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SUBMILLIMETER SATELLITE RADIOMETER

by

R.F. Packard, M. Cohn, R.J. Bauer,
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Prepared under Contract NASw-1000

for

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

ADVANCED TECHNOLOGY CORPORATION
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Technical Supervision: M. Cohn

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ABSTRACT

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This is an interim report on the status of the development of the two critical components needed for a space-borne submillimeter superheterodyne radiometer. They are a 560 GHz fourth harmonic mixer and a 140 GHz third harmonic generator. An experimental model of the latter has been built and tested using a commercial packaged varactor. Results were poor but better performance is expected with epitaxial varactors now being assembled directly in waveguide structures.

Manufacturing techniques have been developed for the construction of precision submillimeter waveguides for use in the 560 GHz mixer and in a signal source built to test the mixer. The mixer has not yet been completed but diode materials for use in it have been studied in millimeter wave devices. Tunnel diodes made with available materials were unsatisfactory in millimeter devices; backward diodes were no better than point contact devices. Data measured with a 140 GHz second harmonic mixer is presented. An analysis of the series parasitic resistance of an idealized diode has been made and suggestions for its reduction are given so that improved mixer performance may be achieved.

Other material included in this report is an analysis of a single oscillator source system for testing harmonic mixers, results of crossed waveguide coupling studies and a comparison of superheterodyne receivers with direct detectors for submillimeter systems. At this stage of the project no conclusions can be drawn as to the choice of 140 GHz for the connection between solid state local oscillator and harmonic mixer or as to future system performance. Experiments are in progress to provide data upon which to base such conclusions in the next period.

Author

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I. STATEMENT OF WORK

This report describes the work accomplished to date on Phase I of a project to develop a satellite-borne submillimeter wavelength radiometer. Phase I (of 18 months duration) includes research, development and testing of components, and a study of system feasibility. The contract Statement of Work calls for the following:

> "The Contractor shall evaluate available components, determine the inadequacies of available components and accomplish the necessary component development to provide a submillimeter radiometer capable of operating in orbit.
>
> Specifically, the study shall be conducted to reduce power requirements, size and weight, increase lifetime, and enable the instrumentation listed below to withstand the rigor of the anticipated environmental conditions.
>
> A. Harmonic Mixer: Develop a rugged, sensitive 600 Gc* harmonic mixer, whose local oscillator power requirements are compatible with the solid state source to be developed. The mixer shall have an IF output which is compatible with available rugged, lightweight, low power drain, sensitive, broadband, solid state IF amplifiers.
>
> B. Solid State Local Oscillator Source: Develop an efficient and rugged millimeter wave harmonic generator which (1) can be driven by available lower frequency solid state sources, and (2) provide sufficient output power to serve as a local oscillator for the harmonic mixer to be developed. The performance of the above harmonic mixer and harmonic generator will be determined with the aid of the ferroelectric bolometer developed under Contracts NASW-259 and NASW-662.

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* GHz is the abbreviation for gigahertzian which is now recommended for use instead of gigacycle per second and will be used throughout this report. A frequency of 600 GHz corresponds to a wavelength of 0.5 millimeter.

C. Ferrite Modulator: Develop a rugged 600 Gc ferrite switch for use as a modulator for the proposed submillimeter wavelength radiometer. It shall be capable of a 1 Kc switching rate. Minimum insertion loss and maximum isolation over a broad bandwidth shall be the design goal.

D. Antenna System: A study shall be conducted to determine the antenna configuration which is most compatible with the proposed 600 Gc radiometer, the space vehicle, and the type of radiometric observation of interest to NASA's Astronomy Program.

E. System Evaluation and Anticipated Performance: Upon determining the feasibility and performance characteristics of such components as the harmonic mixer, harmonic generator, and ferrite modulator, the sensitivity, weight, volume, and primary power requirements of the complete radiometer shall be predicted."

II. APPROACH

Work during this first reporting period has been devoted primarily to the two most critical components of the desired radiometer, namely, a 0.5 millimeter harmonic mixer and a solid state local oscillator. Each is an essential component; each requires development and extension of the best previous designs. The success of the development of these two critical components will determine the requirements for the ferrite modulator and antenna system. Hence, development and study of these latter items has been deferred until the second half of Phase I. Similarly, system evaluation must await the development of the components.

A 0.5 millimeter second harmonic mixer was designed, built, and operated successfully in a superheterodyne radiometer under contract between this laboratory and the U. S. Army Munitions Command, Frankford Arsenal.[1] The minimum temperature difference which could actually be discerned with that radiometer was 26°K with a signal-to-noise ratio estimated to lie between 2.5 and 8.3. That radiometer performance was limited primarily by the harmonic mixer conversion loss, which was deduced from the above-quoted measurement to lie between 23 and 28 db. Operating conditions of the radiometer included 10 sec output integration time, about 4 GHz effective RF bandwidth as determined by the IF amplifier passband, and about 50 milliwatts of local oscillator power.

The crux of the development of an all-solid-state 0.5 millimeter radiometer is to generate the most possible power at the highest possible frequency with a solid-state local oscillator and to design a harmonic mixer of minimum conversion loss to operate with the available LO power. The choice of frequency at which the connection between local oscillator and harmonic mixer is made is critical. The specific approach we have chosen is to develop a fourth harmonic mixer with low conversion loss for minimum LO drive power. The fourth harmonic mixer will be able to use a 2 mm LO source instead of the 1 mm source required for the second harmonic mixer previously referred to. Both fourth harmonic mixer and solid state local oscillator will require further development of previous designs. The choice of 2 mm wavelengths for the interconnection is our estimate of the best compromise for dividing the development problems between the two devices.

Data on the existing state-of-the-art for solid state millimeter wave generators is given in the final report on a previous contract, NASw-662, completed at this laboratory.[2] An output of one milliwatt at 2 mm was achieved with an input of twelve milliwatts at 4 mm. On the present contract effort is being directed toward developing a third harmonic generator at 2 mm wavelengths to use as the final stage in a solid state local oscillator. Varactor diodes are being tested as millimeter harmonic generators with the expectation of achieving better efficiency and, hence, greater output.

The design, construction, test preparation, and development accomplished so far on this project are described in the following sections designated by specific tasks.

III. HARMONIC MIXER

A principal means of obtaining improved mixer performance is to develop improved diodes by using different materials and/or different diode forming techniques. Mixer performance may also be improved by designing and constructing better waveguide structures. Development of a 0.5 millimeter fourth harmonic mixer has been initiated with work in both of these general areas. In addition, means for testing a 0.5 mm mixer have been provided.

A. Harmonic Mixer Testing at 0.5 mm

In order to test an experimental 0.5 mm harmonic mixer simultaneously with the development of a solid-state LO it was necessary to obtain a 2 mm signal source. A Carcinotron* Type COE-20C was selected and purchased for this purpose since it was available promptly, and it would have a sufficient amount of output power for our testing purposes. A high voltage power supply for this tube was also purchased. The particular tube received oscillates between 126 and 141 GHz. During the testing of the mixer it will be operated at 140 GHz where it will generate up to 700 milliwatts of power. However, considerably reduced power will be adequate and should extend the life of the Carcinotron substantially. As a result of the choice of this 2 mm signal source the experimental mixer will detect signals at about 560 GHz.

1. Analysis of Single Source System

A variety of possible test systems at 560 GHz 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 the 140 GHz 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 1.

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

$$i = a_1 e + a_2 e^2 + a_3 e^3 + \cdots \qquad (1)$$

where

i = instantaneous diode current

a_i = amplitude coefficient

e = instantaneous voltage.

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* Compagnie générale de télégraphie Sans Fil, Paris, France, registered trade-mark for backward wave oscillator tubes.

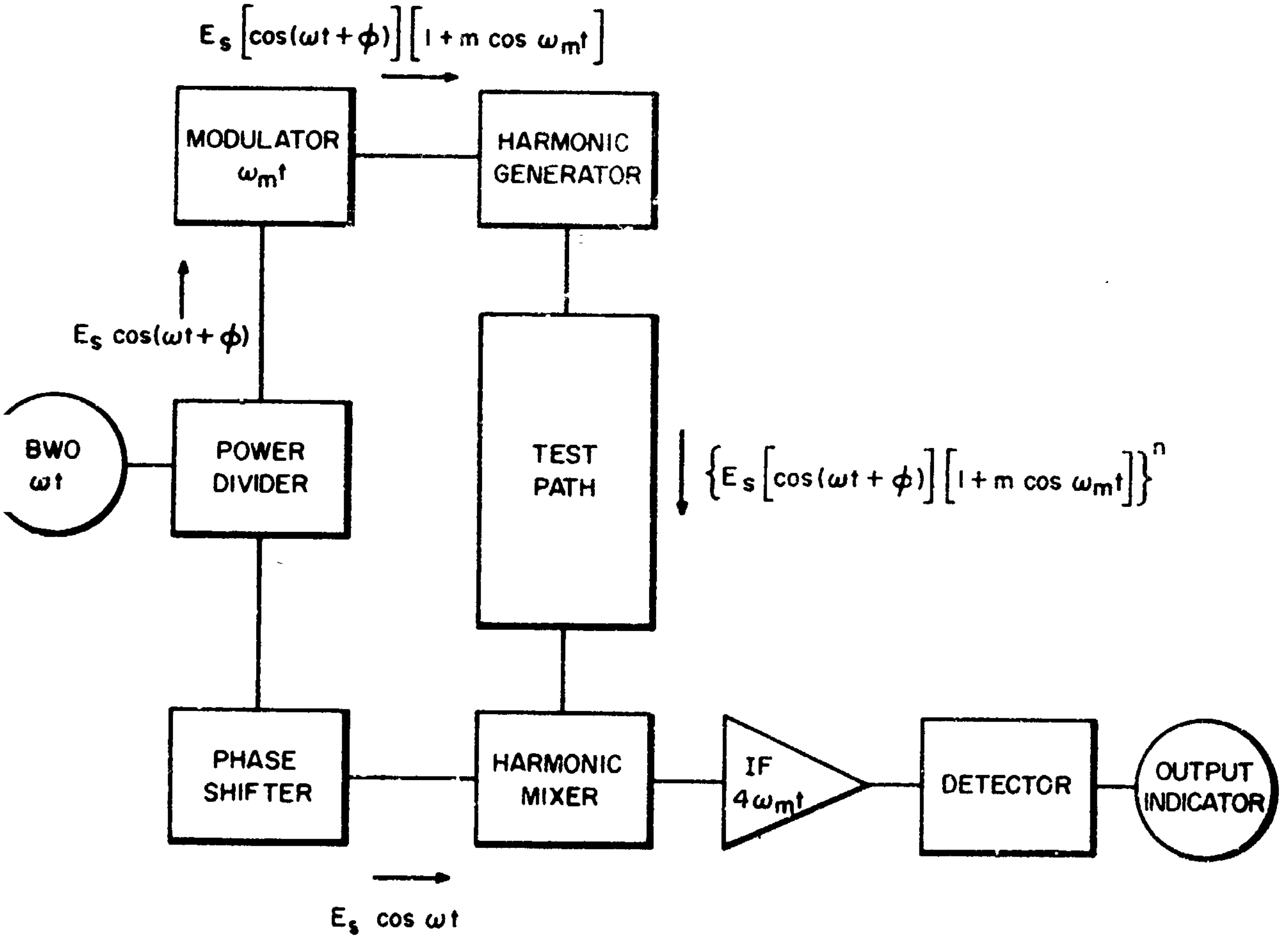

FIG. 1 —SINGLE SOURCE TEST SYSTEM

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

$$e = E_s[\cos(\omega t + \phi)]\,[1 + m\cos\omega_m t] \tag{2}$$

where

E_s = peak carrier voltage

ω = source frequency

ϕ = phase difference

m = modulation index

ω_m = angular modulation frequency.

Substituting this value of e into Equation (1), 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

$$\begin{aligned} i &= a_1E_s[\cos(\omega t + \phi)][1 + m\cos\omega_m t] \\ &+ \frac{a_2E_s^2}{2}[1 + \cos 2(\omega t + \phi)][1 + 2m\cos\omega_m t + m^2\cos^2\omega_m t] \\ &+ \text{similar higher order terms.} \end{aligned} \tag{3}$$

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

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

$$i = \sum_{k=0}^{N}\sum_{\ell=0}^{N} A'_{k\ell}\cos(k\omega t \pm \ell\omega_m t + k\phi) \tag{4}$$

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

$$g = g_o + g_1 \cos \omega t + g_2 \cos 2\omega t + g_3 \cos 3\omega t + \cdot \cdot \cdot \tag{5}$$

Now the currents which will produce signals in the IF amplifier are $i = g\,e$ where g has been defined in Equation (5) and e is the output from the harmonic generator. Assume

$$e = \sum_{k=o}^{N} \sum_{\ell=o}^{N} A_{k\ell} \cos (k\omega t \pm \ell \omega_m t + k\phi) \tag{6}$$

where

$$A_{k\ell} = \eta_k A'_{k\ell} \; . \tag{7}$$

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 (5) and (6) gives

$$i = \left(\sum_{n=o}^{\infty} g_n \cos n\omega t \right) \left(\sum_{k=o}^{N} \sum_{\ell=o}^{N} A_{k\ell} \cos (k\omega t \pm \ell \omega_m t + k\phi) \right) \tag{8}$$

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 $\ell = 4$. 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.,

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

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 $A'_{k\ell}$ coefficients requires that one actually expand Equation (3) and collect terms. The amplitudes of the resulting signals at the IF input, neglecting harmonics higher than the fifth are

$$g_1 \eta_1 A'_{14} = g_1 \eta_1 a_5 \frac{25m^4}{128} E_s^5 + \cdots \quad \text{(contribution due to fundamental)} \tag{9}$$

$$g_2 \eta_2 A'_{24} = g_2 \eta_2 a_4 \frac{m^4}{32} E_s^4 + \cdots \quad \text{(2nd harmonic)} \tag{10}$$

$$g_3 \eta_3 A'_{34} = g_3 \eta_3 a_5 \frac{25m^4}{256} E_s^5 + \cdots \quad \text{(3rd harmonic)} \tag{11}$$

$$g_4 \eta_4 A'_{44} = g_4 \eta_4 a_4 \frac{m^4}{128} E_s^4 + \cdots \quad \text{(4th harmonic)} \tag{12}$$

$$g_5 \eta_5 A'_{54} = g_5 \eta_5 a_5 \frac{5m^4}{256} E_s^5 + \cdots \quad \text{(5th harmonic)} \tag{13}$$

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

$$\frac{g_2 \eta_2 A'_{24}}{g_4 \eta_4 A'_{44}} = 4 \frac{g_2 \eta_2}{g_4 \eta_4} . \tag{14}$$

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 6 db greater than that resulting from fourth harmonic propagation. Clearly, the propagation path must reject any second harmonic signal from the harmonic generator.

Examining the third harmonic to fourth harmonic signal ratio gives

$$\frac{g_3 \eta_3 A'_{34}}{g_4 \eta_4 A'_{44}} = \frac{g_3 a_5 \eta_3}{g_4 a_4 \eta_4} \cdot \frac{25\, E_s}{2} . \tag{15}$$

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

$$\frac{g_5 \eta_5 A'_{54}}{g_4 \eta_4 A'_{44}} = \frac{g_5 \eta_5 a_5}{g_4 \eta_4 a_4} \cdot \frac{5}{2} E_s . \tag{16}$$

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 1 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

$$i = \sum_{j=o}^{N} a_j \left(mE_s \cos \left[(\omega + \omega_m) t + \phi \right] \right)^j . \qquad (17)$$

When expanded, this gives terms of the form

$$A'_k \cos \left[k\omega t + k\omega_m t + k\phi \right] .$$

Then when these are received at the harmonic mixer, all ambiguities are removed, and only the A_4 term (which represents energy propagation at 560 GHz) 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 140 GHz would be a much more difficult and expensive task than the purchase of a second 140 GHz 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 140 GHz 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.

2. Dual Source System

A dual source test system is being set up which will permit positive verification that the mixer is detecting at the fourth harmonic of the local oscillator. The system is diagrammed in Figure 2. The significant feature of the system is that the lowest common multiple frequency of the two oscillators is 560 GHz. The next higher common multiple is 1120 GHz, which is such a high order harmonic of both sources and such a high frequency that it is reasonable to assume the system has relatively no sensitivity to 1120 GHz.

Positive knowledge of the frequency being detected by the system is necessary to evaluate the harmonic mixer. The experimental mixer mount has been designed with a small input waveguide which will not

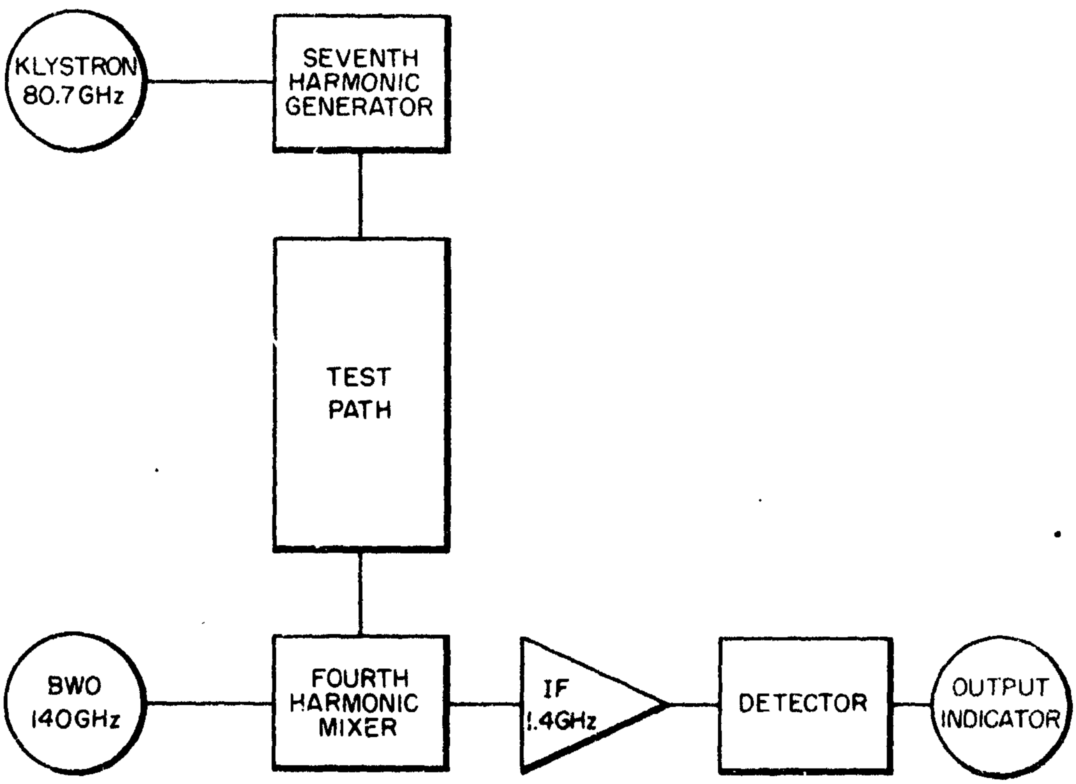

FIG. 2 —DUAL SOURCE TEST SYSTEM

pass energy at the second and third harmonics of the LO. It is desirable to verify that this does occur. It is also desirable to determine the mixer response to signals at the fifth harmonic. This may be done with the two source system of Figure 2 by driving the harmonic generator with a klystron having a frequency some odd fraction of the desired test frequency. For example, one may use 5 x 84 GHz to test for third harmonic mixing or 9 x 77.8 GHz to test for fifth harmonic mixing. Some freedom of choice of klystron frequencies for testing different harmonic orders of mixing is possible by varying the operating frequency of the BWO driving the mixer.

All of the components for the dual source test system are now available except the harmonic mixer itself. An Oki Type 75V10 klystron was purchased to use in this system as well as for other component testing. A Raytheon Type QKK 1150 klystron is also available. The first may be tuned from 70 to 80 GHz, the second from 64 to 74 GHz. Figure 2 shows only the essentials of the test setup. Auxiliary items — such as attenuators, slotted lines, wavemeters, detectors, and a mixer bias box - are available for use as needed. The test path may not be a separate component as shown on the diagram or a free-space path but may be just the mixer signal input waveguide.

3. Seventh Harmonic Generator

A crossed-waveguide mount has been designed and built for use as a 560 GHz seventh harmonic mixer. Figure 3 is a photograph of the device, assembled and disassembled. It has been designed to have several specific features. The two crossed waveguides have been 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 is 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 used on this project smaller than RG-99/U. The harmonic waveguide has been made as short as possible, also to minimize attenuation. A taper transition has been 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 small waveguide, single-mode at the desired harmonic, in order that the harmonic be established in the first order TE_{10} mode.

A potentially important feature of the seventh harmonic generator is that it has been designed so that the diode may be sealed from the

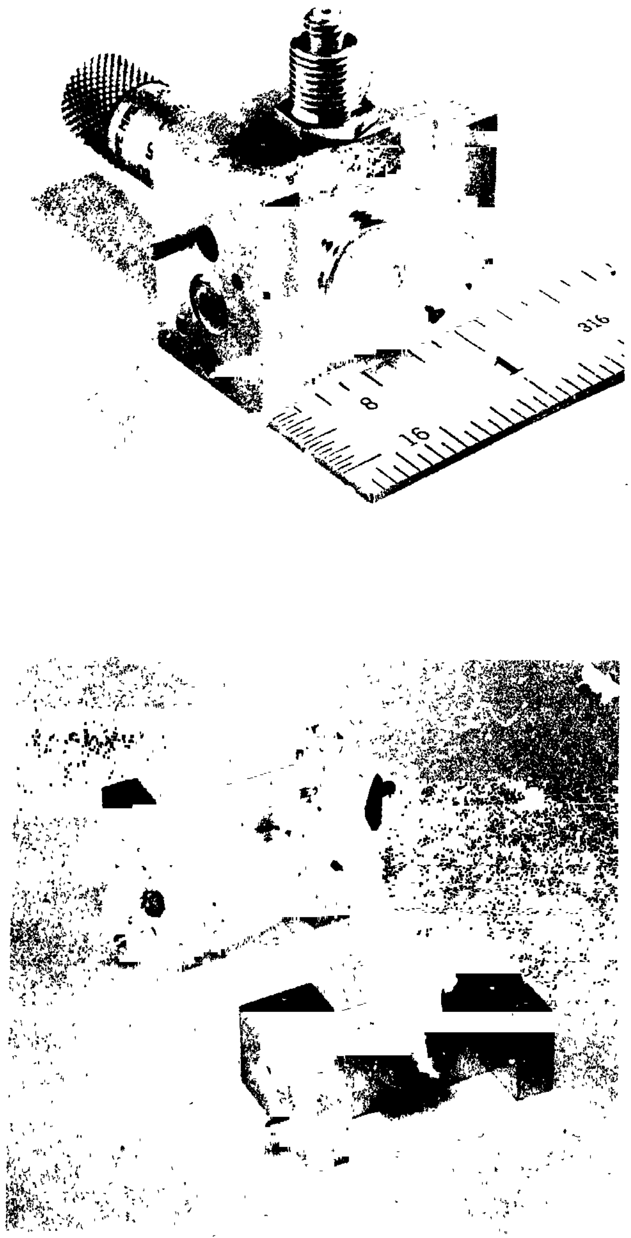

FIG. 3 - 560 GHz SEVENTH HARMONIC GENERATOR

atmosphere. 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 critical requirements on the coaxial port of this device. It is a simple miniature connector which may be sealed in place. The whisker carrier has been 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 mounts in a jig which also holds the differential drive for this process. The waveguide ports may be sealed with thin windows of mica. The losses due to the windows have not yet been evaluated but are expected to be small for thin windows. The micrometer-driven tuners are made as short waveguide sections which attach to the harmonic generator body with the windows between. The tuner for the fundamental guide is not shown in Figure 3. Attachment of the final seals will be done in a dry nitrogen atmosphere. This sealing process should result in longer life for the diode. Later models of the harmonic mixer may be sealed by similar techniques.

There will be no means of testing this device at 560 GHz other than in the dual source system of Figure 2. This is because its output at 560 GHz cannot be distinguished from other harmonics between 560 and 295 GHz by other available detectors, such as bolometers. However, the total harmonic output power above 295 GHz can be 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 is primarily at the lowest harmonic above 295 GHz. This can be verified by measuring the guide wavelength with the harmonic tuner. A diode was formed in this harmonic generator and 90 GHz power applied. An output was detected in the harmonic guide which was determined by measuring guide wavelength to be 360 GHz or the fourth harmonic. The transducer loss was about 30 db.

B. Harmonic Mixer Diode Development

Experimental work has been going on since the beginning of this project to find materials or fabrication techniques resulting in improved millimeter diodes. Various diodes have been compared on the basis of their conversion loss for a range of LO drive power. It has been possible to do these experiments by using existing diode mounts, such as the two crossed-waveguide 2 mm harmonic devices built on Contract NASw-662. The diode development work has been conducted concurrently with the design and construction of an experimental 560 GHz fourth harmonic mixer. The conclusions reached by comparative study of the diode experiments at lower frequency will be a valid basis for the choice of diodes for 560 GHz.

1. Tunnel and Backward Diodes

A tunnel diode is made with highly-doped semiconductor crystal and suitable whisker metal. The parts are assembled first as a point-contact diode. The diode is then pulsed with voltage which alloys a bit of the whisker with the semiconductor to convert it to a tunnel diode. It is now distinguished as a tunnel diode by its current-voltage characteristic curve, shown in Figure 4, which has a current peak followed by a valley for small positive voltage. This is a region of negative resistance. If the pulse-alloying is limited so that a negligible current peak results, a backward diode is 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. They should require less local oscillator power than point contact diodes which do not have as abrupt changes in their I-V curves. This has, in fact, been observed by Eng [3] in studies of 13.5 GHz mixers.

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[4] 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	Carriers per cc	Resistivity in ohm-cm	Matching Whisker Material
M-6	P	GaAs	Zn	58×10^{18}	.002	Sn
3	N	GaAs	Se	7×10^{18}	.0008	Zn
8	N	Ge	As	100×10^{18}	.0065	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 5. 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 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

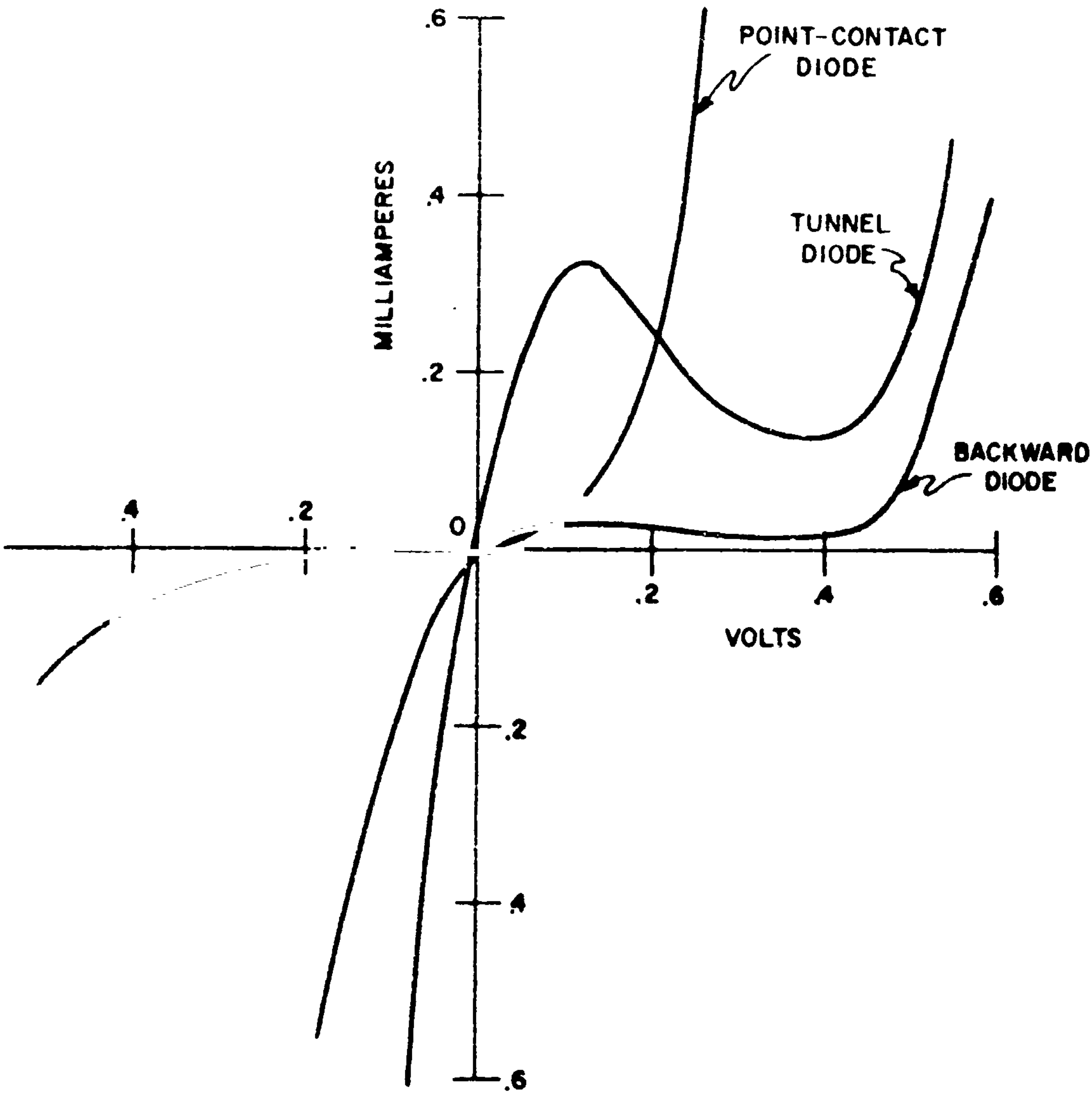

FIG. 4 – VOLTAGE-CURRENT CHARACTERISTICS OF THE POINT-CONTACT, TUNNEL AND BACKWARD DIODES.

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 oscillations at the intermediate frequency (2GHz). 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.

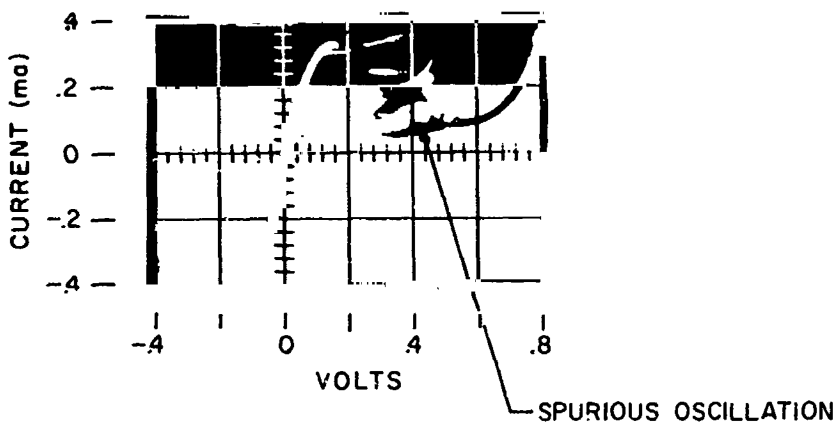

FIG. 5 -TYPICAL MEASURED I-V CURVE FOR TUNNEL DIODE

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 μw 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. It will be necessary to make VSWR measurements on future diodes so that actual conversion loss can be calculated.

2. Point-Contact Diodes

Test and development of point-contact diodes has also been conducted, both to provide data for comparison with newer diode types and to seek improved performance. Figure 6 is a diagram of the complete system assembled for testing point-contact and other diodes in 140 GHz second harmonic mixers. Not all components are used at once, the power monitors and wavemeter being connected as needed.

Some effort was devoted to using this setup to obtain a point-contact diode of minimum conversion loss. Figure 7 is a family of curves plotted from the data measured for a particular diode of conversion loss versus LO power for various bias levels. The VSWR at the signal port could not be measured so that the data is conservative by the amount of any signal mismatch loss. Additional diodes must be tested before it can be determined if these are a typical family of curves. It does appear that the application of a suitable amount of bias voltage will drastically reduce the conversion loss for low LO drive.

3. Considerations of Diode Series Parasitic Resistance

Any semiconductor diode junction has associated with it an effective series resistance (R_s) which dissipates some of the applied signal. If R_s of a mixer diode can be reduced, the conversion loss and the required LO drive will also be reduced. An analysis is given in this section for certain idealized conditions which permits 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 8, 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 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

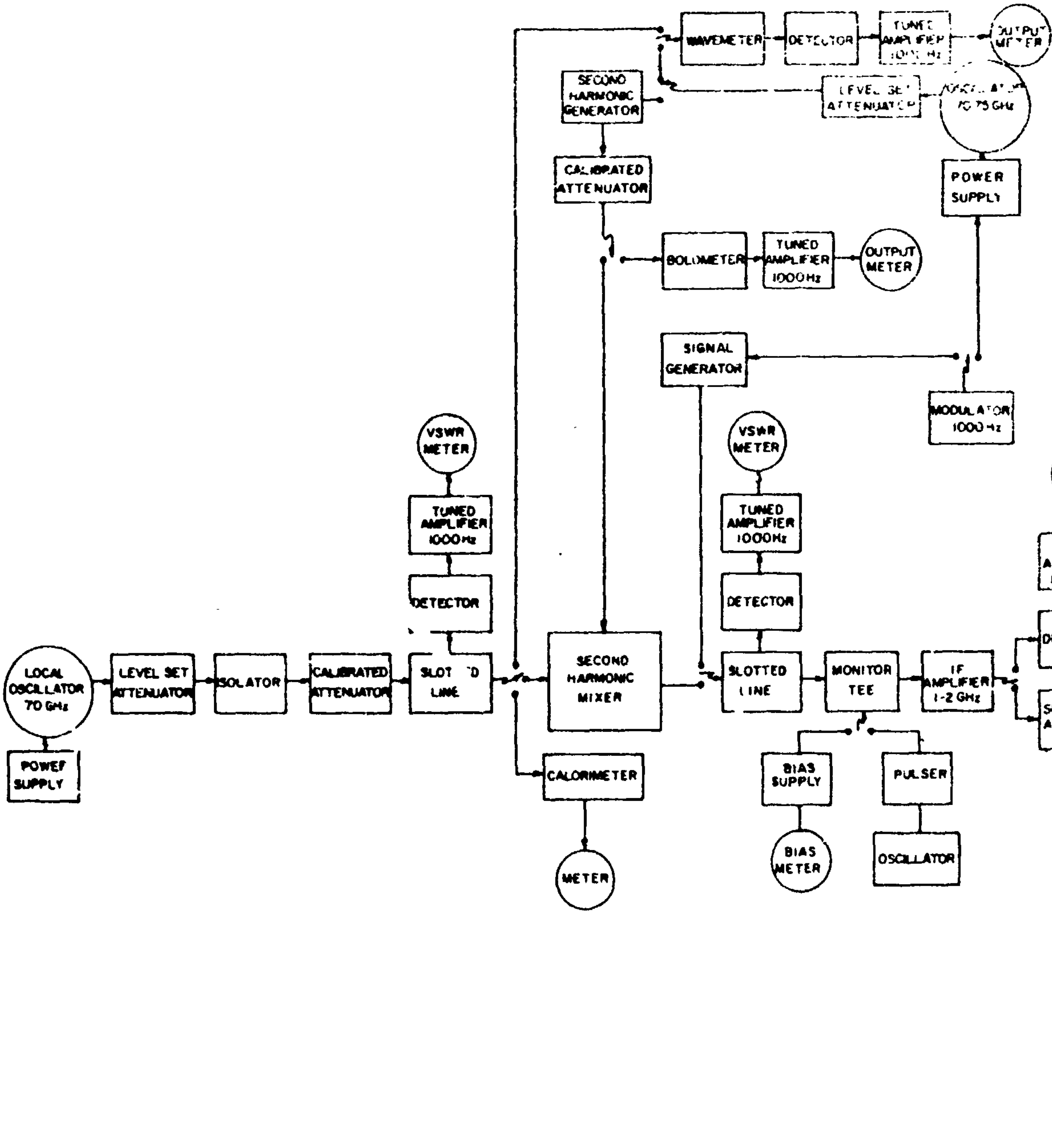

FIG. 6 - 140 GHz SECOND HARMONIC MIXER TEST SYSTEM

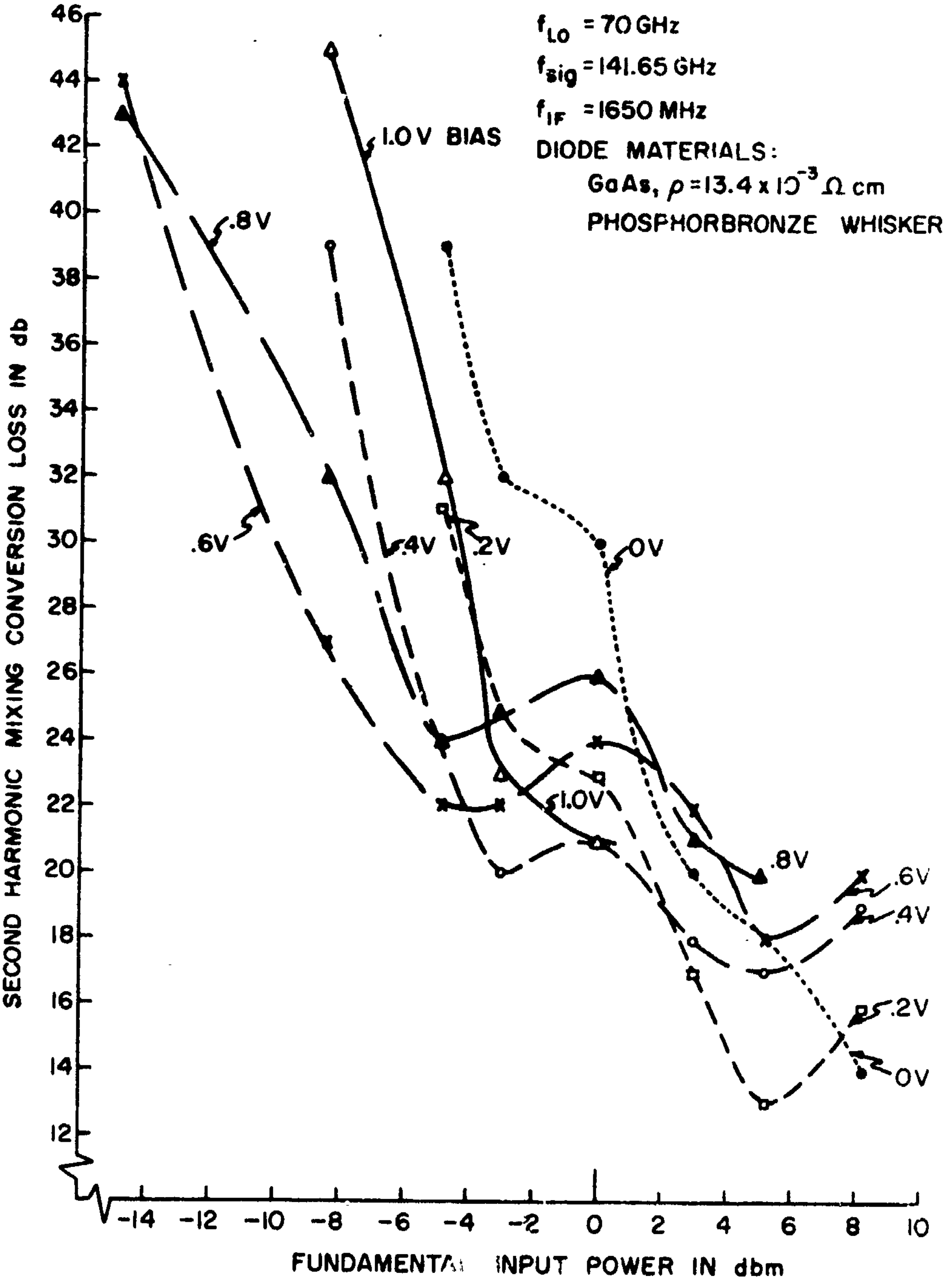

FIG. 7 – MEASURED SECOND HARMONIC MIXING CONVERSION LOSS.

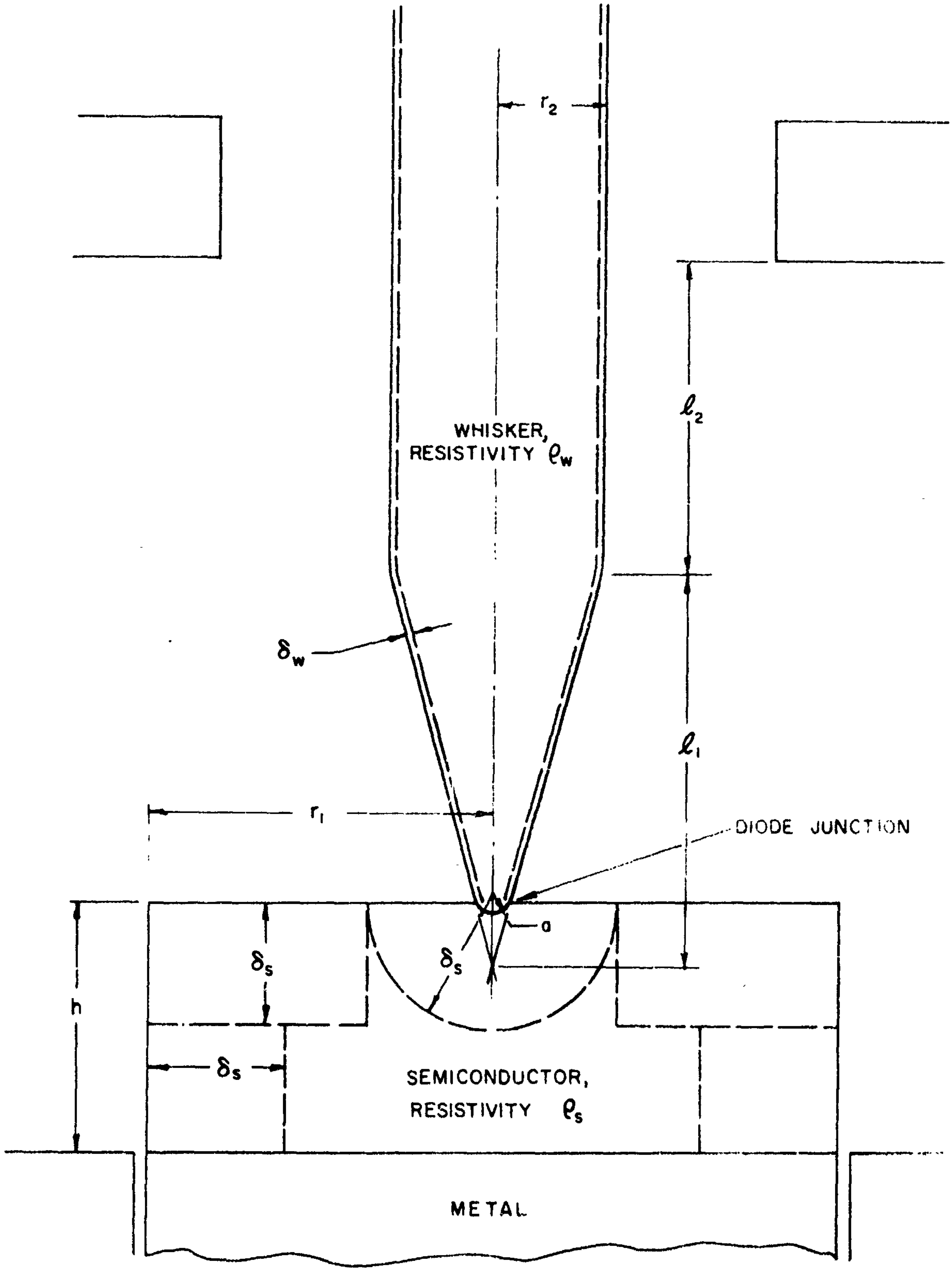

FIG. 8 – GOEMETRY OF POINT-CONTACT JUNCTION

the definition of resistivity, ρ:

$$dR_1 = \frac{\rho_s dr}{2\pi r^2} \tag{1}$$

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

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

Note that for $\delta_s >> a$, (2) reduces to:

$$R_1 = \frac{\rho_s}{2\pi a} \tag{3}$$

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 (3) is seen to be similar to the expression given by Torrey and Whitmer [5] for the resistance of a circular area of contact:

$$R = \frac{\rho}{4a} \tag{4}$$

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

$$dR_2 = \frac{\rho_s dr}{2\pi r \delta_s} \tag{5}$$

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 (5) from δ_s to r_1, the edge of the crystal gives:

$$R_2 = \int_{\delta_s}^{r_1} \frac{\rho_s dr}{2\pi r \delta_s} = \frac{\rho_s}{2\pi \delta_s} \left[\log_e \frac{r_1}{\delta_s} \right] \tag{6}$$

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

$$R_3 = \frac{\rho_s h}{2\pi r_1 \delta_s} \tag{7}$$

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 $r_1 >> \delta_s$).

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:

$$dR_4 = \frac{\rho_w d\ell}{2\pi r \delta_w} \tag{8}$$

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

$d\ell$ = the increment of whisker length.

But by similar triangles

$$\frac{\ell}{r} = \frac{\ell_1}{r_2} \tag{9}$$

Differentiating,

$$d\ell = \frac{\ell_1}{r_2}\, dr \tag{10}$$

Substituting (10) in (8) and integrating from r = a to r = r_2 gives:

$$R_4 = \int_a^{r_2} \frac{\rho_w \ell_1 dr}{2\pi \delta_w r_2 r} = \frac{\rho_w \ell_1}{2\pi\delta_w r_2}\left[\log_e \frac{r_2}{a}\right] . \tag{11}$$

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

$$R_5 \doteq \frac{\rho_w \ell_2}{2\pi r_2 \delta_w} \tag{12}$$

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

Equations (2), (6), (7), (11), and (12) 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.[6]

The skin depth δ is given in meters by [7]

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \tag{13}$$

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

Calculated skin depths in microns (cm x 10^{-4}) are:

	200 GHz	600 GHz
Crystal (δ_s)	13.0	7.5
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:

		200 GHz	600 GHz
Crystal	R_1	12.6	11.4
	R_2	4.8	7.4
	R_3	0.6	2.1
Whisker	R_4	0.5	.8
	R_5	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 x 10^{-4} cm. 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 have detectably better performance.

A number of methods will be described which have been suggested for plating the crystal wafer of a diode to within a few microns of the point-contact. Some experimental attempts have been made to try the suggestions.

a. Preferential Plating

A point-contact junction is formed and the diode placed in a plating bath. A voltage is applied across the diode resulting in a voltage gradient on the surface of the semiconductor. The voltage is adjusted so that there is little electrostatic force attracting ions to the region near the junction. Hopefully, preferential plating will occur. This was tried in a gold cyanide bath but the semiconductor plated non-preferentially immediately upon immersion. This negated any preferential electroplating. Other plating solutions will be tried.

b. Pulse Burning Through Prior Plating

The semiconductor wafer is plated over its entire surface. A pointed whisker is brought into contact with the surface and a voltage is applied to try to burn through the plating. This was tried after applying a 0.0005" gold plating but the whisker would not burn through. A thinner plating might be tried. It is possible that the burning process would blunt the whisker, making it unsuitable for a mm diode.

c. Plating With Junction Masked With Liquid

The whisker is coated with a liquid non-conductor and a point-contact diode is formed. Some liquid flows over a small area of the crystal around the junction. The diode is then plated in a bath after which the masking liquid may be removed with solvents. Seven different liquids such as varnish, lacquer, and silicones were tried. Many would not adhere to the whisker point; those that did washed off in the plating bath. Even if a suitable masking material were found, this method requires that the diode be made and plated in the mount in which it will ultimately be used. If the contact were lost, it would be almost impossible to guide the whisker back to the same spot.

d. Plating Epoxied Junction

A very small amount of epoxy resin is placed on the whisker point. Before the epoxy cures the diode junction is formed. The epoxy is cured; this holds the whisker to the crystal. The crystal and attached whisker are removed from the mount and plated. Diodes have been made successfully by this method. However, sufficient epoxy to hold the whisker to the crystal forms a dot with radius 0.001" to 0.0015". A smaller plated area would be desirable. Curing the epoxy without breaking the diode contact is difficult.

e. Epoxying Pre-plated Junction

By using some convenient masking material, the crystal wafer is plated except for a small area on top. A small amount of epoxy resin is placed on the whisker point. The whisker is observed through a microscope, guided to the unplated area, and a junction is formed. The epoxy is cured. The whisker has been guided to within 0.001" of the plating by this method. Its drawback is that there is no way of knowing if the diode will have low conversion loss since it is not formed in a waveguide mount. It is planned to try illuminating the diode with RF from a horn while forming the junction so that it may be tested as a detector.

C. Harmonic Mixer Structure Design

A fourth harmonic mixer differs from a second harmonic mixer in that the signal waveguide must be small enough to cut off the second and third harmonic, as well as the fundamental frequency. The waveguide cross section for the experimental 560 GHz mixer will be 0.014" x 0.007" for which the calculated cutoff frequency is 420 GHz.

A new method was developed for making a waveguide of very small size such as this which would have smooth surfaces, square corners, and uniform cross section. The method consists of molding a polystyrene mandrel and electroforming a waveguide about the mandrel which is later dissolved. The inside surfaces of the completed waveguide will be as smooth as the original mold. The mold is made of four metal pieces whose molding surfaces can be highly polished. These are clamped together, heated, and polystyrene heated to a plastic state is forced into the cavity under pressure. The finished mandrel is made conductive by vacuum evaporation of gold onto it. Then the waveguide can be built up by electroforming. A one or two mil thickness of gold is applied followed by copper to any desired thickness.

Larger waveguides, such as the local oscillator input, can be electroformed to mandrels made by machining and polishing. The electroformed waveguides are machined on the outside to rectangular shapes for later assembly in the mixer body. This mixer was designed to include a tapered section connecting to the signal waveguide which would serve either as a receiving antenna horn or as a transition from a larger connecting waveguide. The tapered section and the 0.014" x 0.007" signal waveguide were made as one piece by mounting a tapered mandrel and the partially formed waveguide together in a jig and electroforming over all. The mandrels and electroform in various stages of manufacture are shown in Figures 9 to 12.

The reason, of course, for taking such care in making the waveguides is to minimize waveguide attenuation due to surface roughness. Gold is used for the inner surfaces to prevent corrosion which would roughen the surfaces. Special attention to minimizing attenuation is necessary at

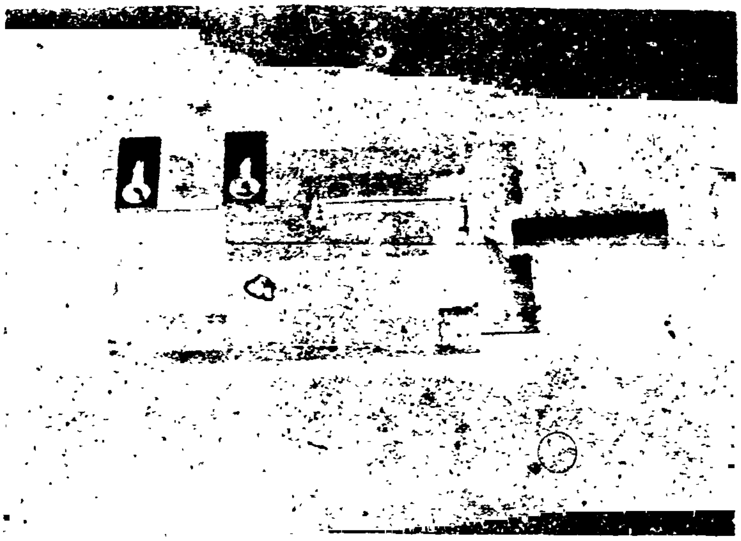

FIG. 9 – WAVEGUIDE MANDREL 0.014" x 0.007 IN JIG

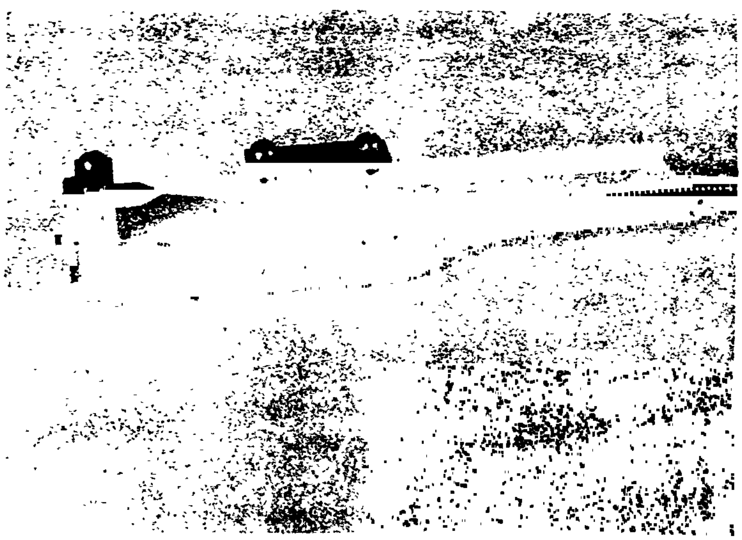

FIG. 10 – PARTIAL ELECTROFORM AND TAPER MANDREL IN JIG

FIG. 11 – WAVEGUIDE AND TAPER AFTER ELECTROFORMING

FIG. 12 – WAVEGUIDE AND TAPER MACHINED FOR MIXER

submillimeter wavelengths because it 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 [8] 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 the experimental 560 GHz mixer the distance from the throat of the horn to the diode whisker will be 0.152 ". This is about five guide wavelengths and should have less than 0.5 db attenuation at the signal frequency.

A special feature of this mixer mount will be the method of supporting the whisker and the construction of the coaxial IF output. It 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 months. The IF output will be built initially with a 50 ohm output and a 100 ohm quarter-wave matching transformer which may be modified later if impedance measurements with the diode in place show this to be desirable.

1. 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 13. Measurements are being made 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 560 GHz harmonic mixer since the available LO power in the satellite radiometer will be small. Previous crossed waveguide devices have been designed on the basis of intuitive reasoning as to coupling effects. The present experiments have been made possible in large measure by the recent fabrication of an epoxy-bonded point-contact millimeter diode which can 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 can be isolated. Figure 14 is a microphotograph 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 15 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

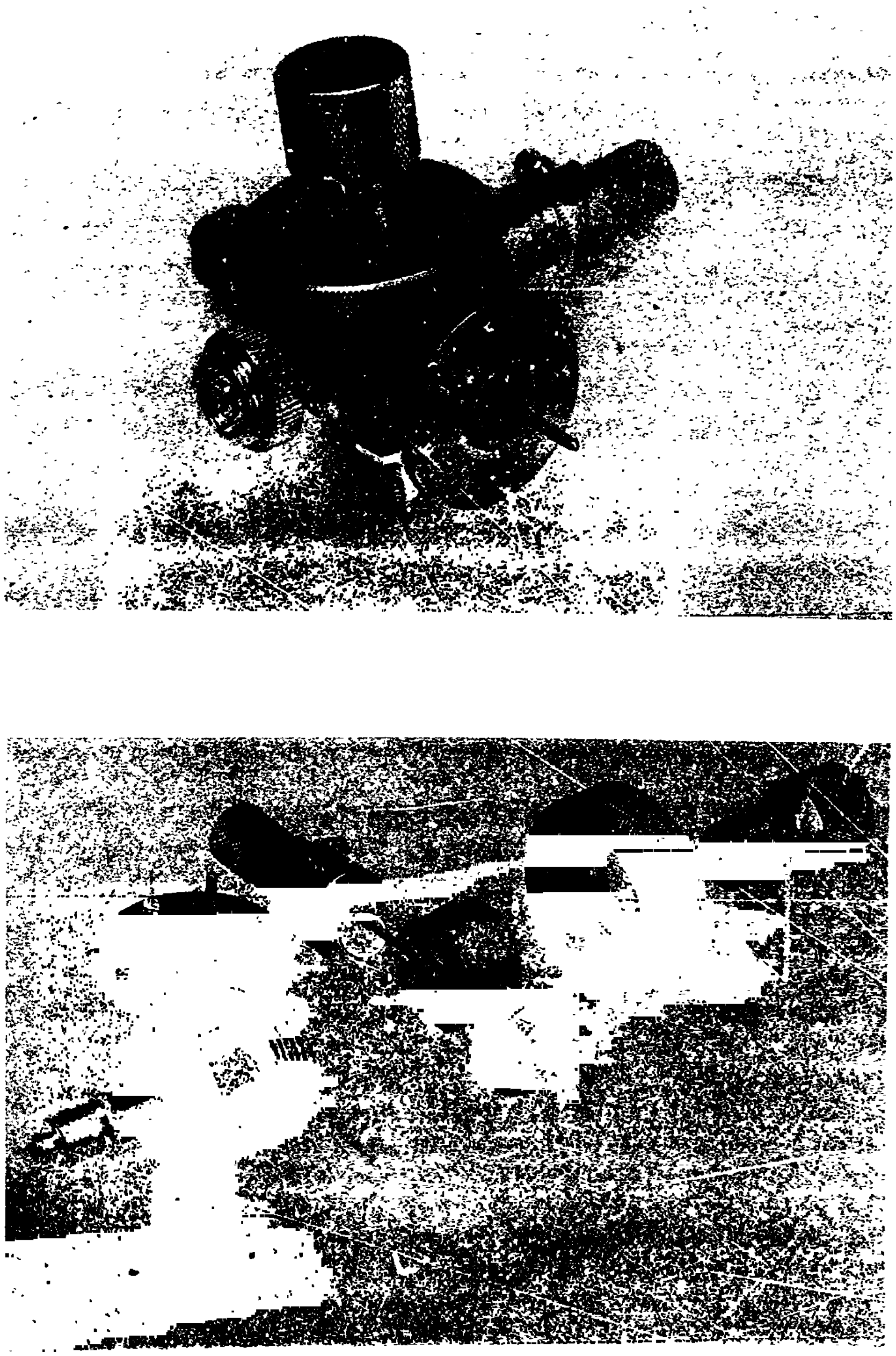

FIG. 13 - FOURTH HARMONIC GENERATOR FOR CROSSED WAVEGUIDE COUPLING STUDIES

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 15 does give some basis for selecting a common wall thickness and was used for the design of the 560 GHz harmonic mixer.

FIG. 14 - EPOXIED DIODE (MICROPHOTOGRAPH)

Another result of the measurements on the 210 GHz fourth harmonic generator was the discovery of an important tuning parameter; namely, the position of the whisker carrier in the fundamental waveguide is critical. In order to determine the optimum position, the mount has been modified to permit the whisker carrier to be moved through a range of positions. Work is presently being carried on to make a more sensitive epoxied diode before continuing harmonic measurements since more dynamic range in the output is needed in order to observe variations. Previous diodes have been adjusted for good dc characteristics before epoxying the junction. The plan now is to measure RF characteristics, also, before epoxying. It is expected that epoxying techniques learned in this work may be applicable later in the development of a rugged 560 GHz diode.

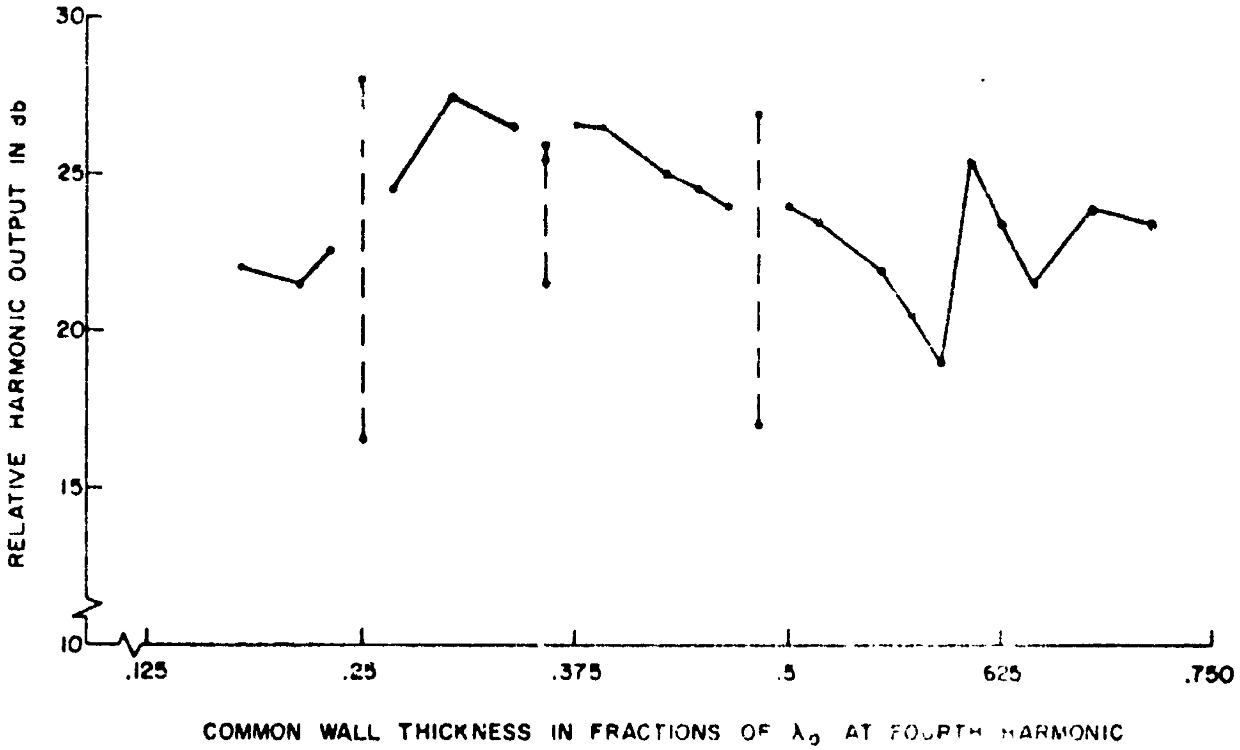

FIG. 15 – DATA FOR FOURTH HARMONIC CROSSED WAVEGUIDE DEVICE

IV. SOLID STATE LOCAL OSCILLATOR

As was explained earlier in this report under Section II titled "Approach," it was decided to make the connection between the harmonic mixer and local oscillator at about 140 GHz. There are various possibilities to be investigated for the generation of 140 GHz power. These include commercial varactors in millimeter waveguide mounts, development of point-contact varactors using epitaxial semiconductor wafers in crossed waveguide mounts, and improvement of point-contact devices previously used as millimeter wave harmonic generators.

The power needed to drive a 560 GHz fourth harmonic mixer will not be known until the experimental mixer is tested. Based upon data for lower frequency mixers such as that given in Figure 16, it appears that 10 or 20 milliwatts is the approximate power needed for optimum operation. As little as 100 microwatts may be usable with perhaps 20 db increased mixer conversion loss. In addition to the requirement of 0.1 to 20 mw output from the 140 GHz harmonic generator, it must operate with the drive power which can be generated at its fundamental frequency. Power available from lower frequency solid state sources determines the efficiency which the 140 GHz harmonic generator must achieve and bears upon the choice of harmonic order.

Present work on the solid state local oscillator is concerned with the 140 GHz output stage. Since its development is more difficult than that of the lower frequency stages, it is being worked on first. Furthermore, we expect to find a commercial product suitable for at least a portion of the solid state local oscillator.

A. Varactor Tripler

Varactors have been used as highly efficient harmonic generators at microwave frequencies. The performance of millimeter wave varactors is limited by the junction capacity, C_j, at a specified bias, the series resistance, R_s, and capacity and resistance of the package in which they are mounted. A varactor may be judged by its cutoff frequency which is defined by

$$f_c = \frac{1}{2\pi R_s C_j} \quad .$$

Efficient harmonic generation can only be achieved when the cutoff frequency is at least several times the operating frequency.

A commercial varactor with one of the highest cutoff frequencies is Texas Instruments' Type TIXVO4, which is rated at 300 GHz for -2v bias

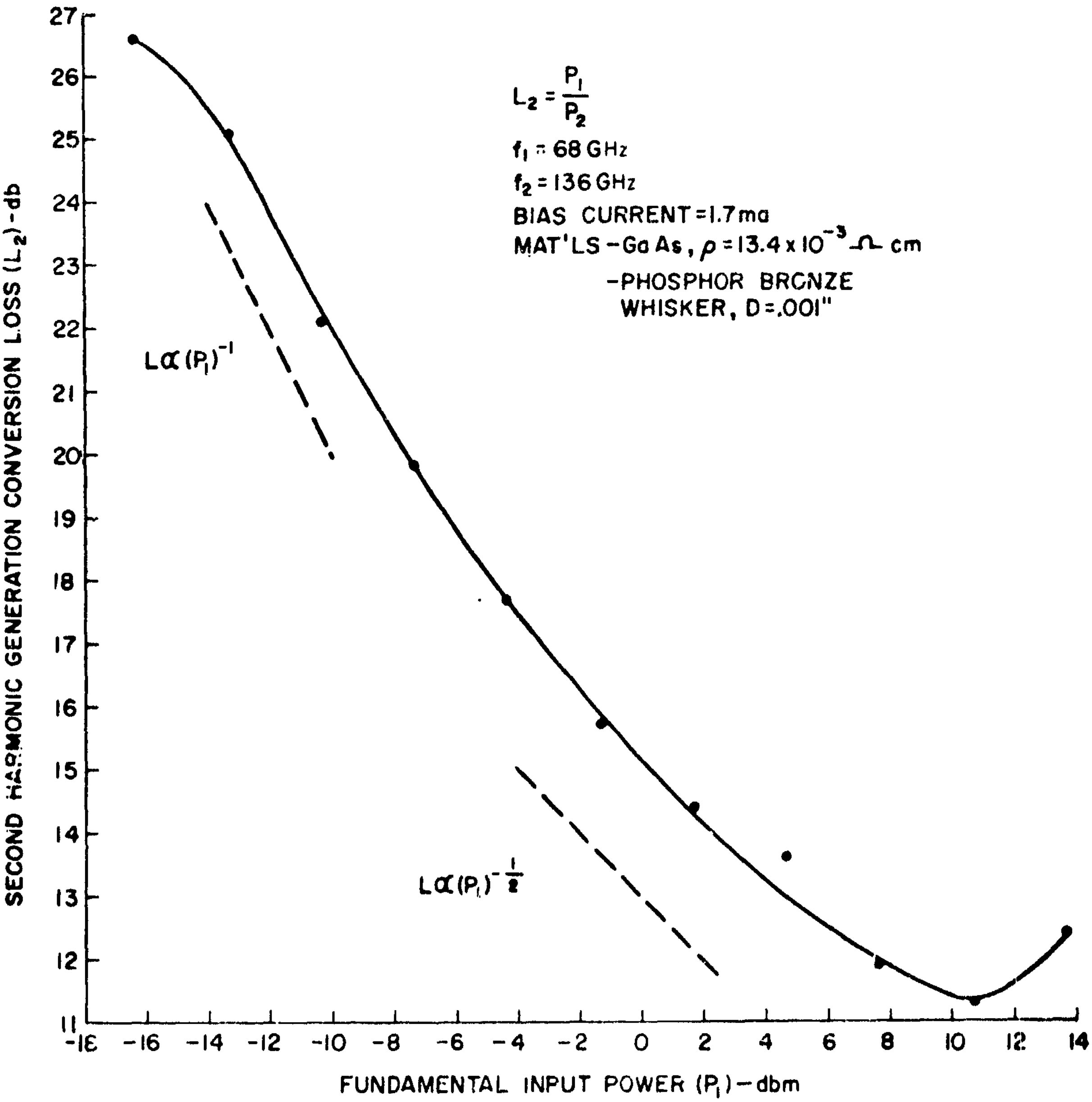

FIG. 16 – MEASURED SECOND HARMONIC GENERATION CONVERSION LOSS.

and breakdown at -6 v. 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 -2 v bias. A waveguide mount was built for it and is shown in Figure 17 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 47 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. The present varactor tripler is certainly inadequate as a 140 GHz local oscillator. A decision to investigate further this or other commercial varactors will depend upon the results obtained in the studies of epitaxial materials described in the next section.

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 for us to try this approach. The TIXVO4 diode now at hand will be saved for possible use in the next lower frequency stage of the 140 GHz harmonic generator chain.

The signal source for the 140 GHz varactor tripler tests was an Oki klystron Type 47V10. For cooling and stability it was mounted in an oil bath. The arrangement is shown in Figure 18. This tube was also used as the signal source for the LO coupling studies with the 210 GHz fourth harmonic generator. A second Oki klystron, Type 75V10, has been purchased and mounted in an oil bath also. It has been used for the harmonic mixer diode development and seventh harmonic generator testing.

The detector shown in Figure 19 was built to check the output frequency of the 140 GHz harmonic generators. It was made with a waveguide size of 0.080" x 0.040" (RG-138/U) which is recommended for the 90 to 140 GHz range. It may be used at higher frequencies, but filters are necessary if one wishes an unambiguous frequency check. The micrometer drive on the tuner was provided so that frequency can be determined by noting half guide wavelength null or maximum positions while adjusting the short. The coaxial port was designed to have a fifty ohm characteristic impedance so that the device might be used efficiently as a mixer if desired.

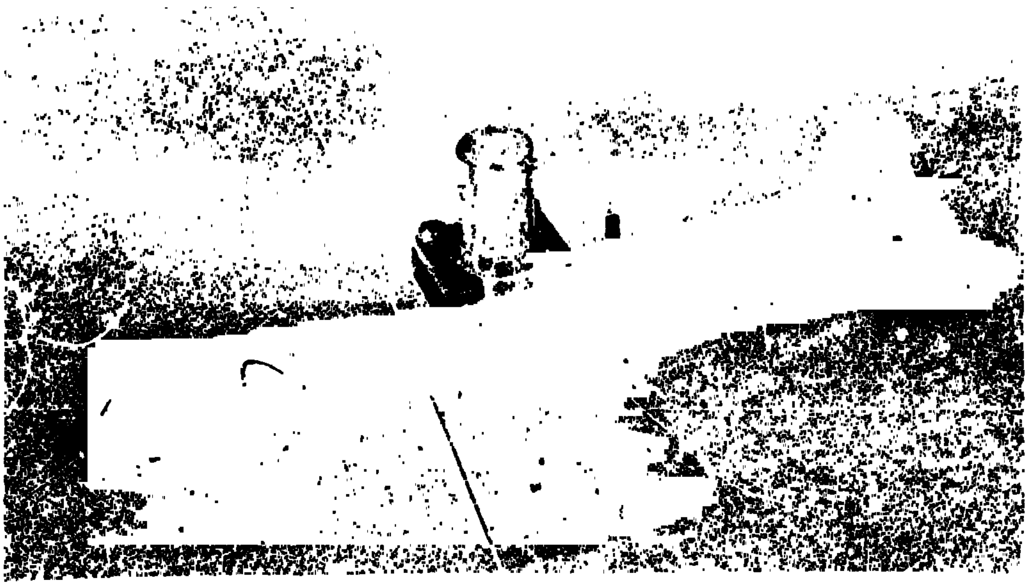

FIG. 17 – VARACTOR AND HARMONIC GENERATOR MOUNT

FIG. 18 – KLYSTRON IN OIL BATH

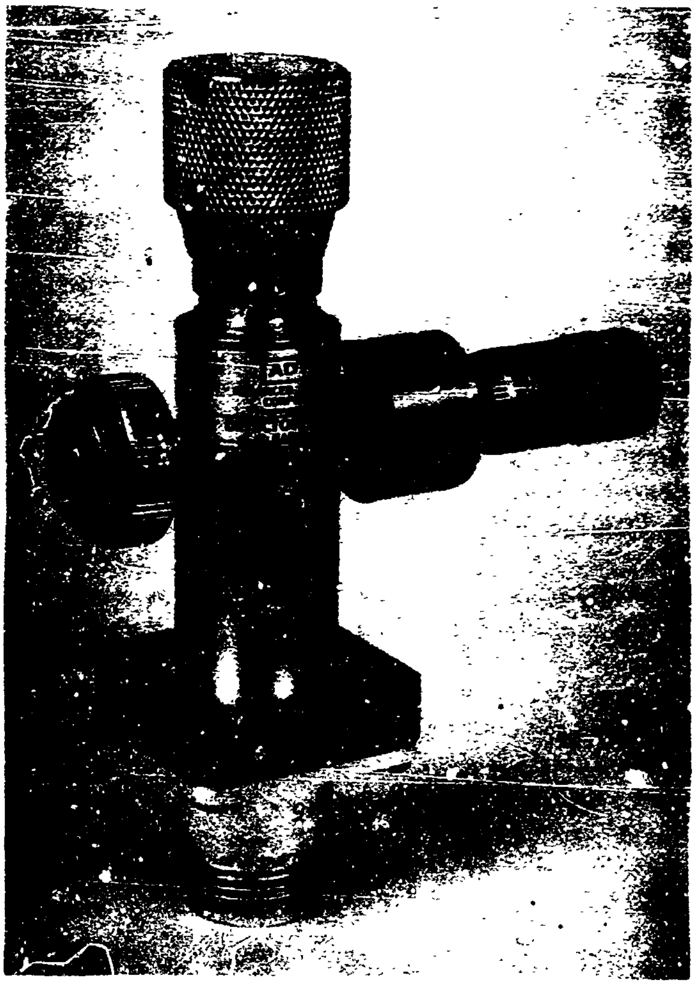

FIG. 19 - DETECTOR FOR 90 TO 140 GHz

B. Epitaxial Diodes

One means of reducing the series resistance in diodes is to use a crystal wafer cut from a double layer semiconductor. A thick layer or "substrate" has a high carrier concentration to give it low resistivity and is the base upon which a thin "epitaxial" layer is deposited which has a lower carrier concentration suitable for the formation of a point-contact or p-n junction. In fact, the TIXVO4 diode described above is made with epitaxial gallium arsenide. Burrus[9] has made epitaxial diodes having cutoff frequencies on the order of 1000 GHz as determined from measurements at 50 to 60 GHz.

Epitaxial material may make excellent diodes for 140 GHz harmonic generation. Three epitaxial samples of gallium arsenide have been obtained for evaluation. Their characteristics are listed in the following table.

	Epitaxial Layer			Substrate
Code No.	Thickness	Carriers per cc	Resistivity ohm-cm	Resistivity ohm-cm
2170	1.2μ	10.9×10^{16}	0.025	0.00099
2171	4.4μ	$125. \times 10^{16}$	0.0025	0.00063
2172	4.0μ	5.9×10^{16}	0 0422	0.00049
Burrus	3.6μ	3.4×10^{16}		

Diodes have been made with these materials using zinc whiskers; typical varactor I-V characteristic curves were obtained. They will be tested next as harmonic generators. This work will be done in part in a lab-owned 105 GHz third harmonic mixer and in part in the crossed waveguide second harmonic devices built on a previous contract, NASw-662, and supplied as GFE. These mounts may also be used to test other non-epitaxial materials which may offer improved conversion loss over previous point-contact devices.

V. 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 section.

Submillimeter radiation has been detected by Putley using an indium antimonide photoconductive detector cooled to 1.5°K.(10), (11), (12) Meredith and Warner(13) also report some of Putley's data. Values of noise equivalent power (NEP) for an output bandwidth of 1 cps are reported as follows:

		Reference
150 GHz	1×10^{-12} w	Table 1 of (13)
300 GHz	5×10^{-12} w	(12)
600 GHz	10×10^{-12} w	(12)
1500 GHz	20×10^{-12} w	(12)

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(10) 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(13) 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 feasibility of constructing such a filter centered on any desired frequency is unknown.

The superheterodyne radiometer has a relatively narrow input band. It is equal to twice the bandwidth of available low noise broadband amplifiers used in the IF. The input bandwidth of a submillimeter superheterodyne radiometer may be 8 GHz, which is 1.4% at 600 GHz. The minimum detectable temperature change, ΔT_{min}, for a superheterodyne receiver is calculated by the formula:

$$\Delta T_{min} = a\, F_{RAD}\, T_o \sqrt{\frac{B_{out}}{B_{IF}}} \tag{1}$$

where

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

F_{RAD} = receiver noise figure for double sidebands

T_o = 300° K, reference temperature for F_{RAD}

B_{out} = output bandwidth.

This definition for ΔT_{min} includes the condition that the receiver output signal-to-noise ratio equal one.

For a direct detection receiver such as Putley's consider the minimum detectable power as occurring for S/N = 1 (defined as the noise equivalent power, NEP) and therefore equal to $k\Delta T_{min} B_{in}$ where

k = Boltzmann's constant

$= 1.38 \times 10^{-23}$ w/°K

B_{in} = input bandwidth.

Here an output bandwidth of 1 cps is understood since that is the condition for which Putley's data is given. Solving for ΔT_{min} one obtains:

$$\Delta T_{min} = \frac{NEP}{k B_{in}} \quad \text{(direct detection).} \tag{2}$$

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

$$\Delta T_{min} = \frac{F_{RAD}\, T_o}{\sqrt{B_{IF}}} \quad \text{(superheterodyne).} \tag{3}$$

In the first case ΔT_{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 ΔT_{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 ΔT_{min} are compared below using a 10% bandwidth for the Putley detector and a 4 GHz IF bandwidth for superheterodyne receivers. Data measured for ΔT_{min} for various ADTEC receivers[14], [15] is used after correcting for S/N = 1 and B_o = 1. Also values for ΔT_{min} for a superheterodyne radiometer at 100 GHz have been included which were calculated using mixer conversion loss measurements for the most recent ADTEC sealed 3 mm diode.[16]

		Putley		ADTEC		
		ΔT_{min}	Ref.	ΔT_{min}	RF losses included	Ref.
100 GHz	I			0.1^o	2 db	(16) median
	II			0.036^o	2 db	(16) selected
150 GHz		4.8^o	(13)	1.1^o	2 db	(14)
600 GHz		12	(12)	$10.^o$	3.6 db	(15)

The ADTEC performance at 150 GHz is for a receiver built three years ago and is based upon 15 db second harmonic crystal conversion loss. The ADTEC 3mm diode represents a 5 db improvement over performance three years ago. A 5 db improvement at 150 GHz would give ΔT_{min} of 0.35^oK. One may conclude that the sensitivities of the Putley detector and the ADTEC 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), ΔT_{min} could be reduced to 1/3.5 of 12 or 3.4^o.

The "ideal" superheterodyne receiver has a theoretical NEP of 2.6×10^{-16}w for 4 GHz IF bandwidth which corresponds to a ΔT_{min} of 0.005^oK. 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.[17] It is a germanium bolometer cooled to 2.0^oK. A commercial germanium bolometer also operated at 2.0^oK is available from Texas Instruments, Inc.[18] The characteristics of these detectors are as follows:

	NEP	B_{in}	ΔT_{min} B_o = 1 cps	Ref.
Low	0.04×10^{-12} w	205-295 GHz	0.047^o	(17)
Texas Instruments	300×10^{-12} w	200-3000 GHz		(18)

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 ΔT_{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 ADTEC diode technology.

Data is not yet available for either Low's bolometer or an all solid-state superheterodyne receiver at 600 GHz. Putley has shown[12] that ideal thermal and photoconductive receivers have similar NEP's at 600 GHz for 10% bandwidth. The 600 GHz radiometer previously built by ADTEC uses a local oscillator tube unsuited for space vehicles. The current work at ADTEC on NASw-1000 is to resolve this problem by developing a fourth harmonic mixer and a solid-state source of 140 GHz power. A radiometer using such components will have a frequency resolution of 8 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 temperatures of $2^{\circ}K$ or lower, obtainable only in cryostats using pumped 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. It is the purpose of this contract to develop components and evaluate the potential of a space-borne superheterodyne receiver using these components.

VI. CONCLUSIONS

One of the problems anticipated at the beginning of this project was the difficulty of fabricating waveguides for submillimeter waves. Practical techniques have now been devised and used to make waveguides of 0.020" x 0.010" and 0.014" x 0.007" internal dimensions with smooth, gold internal surfaces. It should be possible to make waveguides as small as 0.006" x 0.003" by these methods. Such a waveguide would pass frequencies of about 1000 GHz and above. Small, single mode submillimeter waveguides have high attenuation per unit length but the total attenuation in actual mixers for use in superheterodyne receivers can be kept reasonably low by using short lengths having smooth surfaces. The signal attenuation should compare favorably to that imposed by optical type filters at the input of direct detection receivers.

During this reporting period various means of improving submillimeter diode performance have been studied. These studies will continue. Present indications are that tunnel diodes cannot be used to advantage for lack of suitable semiconductor materials. Backward diodes have not yet exhibited any better performance than point-contact diodes.

improvements in the latter appear possible if a method of plating a formed diode to reduce its series resistance can be devised. Experiments to this end are in progress. Methods of applying epoxy resin to the diode junction are being tried in order to seal the diode from atmospheric effects and to secure the whisker to the crystal, resulting in a rugged, durable diode.

A decision was made at the beginning of the project to design the harmonic mixer to be driven by a fundamental LO at 140 GHz and to develop a solid state source of 140 GHz power. It remains to be demonstrated that this is a suitable choice of frequency at which to connect the local oscillator to the harmonic mixer. When the mixer is completed its conversion loss as a function of LO drive power will be measured. Measurements are being made to determine the amount of LO power which can be generated with both commercial packaged varactor diodes and with point-contact varactor diodes made with epitaxial crystal wafers mounted directly in waveguide. Preliminary results indicate the latter will give better results.

Plans have been made and equipment secured to test the submillimeter mixer with a dual source system. A 140 GHz backward wave oscillator has been purchased to use as the test LO source and a 560 GHz seventh h monic generator has been built to use as a signal source. The latter will be driven by a klystron which is also now on hand. Another klystron has been obtained to drive the varactor multipliers being evaluated for the solid state LO. Other components built or obtained for test setups have been mentioned in Sections III and IV of this report. The design and construction of test and prototype components is now complete except for the 560 GHz mixer which will be finished very soon. Work during the next six-month period will be concentrated on testing and improving the two most critical components of the envisioned submillimeter radiometer, namely, the harmonic mixer and the millimeter wavelength stages of the solid state local oscillator. The evaluation of these devices will be performed as rapidly as possible so that the performance of a complete radiometer may be judged and plans made for later phases of this program leading to a satellite-borne radiometer.

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