ports were designed to be sealed with

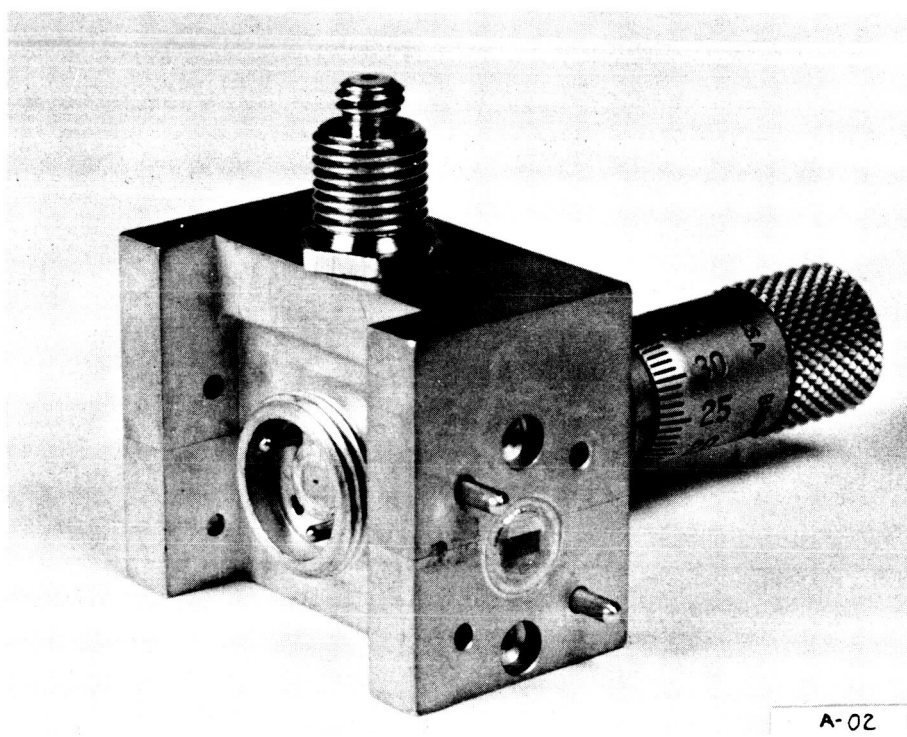


FIG. 4-1 - Submillimeter Harmonic Generator

dielectric windows. The micrometer-driven tuners were made as short waveguide sections which attach externally to the sealed harmonic generator body. Only the tuner for the signal guide is shown in Figure 2-8.

Considerable experimentation was necessary to develop the technique of cementing windows on the four waveguide ports. Mica windows one mil thick were cemented to the larger waveguide. Mica could not be applied to the smaller waveguide ports without some cement flowing into the waveguide. Teflon tape two mils thick with a one mil silicone adhesive backing was applied to the smaller waveguide ports. It was held in place by the mating waveguide tuner on one end and by the waveguide taper transition on the other end. These windows withstood pressures up to 46 psig. The insertion loss of one teflon tape window was measured to be 1 dB at 320 GHz. However, the harmonic generator was not used as a sealed device during this project. Before the sealing techniques could be fully developed, the device was pressed into service as a signal source to test the submillimeter mixer. It worked well in that service although it occasionally required the formation of a new diode junction. This was most easily performed with the diode left unsealed.

The diode junctions consisted of a wafer of n-GaAs, 0.008" dia. by about 0.004" thick, and an electrolytically-pointed one mil diameter whisker. Phosphor bronze was used in most cases as the whisker metal, but some of the later diodes were made with gallium-doped gold wire. The GaAs crystals were prepared from the same sample (Code No. K-4) as was used for the submillimeter mixer. A single attempt was made to use epitaxial gallium arsenide and zinc whiskers in the submillimeter multiplier, but it failed because the gallium arsenide was damaged during the mounting process. Since this device was a tool, not an end product of itself, efforts to develop maximum output were limited.

B. Test Procedures and Data

With a bolometer connected directly to the multiplier, the total harmonic output power above 295 GHz was measured for various fundamental frequencies. Since the power in each successively higher harmonic from a harmonic generator is typically reduced by several decibels, it is reasonable to assume that the power measured was primarily at the lowest harmonic above 295 GHz. This was verified by measuring the guide wavelength with the harmonic tuner.

A diode was formed in the submillimeter harmonic generator, and 90 GHz power applied. An output was detected in the harmonic guide which was determined by measuring the guide wavelength to be 360 GHz or the fourth harmonic. The transducer loss was about 30 dB. On another occasion a signal at 75 GHz was applied, and the output was identified as 300 GHz or the fourth harmonic. The conversion loss was 30 dB. Later, a new diode junction was made and the fourth harmonic at 300 GHz was not detected but the fifth harmonic at 375 GHz was identified. The fourth harmonic is believed to have been minimized by forming and tuning the diode to optimize the odd harmonics. The fifth harmonic transducer conversion loss was measured to be about 40 dB on that occasion. In other measurements the fundamental input was set at 80 GHz; the fourth harmonic conversion loss to 320 GHz was 33.5 to 37.5 dB. Measurements of the seventh harmonic of 80 GHz were obtained by using the high pass filter described in Section II, which had a cutoff frequency of 500 GHz.

In Table 4-1 are listed the best conversion loss measured for each combination of input frequency and harmonic order of multiplication. The data are listed in order of increasing output frequency. With the exception of the fourth and seventh harmonics of 80 GHz, each measurement was made with a different diode junction. In each case the tuning and diode operating conditions were optimized for the particular output desired. The input power was set between +9 and +13 dBm. In some cases +9 dBm was the maximum which could be applied to the diode without causing an unstable output; in the other cases +13 dBm was the maximum available fundamental power. The transducer conversion losses are seen to increase with output frequency.

The data of Table 4-1 may be compared to the curves of typical harmonic generation conversion loss for various harmonic orders as drawn in Figure 3-1. The starred values in Table 4-1 were used together with data from other sources in preparing Figure 3-1. The values measured later for fourth harmonic generation of 320 and 360 GHz appear to be 3 or 4 dB higher than expected from extrapolation of the quadrupler curve of Figure 3-1. Similarly, the fifth harmonic generation of 375 GHz appears to be about 4 dB more lossy than expected. The conversion loss of the 3X 140 GHz multiplier is 14 dB poorer than the value of the tripler curve of Figure 3-1 which was based upon data measured for another device designed specifically for this purpose. There are two possible reasons for the poorer conversion loss. First, the input waveguide was not limited to a single mode of propagation at 140 GHz; there may have been losses in conversion to a higher mode. Second, the output waveguide probably had abnormally high dissipative

Fundamental Input Frequency	Harmonic Order	Output Frequency	Input Power Level	Output Power Level	Transducer Conversion Loss	Conversion Loss Predicted from Figure 3-1
GHz		GHz	dBm	dBm	dB	dB
75	4	300	+9	-21	30*	30
80	4	320	+12.5	-21	33.5	30.5
70	5	350	+9	-26	35*	35
90	4	360	+13	-22	35	31
75	5	375	+9	-31	40	36
140	3	420	+10	-32	42	28
80	7	560	+12.5	-48	60.5	
140	5	700	+10	S/N = 15 dB with submillimeter mixer		

* Data that was used as part of the plots of Figure 3-1.

TABLE 4-1
Measured Data for Submillimeter Harmonic Generators

losses resulting from some adhesive material which had leaked into the waveguide during the experimentation with seals and which could not be cleaned out.

Values of conversion loss as a function of external bias voltage were not recorded. It was found with this multiplier, as with the devices tested in the solid state local oscillator development, that external bias did not increase the output. The output VSWR was not measured for lack of instrumentation at the frequencies involved. The input VSWR was not measured, but the input was matched to the source of fundamental power with an E-H tuner.

Most of the diodes were operated in the variable resistance mode by using a low load resistance, say 1000 ohms, on the coaxial port in which case about 2 mA of self bias current was generated. Some diodes that were formed with gallium-doped gold whiskers required a 20 K ohm load for minimum conversion loss and generated 0.3 mA of self bias current. It is thought that these diodes were operating primarily in a variable reactance mode. The two modes gave about the same conversion loss. As was explained in Section III describing the SSLO, the varactor mode is capable of better performance, but in the submillimeter multiplier no provision was made for efficient matching into a varactor diode and its superior potential could not be realized.

C. Conclusions

A versatile submillimeter harmonic generator was constructed and was used as a signal source for measuring the submillimeter harmonic mixer conversion loss. It was operated with input frequencies ranging from 70 to 140 GHz. Harmonics such as the third, fourth, fifth, and seventh at frequencies from 300 to 700 GHz were identified with frequency selective detection schemes. The output of the harmonic generator itself contained all harmonics of the input which occurred above 295 GHz. The measured conversion loss was poorer than expected by 3 or 4 dB in three cases and by 14 dB in one case. The latter case was that of generation of 560 GHz by tripling from 140 GHz. It is believed that dissipative losses in the output waveguide caused extra loss at 560 GHz for this case and for 7X 80 GHz operation also. However, the multiplier was used successfully to measure the submillimeter harmonic mixer performance.

Several diode junctions were formed in the submillimeter harmonic generator. They were usually operated in the variable resistance mode but some were operated in the varactor mode and were equally successful. The potential for better performance is believed greater with

varactor mode operation. The most likely means of realizing some of this potential would be to design a structure with reduced height waveguides in order to obtain a better impedance match to a diode in the varactor mode.

The submillimeter harmonic generator was designed to be sealed in an atmosphere of dry nitrogen and was used to develop techniques for accomplishing this. It was not used as a sealed device in the mixer measurements, however, because it was needed before an electrically-suitable sealed diode was obtained. The structure was successfully sealed using mica and teflon tape windows and withstood pressures up to 46 psig. The sealing methods may be applied to future harmonic mixers and multipliers in solid state local oscillators.

V DIODE DESIGN FOR SPACE ENVIRONMENT

A. General Requirements

The components of a space-borne radiometer must withstand environmental conditions much more severe than the operating conditions for an experimental laboratory radiometer. The mixer diode and solid state local oscillator diodes are essential parts of key radiometer components. They must meet the following general requirements:

- 1) The system must meet standards of reliability for a useful operating lifetime with sufficient electrical performance to permit a worthwhile experiment.
- 2) The system must withstand the conditions of a vehicle launch environment including preparations for launch.
- 3) The system must withstand the conditions of the space environment.

A number of diode fabrication techniques were investigated which have shown promise of helping to meet the above requirements. The basic idea behind all of the techniques was to make more rugged diodes by sealing them from the atmosphere. Methods of sealing the waveguide structure and methods of sealing the diode junction locally were investigated. Specific techniques are described below.

B. Specific Techniques

1. Hermetically Sealed Diode Mounts

Welded junction diodes of phosphor-bronze and gallium arsenide have been built at ADTEC and sealed in fixed-tuned single waveguide mounts for use as fundamental mixers from 35 to 170 GHz. Models with five different waveguide sizes are illustrated in Figure 5-1. Each was sealed at the coaxial IF port with epoxy resin and at the waveguide port with a mica window cemented to the brass body. The opposite end of each waveguide has a fixed short epoxied in place. The diodes in the mounts were formed with an electrolytically-pointed one mil whisker wire and a semiconductor wafer in the same manner as the diodes in the submillimeter harmonic mixer and harmonic generator. An external differential drive was used with these mounts as with the submillimeter harmonic generator in order to position the whisker carrier which was sealed by epoxying in place after forming the junction diode. A subtle but possibly significant difference in the formation of diodes in the submillimeter devices is that the governing electrical performance

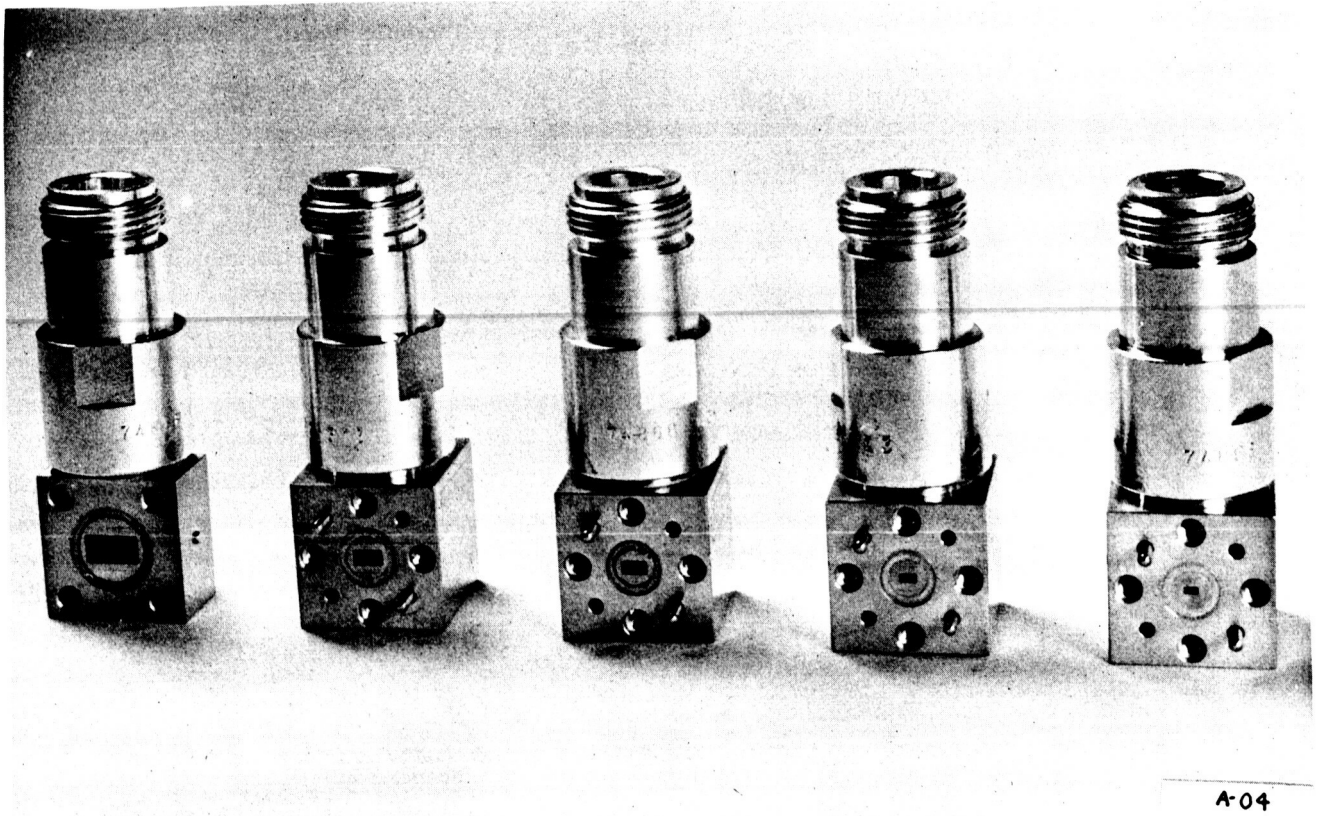


FIG. 5-1 - Hermetically Sealed Fundamental Mixers (26 to 170 GHz)

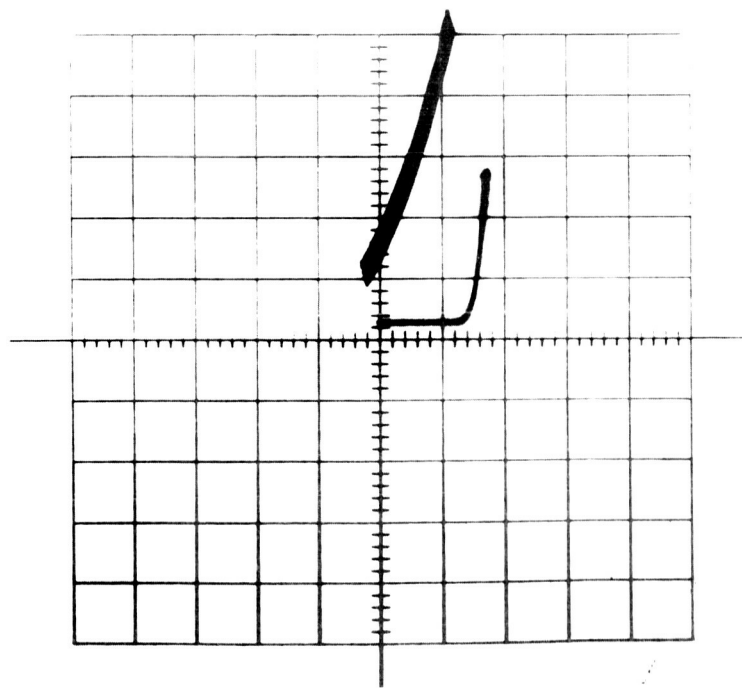
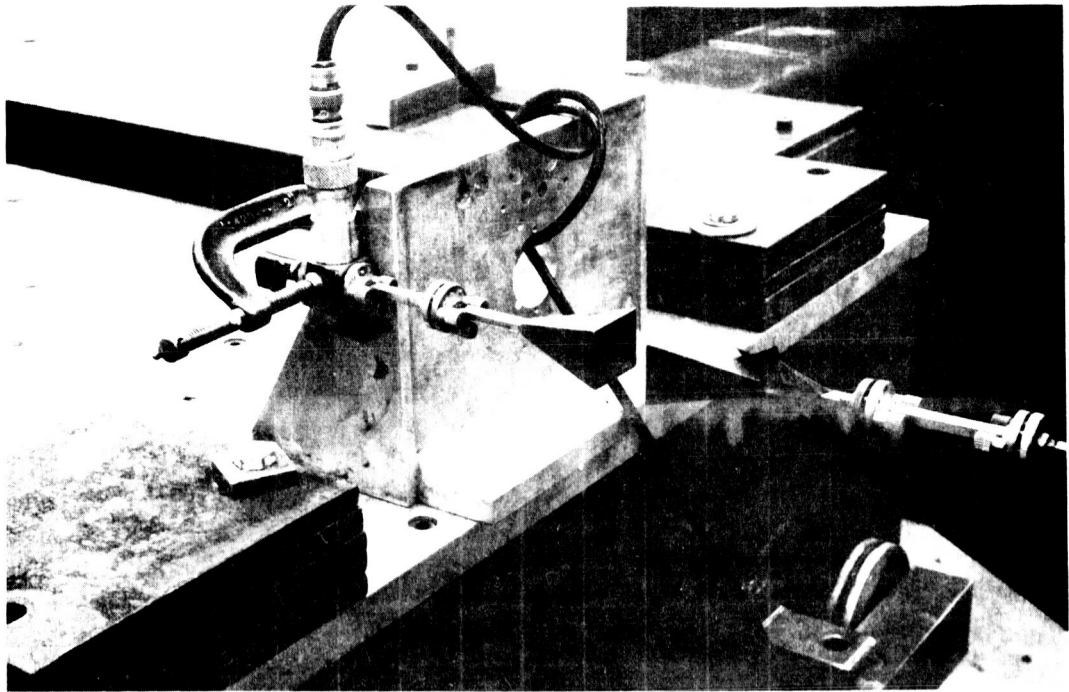
may result from a lighter contact than needed for the millimeter wave devices.

The diode mixers of Figure 5-1 were sealed in an atmosphere of dry nitrogen. Over 100 diode mixers of this type have been constructed. Unlike diodes exposed to the atmosphere in unsealed structures, these diodes do not appear to deteriorate or fail due to aging. Several (10) diodes have been kept in the laboratory for over a year with little or no change in their electrical properties.

There have been some environmental tests made on this type of sealed diode which are most encouraging. For example, a group of 94 GHz mixers has been tested⁽⁵⁻¹⁾ over a temperature range from -25°C to $+70^{\circ}\text{C}$ without any degradation in performance (one diode failed at $+75^{\circ}\text{C}$). These diodes were cycled over this temperature range a number of times to insure that there would be no gradual fatigue in either the welded junction or the mica window waveguide bond.

Vibration tests have been run on a number of diodes with only one failure.⁽⁵⁻²⁾ Figure 5-2 shows the diode mounted on the vibration table with a waveguide horn so that the diode could be illuminated with RF energy for continuous monitoring. The monitoring technique consisted of the same procedure that was used for initially forming the diode, i.e., a dual trace presentation was maintained on a scope while the diode was being vibrated in each axis. The bottom portion of Figure 5-2 shows the dual trace presentation that was observed. The lower curve is the usual I-V curve of the gallium arsenide diode; it manifests the characteristic sharp break of forward voltage at approximately 0.7 volts. The upper curve represents RF rectification of the diode. It can be observed that this upper curve has a slight blur, which was caused by the vibration of the horn that received the RF energy. It is apparent that the blurred upper trace does not represent a change in the diode characteristics since the lower trace, which is not a function of the RF power level, remained steady. The diodes were subjected to vibration at 10 to 60 cps on each of three mutually perpendicular axes in both a continuous changing frequency mode and a step changing frequency mode. Each diode was tested for approximately 15 minutes with a maximum force of 8.64 G's at 30 cps and 7 G's at 60 cps. These vibration tests were made at levels comparable to the low frequency vibration requirements in OGO experiment design qualification tests.⁽⁵⁻³⁾

The submillimeter second harmonic mixer and multipliers in the solid state local oscillator of a satellite-borne radiometer will be crossed-waveguide devices. Development models will have a tuner in each waveguide while the final models may be fixed tuned. The



A-17

FIG. 5-2 - Vibration Testing of 94 GHz Mixer Diode

methods of sealing the fundamental mixers shown in Figure 5-1 may be extended to seal crossed-waveguide devices with the added complications of more ports to seal and more time required for rework when necessary for experimental devices.

Section IV of this report discusses experimentation with a crossed-waveguide submillimeter harmonic generator designed to be sealed. The principal problem in sealing this structure has been the attachment of mica windows to the smaller waveguide without spreading any cement on the inner surfaces of the waveguide. A partial solution to this problem was the use of adhesive-backed teflon windows. A different mount design has been suggested which would eliminate the losses of the windows between the diode and the tuners. The present contacting tuner shorts would be replaced with an appropriately dimensioned bellows sealed with a short on the far end. The mechanism for actuating the bellows would be external to the sealed region.

While fabrication details remain to be worked out, it appears that sealed crossed-waveguide harmonic mixer and harmonic generators can be built. The techniques are applicable to structures designed for different frequency bands and for different types of diodes. The successful operation of sealed fundamental mixers under conditions of varying temperature and vibration is evidence that sealed structures will meet, in part or completely, the requirements for use in a space-borne radiometer.

2. Pin Diode Structure

An alternate or supplemental method of ruggedizing the millimeter and submillimeter diodes described herein may be mentioned.* The technique is to pot the entire diode on the end of a 40 mil diameter metal pin. Hence, it may be called a "pin" diode. Figure 5-3 is a cross-section drawing of the structure, and Figure 5-4 shows photographs of an experimental unit. After the diode is formed on the end of a pin, it is potted in epoxy resin within a ceramic cylinder 0.010" high x 0.030" O.D. and 0.012" I.D. The whisker is in contact with the metallized top of the cylinder and a metal cap. The diode is inserted into the circuit through the waveguide wall and makes contact via a low inductance post. This construction gives the diode a very high self-resonant frequency.

*The technique was developed by a member of the staff of this laboratory during his prior employment at the Carlyle Barton Laboratory of The Johns Hopkins University under Contract NAS 5-3546 for NASA, Goddard Space Flight Center.

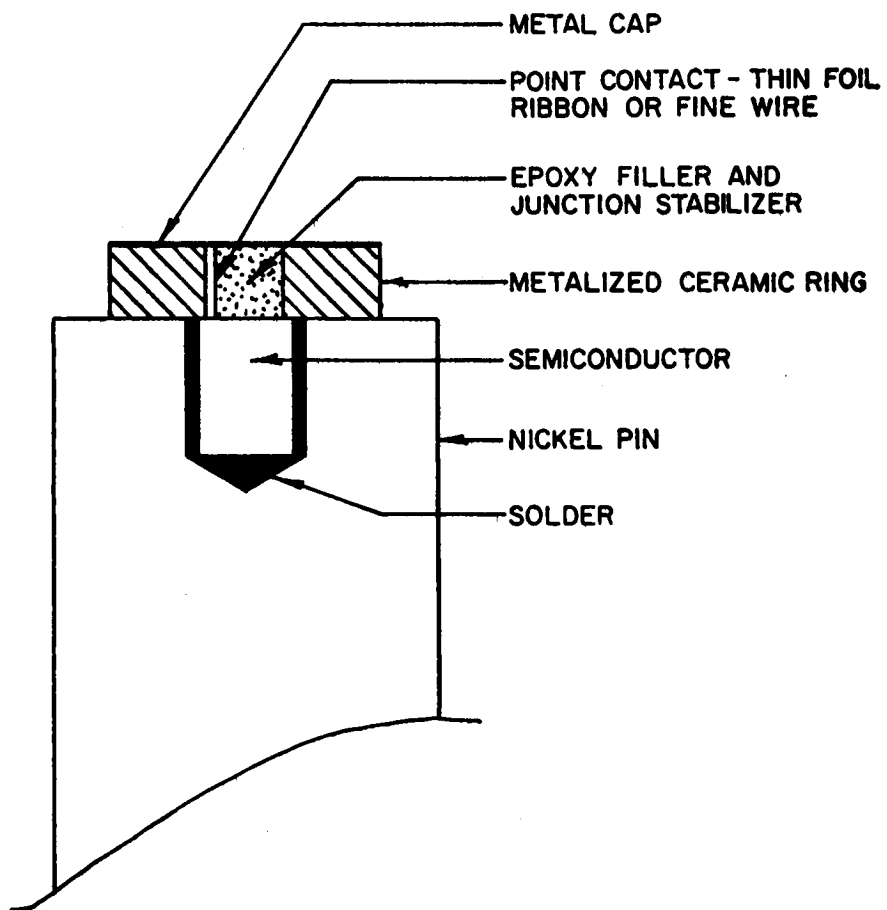
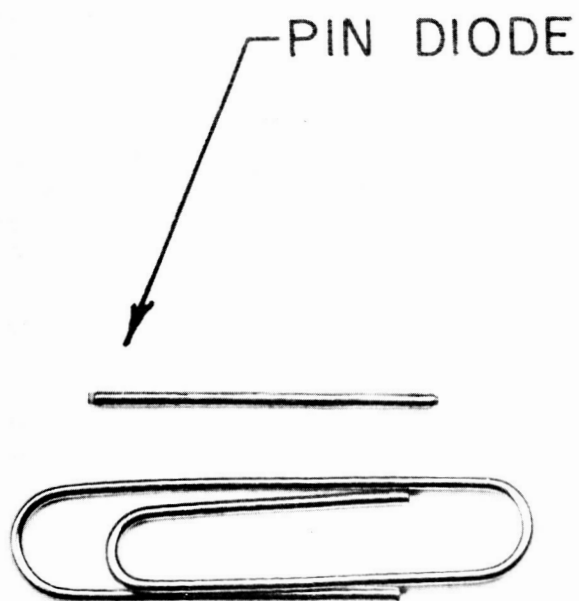


FIG. 5-3 - Millimeter Wave Diode (Pin Construction)



Magnified View of Pin Diode Showing Metal Cap,
Ceramic Ring, and Nickel Pin



A-18

Relative Size of Pin Diode (the Nickel Support Post
Length is Optional)

FIG. 5-4 - Millimeter Wave Back Diode (Pin Construction)

The support given the whisker and the sealing of the diode with epoxy makes the pin diode inherently stable. The limiting environmental conditions it will withstand will be determined by test.

The pin diode configuration has been proven electrically with the fabrication and test of germanium and gallium arsenide back diodes as video detectors at both 25 and 70 GHz. Sensitivity was as good at 70 GHz as at the lower frequency indicating that the packaged diode has not begun to reach its high frequency cutoff. These diodes were as sensitive as Burrus reported⁽⁵⁻⁴⁾ for unpotted junctions. This performance leads to the belief that the pin diode structure, possibly in a smaller size, can be applied to 140 GHz harmonic generators with satisfactory electrical performance. The epoxy sealing but not the ceramic housing may be applicable to the design of the submillimeter harmonic mixer. The pin diode fabrication process is still in development for its original application to back diode detectors and will require modification for these proposed applications.

3. Epoxied Diodes

Some work on epoxied diodes is reported in Appendix C. Diodes were fabricated which could be installed and removed any number of times. However, the diodes had poor RF characteristics. Different epoxy resins and techniques for monitoring RF properties while forming the diodes need to be investigated. No environmental tests of the epoxied diode junction have been made, but it is a mechanically stronger junction and may be a means of meeting the requirements of a space-borne radiometer.

C. Conclusions

Several techniques for making diodes more rugged by sealing them from the atmosphere have been discussed. These techniques have been developed on this and other programs for diodes of various types. Similar methods of fabrication can be used for submillimeter mixer diodes and for varactor harmonic generator diodes for solid state local oscillators. Now that the electrical feasibility of these key radiometer components has been proven it is recommended that the next phase of this program include the application of sealing methods to the diodes for the key components. Environmental tests of the diode devices would be an essential part of the next phase in order to determine that the general requirements listed at the beginning of this section would be met.

VI. CONCLUSIONS AND RECOMMENDATIONS

A. Feasibility of Submillimeter Radiometer

Experiments with semiconductor diode harmonic mixers and multipliers have demonstrated that fourth harmonic mixing at 0.5 mm wavelength can be achieved with sensitivity useful for radio astronomical experiments from a space platform and with a local oscillator power requirement that can be met with a solid state local oscillator. In tests at 560 GHz a mixer conversion loss of 29.5 dB was measured for an LO drive power of 10 mW. In tests of a 140 GHz harmonic generator an output power of 10 mW was measured for an input power of 40 mW (25% efficiency). With devices having these characteristics a submillimeter radiometer receiver such as is given in block diagram form in Figure 1-1 would be built. It would have a calculated minimum detectable temperature, $(\Delta T)_{\min}$, of 7°K for a signal-to-noise ratio of unity. This calculated sensitivity is based upon: RF losses of 3 dB, mixer diode noise ratio of 2, IF noise figure of 4 dB, IF bandwidth of 1.6 GHz, and output integration time of 10 seconds. These receiver component characteristics are realistic values which have been achieved.

A minimum detectable temperature of 7°K is a value which would permit useful observations to be made with a space-borne submillimeter radiometer. If the sun is assumed to have a brightness temperature on the order of 7000°K at 0.5 mm as has been measured at 4.3 mm⁽¹⁻⁵⁾ and is assumed to fill the main beam of a radiometer antenna having negligible sidelobes, then signal-to-noise ratios on the order of 30 dB should be attainable. If Mars or the Earth's moon is the object of interest and fills the main beam of the antenna, then a brightness temperature on the order of 200°K may be assumed⁽¹⁻⁷⁾ and signal-to-noise ratios on the order of 14 dB are possible.

Although these are simplified examples, they should not be considered too optimistic as they are based upon the performance of first models of the harmonic mixer and the harmonic generator. The next model of the 140 GHz harmonic generator should be built in reduced height waveguide, because with such a structure there would be a strong probability of allowing greater output power with even better efficiency. Measurements of the mixer have already shown that a 6 dB increase in LO power can result in a 5.5 dB reduction in mixer conversion loss. The calculated $(\Delta T)_{\min}$ would then be 2°K . Even for the same LO power, additional mixer improvement is expected to result from extended experience in working with the mixer and the gradual evolution of the structure through prototype to flight model.

Of course, some mixer sensitivity and harmonic generator output may have to be given up in designing the flight models of these devices to withstand the environmental conditions of launch and space flight. However, it is believed that flight models will be built and assembled in an all solid state submillimeter radiometer which will have better than 7°K minimum detectable temperature for unity signal-to-noise ratio and 10 seconds integration time.

Certain options are available to the designer of a submillimeter radiometer and should be considered in the light of specific system requirements. Any increase in the output integration time of the radiometer circuits will result in an increase in sensitivity as the square root of the ratio of the integration times. The mixer and harmonic generator designs and diode fabrication techniques may be applied at a variety of frequencies. Radiometers could be built to operate at any frequency from 300 to 700 GHz with generally comparable performance. Somewhat poorer sensitivity would result at the higher frequencies but the feasibility of using waveguide devices up to 700 GHz has been demonstrated by operation of the signal source and mixer in a test setup on this project.

The proposed radiometer would have a frequency resolution equal to twice the high end of the IF band, i.e., 2×4 GHz or 8 GHz. This is 1.4% of the nominal center frequency, 560 GHz. The narrow band frequency response is accomplished without narrow band filters, whose construction is problematical at submillimeter wavelengths but which would be required to give such resolution with thermal detection radiometers.

Such frequency resolution would be very useful in identifying absorption peaks in the emission spectra of astronomical bodies. Multiple radiometers could be built using the design techniques developed at 560 GHz to operate at several frequencies of interest in the submillimeter spectrum. A radiometer with multiple RF heads could also be constructed to achieve a diversity of measurements with fewer instrument components. A single tunable RF head is possible in principle but would not be practical in an unmanned space vehicle without much development to reduce the number of tuning adjustments that present state-of-the-art components would require.

The design of an antenna for a submillimeter satellite radiometer will be a straightforward engineering problem. Quasi-optical types such as the parabolic reflector, the horn-paraboloid, or the horn-lens combination are suitable. An example of the latter is the antenna system used in the 600 GHz radiometer previously referenced.⁽¹⁻²⁾ It consists

of a focal horn built in the mixer body and a 10-inch diameter teflon lens having a 30-inch focal length. Calculating the antenna beamwidth for this aperture size with an allowance for the amplitude taper due to the horn antenna receiving pattern gives a half-power beamwidth of one-seventh degree or 2.5 milliradians. The tenth-power beamwidth is estimated to be 3.5 milliradians. The effectiveness of this antenna beamwidth is seen, for example, by the fact that Mars would be within the tenth-power beamwidth at ranges of 1,200,000 miles or less. Closer ranges or larger lenses would permit resolution of areas of Mars. In summary, the radiometer antenna would be designed with consideration for the vehicle trajectory and the body to be observed.

The estimates of radiometer sensitivity which have been presented have been based upon an allowance of 3 dB for RF losses such as lens transmission and reflection loss, focal horn side lobes, and circulator or chopper insertion loss. The teflon lens mentioned above introduced 1.6 dB loss in the radiometer. The dissipative losses in the focal horn and input waveguide are included in the measured values of mixer conversion loss upon which the estimate of radiometer noise figure and minimum detectable temperature were based.

Measurements of the conversion loss of the submillimeter mixer were made possible by the development of a versatile submillimeter harmonic generator. Its conversion losses for various input frequencies and harmonic orders have been listed in Table 4-1. There is evidence that the values quoted could be improved, especially with the construction of a reduced height waveguide mount. However, the present device was used with the submillimeter mixer in test setups having a dynamic range of 42 dB at 420 GHz, 30 dB at 560 GHz and 15 dB at 700 GHz. Such test setups can be useful for measurements of materials and components.

B. The Next Phase of Development

In order to achieve operational, space-borne submillimeter wavelength receivers and radiometers, further work is necessary to assure that the most critical components, the mixer and the solid state local oscillator, can survive launching conditions and operate reliably in a space environment for a sufficient length of time. It is recommended that the following work be accomplished in the next phase:

- 1) Perform environmental tests on the harmonic mixer during the development and testing of this unit. The results of the environmental tests will determine if structural changes are required and, if so, the electrical tests will be repeated on the modified harmonic mixer.

2) Continue the development and electrical testing of the latter stages of the solid state source utilizing varactor harmonic generators (output at a wavelength of 2 mm) concurrently with the performance of environmental tests on the SSLO. The objective is to develop rugged units with sufficient power at the output so as to satisfy the LO power requirements of the fourth harmonic mixer. High cutoff frequency, integral waveguide or pin structure varactors should be developed and tested.

The estimate of a 7°K minimum detectable temperature which should be attainable in a radiometer with the present performance of the 0.5 millimeter harmonic mixer is sufficiently low that the mixer could be tested in a laboratory radiometer looking at a thermal source. Thermal sources of high emissivity which can be heated to 170°C have been built and used at this laboratory to measure submillimeter harmonic mixing radiometer performance.⁽¹⁻²⁾ A thermal source capable of temperatures up to 1000°C was recently obtained. It is a flat panel, 16 cm by 36 cm, of unglazed ceramic with a heater element imbedded in it. With this source close to a small horn and with an optical chopper a radiometer could be operated with a dynamic range of about 20 dB. It is recommended that such a radiometer test bed be constructed to use in testing other components in the next and later phases of the development of a satellite radiometer.

An improved test method as described above would lead to improved diode performance. It would permit rapid and definitive measurement of the effects of changes in diode materials and forming techniques. It has not yet been possible to make the repeated tests which are necessary to develop optimum diodes. It is expected that improved mixer diodes will also result from a continuation of experimentation with lower frequency mixers and the application of resultant techniques to the submillimeter mixer. It is only at lower frequencies that slotted lines are available to measure impedances as a guide to structure optimization. Continued development of fabrication techniques which would permit construction of improved submillimeter mixers is also recommended.

C. Later Phases

Thus far, this engineering report has set forth the development of the critical components of a satellite submillimeter radiometer to meet electrical requirements. Methods have been described for modifying these components to meet the expected environmental conditions. In this section the additional work will be described that will be necessary

to provide a flight model and a flight spare of a submillimeter satellite radiometer. Additional work of a similar nature could also be accomplished to provide a submillimeter communications receiver based upon the component development reported herein.

The first step would be to prepare a detailed system design and specific requirements for the other radiometer components. Commercial items meeting these requirements would be selected where possible. For planning purposes the system design may be considered to include the components shown in the block diagram of Figure 1-1. Some components such as the antenna assembly, ferrite switch or mechanical chopper, driver stage of the solid state local oscillator, and signal processing circuits may have to be developed to meet electrical or environmental requirements or both. Purchased components not tested by the vendor and ADTEC-developed components would be qualified for the anticipated environment.

As qualified prototype components become available they would be substituted in the laboratory radiometer setup thereby converting it to a breadboard of the satellite system. Minimum detectable temperature and other system properties would be measured. Problems of electromagnetic compatibility would be resolved. Design changes necessary to minimize size and weight would be established. Component performance records would be maintained throughout breadboard tests to serve as a basis for initial reliability evaluation of the components and system.

The breadboard radiometer, in turn, would be converted to a prototype of the flight radiometer by installing any components redesigned for the flight experiment. When it has the final form and is truly a prototype system it would be given the environmental design qualification tests as applicable for the intended mission. Redesign and retest would be accomplished as necessary. At this time detailed consideration would be given to the integration of the radiometer experiment into a complete experimental mission. Of particular importance would be the plans for the antenna installation and for handling the output data. Operation of the prototype might be continued after the design qualification tests to obtain lifetime reliability data.

Concurrently with the assembly and test of the breadboard and prototype radiometers, the requirements for auxiliary test equipment at the launch site would be established and a set of such equipment assembled. Existing available government-owned equipment would be used where possible, items purchased commercially, or, if necessary, special items would be designed and constructed.

After prototype qualification tests one or more flight models and one or more flight spares would be assembled using completely new components. Each system would pass environmental acceptance tests by itself or as a part of the vehicle or both as required by the contracting agency. Field support during preparation for flight, launch, and data reception would be required.

Appendix A

COMPARISON OF SUBMILLIMETER RADIATION DETECTORS

Various methods have been used by different investigators to detect submillimeter radiation. A survey of the literature has been made and references studied. The most sensitive detectors have been selected for comparison with superheterodyne circuits using point-contact diodes in harmonic mixers. A photoconductive detector and a bolometric thermal detector are discussed in this appendix.

Submillimeter radiation has been detected by Putley using an indium antimonide photoconductive detector cooled to 1.5° K. (A1, A2, A3) Meredith and Warner^(A4) also report some of Putley's data. Values of noise equivalent power (NEP) for an output bandwidth of 1 Hz are reported as follows:

		<u>Reference</u>
150 GHz	1×10^{-12} W	Table 1 of (A4)
300 GHz	5×10^{-12} W	(A3)
600 GHz	10×10^{-12} W	(A3)
1500 GHz	20×10^{-12} W	(A3)

Putley states that this performance is limited by the noise of the associated amplifier, but he does not give the amplifier's noise figure. The NEP for an "ideal" photoconductive detector with a 10% bandwidth centered at 600 GHz Putley gives as 1.84×10^{-14} W. The "ideal" detector is one in which signal fluctuations are much larger than internal detector and amplifier noise.

Values of NEP are not generally useful for comparing radiometers of different types for applications where the energy source to be detected radiates over a broad frequency band. The total power reaching the detector in any radiometer will depend upon the input pass band; hence, the sensitivity will depend upon the input bandwidth.

Putley's detector is inherently a broadband device which responds to energy at all frequencies in the submillimeter band on up through the infrared. It is a difficult design problem to devise filters which will limit the frequency band reaching the detector in applications where frequency resolution is required. Putley^(A1) reported that he used

"a black paper filter at room temperature and a black polythene filter in the helium to remove short-wave radiation." No bandwidth is stated. Meredith and Warner^(A4) assume a 10% bandwidth in order to calculate minimum detectable temperature changes. However, their calculations assume no loss of performance due to the insertion loss of a 10% bandpass filter. Also, the construction of such a filter centered on any desired frequency may be impractical.

The superheterodyne radiometer has a relatively narrow RF input band, namely twice the bandwidth of the IF amplifiers. The input bandwidth of a submillimeter superheterodyne radiometer may be 8 GHz, which is 1.4% at 600 GHz. The minimum detectable temperature change, $(\Delta T)_{\min}$, for a superheterodyne receiver is calculated by the formula:

$$(\Delta T)_{\min} = a F_{\text{RAD}} T_o \sqrt{\frac{B_o}{B_{\text{IF}}}} \quad (\text{A1})$$

where

a is a constant dependent upon circuit details and is on the order of 1

F_{RAD} is the receiver noise figure (two sidebands)

T_o is 290° K, reference temperature for F_{RAD}

B_o is the output bandwidth (equals the reciprocal of the output integration time)

B_{IF} is the IF bandwidth

This definition for $(\Delta T)_{\min}$ includes the condition that the receiver output signal-to-noise ratio equals one.

For a direct detection receiver such as Putley's consider the minimum detectable power as occurring for a signal-to-noise ratio of unity (defined as the noise equivalent power, NEP) and therefore equal to $k(\Delta T)_{\min} B_{\text{in}}$ where

k is Boltzmann's constant = $1.38 \times 10^{-23} \text{ W/}^\circ\text{K}$

B_{in} is the input bandwidth.

Here an output bandwidth of 1 Hz is understood since that is the condition for which Putley's data is given. Solving for $(\Delta T)_{\min}$ one obtains:

$$(\Delta T)_{\min} = \frac{NEP}{kB_{\text{in}}} \quad (\text{direct detection}) . \quad (\text{A2})$$

This may be compared, for $B_o = 1$, to:

$$(\Delta T)_{\min} = \frac{F_{\text{RAD}} T_o}{\sqrt{B_{\text{IF}}}} \quad (\text{superheterodyne}) . \quad (\text{A3})$$

In the first case $(\Delta T)_{\min}$ varies inversely as first power of input bandwidth. Increasing the input bandwidth improves performance more rapidly for the direct detection than increasing B_{IF} in the superheterodyne where $(\Delta T)_{\min}$ varies inversely as the one-half power of B_{IF} . However, if in the two cases, B_{in} and B_{IF} are both decreased, the superheterodyne performance is degraded more slowly.

Some actual numbers for $(\Delta T)_{\min}$ are compared below using a 10% bandwidth for the Putley detector and a 4 GHz IF bandwidth for superheterodyne receivers. Data measured for $(\Delta T)_{\min}$ for a 600 GHz superheterodyne receiver^(A5) is used after correcting for $S/N = 1$ and $B_o = 1$ Hz. Also, values for $(\Delta T)_{\min}$ for superheterodyne radiometers at 100 GHz and 150 GHz have been included which were calculated using mixer conversion loss measurements for sealed 3 mm and 2 mm diodes^(A6) as currently made at this laboratory. A value of $(\Delta T)_{\min}$ is given for a 560 GHz receiver based upon the data presented in Section I.D. of this report for the submillimeter fourth harmonic mixer. It has been corrected for $B_{\text{in}} = 4$ GHz and $B_o = 1$ Hz.

	Putley		Superheterodyne Receivers		
	<u>$(\Delta T)_{\min}$</u>	<u>Ref.</u>	<u>$(\Delta T)_{\min}$</u>	<u>RF losses included</u>	<u>Ref.</u>
100 GHz					
Median Case			0.1° K	2 dB	(A6)
Selected Case			0.036° K	2 dB	(A6)
150 GHz	4.8° K	(A4)	0.44° K	2 dB	*
560 GHz			14.° K	3 dB	Sec. I.D.
600 GHz	12.° K	(A3)	10.° K	3.6 dB	(A5)

* Sealed 2 mm diodes currently being made at this laboratory typically have conversion losses of 11 dB.

One may conclude that the sensitivities of the Putley detector and the superheterodyne are of the same order of magnitude for applications requiring frequency resolution. Note that in this comparison the Putley type has B_{in} of 60 GHz at 600 GHz. If it had a 35% bandwidth (210 GHz), $(\Delta T)_{min}$ could be reduced to 1/3.5 of 12 or 3.4°K.

An "ideal" superheterodyne receiver is one which has a noise figure of unity. Therefore, by Equation A1, $(\Delta T)_{min}$ for an ideal superheterodyne radiometer receiver is $T_o \sqrt{B_o/B_{IF}}$. For the case considered in the preceding table, $B_o = 1$ Hz, $B_{IF} = 4$ GHz, and $(\Delta T)_{min} = 0.005^\circ K$ for a receiver of unity noise figure. For the case discussed in Section IA2, $B_o = 0.1$ Hz, $B_{IF} = 1.6$ GHz, and $(\Delta T)_{min} = 0.002^\circ K$. While this sensitivity can only be approached, the potential for superheterodynes is greater than for photo-conductive receivers for radiometers capable of frequency resolution.

A more sensitive thermal detector has recently been reported by Low. (A7, A8) It is a germanium bolometer cooled to 2.0°K. A commercial germanium bolometer also operated at 2.0°K is available from Texas Instruments, Inc. (A9) The characteristics of these detectors are as follows:

	NEP	B_{in}	$B_o = 1$ cps $(\Delta T)_{min}$	Ref.
Low	0.04×10^{-12} W	205 - 295 GHz	$0.047^\circ K$	(A7)
Texas Instruments	300×10^{-12} W	200 - 3000 GHz		(A8)

Because of the different input bands these are not strictly comparable. Also, the NEP depends upon the field of view, which is a 15° cone for the TI bolometer but is unknown for the Low bolometer. The smaller the field of view, the smaller the NEP and hence the smaller the minimum detectable power. If Low's bolometer were restricted to 10% input bandwidth with filters having no more insertion loss than at present, then $(\Delta T)_{min}$ would be about $0.16^\circ K$. This is the same order of magnitude as shown in the earlier paragraph describing estimated performance for superheterodynes at 150 GHz with the current sealed millimeter wave diode technology. Data is not yet available for a germanium bolometer at 600 GHz.

Both the photoconductive and the bolometer detector have the property of being sensitive from millimeter wavelengths up through the entire infrared band and beyond, or at least three decades of the spectrum. Furthermore, thermal radiators have more intense radiation at the higher frequencies. It is a severe problem to devise filters which will give radiometers using these detectors a degree of frequency resolution. The low frequency limit of the desired pass band can be determined by using sufficiently small waveguide. The high frequency limit of the desired pass band should be determined by a suitable low pass filter. For the bulk detectors the low pass filter must reject energy over at least two decades of the spectrum. The superheterodyne is susceptible to spurious responses at discrete frequencies only, for which narrow band filters may be designed. There is no known technique for designing a low pass filter with a very broad stop band and any desired cutoff frequency in the submillimeter band. Filter materials in current use by infrared physicists have cutoff frequencies in the near infrared. It also should be noted that the filter requirement will depend upon the absolute temperature of the source to be observed since the frequency of maximum output varies with temperature.

The comparison of submillimeter wave detectors cannot be limited to performance data for ground-based laboratory instruments. One must consider also the problems of developing instruments for space-vehicle use. As mentioned above, a solid-state local oscillator must be developed for superheterodyne systems. Also the mixer diode must be made more rugged and reliable. The bulk detectors have a clear advantage in this respect in having no delicate point-contact junction. However, their sensitivity is achieved only at extremely low temperatures obtainable only in cryostats using liquid helium. A further drawback to the Putley photoconductive detector is that it requires a magnetic field of some thousands of gauss for operation. A consideration when comparing superheterodynes and bulk detectors for space use is their relative susceptibility to loss of sensitivity due to space radiation. At present, no information is at hand with which to make such a comparison.

To summarize the foregoing discussion, the superheterodyne receiver for submillimeter wavelengths may be expected to have the best sensitivity for applications using relatively narrow bandwidths to achieve high frequency resolution whereas direct detection systems using bulk detectors can have greater sensitivity where the bandwidth can be permitted to be very large. Direct comparisons between bulk detectors and superheterodyne receivers are difficult because of the many factors entering the comparison. There is little experimental

data because so few systems have been built and operated at submillimeter wavelengths. Furthermore, the systems which have been built have been for earth-bound use and there are severe problems in the development of any of these for use in space vehicles.

Appendix B

PROPAGATION IN WAVEGUIDES BELOW CUTOFF

During the testing of the submillimeter harmonic mixer certain unusual results were noted. When the mixer mount was first being tested, it was also tried as a harmonic generator. For this test a diode detector having a 0.080" x 0.040" waveguide and micrometer driven tuner was attached to the mixer signal port. (The mixer signal port tapers from 0.080" x 0.040" at the input to 0.014" x 0.007" for the input filter.) When a 140 GHz signal was applied to the mixer diode via the crossed waveguide, an output was detected with the external diode detector. The wavelength of the output was measured with the detector tuner and was found to be the guide wavelength of 140 GHz. The mixer insertion loss at 140 GHz between LO port and signal port was about 65 dB. Further evidence that the output was a portion of the input 140 GHz and not a harmonic is the fact that the output was still detected when the mixer diode contact was broken. Even with a good diode in the mixer any harmonic energy that may have been present was masked by the fundamental.

The mixer signal waveguide is nominally 0.014" x 0.007" and is 0.148" long from the edge of the diode crystal to the beginning of the input horn flare. The waveguide was electroformed of gold and copper on a precision plastic mandrel and is believed to be dimensionally accurate within 0.0001". The theoretical cutoff frequency is 420 GHz, and the theoretical attenuation at 140 GHz of a 0.148" length is 320 dB as calculated by Equation 8-21 of Reference (B1). The attenuation at 415 GHz was calculated by Equation 5.2-18 of Reference (B2), an equation which is useful for calculations at or near the cutoff frequency. The result for 415 GHz was 53 dB for the 0.148" length of gold. A calculation with the same equation gave 4.8 dB at the cutoff frequency, 420 GHz.

Although considerable thought has been given to the problem, no satisfactory explanation of the measured results can be suggested. On several occasions energy at 140 GHz was passed through the submillimeter mixer and detected at the signal port when it should have been attenuated far below a detectable level. This happened both with and without a diode junction in the mixer although the presence and condition of a diode affected the value of the transmission loss.

The initial method of construction of the mixer was thought to be a possible cause of the 140 GHz transmission but later proved not to be. The local oscillator and signal crossed waveguides are separate parts which are clamped together by the mixer body. In the region in which they have a common wall, the wall of one is cut away and the wall of the other serves both. As first constructed, the common wall of the 0.014" x 0.007" signal waveguide was cut away. After test, it was thought that the other waveguide might not be covering the signal waveguide tightly enough to prevent lower frequencies from propagating in the cracks. The mixer was rebuilt with a new signal waveguide. This time the common wall of the LO waveguide was removed leaving the signal waveguide intact. The mixer then transmitted 140 GHz to the signal port but with about 50 dB insertion loss as compared to 65 dB before. The exact value varies, as before, with the position of whisker and formation of a diode junction. It appears that the cracks were not the cause of the unexpected propagation of 140 GHz through the signal waveguide. Instead, the cracks caused some added dissipative losses. By eliminating the cracks the dissipative losses were reduced. This resulted in the lower mixer conversion loss that was observed.

The submillimeter signal source which was built to use in testing the mixer has been found to operate in a predictable manner, that is, it does not propagate signals below its waveguide cutoff frequency. It is a crossed waveguide device with an electroformed output waveguide 0.020" x 0.010" in cross section and 0.193" long from the edge of the diode to the output flange. Its theoretical cutoff frequency is 295 GHz. With inputs of 75, 80, and 140 GHz no energy below 300 GHz has been detected at its output. It may be that the degree of coupling of the fundamental from the larger input waveguide to the crossed output waveguide is less than in the submillimeter mixer and thereby prevents any fundamental from being detected at the output. More definitive measurements with single waveguides would be desirable.

It appears that analysis of millimeter wave propagation based upon theoretical idealized conditions does not explain the anomalies that have been observed. Factors which may need to be considered at millimeter and submillimeter wavelengths are waveguide dimensional tolerances, corner radii, surface finish and condition, and material conductivity. The calculated insertion loss for waveguides beyond cutoff may be significantly in error due to end effects causing unexpected coupling. It may be that proportionately longer high pass filter sections must be used to overcome end effects at these frequencies. If a longer 0.014" x 0.007" signal waveguide were to be built, revisions to the mandrel molding technique would be necessary. Perhaps the position of the diode in

the signal waveguide could be moved to leave less waveguide for the tuner and more for the input filter. However, it should be pointed out that in addition to 0.148" of signal waveguide in the submillimeter mixer, 0.158" of the 0.300" length of the input horn is also cut off at 140 GHz.

The fact that 140 GHz energy was transmitted through the mixer signal waveguide unexpectedly indicates that care must be taken to identify the frequency in harmonic mixer measurements. It has already been reported in Section II that harmonic mixing was observed and distinction made between third harmonic at 420 GHz and fourth harmonic at 560 GHz by means of a double conversion system and narrow IF band. The guide wavelengths of the two frequencies were positively identified by using the mixer signal waveguide tuner. In a radiometer system the signal waveguide cutoff frequency would have to be adjusted to reject signals at the next lower harmonic below the desired harmonic of the fundamental LO frequency. This is more difficult for higher order mixing. The frequency being received would be verified by measuring its wavelength with the signal tuner. This was done successfully with a 600 GHz radiometer built for the U. S. Army Munitions Command⁽¹⁻²⁾ although a second harmonic mixer rather than a fourth harmonic device was used.

Appendix C

CROSSED WAVEGUIDE COUPLING STUDIES

A crossed waveguide device was built which permits modification of the common wall between the waveguides. Its parts and the complete assembly are shown in Figure C1. Measurements were made with it to determine design criteria for the optimum coupling of local oscillator energy to the semiconductor junction in a fourth harmonic crossed waveguide mixer. Optimum LO coupling is very important for the submillimeter harmonic mixer since the available LO power in the satellite radiometer will be small. These experiments were made possible in large measure by the fabrication of an epoxy-bonded point-contact millimeter diode which could be removed from a crossed waveguide mount and reinstalled many times without changing its properties. Thus, the effect of varying wall thickness and coupling hole diameter could be studied. Figure C2 is a micro-photograph of an epoxied junction with an etched whisker beside it for comparison.

The device was designed as a 210 GHz fourth harmonic generator and a set of shims supplied so that tests could readily be made with various common wall thicknesses and in any sequence. Using one epoxy-bonded diode throughout the experiment, 30 measurements were made of conversion loss for 24 different wall thicknesses. Figure C3 is a plot of the data. Only some of the points lie on a smooth curve. The test was repeated for three wall thicknesses with variable results as indicated by the vertical dashed line. There was also an abrupt change in output at the right hand side of the plot. While trying to resolve these anomalies it was determined that slight motion of the coaxial connector and whisker carrier caused significant changes in the output. The data of Figure C3 does give some basis for selecting a common wall thickness and was considered in the design of the submillimeter fourth harmonic mixer.

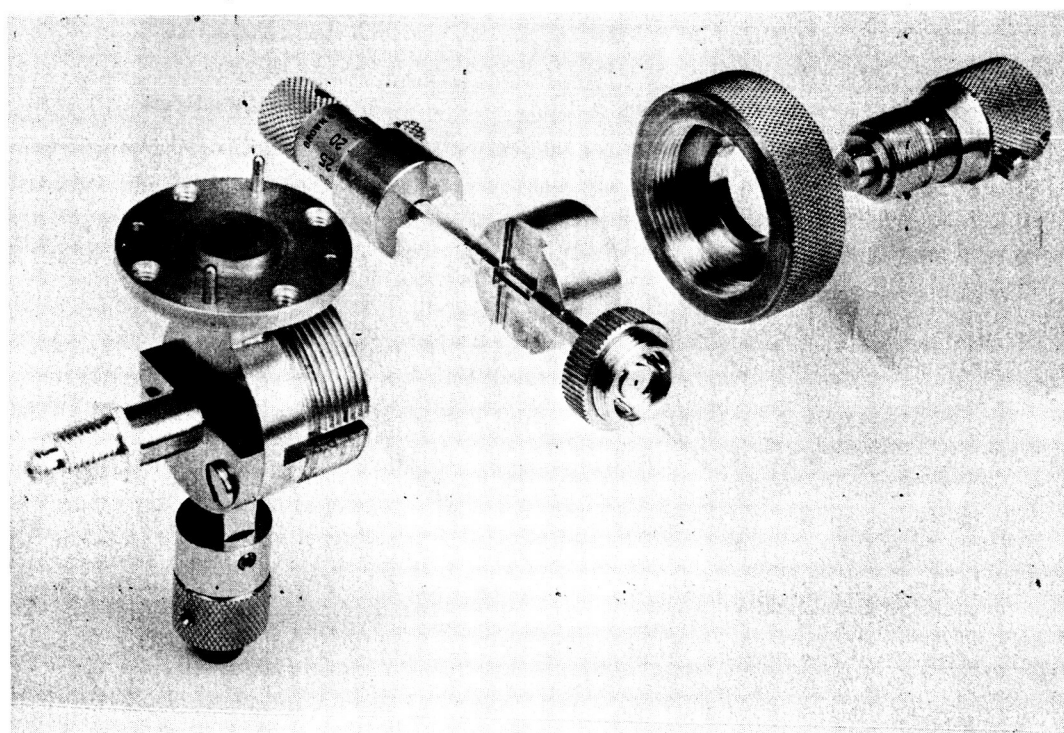
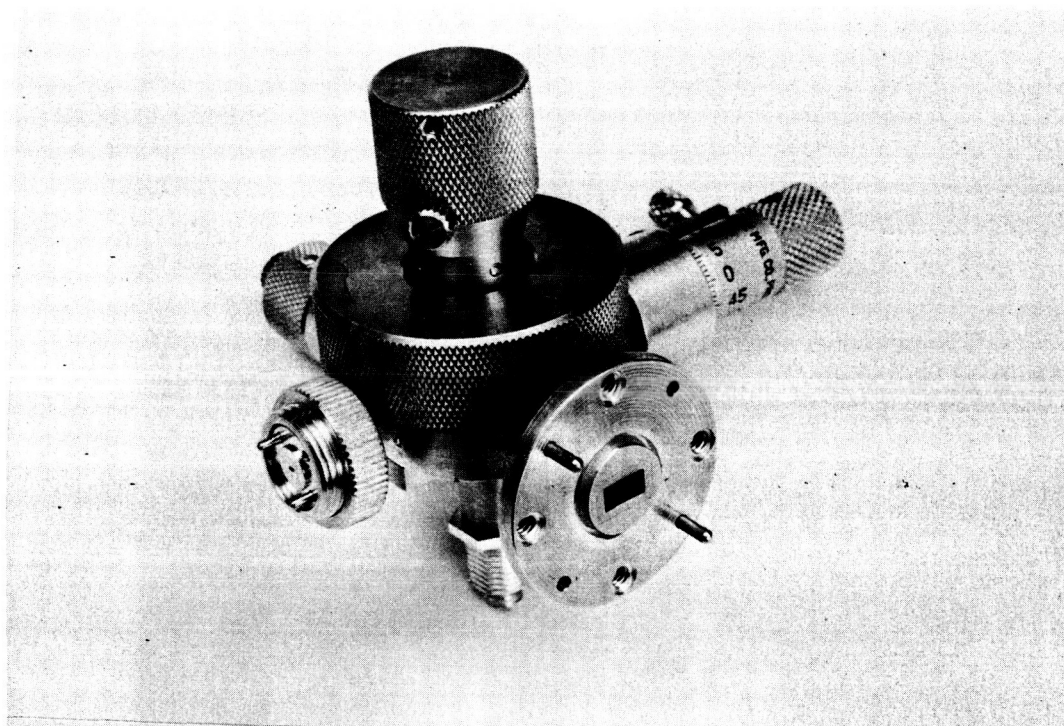


FIG. C1 - Fourth Harmonic Generator for Crossed Waveguide
Coupling Studies

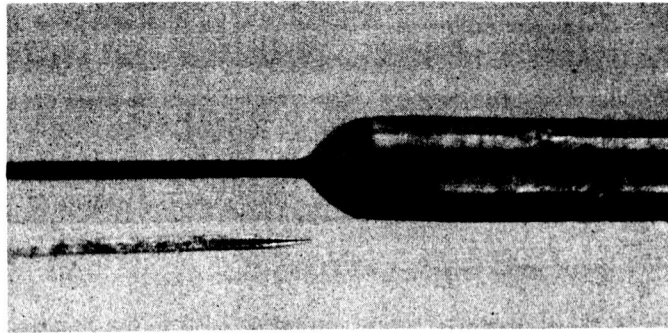


FIG. C2 - Epoxied Diode, Microphotograph

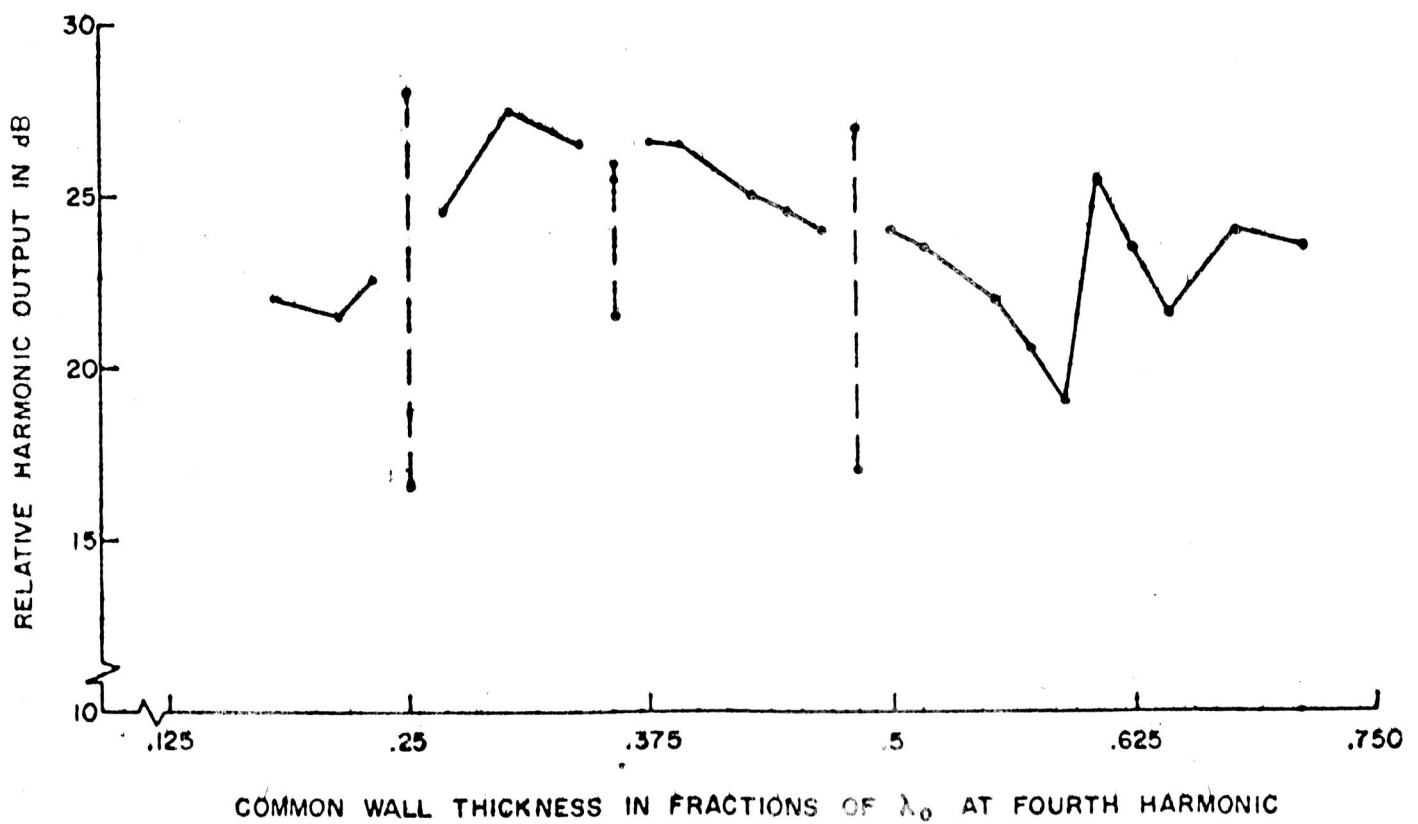


FIG. C3 - Data for Fourth Harmonic Crossed Waveguide Device

Appendix D

ANALYSIS OF DIODE LOSS FACTORS

If a semiconductor diode is to perform well at millimeter and submillimeter wavelengths, the diode parasitics [spreading resistance (R_s) and barrier capacitance (C_b)] must be low. The measurements made at this laboratory indicate that the well known equivalent circuit for a semiconductor diode (shown in Figure D1) is still a good model at frequencies up to at least 600 GHz. This equivalent circuit and the equations derived from it have been used to predict the increase in mixer conversion loss, $L_c(\omega)$, as a function of frequency. This equivalent circuit can also explain the degradation in performance of harmonic generators and video detectors as the frequency is increased. In the case of a mixer application, it has been shown that^{(D1)(D2)(D3)}

$$L_c(\omega) = L_o L_1 = L_o \left(1 + \frac{R_s}{R_b} + \omega^2 C_b^2 R_s R_b \right), \quad (D1)$$

where:

$L_c(\omega)$ is the frequency dependent conversion loss,

L_o is the conversion loss with zero spreading resistance,

R_b is the nonlinear barrier resistance, and

ω is the angular frequency = $2\pi f$.

Performing the operation

$$\frac{\partial L_c}{\partial R_b} = 0, \quad (D2)$$

one obtains the following equation for the value of R_b which minimizes the conversion loss

$$R_b = \frac{1}{\omega C_b}. \quad (D3)$$

Substituting D3 into D1 yields

$$L_c(\omega)_{\min} = L_o (1 + 2\omega R_s C_b). \quad (D4)$$

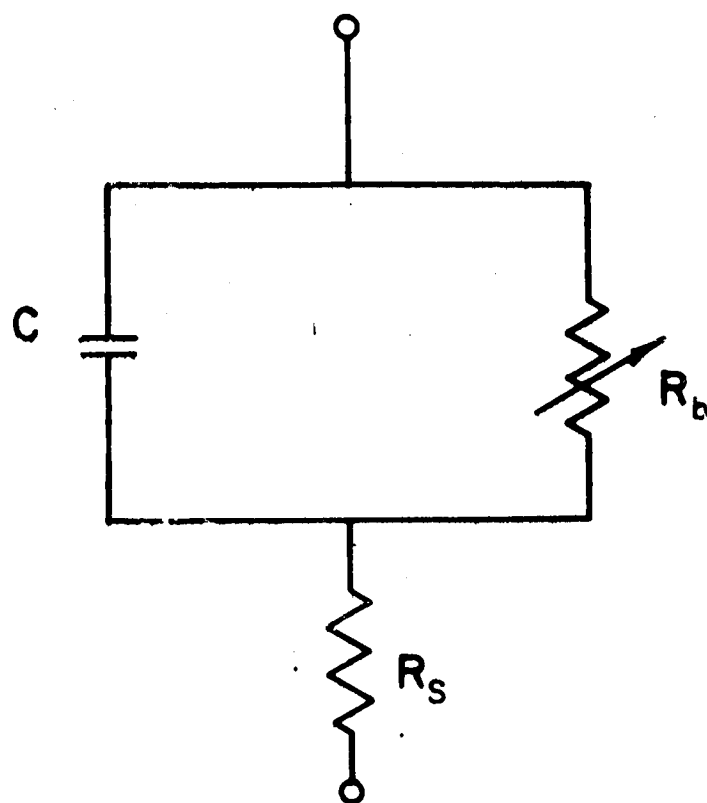


FIG. D1 - Diode Equivalent Circuit

Since R_b is a monotonically decreasing function of the LO drive level, Equation D3 explains the commonly observed fact that the optimum LO drive level increases with increasing frequency and often is not reached even at diode burnout levels. At sufficiently high frequencies, the last term within the parentheses of Equations D1 and D4 becomes dominant. It is thus desirable to minimize the barrier capacitance and if possible provide sufficient LO power to satisfy Equation D3. In that way $L_c(\omega)$ is proportional to ω rather than to ω^2 . Both Equations D1 and D4 illustrate the necessity of minimizing the $R_s C_b$ product.

Gallium arsenide, because of its high mobility and relatively low dielectric constant, facilitates the development of diodes in which the values of the parasitic elements can be minimized. ^(D4) A method of forming the diode while observing its RF performance has been developed. The diode is formed in its mount with square wave modulated RF power impinging upon it. Simultaneously, a sweeping bias voltage derived from a transistor curve tracer is applied to the diode. The diode current is displayed as the ordinate on the CRT of the transistor curve tracer. A photograph of the resulting I-V characteristics is shown in Figure 5-2. If the square wave modulation frequency is harmonically related to the sweep frequency, dashed curves result; otherwise, the two curves appear to be continuous. The lower curve is the usual I-V curve of a gallium arsenide diode; it manifests the characteristic sharp break at a forward voltage of about 0.7 volts. The separation of the curves is a function of the high level rectification efficiency, which in turn is a function of and hence provides an indication of the $R_s C_b$ product.

An analysis is now given for certain idealized conditions which permit the effective resistance at some operating frequency to be estimated. The contributions to R_s which are considered are that due to the resistance of the whisker and that due to the resistance of the semiconductor between the active junction and the metal crystal post. The skin effect due to the high frequencies is taken into account since at millimeter wavelengths the skin depth is much less than the crystal wafer thickness or radius. All surfaces are assumed to be perfectly smooth. Additional resistance due to roughness of real parts has not been estimated.

Referring to Figure D2, assume the diode whisker tip is a hemisphere of radius a located just within the semiconductor crystal which is a flat disk of radius r_1 and thickness h . Also assume that the current flows radially in the semiconductor to a distance equal to one skin depth, δ_s . Then assume a transition region until the current is flowing in a layer of one skin depth to the edge of the crystal and

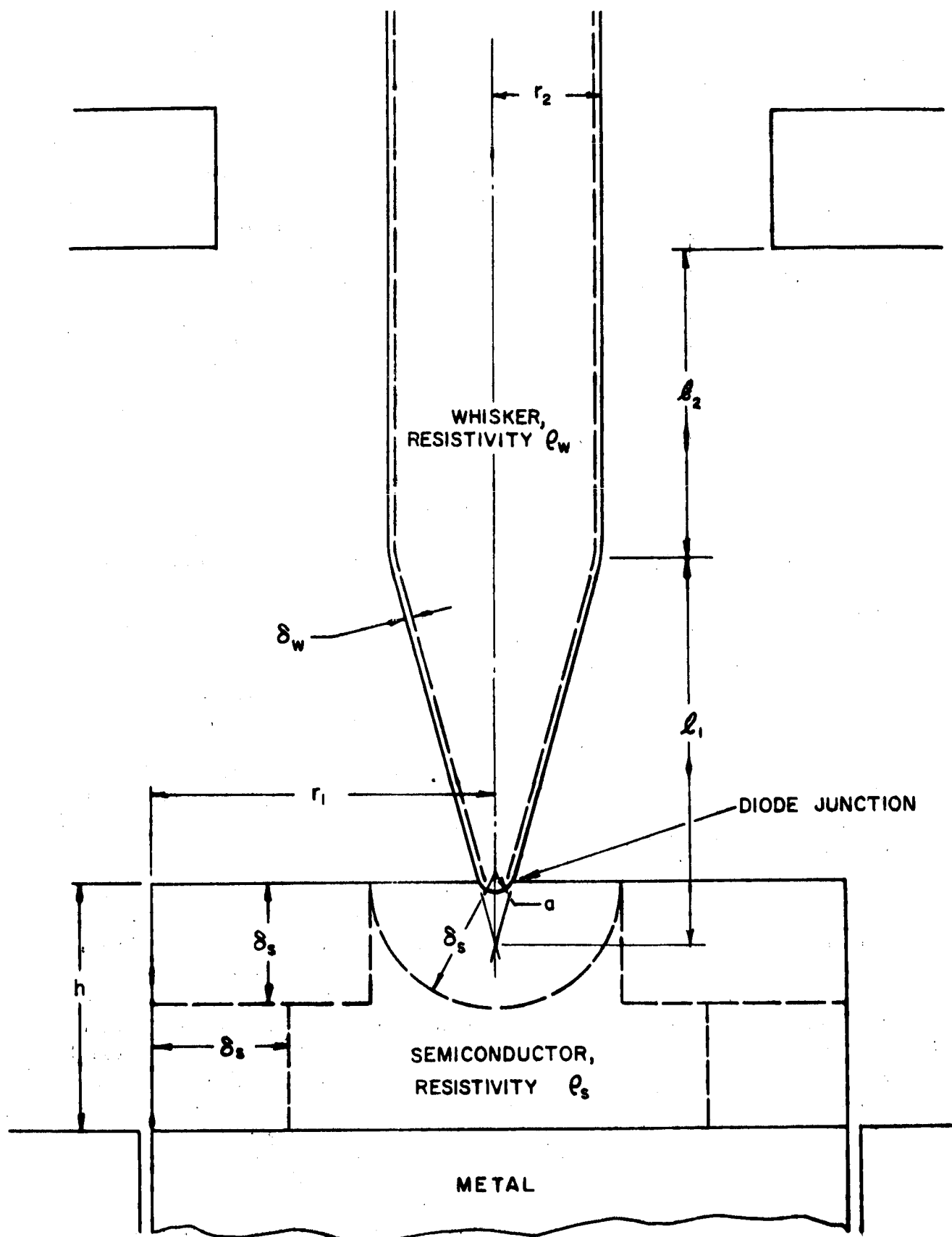


FIG. D2 - Geometry of Point-Contact Junction

down the sides to the metal post. An expression for the resistance of each of these regions will be obtained.

In the hemispherical region under the whisker point let dR_1 be the resistance of a hemispherical shell of radius r and thickness dr . From the definition of resistivity, ρ :

$$dR_1 = \frac{\rho_s dr}{2\pi r^2} \quad (D5)$$

where $2\pi r^2$ = the cross-sectional area, here the surface area of a hemisphere.

Integrating from the whisker tip to a radius of one skin depth one obtains

$$R_1 = \int_a^{\delta_s} \frac{\rho_s dr}{2\pi r^2} = \frac{\rho_s (\delta_s - a)}{2\pi a \delta_s} \quad (D6)$$

Note that for $\delta_s \gg a$, D6 reduces to:

$$R_1 = \frac{\rho_s}{2\pi a} \quad (D7)$$

This means that the outer shells contribute a negligible amount to the series resistance compared to the contribution from the restricted region about the whisker point. Equation D7 is seen to be similar to the expression given by Torrey and Whitmer^(D5) for the resistance of a circular area of contact:

$$R = \frac{\rho}{4a} \quad (D8)$$

In the next region the current flows through cylindrical shells of radius r and thickness dr . The resistance, dR_2 , of each is:

$$dR_2 = \frac{\rho_s dr}{2\pi r \delta_s} \quad (D9)$$

where $2\pi r \delta_s$ = the surface area of a shell.

The beginning of the cylindrical region is arbitrarily taken at a radius equal to δ_s which was the outer limit of integration for the hemispherical region. Integrating D9 from δ_s to r_1 , the edge of the crystal, gives:

$$R_2 = \int_{\delta_s}^{r_1} \frac{\rho_s dr}{2\pi r \delta_s} = \frac{\rho_s}{2\pi \delta_s} \left[\log_e \frac{r_1}{\delta_s} \right] \quad (D10)$$

Finally, R_3 , due to current flow down the sides of the crystal is given by

$$R_3 = \frac{\rho_s h}{2\pi r_1 \delta_s} \quad (D11)$$

where h = the height of the crystal, i.e., the length of the current path;

$2\pi r_1 \delta_s$ = the cross-sectional area (since $r_1 \gg \delta_s$).

The cross-sectional area is constant and no integration is necessary to obtain R_3 .

Any resistance in the whisker becomes a part of the diode series resistance. In order to evaluate the magnitude of this contribution, expressions are derived below for the resistance in the tapered portion and in the cylindrical portion of the whisker.

Consider that the idealized whisker point is a right circular cone of length l_1 and base diameter $2r_2$ whose apex has been truncated leaving a circular contact of diameter $2a$. Let δ_w be the skin depth in the whisker material being used at the frequency of interest. The current will flow through a series of rings of radius r . The resistance, dR_4 , of each is:

$$dR_4 = \frac{\rho_w dl}{2\pi r \delta_w} \quad (D12)$$

where $2\pi r \delta_w$ = the cross-sectional area of a ring (since $r \gg \delta_w$)

dl = the increment of whisker length.

But by similar triangles

$$\frac{\ell}{r} = \frac{\ell_1}{r_2} \quad (D13)$$

Differentiating,

$$d\ell = \frac{\ell_1}{r_2} dr \quad (D14)$$

Substituting D14 in D12 and integrating from $r = a$ to $r = r_2$ gives:

$$R_4 = \int_a^{r_2} \frac{\rho_w \ell_1 dr}{2\pi \delta_w r_2 r} = \frac{\rho_w \ell_1}{2\pi \delta_w r_2} \left[\log_e \frac{r_2}{a} \right]. \quad (D15)$$

The cylindrical portion of the whisker, length ℓ_2 , has a resistance, R_5 , which is

$$R_5 = \frac{\rho_w \ell_2}{2\pi r_2 \delta_w} \quad (D16)$$

where $2\pi r_2 \delta_w$ = the cross-sectional area (since $r_2 \gg \delta_w$).

Equations D6, D10, D11, D15, and D16 have been used to estimate the series resistance at 200 and at 600 GHz of a point-contact diode made of materials with the following properties:

Semiconductor Wafer

gallium arsenide, $\rho_s = 13.4 \times 10^{-3}$ ohm-cm
(Code No. K-3)

0.010" radius (r_1) x 0.004" thick (h) for 200 GHz

0.004" radius (r_1) x 0.003" thick (h) for 600 GHz

Whisker

phosphor bronze, $\rho_w = 11 \times 10^{-6}$ ohm-cm

0.0005" radius (r_2)

0.0025" tapered length (ℓ_1)

0.0105" straight length (ℓ_2) for 200 GHz

0.0025" straight length (ℓ_2) for 600 GHz

Conductive Plating

gold, $\rho = 2.4 \times 10^{-6}$ ohm-cm

The radius a of the point contact region was taken to be 1.5×10^{-4} cm based upon microscopic measurements of whisker points made on another project. (D6)

The skin depth δ is given in meters by (D7)

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \quad (D17)$$

where f = frequency in hertz, $\mu = 4\pi \times 10^{-7}$ for these materials, and $\sigma = 1/\rho$, conductivity in mhos per meter.

Calculated skin depths in micrometers are:

	<u>200 GHz</u>	<u>600 GHz</u>
Crystal (δ_s)	13.0 μm	7.5 μm
Whisker (δ_w)	0.37	0.22
Gold	0.17	0.10

The values for series resistance in ohms which were obtained are as follows:

		<u>200 GHz</u>	<u>600 GHz</u>
Crystal	$\left\{ \begin{array}{l} R_1 \\ R_2 \\ R_3 \end{array} \right.$	12.6 Ω	11.4 Ω
		4.8	7.4
		0.6	2.1
Whisker	$\left\{ \begin{array}{l} R_4 \\ R_5 \end{array} \right.$	0.5	.8
		1.0	0.3
Total	R_s	19.5	22.0

Keeping in mind that these numbers are calculated for a simplified, idealized model and are estimates only for a real diode, nevertheless, one may use them as a guide to show where improvement is most needed. The equations may be used to calculate the resistance of diodes modified for possible improvement. For example, applying gold plating at least a fraction of a micron thick to the sides of the crystal wafer will reduce R_3 to a negligible value. Plating the top of crystal will reduce R_2 but will not affect R_1 unless the plating can be applied within one δ_s of the whisker contact. This top plating approach to reducing R_s is being tried on this project. The closest it has been possible so far to apply the plating has been 0.001" or 25 μm . This would reduce R_2 by 3.8 or 4.0 ohms at 200 or 600 GHz respectively. Combined with the elimination of R_3 , this is a reduction of better than 20% in either case. Diodes plated this way should be detectably better performance.

Appendix E

TUNNEL AND BACKWARD DIODES

Experimental diodes were made with highly doped semiconductor crystals and whisker wires of suitable metal. The parts were assembled first as a point-contact diode. The diode was then pulsed with voltage which alloyed a bit of the whisker with the semiconductor to convert it to a tunnel diode. It was then distinguished as a tunnel diode by its current voltage characteristic curve, shown in Figure E1, which has a current peak followed by a valley for small positive voltage. This is a region of negative resistance. If the pulse-alloying was limited so that a negligible current peak results, a backward diode was formed. The tunnel diode has a narrow region of nonlinearity before the current peak, and the backward diode has an abrupt change at zero bias. These nonlinearities make them especially well-suited for use as mixers. Eng observed (E1) in studies of 13.5 GHz mixers that tunnel diodes do not require as much local oscillator power as point contact diodes.

Another factor of interest in the application of tunnel or backward diodes to millimeter wavelengths is that the parasitic series resistance may be expected to be low because of the low resistivity of the highly doped semiconductors used. Burrus (E2) has successfully fabricated and used backward diodes at 59 and 115 GHz. His materials were n-type germanium having a resistivity of 0.00056 ohm-cm and gallium coated whiskers. The materials available for testing on this project were as follows:

Code No.	Type	Semi- conductor	Dopant	Carrier Concentration (atoms/cm ³)	Resistivity (ohm-cm)	Matching Whisker Material
M-6	p	GaAs	Zn	580×10^{17}	$2. \times 10^{-3}$	Sn
3	n	GaAs	Se	70×10^{17}	0.8×10^{-3}	Zn
8	n	Ge	As	1000×10^{17}	6.5×10^{-3}	Ga

Tunnel diodes were formed with both p- and n-type gallium arsenide. A photograph of a typical I-V curve as observed on an oscilloscope is reproduced in Figure E2. The diodes were formed in a section of RG-99/U waveguide and tested as fundamental 94 GHz mixers. The most sensitive diodes were also the least stable, and they could not be measured under

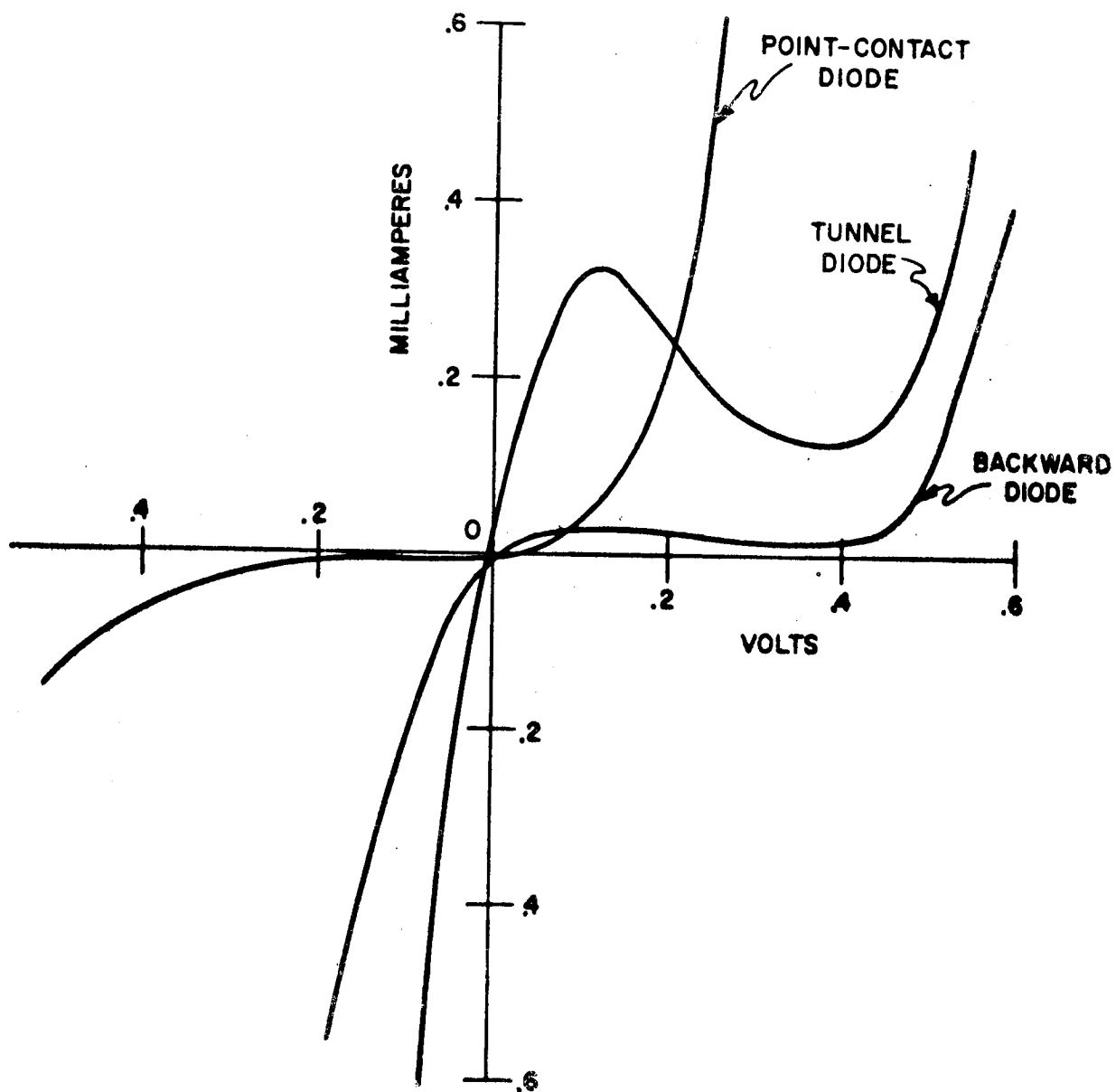


FIG. E1 - Current-Voltage Characteristics of the Point-Contact, Tunnel, and Backward Diodes

a variety of conditions. Measurements were made of the "transducer" loss of the devices which includes the conversion loss and losses due to signal and IF impedance mismatches. Transducer loss values ranged from 13 to 25 dB. Minimum transducer loss occurred for the maximum LO power that the diode could withstand without changing its I-V curve. A relatively insensitive but stable tunnel diode was tested over a range of bias voltages and LO powers. When biased in the negative resistance region and driven with less than 1 mW of LO power, the diode generated

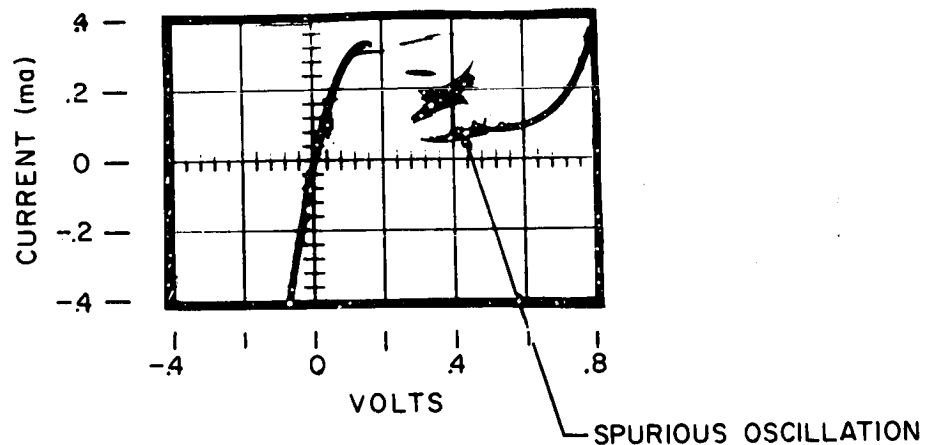


FIG. E2 - Typical Measured I-V Curve for Tunnel Diode

oscillations at the intermediate frequency (2 GHz). However, the minimum transducer loss occurred for 0.3 V bias when sufficient LO drive was applied to prevent IF oscillations. It did not appear possible to bias near the nonlinear current peak and achieve low transducer loss for low LO drive without IF oscillation.

Effort was then shifted to forming and testing backward diodes. Both the n-type gallium arsenide and n-type germanium were used, and typical backward diode I-V curves obtained. Diode stability was poor at first but improved with experience in forming this type. Each diode was tested as a point-contact diode before pulse alloying to obtain the backward diode characteristic. In all cases the point-contact diode had the lower transducer loss. The backward diodes were all tested as second harmonic devices with 70 GHz drive power, some as harmonic generators and others as harmonic mixers. The drive power was between

4 and 23 mW. The LO power and bias voltage were varied with some of the diodes. The best backward diodes had the lowest transducer loss for zero bias, measuring 28 dB for 1.6 mW LO drive. For one diode, as LO drive was reduced to 200 μ W the transducer loss increased only to 31.5 dB. This relative insensitivity of transducer loss to LO drive is highly desirable if low transducer loss can also be achieved.

Experimentation with tunnel and backward diodes was discontinued at this point since no immediate application to submillimeter mixing appeared to be forthcoming.

Appendix F

140 GHz SECOND HARMONIC MIXER

A crossed waveguide mount, Type M-9838, was used to test as harmonic mixers a number of diodes formed of n-GaAs crystals and phosphor bronze whiskers. The purposes of the tests were to develop techniques at millimeter wavelengths which might be applicable at sub-millimeter wavelengths and to obtain mixer data for comparison with test results of other mixer diode types being investigated.

Second harmonic mixing was tested with the local oscillator set at 70 GHz. The test procedures were similar to those described in Section II.C. for the submillimeter mixer except that the instrumentation was simpler at these longer wavelengths. Signal power was measured with a bolometer. The signal input VSWR could not be measured, however, and so the conversion loss data is conservative by an amount representing the mismatch loss.

A family of curves of mixer conversion loss versus LO drive power for different bias levels is plotted in Figure F1. The lowest conversion loss was 13 dB and occurred for 0.2 V external bias and 3.3 mW LO drive. It appears that if LO drive could have been increased above 6.6 mW, a lower conversion loss could have been obtained for zero bias. Mixer conversion loss at 140 GHz was measured on a previous contract^(F1) but no extended set of data such as this was obtained. The previous measurement was "on the order of" 10 dB conversion loss for 13 mW LO drive. This appears consistent with the data of Figure F1 although it cannot be verified because the LO drive was not extended to 13 mW in the more recent tests.

It is of interest to compare the curves of Figure F1 to the data in Figures 2-10, 2-14, 2-17, and 2-18 for fourth and third harmonic mixing at submillimeter wavelengths. All the families have similarities, particularly inflection points at low LO drive for most bias levels. Also, in general, when LO drive power is limited, the conversion loss can be minimized by a judicious choice of external bias voltage, but when optimum LO drive is available then zero bias is required for minimum conversion loss.

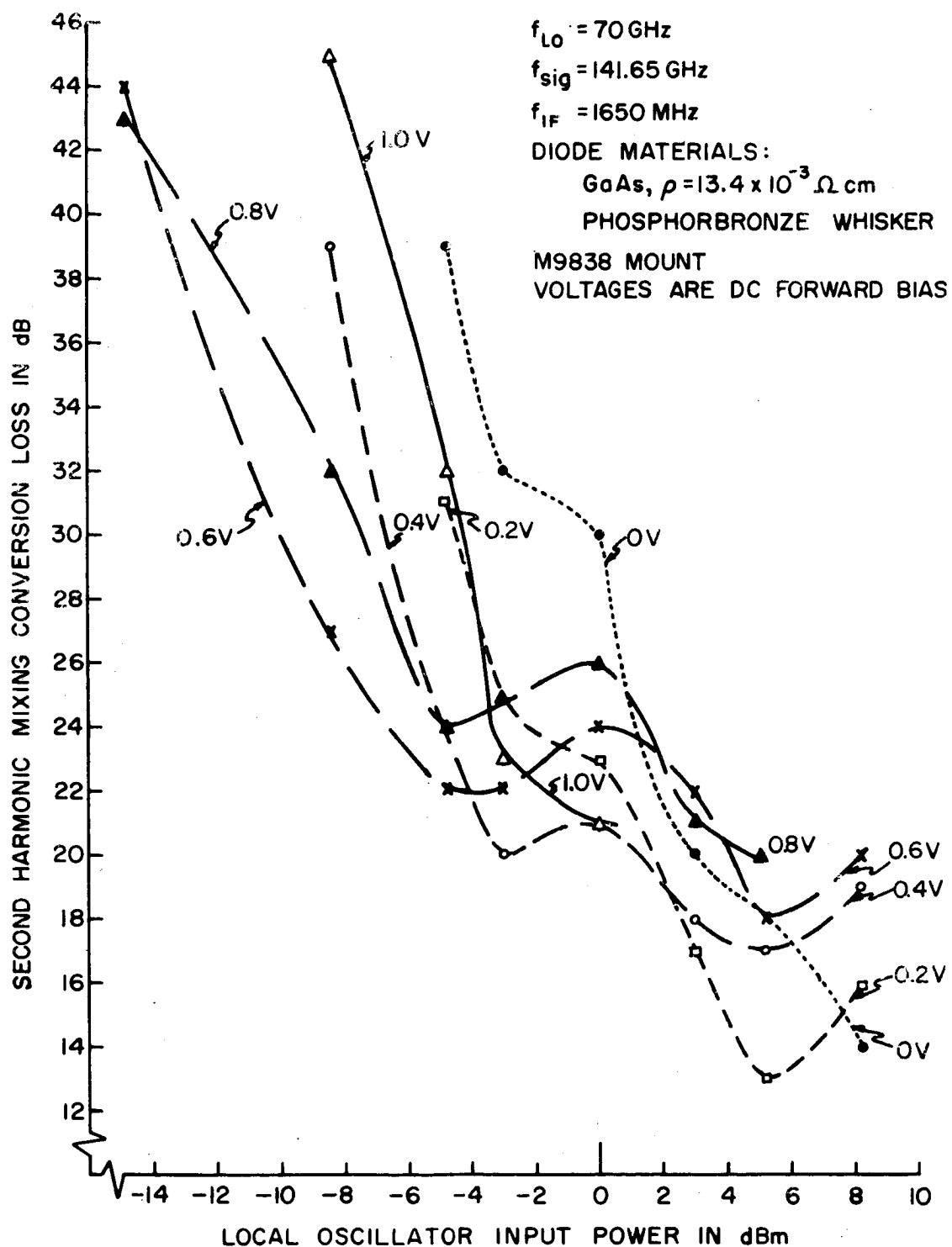


FIG. F1 - Measured Second Harmonic Mixing Conversion Loss vs Local Oscillator

Appendix G

ANALYSIS OF A SINGLE SOURCE MIXER TEST SYSTEM

A variety of possible test systems at 0.5 mm wavelengths were considered. These systems may be divided into two major categories: single source systems and two source systems. For reasons of economy, the former was of primary interest, and a potential system was explored in some detail. A summary of this study follows.

The single source system of interest involved using a 2 mm backward wave tube as both signal source and receiver local oscillator. The signal power was to be modulated and then applied to a harmonic generator. The local oscillator power was to be used to drive a harmonic mixing receiver crystal. It was hoped that the signal due to the fourth harmonic could be isolated by use of some simple waveguide filtering and a receiver IF tuned to the fourth harmonic of the modulation frequency. This system is indicated in block diagram form in Figure G1.

Analytically, one can represent a diode voltage-current characteristic in a power series of the form

$$i = a_1 e + a_2 e^2 + a_3 e^3 + \dots \quad (G1)$$

where

i = instantaneous diode current

a_i = amplitude coefficient

e = instantaneous voltage.

(To avoid confusion in later interpretations, it should be noted immediately that the coefficients here are not dimensionless, i. e., $\text{dim.}[a_i] = \text{ohm}^{-1} \text{volts}^{1-i}$.) The modulator output signal may be written

$$e = E_s [\cos(\omega t + \phi)] [1 + m \cos \omega_m t] \quad (G2)$$

where

E_s = peak carrier voltage

ω = source frequency

ϕ = phase difference

m = modulation index

ω_m = angular modulation frequency.

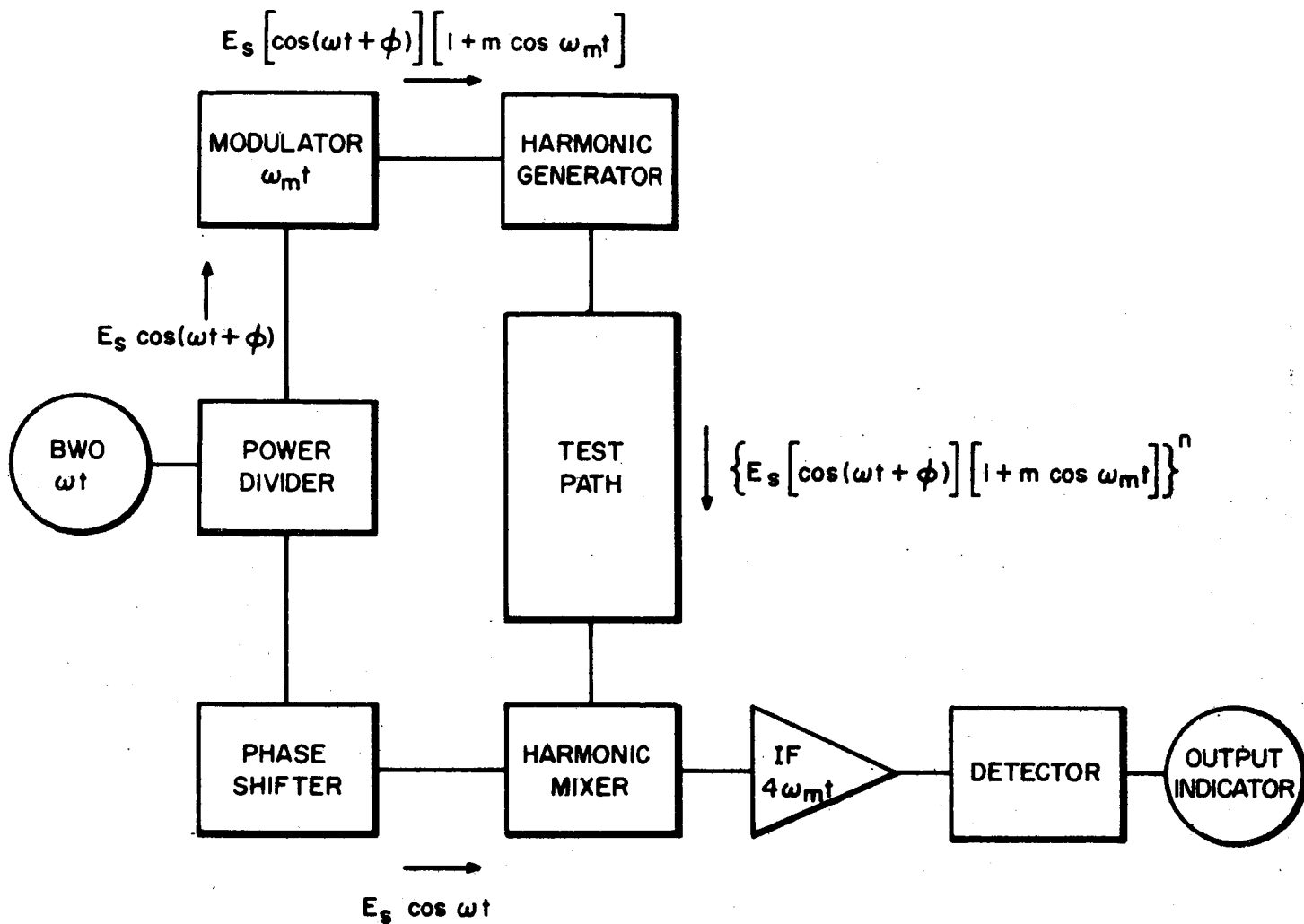


FIG. G1 - Single Source Test System

Substituting this value of e into Equation G1, which is taken as the diode transfer characteristic, there results an expression for the sum of all possible harmonics generated in the diode. Initially, this becomes

$$\begin{aligned} i = & a_1 E_s [\cos(\omega t + \phi)] [1 + m \cos \omega_m t] \\ & + \frac{a_2 E_s^2}{2} [1 + \cos 2(\omega t + \phi)] [1 + 2m \cos \omega_m t + m^2 \cos^2 \omega_m t] \\ & + \text{similar higher order terms.} \end{aligned} \quad (G3)$$

From each of the product terms ($\cos x \cos y$) terms of the form $A'_{kl} \cos(k\omega t \pm l\omega_m t + k\phi)$ result.

Defer temporarily the evaluation of the coefficients A'_{kl} and assume only that the output of the harmonic generator may be represented as the power resulting from the currents

$$i = \sum_{k=0}^N \sum_{l=0}^N A'_{kl} \cos(k\omega t \pm l\omega_m t + k\phi) \quad (G4)$$

where N may be made arbitrarily large.

Consider the harmonic mixer and a transconductance representation of its outputs. The instantaneous mixer diode conductance, which is determined by bias conditions and local oscillator drive, can be represented as

$$g = g_0 + g_1 \cos \omega t + g_2 \cos 2\omega t + g_3 \cos 3\omega t + \dots \quad (G5)$$

Now the currents which will produce signals in the IF amplifier are $i = g e$, where g has been defined in Equation G5 and e is the output from the harmonic generator. Assume

$$e = \sum_{k=0}^N \sum_{l=0}^N A_{kl} \cos(k\omega t \pm l\omega_m t + k\phi) \quad (G6)$$

where

$$A_{kl} = \eta_k A'_{kl} \quad (G7)$$

The η_k represents the total efficiency of coupling the harmonic generator diode currents to propagating waveguide modes in the test path and the coupling of this resultant energy into the mixer diode. It also includes the dimensions of impedance necessary to convert the current representation to a voltage representation.

Combining G5 and G6 gives

$$i = \left(\sum_{n=0}^{\infty} g_n \cos n\omega t \right) \left(\sum_{k=0}^N \sum_{l=0}^N A_{kl} \cos (k\omega t \pm l\omega_m t + k\phi) \right) \quad (G8)$$

as the signal available to the IF. With the IF amplifier tuned to the frequency $4\omega_m$, the only product terms which will result in a detected output will be those difference terms where $n = k$ and $l = 4$. Therefore, the magnitude and phase of terms of the form

$$g_k A_{k4} \cos (\pm 4\omega_m t + k\phi)$$

are of primary interest.

Note that this actually represents two terms, i. e.,

$$g_k A_{k4} [\cos (4\omega_m t + k\phi) + \cos (4\omega_m t - k\phi)].$$

Hence, if the phase difference ϕ equals $\frac{\pi}{2k} (2n - 1)$, where n is any integer; then, there are proper conditions for complete phase cancellation of this k^{th} term. Without any further investigation, the practicality of the single source system seems doubtful since the above analysis points to the necessity for excellent frequency stability in the source.

However, the amplitude of the various terms involved will be examined next. This is determined by the magnitude of $g_k \eta_k A'_{k4}$ for each term. The k now indicates the order of the harmonic involved.

Evaluating the A'_{kl} coefficients requires that one actually expand Equation G3 and collect terms. The amplitudes of the resulting

signals at the IF input, neglecting harmonics higher than the fifth are

$$g_1 \eta_1 A'_{14} = g_1 \eta_1 a_5 \frac{25m^4}{128} E_s^5 + \dots \text{ (contribution due to fundamental)} \quad (G9)$$

$$g_2 \eta_2 A'_{24} = g_2 \eta_2 a_4 \frac{m^4}{32} E_s^4 + \dots \text{ (2nd harmonic)} \quad (G10)$$

$$g_3 \eta_3 A'_{34} = g_3 \eta_3 a_5 \frac{25m^4}{256} E_s^5 + \dots \text{ (3rd harmonic)} \quad (G11)$$

$$g_4 \eta_4 A'_{44} = g_4 \eta_4 a_4 \frac{m^4}{128} E_s^4 + \dots \text{ (4th harmonic)} \quad (G12)$$

$$g_5 \eta_5 A'_{54} = g_5 \eta_5 a_5 \frac{5m^4}{256} E_s^5 + \dots \text{ (5th harmonic)} \quad (G13)$$

Note, for instance, that the ratio of the voltage contribution due to the second harmonic signal to that due to the fourth harmonic signal is

$$\frac{g_2 \eta_2 A'_{24}}{g_4 \eta_4 A'_{44}} = 4 \frac{g_2 \eta_2}{g_4 \eta_4} \quad (G14)$$

Since the coefficients g_i converge, it is clear that $g_2/g_4 > 1$ and probable that $\eta_2/\eta_4 > 1$ (although this can be influenced by the position of the shorting plungers in the waveguide at the harmonic generator and mixer). If it is assumed, however, that this voltage ratio is at least 4, then the voltage in the IF amplifier as a result of second harmonic propagation through the test path will be at least 6dB greater than that resulting from fourth harmonic propagation. Clearly, the propagation path must reject any second harmonic signal from the harmonic generator.

An examination of the third harmonic to fourth harmonic signal ratio gives

$$\frac{g_3 \eta_3 A'_{34}}{g_4 \eta_4 A'_{44}} = \frac{g_3 a_5 \eta_3}{g_4 a_4 \eta_4} \cdot \frac{25 E_s}{2} \quad (G15)$$

A reasonable approximation is that the ratio is about $\frac{25}{2} E_s$. (Here again note that (a_5/a_4) has units of volts⁻¹, which combines with the volts of E_s to give a dimensionless ratio.) Since E_s is the peak voltage in the carrier, it will be of the order of a volt at the milliwatt power level, and again the ratio will certainly be greater than one. Hence, the third harmonic must also be prevented from reaching the mixer.

The ratio of fifth to fourth harmonic contributions is

$$\frac{g_5 \eta_5 A'_{54}}{g_4 \eta_4 A'_{44}} = \frac{g_5 \eta_5 a_5}{g_4 \eta_4 a_4} \cdot \frac{5}{2} E_s. \quad (G16)$$

The $5/2 E$ will be greater than one; its multiplier, it may reasonably be assumed will be less than one. Whether the product will be sufficiently less than one to be completely negligible is not altogether clear without making some approximations for the a_i and g_n coefficients and the efficiencies. It is entirely possible, however, that under certain conditions of shorting plunger positions, phase differences, etc., the fifth harmonic contribution could be appreciable.

So far analysis has shown that the single source system of Figure G1 has a variety of drawbacks. Even with the fundamental and second and third harmonics filtered from the test path, there is some doubt about the system performance unless the fifth harmonic is also filtered out. There is also the problem of spurious paths to the receiver. The isolation from transmitter to receiver at the fundamental frequency can be made to be some 40 or 50 dB in the power divider. However, the second and third harmonics from the harmonic generator can propagate back through the fundamental waveguide in any of several possible mode combinations, and they will not, in general, be well isolated from the receiver branch of the system. It is conceivable, then, that these signals could reach the receiver mixer at a sufficiently high level that they could still be bothersome, particularly if the fourth harmonic test path had appreciable losses.

A brief inspection shows that a vast improvement can be made in the single source system by employing a single sideband modulator. The currents in the harmonic generator would then simply be

$$i = \sum_{j=0}^N a_j \left(m E_s \cos [(\omega + \omega_m)t + \phi] \right)^j. \quad (G17)$$

Then when these are received at the harmonic mixer, all ambiguities are removed, and only the A_4 term (which represents energy propagation at 0.5 mm) will furnish input to the amplifier tuned to $4\omega_m$.

However, the design and development of a good and efficient single sideband modulator at 2 mm would be a much more difficult and expensive task than the purchase of a second power source. Since the original reason for considering the single source system was economy, the single sideband approach was immediately rejected.

The conclusion to be drawn is that the use of a single source system is entirely practical provided a single sideband modulator is available. With the types of modulators available at 2 mm wavelengths which pass both sidebands and the carrier, RF preselection is necessary. This preselection must suppress all lower harmonics, and should, if possible, suppress higher harmonics as well. With this accomplished, there are still potential problems with spurious leakage paths for the lower harmonics, as well as a problem of frequency stability due to the phase dependency. A single source system was rejected for testing components developed on this project.

Appendix H

STRONTIUM TITANATE HARMONIC GENERATOR

1. General Description

An investigation was conducted to determine if a millimeter wave harmonic generator could be built using monocrystalline strontium titanate as the nonlinear element. The investigation was limited to the construction of a single device to function as a tripler of 47 GHz energy. It was operated at the temperature of liquid nitrogen. The goal was to demonstrate harmonic generation. Further development of a strontium titanate harmonic generator would require modification of the waveguide structure and the strontium titanate form and possibly operation at lower temperatures.

2. Nonlinear Properties of Strontium Titanate

Strontium titanate is one of several materials known as perovskites which have a chemical structure of the form ABO_3 . Single crystals of strontium titanate are available in boules weighing up to 80 grams and possibly more. Ceramic mixtures of various proportions of lead and strontium titanates are also available. The latter have been tested for electrical properties in some detail. These materials are all ferroelectric at temperatures below their Curie temperatures which fact makes them of particular interest in some applications. They also have a nonlinear permittivity as a function of applied electric field above their Curie temperatures; this is the characteristic to be used in harmonic generation. The dependence of the dielectric constant (permittivity) and loss tangent on temperature and on dc electric field of ferroelectric ceramics is shown in Figures H1 and H2. Similar characteristics are observed with single crystal $SrTiO_3$ except that the Curie temperature is on the order of $35^\circ K$. These figures illustrate the desirability of maintaining the material at a temperature near the Curie point. As the temperature departs from the Curie point the rate of change of permittivity ($\Delta\epsilon_r$) as a function of voltage change (ΔE_{dc}) decreases markedly.

Since the nonlinear permittivity of $SrTiO_3$ is a bulk rather than a junction phenomenon, its use in high power systems is an attractive alternative to conventional varactors. The nonlinear properties are greatest near its Curie temperature (30° to $40^\circ K$) but still very large at liquid nitrogen temperatures ($77^\circ K$). For the intended harmonic generator application, the large signal characteristics of $SrTiO_3$ are of interest. Some large signal data are available on the properties of ceramic ferroelectrics (H1, H2, H3). These limited data indicate that while the dependence of the dielectric constant on the magnitude of the

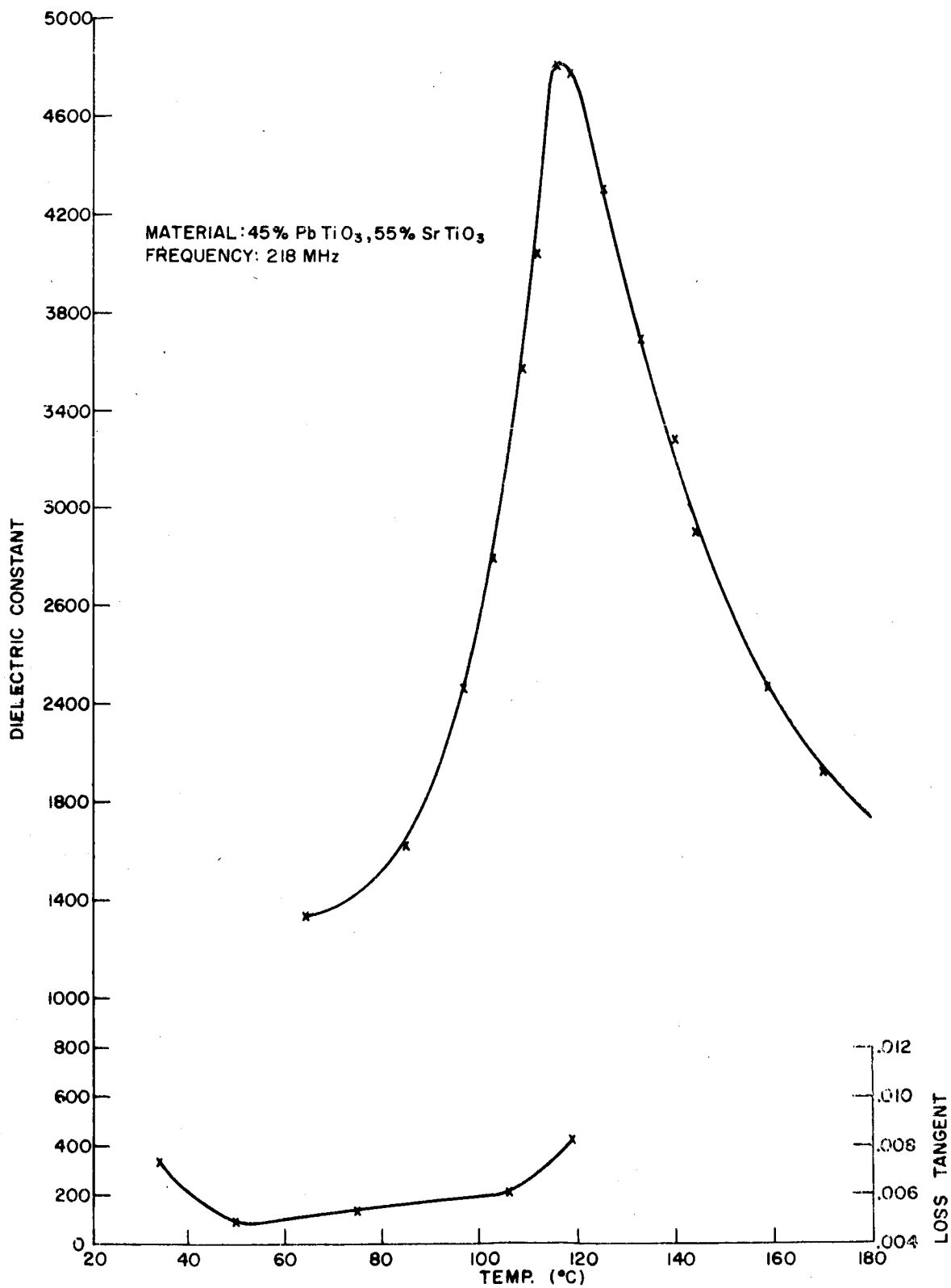


FIG. H1-DIELECTRIC CONSTANT AND LOSS TANGENT vs TEMP-
ERATURE FOR A FERROELECTRIC CERAMIC

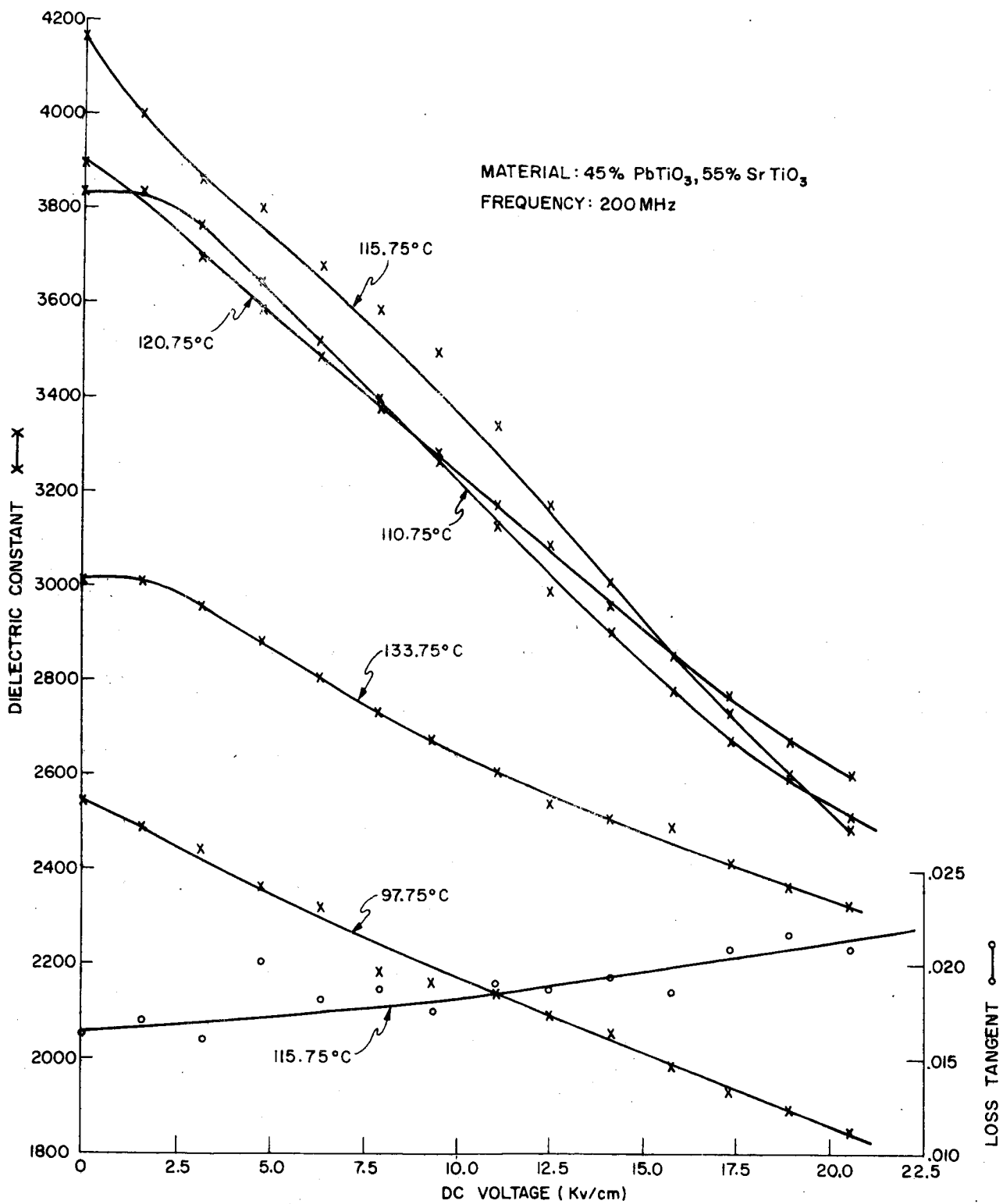


FIG.H.2-DIELECTRIC CONSTANT AND LOSS TANGENT vs DC VOLTAGE FOR A FERROELECTRIC CERRAMIC

RF electric field intensity (E_{RF}) is substantial, it is less than the dc electric field dependence. Recent large signal measurements performed on the material of Figures H1 and H2 yielded the following dependence of large signal dielectric constant on E_{RF} :

$$\epsilon(E_{RF}) = \epsilon_0 (4806 - 3.82 \times 10^{-4} |E_{RF}|), \quad (H1)$$

where E_{RF} is given in volts per meter, and ϵ_0 is the permittivity of free space. Comparison of the above equation with the curves of Figure H2 indicates that $(\Delta\epsilon_r/\Delta E_{RF})$ is nearly comparable to $(\Delta\epsilon_r/\Delta E_{dc})$.

The variation of the electrical properties of ferroelectric materials with frequency is also of interest, but unfortunately very little pertinent data exist. That is particularly true at millimeter wavelengths. (H4) The usual dependence of the electrical properties of ferroelectrics on frequency is illustrated by the curves of Figure H3. For most ferroelectric materials the relaxation frequency, where both dielectric constant and the voltage tunability decrease markedly, occurs in the microwave region. Recent studies of single crystal perovskite structures (strontium-titanate, barium titanate, and various tantalates and niobates) have indicated that the relaxation frequency is above 200 GHz and may extend well into the submillimeter wave region. (H5, H6)

Although nonlinear ferroelectric materials have been used in device applications at microwave frequencies, the fact that their losses increase and tunability decreases with increasing frequency, limits their practical use to the HF, VHF, and UHF bands. An exception to the above is single crystal strontium titanate. (H7, H8) This material has a dielectric loss tangent of less than 0.001 at a frequency of 35 GHz. In terms of varactor terminology, these figures correspond to an unprecedented cutoff frequency of 35,000 GHz. It appears that a significant extension of the frequency range in which nonlinear devices are feasible can be achieved with this material.

Single crystal strontium titanate is not only a unique ferroelectric material, but it is also an electro-optic material. Its transparency at visible wavelengths is comparable to that of high quality optical glass, and its electro-optic coefficient at room temperature is comparable to that of KDP (potassium di-hydrogen orthophosphate) and ADP (ammonium di-hydrogen orthophosphate). At 77°K its electro-optic coefficient is significantly greater than that of the above cited phosphates. The combined characteristics of low optical losses, low microwave losses, and a high

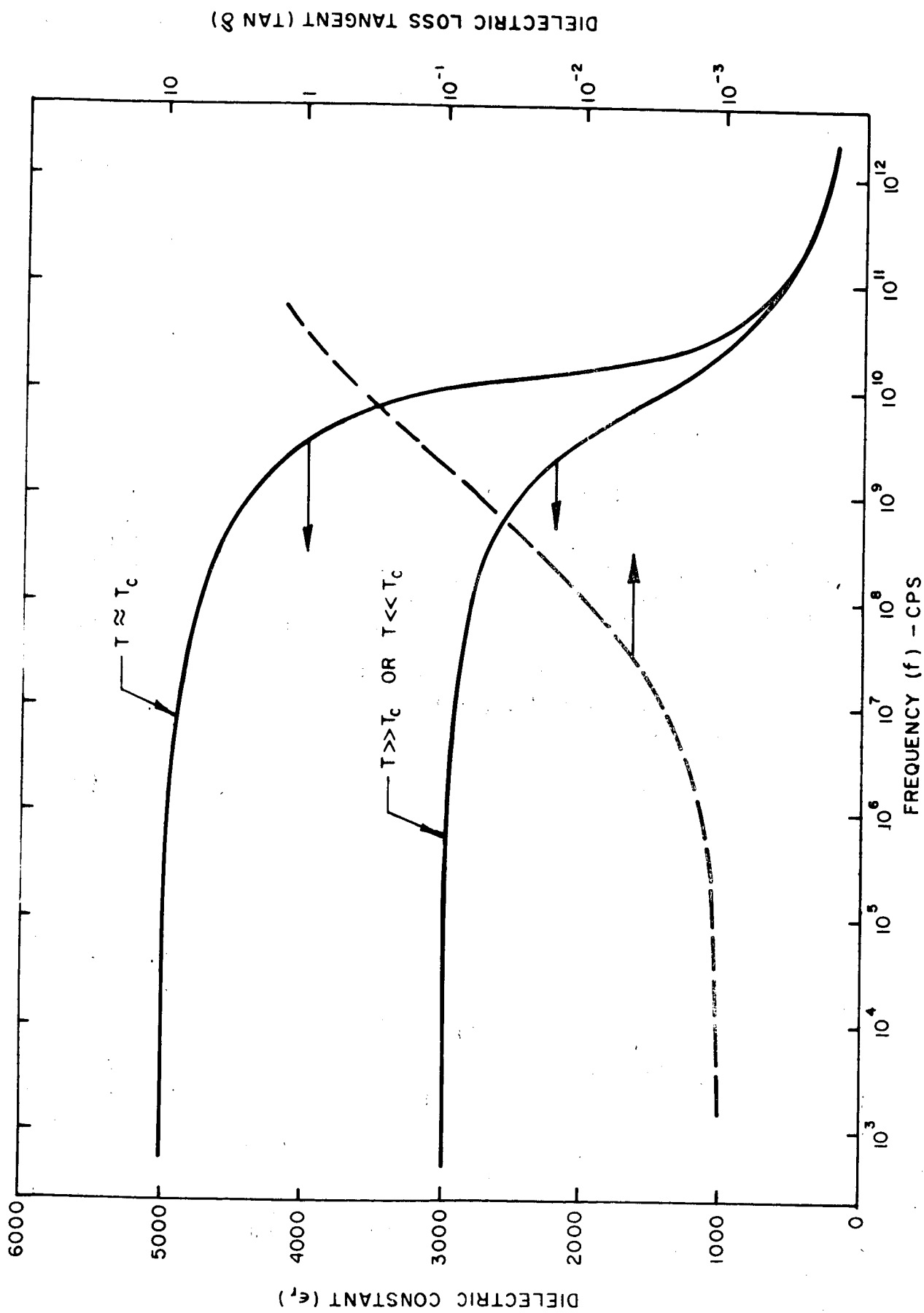


FIG.H.3-TYPICAL DIELECTRIC CONSTANT AND LOSS TANGENT VARIATIONS vs FREQUENCY FOR FERROELECTRICS

electro-optic coefficient indicate that high performance electro-optic devices can be achieved utilizing this material.

Although there are little data available on the electrical properties of SrTiO_3 at millimeter and submillimeter wavelengths, its excellent characteristics at both microwave and visible wavelengths indicate that excellent properties can be anticipated at millimeter wavelengths.

Without any loss of generality, the permittivity dependence on the RF electric field $\epsilon(E_{\text{RF}})$ can be expressed as a power series:

$$\epsilon(E_{\text{RF}}) = \epsilon_0 (K + K_1 E_{\text{RF}} + K_2 E_{\text{RF}}^2 + \cdots K_n E_{\text{RF}}^n + \cdots), \quad (\text{H2})$$

where ϵ_0 is the permittivity of free space and K is the small signal dielectric constant. All of the K_n 's are functions of temperature, dc electric field, and frequency. Previously reported work performed at this laboratory^(H3) on a high power ferroelectric limiter using a ceramic material has shown that the principal nonlinear effect was due to the linear term ($K_1 E_{\text{RF}}$). Other investigations of third harmonic generation^(H2, H9) at microwave frequencies indicated that the results could be explained by the nonlinearity induced via the quadratic term ($K_2 E_{\text{RF}}^2$).

For an unbiased ferroelectric material in its paraelectric phase ($T > T_c$ where T_c is the Curie temperature) the crystal is symmetric; and hence the permittivity is an even function of E_{RF} .

$$\epsilon(E_{\text{RF}}) = \epsilon(-E_{\text{RF}}). \quad (\text{H3})$$

$$K_n = 0, \text{ for } n \text{ odd.}$$

If T is less than or even not sufficiently greater than T_c , or if dc bias is applied, the symmetry condition will not exist. When the symmetry condition exists, only odd harmonics are generated; otherwise, all harmonics are possible. The symmetrical condition is analogous to the nonlinear capacitance-voltage relationship obtained through the use of oppositely biased parallel varactor diodes,^(H10) $C(V) = C(-V)$. An example of the potential benefits obtained via the symmetry of unbiased SrTiO_3 in the paraelectric phase occurs in third harmonic generation, for which there is no longer a need for a tuned circuit at the second harmonic idler frequency. This is the case of interest here for the solid state local oscillator needed for the submillimeter satellite radiometer.

Since harmonic generation in SrTiO_3 is a nonlinear reactive phenomenon, the Manley-Rowe relations apply. Due to the very low

dielectric loss tangent ($\tan \delta$) of single crystal SrTiO_3 , even at millimeter and probably at submillimeter wavelengths, it is nearly a perfect reactance, and hence circuit losses can be the controlling factor in harmonic generation efficiency. The above is analogous to the situation with varactor harmonic generators when $f_{\text{out}}/f_c \ll 1$.

3. Design Description of Experimental SrTiO_3 Harmonic Generator

In addition to the properties of the nonlinear material itself, the properties of the complete harmonic generator structure must be considered if an efficient source of power is to be achieved. An ideal structure (1) permits the application of dc voltage bias if it is required, (2) has provision for efficiently injecting power at the fundamental frequency and extracting the desired harmonic output, (3) takes maximum advantage of the nonlinear characteristics of the active material and (4) adds a minimum of loss to the irreducible amount determined by properties of the material. It should also have provision for controlling the temperature of the active material and extracting heat developed by the harmonic generation process.

An obvious method of applying RF power to a piece of strontium titanate crystal is to mount a crystal as a capacitor in a waveguide. By orienting the leads across the center of the narrow dimension of the waveguide they act as an antenna to receive the RF power and apply it to the capacitor. Similarly, harmonic power generated in the nonlinear capacitor is reradiated into the waveguide. This physical arrangement can conveniently be achieved with the types of waveguide structure used for point-contact diode harmonic generators. The strontium titanate replaces the semiconductor wafer. One lead of the capacitor is the crystal post; the other lead takes the place of the whisker and may, in fact, be of the same form as the whisker.

The variation of the dielectric constant of crystalline strontium titanate as a function of RF field strength at millimeter wavelengths is not known but is assumed to have the general character of the curves of Figure H2. Note that in Figure H2 the curves at some temperatures removed from the Curie temperature have a region of little slope for low values of field strength. It is possible that the dielectric constant has less variation with RF field than with dc field. Therefore, it is desirable to design an experiment to maximize the RF field in the SrTiO_3 capacitor.

Three factors determine the field strength in the strontium titanate capacitor - the magnitude of the applied RF power, the impedance match of the capacitor and leads to the waveguide, and the form of the capacitor.

The experimental mount was designed to use a fundamental input at 6 mm wavelengths where a klystron source having 100 milliwatts of output was available. An order of magnitude increase in applied RF power would require a different type of millimeter wave oscillator; such was not available for this experiment. In order to transfer the maximum power from the waveguide to the capacitor, tuning adjustments were incorporated in the structure. The form of the capacitor could not be chosen solely to maximize the power transfer since the form was affected by other considerations also. The field strength in the dielectric was maximized by minimizing the dielectric thickness. In addition the field was concentrated in the dielectric near one electrode by making that electrode a point contact.

By developing special techniques it was possible to prepare disks of strontium titanate mounted on brass posts of 0.010" and 0.020" diameter and to polish the disks to a thickness as low as 0.0006". A slice of strontium titanate about 0.015" thick was first cut from the bulk material. It was then mounted in salol (phenyl salicylate) on a metal block, lapped on one surface, remounted and lapped on the other surface to about 0.002" thickness. A razor blade was used to cleave the slice into pieces about 1/32" across. DuPont gold paste No. 6976 was applied and fired in several steps to build up a layer about 0.002" thick. The chips were remounted with salol and the gold paste dressed off parallel to the crystal face. It was then necessary to mount the pieces on the posts to be used in the harmonic generator before finishing the crystal to its final thickness. The posts were prepared for soldering by making three diametric vee cuts about 0.0005" deep in the end of the post. The crystal pieces and posts were held in a jig during soldering to maintain parallel surfaces. The solder used was 50% tin, 50% indium to which 10% gold had been added. The crystals were finished to the post diameter by stoning by hand at first and then by stoning in a lathe. The post was placed in a fixture to hold it vertical for the final lapping. Extra fine silicon carbide powder was used with water as the vehicle.

The above procedure resulted in only part of the desired capacitor. One capacitor plate, the gold, was attached to the dielectric and to a conductive lead, the brass post. The second capacitor plate was missing. Gold paste could not then be fired on the other face of the crystal because the solder would not withstand the firing temperature. Furthermore, the gold paste could not have been applied to the other side of the crystal before the solder because the crystal had to be lapped to its final thickness after it was mounted on the post on which it was to be used in the harmonic generator. Since no method could be devised for

bonding a conductor to the unplated side of the mounted crystal, the mount was designed to utilize a pressure contact as the second capacitor electrode.

While the technique for preparing the strontium titanate was being developed, the waveguide structure for the harmonic generator was being designed. The principal problem was to design a structure which could be operated at low temperatures. Although maximum nonlinearity of the SrTiO_3 permittivity is to be expected near the Curie temperature, about 350°K, it was decided to make the first feasibility tests of harmonic generation in a liquid nitrogen bath at 77°K. The two principal methods of achieving lower temperatures – a liquid helium bath or mechanical refrigeration – are much more expensive. Construction and test of a harmonic generator in a liquid nitrogen bath was chosen as a practical first approach. The experiment could then be extended to lower temperatures.

Certain special requirements had to be met by the harmonic generator structure in order that it could be used at low temperatures. The strontium titanate had to be mounted with a good heat-conducting path to the liquid nitrogen. Heat leaks from the bath to the surroundings had to be kept low enough that excessive amounts of liquid nitrogen would not be consumed. Water vapor condensation within the waveguide had to be prevented or high RF losses would have resulted. These requirements meant that an existing waveguide structure could not be used.

An insulated cylindrical flask was selected for the nitrogen bath which has an inside diameter of 4" and an outside diameter of 7 1/2". Its capacity is one liter. Various waveguide layouts and structures were considered but an inline structure was selected for the following reasons: 1) the input and output ports could also be used for introducing tuner motion, thus minimizing heat leakage 2) the structure could be installed easily in the flask by cutting two holes in opposite walls of the flasks 3) the lengths of input and output waveguides could be minimized, thus minimizing RF attenuation.

The electrical features of the mount can be understood by reference to Figure H4. The strontium titanate capacitor is mounted across a waveguide having inside dimensions of 0.224" x 0.112" (equivalent to RG-97/U). The input waveguide slides into this main waveguide from the left. Its inside dimensions are 0.148" x 0.074" (equivalent to RG-98/U). The input waveguide couples to the main waveguide through an iris consisting of a rectangular slot 0.126" x 0.020" x 0.002" thick. The iris was designed to pass 47 GHz and reflect 140 GHz. A test iris measured

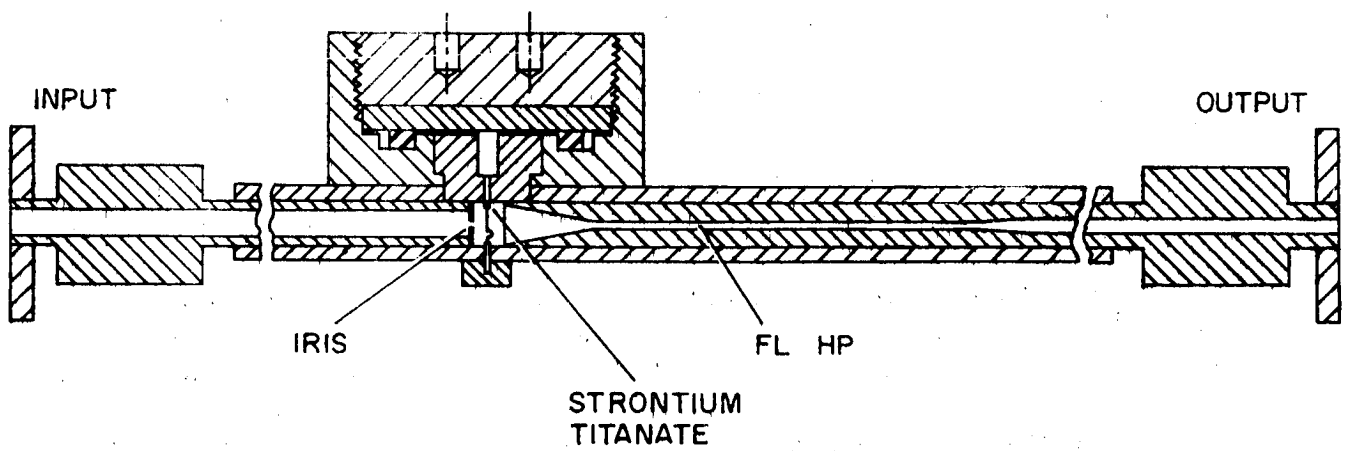


FIG.H.4 - EXPERIMENTAL STRONTIUM TITANATE HARMONIC GENERATOR.

<0.1 dB insertion loss at 47 GHz and 8.8 dB loss at 140 GHz. By sliding the input waveguide in the main waveguide a tuning effect on the harmonic output impedance is achieved.

The output waveguide slides into the main waveguide from the right. Its inside dimensions are 0.080" x 0.040" (equivalent to RG-138/U) at the output. However, it includes a smaller section, 0.051" x 0.0255" (equivalent to RG-135/U), which has a nominal cutoff frequency of 116 GHz and serves as a filter to reflect any fundamental and second harmonic power propagating beyond the strontium titanate. The filter section is 1.00" long and is matched to the main waveguide and to the 0.080" x 0.040" output portion by tapered transitions. The tapers, filter, and output were all electroformed as one piece. A tuning effect on the input impedance is obtained by sliding the output waveguide within the main waveguide.

Water vapor condensation was prevented in the mount by sealing it from the atmosphere and evacuating it. The input and output waveguides were sealed with Teflon tape sandwiched between mating flanges. The sliding joints were sealed with cylindrical metal bellows soldered at the ends thus permitting motion for tuning. The mounting hole for the strontium titanate capacitor was sealed with a polyurethane ring gasket. Although this seal is subjected to the low temperature of the bath it worked satisfactorily.

The crystal post in a diode harmonic generator usually has a differential drive. This feature was not included in this mount because of the difficulty of sealing it and because of the need for a good heat conduction path from the strontium titanate. The structure as shown in Figure H4 assures a good conduction path by applying positive pressure to the crystal post when the threaded retainer is tightened against the pressure plate. The crystal post protrudes slightly above the crystal post housing. The pressure forces the crystal post into intimate contact with the crystal post housing and it in turn with the structure which is immersed in the liquid nitrogen bath.

A pressure contact to serve as the second electrode of the strontium titanate capacitor is provided by a 1 mil whisker wire having a C bend. It is soldered in a 0.020" diameter whisker carrier and inserted in a blind hole in the lower side of the main waveguide. The whisker extends across the waveguide to contact the SrTiO_3 crystal when it is installed. The whisker must be precut to the correct dimension before installation as there is no position adjustment. Note that in this first experimental mount both electrodes are grounded to the waveguide with no provision for an insulated dc lead to either. Such a lead would have introduced

additional sealing problems. Since this device was designed as a third harmonic generator, no dc bias was required. A means of measuring if the whisker were shorted would have been convenient, however.

Figure 5 is a photograph of the complete inline structure assembled in the insulated flask. The main waveguide extends through the walls of the flask so that it may be sealed in place. The bellows which seal the input and output waveguides are mounted outside the flask. They are not visible in the photograph, however, as they are surrounded by cylindrical parts which are threaded internally and serve to position the sliding waveguides.

In the structure which has been described the heat leaks which have been introduced in the bath are the waveguides. These leaks were minimized by making the waveguides with thin wall sections where they pass through the walls of the flask. All three waveguides were electroformed to precision polystyrene mandrels. At the beginning of the process, after about 0.001" of copper had been deposited, sections were masked off where a thin metallic wall was desired. After the rest of the wave had been built up to the desired thickness, the thin wall sections were reinforced with epoxy resin and, in one case, fiberglass cloth. A second output waveguide had to be made as the first ruptured in the thin wall section during the process of dissolving the mandrel. As the time for rework was limited, the second section was made without a thin wall section. This resulted in the formation of frost on the outside of the output waveguide when liquid nitrogen was placed in the flask. However, it did not freeze the tuning adjustment. The total heat leakage of the assembly was such that one filling of the flask with liquid nitrogen would cool it and maintain it at low temperature for an hour or two before the liquid had boiled away to a level below the waveguide structure.

4. Experimental Data

When the construction and assembly of the strontium titanate harmonic generator was complete there was time enough left on the project to begin its evaluation. Some preliminary measurements of the dielectric constant of SrTiO_3 were made at 1 to 2 GHz at room temperature and at about 77°K.

In the measurement of dielectric constant a pellet of SrTiO_3 , 0.0475" dia. and 0.0101" thick, was positioned in a coaxial test fixture forming a capacitance termination to the transmission line. The value of the capacitor was computed from the distance the standing wave null shifted from its position when a short circuit terminated the transmission line. Since the dimensions of the pellet were known, its dielectric constant was determined.

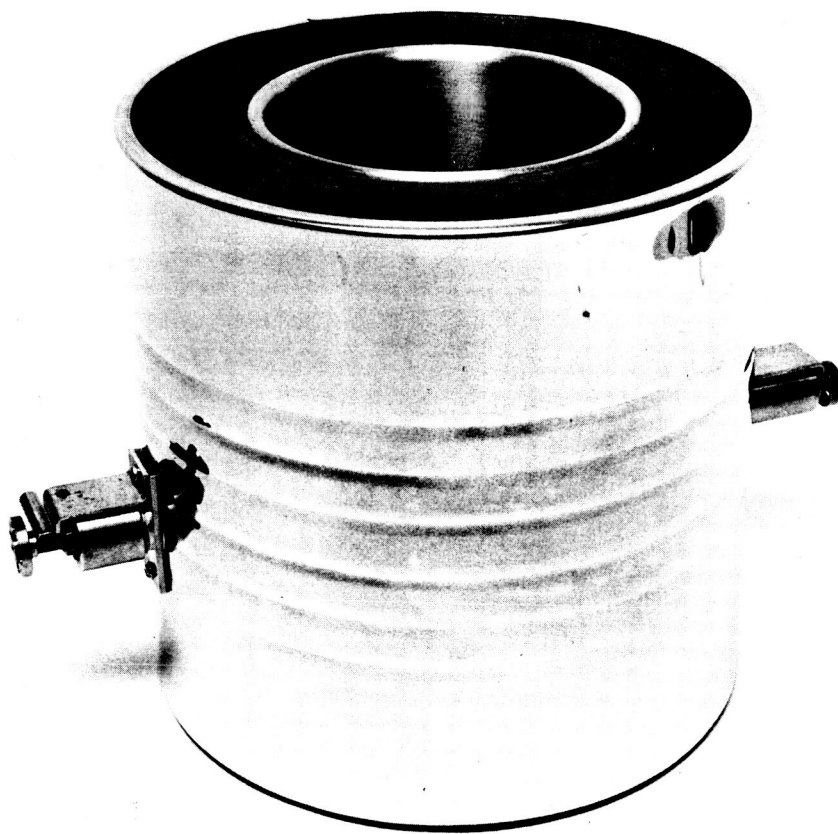


FIG. H-5 - Strontium Titanate Harmonic Generator

The initial measurements proved that the capacitor formed by the test fixture contacts and the pellet was not always reproducible due to variation in the small air gaps between the contacts and the pellet. This difficulty was overcome by depositing gold capacitor plates on the two faces of the test pellet. The test pellet was thereafter positioned in and out of the test fixture several times showing that repeatability of the experimental data had been achieved for room temperature measurements.

When the test fixture was immersed in liquid nitrogen it was initially assumed that the pellet would reach an equilibrium temperature near 77° K; however, the measurement data was found to be quite sensitive to small differences in immersion depths of the test fixture. It was apparent that the temperature of the pellet was in a region where the dielectric constant was changing rapidly as a function of temperature; since a means of measuring the temperature at the pellet had not been provided, it was not possible to relate the low values of temperature to dielectric constant. The following table summarizes the dielectric constant measurement results. Note that at the low temperatures a second, higher value of ϵ_r is given for 1500 MHz and 2000 MHz. These values were achieved by increasing the immersion depth by approximately 1/2 inch over that for the first set of values, and they serve only to indicate the steepness of slope for dielectric constant as a function of temperature.

Frequency, MHz	ϵ_r @ 296° K	ϵ_r @ $\approx 77^\circ$ K
1200	343	1016
1500	343	1220 (2620)
2000	321	1470 (2100)

Measurements were also made at 296° K of dielectric constant as a function of applied power. For power levels between 0.0001 and 7 watts, no significant difference in dielectric constant was detectable. Similar measurements were planned at 77° K but were not carried out when it was found that the experimental setup gave highly variable results due to lack of precise temperature control.

The assembly of the strontium titanate harmonic generator in an insulated flask was described in the previous section. No measurements

were made of the individual waveguides before assembly because they had no flanges or the flanges were attached after assembly. Some VSWR measurements of the assembly were made before the strontium titanate was installed. The VSWR of the input port at 47 GHz varied from 8.5 to 11 and the standing wave pattern moved along the waveguide as the input and output waveguides were adjusted through their ranges of travel. This is evidence that the iris in the input waveguide was transmitting the 47 GHz and the output waveguide was acting as a short circuit at 47 GHz. The VSWR looking in the output port at 140 GHz varied from 1.1 to 5.2 depending upon the tuner settings. The overall insertion loss at 140 GHz from output to input varied from 6.5 to 20 dB as the waveguides were adjusted. These data suggest that the region between sliding waveguides was acting as a tunable cavity due to discontinuities at the iris and at the taper transition. Since the attenuation of the output waveguide alone at 140 GHz is unknown, it is impossible to eliminate its effect on the reflection from the iris.

A wafer of strontium titanate 0.010" x 0.0012", mounted on a post, was installed in the harmonic generator together with a phosphor bronze whisker for the second electrode. The capacitor cannot be inspected after assembly nor is there any test which can be made to check that the whisker is making contact. However, when the whisker was later removed and examined, it was observed that the tip was deformed. Therefore, it must have made contact with the crystal or the adjacent waveguide wall. A vacuum pump was connected to the waveguide and turned on. The flask was filled with liquid nitrogen. Unfortunately, there is no way of knowing if the whisker is in contact with the crystal when cooled to the bath temperature except by observation of harmonic generation. Approximately 100 mW average power at 47 GHz was applied to the input port. No harmonic output could be detected. The tuners were adjusted and the input frequency varied to no avail.

During a second experiment, a third harmonic output was detected but it remained present when the strontium titanate was removed and proved to be harmonic energy generated in the klystron. The power level at the output of the harmonic generator assembly was 56 dB below the fundamental level when the waveguide tuners were optimized for maximum harmonic output.

No time was available for further attempts to measure harmonic generation in strontium titanate. The tests that were performed were in no way exhaustive. The question of the feasibility of harmonic generation in strontium titanate is unresolved.

5. Conclusions and Recommendations

A start has been made on an experimental investigation of harmonic generation in strontium titanate. Techniques for preparing thin, mounted wafers of crystalline SrTiO_3 have been developed. An experimental third harmonic generator structure was constructed which could be operated in a liquid nitrogen bath. Its major drawback is the difficulty of making a second connection to the strontium titanate wafer and knowing if the contact has been made. For this reason it is necessary that a series of trial measurements be made in order to attain a reasonable probability that contact was achieved.

The possible presence of harmonics from the fundamental drive source must be guarded against. A source of greater power would be desirable so that some isolating attenuation could be inserted between the source and the harmonic generator. This would prevent interaction between the harmonic generator and the source which could increase the source harmonic output. Greater fundamental power would also make it easier to detect any third harmonic generated in the strontium titanate. The first goal, of course, is to detect any harmonic generation. Once the presence of the effect is established it will be possible to vary the conditions to maximize the efficiency.

Any air gap between the whisker and the strontium titanate introduces series capacity and seriously reduces the field strength in the strontium titanate. It might be advantageous to vacuum deposit a grid of small conductive spots on the mounted crystal with the expectation that on assembly the whisker would make contact with one of them. The dot contacted would be the second plate of the capacitor; there would be no possibility for a series air gap between the dot and the crystal.

Further investigation of strontium titanate harmonic generation should include provision for modifying the waveguide structure and bath. Better access for easier installation of the SrTiO_3 capacitor would be a big time-saver. A polyfoam container for the liquid nitrogen that could be cut open and repaired or replaced after each experiment is a possibility to consider. If a mechanical cryostat with a cold finger could be obtained for use on the project the flask might be dispensed with and different mount configurations would become feasible. Also, lower temperatures could be achieved. Lower temperatures may be essential in order to achieve a detectable amount of harmonic generation.

The experiments which were performed on this project were insufficient in number and in variety of conditions to determine the feasibility of harmonic generation in strontium titanate at millimeter wavelengths. It is recommended that work be continued with the structure already built and with improved structures and experimental techniques.

Appendix I

140 GHz TRIPLER USING COMMERCIAL VARACTOR

A commercial varactor with one of the highest cutoff frequencies is Texas Instruments Type TIXVO4, which is rated at 300 GHz for -2V bias and breakdown at -6V. It is built in a beryllium oxide miniature pill package. One of these was purchased for evaluation. The particular diode received was stated to have $f_{co} = 477$ GHz at -2V bias. A waveguide mount was built for it and is shown in Figure I1 together with the varactor. It is designed to be operated as a third harmonic generator. Its input waveguide is RG-97/U, which is used for 33 to 50 GHz. Its output is RG-138/U size, but it has an internal section of smaller waveguide which has a low frequency cutoff of approximately 100 GHz. The mount was tested with the TIXVO4 varactor and 50 milliwatts of power at 57 GHz applied. The third harmonic was detected at a very low level. The transducer loss was 49 dB. The poor over-all efficiency may be due to losses in the waveguide mount and in the pill package which degrade the inherent conversion efficiency of the varactor junction. This varactor tripler is certainly inadequate as a 140 GHz local oscillator.

It should be noted that the manufacturers of these commercial varactors do not claim or expect they will work at millimeter wavelengths. It is possible that selected diodes might work, but they are much too expensive to try this approach. The TIXVO4 diode or equivalent might be used in the next lower frequency stage of the 140 GHz harmonic generator chain.

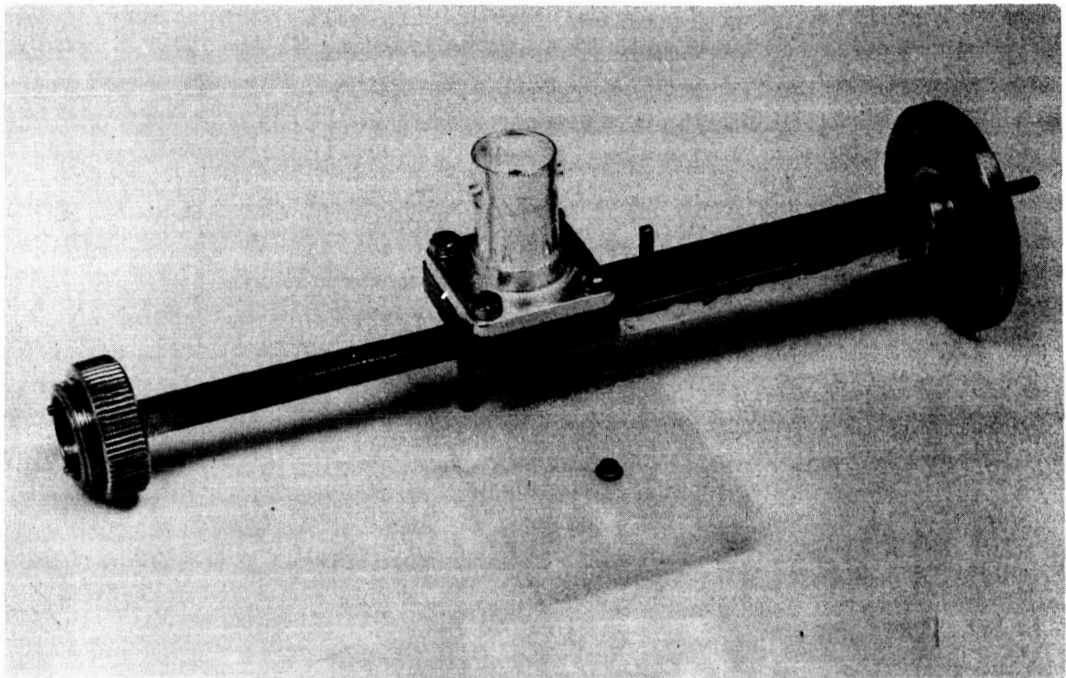


FIG. I-1 - 140 GHz Tripler Using Commercial Varactor

REFERENCES

- (1-1) S.Y. Chang and J.D. Lester, "Radiometric Measurement of Atmospheric Absorption at 600 Gc," Proc. IEEE, vol. 54, pp. 459-461; April, 1966.
- (1-2) R.F. Packard, "Submillimeter Radiometer for the Detection and Acquisition of Ground Targets," Advanced Technology Corp., Final Rept. on USAMC Contract No. DA 36-034 ORD-3765-RD, October 30, 1964; AD 464 838.
- (1-3) E.S. Rosenblum, "Atmospheric Absorption of 10-400 kMcps Radiation: Summary and Bibliography to 1961," Microwave J., vol. 4, pp. 91-96; March, 1961.
- (1-4) J. Strong and G.A. Vanasse, "Interferometric Spectroscopy in the Far Infrared," J. Opt. Soc. Am., vol. 49, pp. 844-850; September, 1959.
- (1-5) R.J. Coates, "Measurement of Solar Radiation and Atmospheric Attenuation at 4.3-millimeter Wavelength," Proc. IRE, vol. 46, pp. 122-126; January, 1958.
- (1-6) M. Cohn, F.L. Wentworth, J.C. Wiltse, "High Sensitivity 100 to 300 Gc Radiometers," Proc. IEEE, vol. 51, pp. 1227-1232; September, 1963.
- (1-7) C.H. Mayer, "Thermal Radio Radiation from the Moon and Planets," IEEE Trans., vol. AP-12, pp. 902-913; December, 1964.
- (2-1) A.F. Harvey, "The Electroforming of Components and Instruments for Millimeter Wavelengths," Proc. IEE (London), vol. 102, pt. B, pp. 223-230; March, 1955.
- (2-2) J.D. Rodgers, editor, "Techniques for Superheterodyne Receivers Above 100 Gigacycles," Research Div., Electronic Communications, Inc., (now Advanced Technology Corp.), Final Rept. on NASA Contract No. NASw-662; April, 1964.
- (2-3) J.M. Cotton, "A Submillimeter Superheterodyne Receiver," IEEE Trans., vol. MTT-11, pp. 385-389; September, 1963.
- (2-4) H.C. Torrey and C.A. Whitmer, Crystal Rectifiers, M.I.T. Radiation Laboratory Series, vol. 15, pp. 30-32, McGraw Hill Book Co., Inc., New York, N.Y.; 1948.

- (3-1) A. Uhlir, Jr., "The Potential of Semiconductor Diodes in High Frequency Communications," Proc. IRE, vol. 46, pp. 1099-1115; June, 1958.
- (3-2) J.M. Manley and H.E. Rowe, "Some General Properties of Non-linear Elements - Part I. General Energy Relations," Proc. IRE, vol. 44, pp. 904-913; July, 1956.
- (3-3) Catalogue SF 4002, Microwave Associates, Burlington, Mass.; February, 1965.
- (3-4) Demonstration by Sylvania Electric Products, Inc., at the 1964 International Solid State Circuits Conference, Philadelphia, Pa.
- (3-5) "Solid-state Source Reaches 2 mW at 60 Gc," Electronic Design, vol. 13, pp. 72-73; September 27, 1965.
- (3-6) B.C. DeLoach, "A 54 Gc Parametric Amplifier and Doubler," Proc. IEEE, (Correspondence), vol. 51, p. 1153; August, 1963.
- (3-7) See Reference (2-2), pp. 66-75.
- (3-8) R.H. Pantell, "General Power Relationships for Positive and Negative Nonlinear Resistive Elements," Proc. IRE, vol. 46, pp. 1910-1913; December, 1958.
- (3-9) C.M. Johnson, D.M. Slager, D.D. King, "Millimeter Waves from Harmonic Generators," Rev. Sci. Instr., vol. 25, pp. 213-217; March, 1954.
- (3-10) See Reference (2-4), pp. 68-107.
- (3-11) C.T. Sah, R.N. Noyce and W. Shockley, "Carrier Generation and Recombination in P-N Junctions, and P-N Junction Characteristics," Proc. IRE, vol. 45, pp. 1228-1243; September, 1957.
- (3-12) E. Spence, Electronic Semiconductors, pp. 84, 111, McGraw Hill Book Co., Inc., New York, N.Y.; 1958.
- (3-13) W.M. Sharpless, "Wafer-Type Millimeter Wave Rectifiers," Bell Sys. Tech. J., vol. 35, pp. 1385-1402; November, 1956.
- (3-14) W.M. Sharpless, "Gallium-Arsenide Point-Contact Diodes," IRE Trans., vol. MTT-9, pp. 6-10; January, 1961.
- (3-15) R.J. Bauer, M. Cohn, J.M. Cotton, Jr., R.F. Packard, "Millimeter Wave Semiconductor Diode Detectors, Mixers, and Frequency Multipliers," Proc. IEEE, vol. 54, pp. 595-605; April, 1966.

- (3-16) F.L. Wentworth, J.W. Dozier, J.D. Rodgers, "Millimeter Wave Harmonic Generators, Mixers, and Detectors," Microwave J., vol. 7, pp. 69-75; June, 1964.
- (3-17) D. Kahng, "Au-n-Type GaAs Schottky Barrier and Its Varactor Application," Bell Sys. Tech. J., vol. 43, pp. 215-224; January, 1964.
- (3-18) P. Penfield, Jr., and R.P. Rafuse, Varactor Applications, p. 329, The M.I.T. Press, Cambridge, Mass.; 1962.
- (3-19) L.K. Staley, "A 'Goodness' Criterion for any Varactor Diode as a High Order Frequency Multiplier," Digest, 1963 International Solid State Circuits Conference, Philadelphia, Pa.; February, 1963.
- (3-20) J.J. Wortman, "Effect of Mechanical Strain on P-N Junctions," NASA CR-275, Prepared under Contract No. NASr-222 by Research Triangle Institute, Durham, N.C.; August, 1965.
- (3-21) J.A. Morrison, "Maximization of the Fundamental Power in Nonlinear Capacitance Diodes," Bell Sys. Tech. J., vol. 41, pp. 677-722; March, 1962.
- (3-22) B.C. DeLoach, "A New Microwave Measurement Technique to Characterize Diodes and an 800 Gc Cutoff Frequency Varactor at Zero Volts Bias," IEEE Trans., vol. MTT-12, pp. 15-20; January, 1964.
- (3-23) C.A. Burrus, "Formed-Point-Contact Varactor Diodes Utilizing a Thin Epitaxial Gallium Arsenide Layer," Proc. IEEE (Correspondence), vol. 51, pp. 1777-1778; December, 1963.
- (3-24) See Reference (2-2), pp. 58-62.
- (5-1) R.J. Bauer and J.W. Dozier, "Diode, Millimeter Wave Type Mixer (3.2 mm)," Advanced Technology Corp., Sixth Quarterly Prog. Rept. on USAERDL Contract No. DA 36-039 AMC-02347(E); 31 December 1964, AD 615 743.
- (5-2) R.J. Bauer and L.E. Dickens, "Diode, Millimeter Wave Type Mixer (3.2 mm)," Advanced Technology Corp., Ninth Quarterly Prog. Rept. on USAERDL Contract No. DA 36-039 AMC-02347(E); 28 February 1966.
- (5-3) W.H. Hord, Jr., and T.J. Greenwell, "Environmental Design Qualification Test, OGO Experiments," Goddard Space Flight Center, T and E Specification No. S-4-101, Rev. A, February 8, 1963.

- (5-4) C.A. Burrus, Jr., "Backward Diodes for Low-Level Millimeter-Wave Detection," IEEE Trans., vol. MTT-11, pp. 357-362; September, 1963.
- (A1) E.H. Putley, "Impurity Photoconductivity in n-type InSb," Proc. of the Physical Society (Correspondence), vol. 76, pp. 802-805; November, 1960.
- (A2) E.H. Putley, "Impurity Photoconductivity in n-type InSb," J. of the Physical Chemistry of Solids, vol. 27, pp. 240-247; December, 1961.
- (A3) E.H. Putley, "The Detection of Sub-mm Radiation," Proc. IEEE, vol. 51, pp. 1412-1423; November, 1963.
- (A4) R. Meredith and F.L. Warner, "Superheterodyne Radiometers for Use at 70 Gc and 140 Gc," IEEE Trans., vol. MTT-11, pp. 397-411; September, 1963.
- (A5) See Reference (1-2).
- (A6) R.J. Bauer, "A Low Noise Figure 94 Gc Gallium Arsenide Mixer Diode," Digest, 1965 G-MTT Symposium, Clearwater, Fla.; May, 1965.
- (A7) F.J. Low, "Performance of Thermal Detection Radiometers at 1.2 mm," Proc. IEEE (Correspondence), vol. 53, p. 516; May, 1965.
- (A8) F.J. Low, "Thermal Detection Radiometry at Short Millimeter Wavelengths," Proc. IEEE, vol. 54, pp. 477-484; April, 1966.
- (A9) "Low Temperature Gallium Doped Germanium Bolometer Systems," Product Bulletin C163, Apparatus Div., Texas Instruments, Inc., Houston, Texas.
- (B1) T. Moreno, Microwave Transmission Design Data, p. 140, McGraw-Hill Book Company, Inc., New York, N.Y.; 1948.
- (B2) G.C. Southworth, Principles and Applications of Waveguide Transmission, pp. 110-112, D. Van Nostrand Company, Inc. Princeton, N.J.; 1958.
- (D1) See Reference (3-13).
- (D2) G.C. Messenger and C.T. McCoy, "Theory and Operation of Crystal Diodes as Mixers," Proc. IRE, vol. 45, pp. 1269-1283; September, 1957.

- (D3) R. Meredith, F.L. Warner, Q.V. Davis, J.L. Clarke, "Superheterodyne Radiometers for Short Millimeter Wavelengths," Proc. IEE (London), vol. 111, pp. 241-256; February, 1964.
- (D4) D.A. Jenny, "A Gallium Arsenide Microwave Diode," Proc. IRE, vol. 46, pp. 717-722; April, 1958.
- (D5) See Reference (2-4), p. 24.
- (D6) R.J. Bauer, "Diode, Millimeter Wave Type Mixer (3.2 mm)," Advanced Technology Corp., Fourth Quarterly Prog. Rept. on USAERDL Contract No. DA 36-039 AMC-02347(E), p. 21; 30 June 1964, (AD 447 083).
- (D7) See Reference (2-4), p. 421.
- (E1) S.T. Eng, "Low-Noise Properties of Microwave Backward Diodes," IRE Trans., vol. MTT-9, pp. 419-425; September, 1961.
- (E2) See Reference (5-4).
- (F1) See Reference (2-2), pp. 78-80.
- (H1) D.A. Johnson, "Microwave Properties of Ceramic Nonlinear Dielectrics," Microwave Lab. Rept. No. 825, Stanford Univ., Contract No. AF 49(628)-514; July, 1961.
- (H2) M. DiDomenico, D.A. Johnson, R.H. Pantell, "A Ferroelectric Harmonic Generator and the Large Signal Microwave Characteristics of a Ferroelectric Ceramic," Internal Memorandum M.L. No. 862, Stanford Univ.; October, 1961.
- (H3) M. Cohn and A.F. Eikenberg, "A High Power Ferroelectric Limiter," IEEE Trans., vol. MTT-13, pp. 47-54; January, 1965.
- (H4) D.A. Johnson, K.B. Mallory, R.H. Miller, R.H. Pantell, P.A. Szente, "Small Signal Characteristics of Ferroelectric Ceramics at Millimeter Wavelengths," Proc. IEEE, vol. 51, pp. 332-339; February, 1963.
- (H5) J.E. Geusic, S.N. Kurtz, L.G. Van Uitert, S.H. Wemple, "Electro-Optic Properties of Some ABO_3 Perovskites in the Paraelectric Phase," Applied Physics Letters, vol. 4, pp. 141-143; April 15, 1964.

- (H6) F.S. Chen, J.E. Geusic, S.K. Kurtz, J.G. Skinner, S.H. Wemple, "The Use of Perovskite Paraelectrics in Beam Deflectors and Light Modulators," Proc. IEEE (Correspondence) vol. 52, pp. 1258-1259; October, 1964.
- (H7) G. Rupprecht, "Investigation of the Microwave Properties of Ferroelectrics," Res. Div. Raytheon Co., Scientific Rept. No. 1, Contract No. AF 19(604)-4085; June, 1960.
- (H8) G. Rupprecht, B.D. Silverman, R.J. Bell, "Investigation of the Microwave Properties of Ferroelectrics," Res. Div. Raytheon Co., Final Rept., Contract No. AF 19(604)-4085; October, 1960.
- (H9) D.A. Johnson, "Microwave Properties of Ceramic Nonlinear Dielectrics," Microwave Lab. Rept. No. 825, Contract No. AF 49(638)-514, Stanford University; July, 1961.
- (H10) W.R. Avellino, "Efficient Odd Harmonic Generation With Oppositely Biased Varactor Diodes," Proc. IEEE (Correspondence), vol. 52, pp. 868-869; July, 1964.