Low-Noise Room-Temperature and Cryogenic Mixers for 80-120 GHz

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Abstract—A description is given of two new mixers designed to operate in the 80–120-GHz range on the 36-ft radio telescope at Kitt Peak, Ariz. It is shown that for a hard-driven diode the parasitic resistance and capacitance are the primary factors influencing the design of the diode mount. A room-temperature mixer is described which achieves a single-sideband (SSB) conversion loss (L) of 5.5 dB, and a SSB noise temperature (T_m) of 500 K (excluding the IF contribution) with a 1.4-GHz IF. A cryogenically cooled version, using a quartz structure to support the diode chip and contact whisker, achieves values of $L = 5.8$ dB and $T_m = 300$ K with a 4.75-GHz IF. The mixers use high-quality Schottky-barrier diodes in a one-quarter-height waveguide mount.

I. INTRODUCTION

THIS PAPER describes the results of a program of mixer development aimed at producing more sensitive millimeter-wave receivers for the National Radio Astronomy Observatory's 36-ft radio telescope at Kitt Peak, Ariz.

The most significant development in millimeter-wave mixers since Sharpless [1] introduced the wafer diode mount in 1956, has been the introduction of the Schottky-barrier diode. The nearly ideal exponential characteristic of the Schottky diode led Barber [2] to approximate the device by a switch in series with a small resistance; the conversion loss of a mixer is then a function of the pulse duty ratio (PDR) of the switch. Dickens [3] has achieved good agreement between Barber's theory and experimental results at 60 and 95 GHz. Leedy *et al.* [4] demonstrated good agreement between theory and experiment when they assumed, following Torrey and Whitmer [5], a sinusoidal LO voltage at the diode. Although this assumption is unlikely to be strictly valid [6], [7], it is consistent at high LO levels with Barber's switching model.

More recently, nonlinear analysis techniques have been applied to the mixer problem in an effort to achieve a more accurate understanding of the mixing process [6]–[9]. However, these attempts have been limited to cases in which the diode has a fairly simple embedding network. The difficulty of characterizing the embedding network at the harmonics of the LO frequency has so far prevented these methods from being used to give an accurate solution for the case of a waveguide-mounted diode.[1]

In this paper the approach taken to mixer design is to consider the mixer as three interconnected networks as shown in Fig. 1:

- N_1 the embedding network, or diode mount;
- N_2 the network containing the diode's parasitic capacitance and resistance, which connects the ideal diode to the embedding network;
- N_3 the ideal exponential diode.

These three networks are optimized to obtain maximum power transfer between the embedding network and the periodically varying junction resistance at the input (RF) and output (IF) frequencies. The embedding network is assumed reactive at the harmonics of the LO frequency.

The single-sideband (SSB) noise temperature of a mixer receiver can be written as

$$T_R = T_M + LT_{\mathrm{IF}} \tag{1}$$

where T_M is the noise contribution of the mixer itself, L is the SSB conversion loss of the mixer, and T_{IF} is the noise temperature of the IF amplifier. Following the argument given by Weinreb and Kerr [12], T_M can be expressed in terms of an average temperature associated with the diode $T_{D\mathrm{AV}}$ and the conversion loss, thus

$$T_M = (L - 2)T_{D\mathrm{AV}}. \tag{2}$$

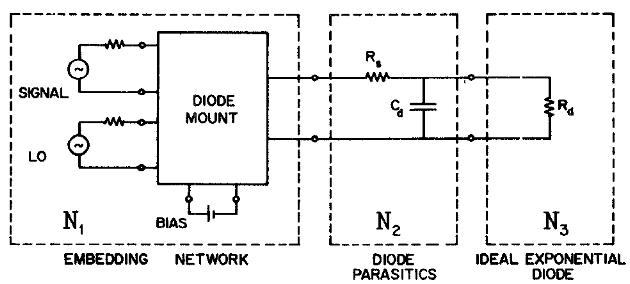

Fig. 1. A mixer represented as three interconnected circuits: N_1—the embedding network; N_2—the diode parasitic resistance and capacitance; and N_3—the ideal exponential diode.

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[1] Eisenhart and Khan [10] and Eisenhart [11] have made an analysis of the simple waveguide mount, which is accurate up to many times the normal operating frequency of the waveguide. In practice, however, the waveguide mount deviates from the simple model in such details as the nonideal RF choke, an input waveguide transformer, and a nonplanar short circuit behind the diode.

The mixer is assumed to be of the broad-band type for which the conversion loss is the same at the signal and image frequencies. It has been found in practice that if the tuning, LO drive, or bias of a mixer is varied, $T_{D_{AV}}$ generally changes to some extent, but that the change in $(L - 2)$ dominates the right-hand side of (2). Reducing L therefore reduces both the mixer and IF contributions to the receiver noise temperature as given by (1).

The object of this paper is to show that with careful attention to the mixer design, it is possible to achieve low noise and conversion loss at frequencies up to 120 GHz. Two mixers are described which are tunable from 80 to 120 GHz; one is intended for room-temperature operation, and the other, a variation of the first, uses a quartz diode mount which is suitable for cryogenic operation.

II. MIXER THEORY

A. *Ideal Exponential Diode*

The junction resistance of a practical Schottky-barrier diode behaves as an ideal exponential element over many decades of current. The current i and voltage v are related by

$$i = i_0(e^{\alpha v} - 1) \tag{3}$$

where $\alpha = q/\eta kT \simeq 35\ \mathrm{V}^{-1}$ at room temperature. For practical purposes $i_0 \ll i$, and the incremental conductance of the diode may be written as

$$g = \frac{\partial i}{\partial v} \simeq \alpha i. \tag{4}$$

The behavior of the diode as a mixer depends both on the waveform of $g(t)$ produced by the LO, and on the embedding network seen by the diode.

If a transformer of ratio n is inserted between an ideal exponential diode, operating as a mixer, and its embedding network, the properties of the mixer will remain unchanged provided the dc bias and LO power are changed according to

$$V_{\text{bias}} \rightarrow V_{\text{bias}} - \ln(n^2)/\alpha \tag{5a}$$

and

$$P_{\text{LO}} \rightarrow P_{\text{LO}}/n^2. \tag{5b}$$

It follows that the ideal exponential diode has no preferred impedance level and can perform equally well as a mixer at any impedance. Thus, in optimizing the three networks of Fig. 1 for maximum signal-frequency power transfer to the ideal diode, N_3 imposes no constraint on the impedance levels of N_1 and N_2. If the parasitic elements of N_2 are fixed for the available diodes, the impedance levels of N_1 and N_3 can be chosen to minimize the conversion loss.

B. *Diode Capacitance and Series Resistance*

The signal frequency equivalent circuit of a Schottky-barrier diode, operating as a mixer, is shown in Fig. 1. R_d is the input impedance of the time-varying junction resistance, and C_d and R_s are the mean values of the junction capacitance and series resistance, assumed equal to their values at the dc bias voltage when no LO power is applied. For a given semiconductor sample C_d and R_s depend primarily on the area A of the diode and on the doping and thickness of the epitaxial layer on which the diode is formed [13]. R_s includes contributions from skin effect in the semiconductor and contact wire. The cutoff frequency is defined as $\omega_c = 1/(R_sC_d)$.

The effects of R_s and C_d on the mixer performance are threefold.

1) They contribute to the conversion loss because of power dissipated in R_s at the RF and IF frequencies.

2) They affect the waveform $g(t)$ of the mixing element by changing the termination of the LO harmonics.

3) They affect the terminations seen by the frequencies $nf_{\text{LO}} \pm f_{\text{IF}}$, $n > 1$.

These effects may be further elaborated as follows.

1) The degradation of the conversion loss caused by power dissipated in R_s at the signal frequency ω, is

$$\delta_{\text{RF}} = 1 + \frac{R_s}{R_d} + \frac{R_d}{R_s}\frac{\omega^2}{\omega_c^2} \geq 1. \tag{6}$$

At the IF frequency R_s appears in series with the output impedance R_o of the exponential element. The loss due to R_s is

$$\delta_{\text{IF}} = 1 + \frac{R_s}{R_o} \geq 1. \tag{7}$$

The combined RF and IF loss due to R_s and C_d is $\delta = \delta_{\text{RF}} \times \delta_{\text{IF}}$; this is shown in Fig. 2 as a function of normalized frequency. The parameter $K = R_o/R_a$ is the quotient of the output (IF) impedance and the input (RF) im-

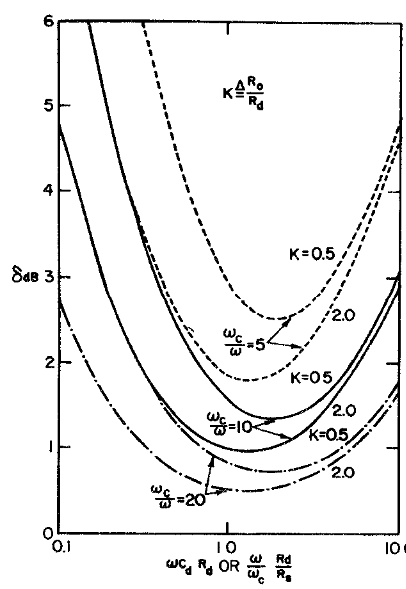

Fig. 2. Loss δ due to RF and IF dissipation in the diode series resistance R_s. The RF (signal) frequency is ω, and $K \triangleq R_o/R_d$ is the ratio of IF impedance to RF impedance.

pedance. It will be shown in the following that for a broad-band mixer K is expected to lie between 0.5 and 2.

2) Barber [2] has used the concept of an equivalent PDR to characterize the mixer properties of a diode with a conductance waveform $g(t)$. The PDR is a function of LO power and bias voltage, and can be maintained constant along with the conversion loss despite changes in $g(t)$ caused by variation of the embedding impedance at frequencies nf_{LO}, $n > 1$.

3) Saleh [14] has shown that for Barber's equivalent PDR to uniquely define the conversion loss it must be dependent not only on the $g(t)$ waveform but also on the embedding impedance seen by the diode at frequencies $nf_{\mathrm{LO}} \pm f_{\mathrm{IF}}$, $n \geq 1$. A change in the reactive termination at some sideband frequency $nf_{\mathrm{LO}} \pm f_{\mathrm{IF}}$, $n > 1$, affects both the PDR, which can be restored by appropriate LO and bias adjustments, and the optimum RF and IF impedances of the mixer. It is assumed here that loss in R_s at these sideband frequencies is small, an assumption which is likely to hold for a practical mixer.

C. IF Impedance

Saleh [14] has made an extensive investigation of the effects on mixer performance of the diode's conductance waveform $g(t)$ and of the embedding impedances at the harmonics of the LO. It is observed from his results that for a broad-band mixer the optimum source (RF) and load (IF) impedances never differ by a factor of more than 2, regardless of LO drive level or bias. Although this is not generally proven for all combinations of terminations of the higher frequency sidebands, $nf_{\mathrm{LO}} \pm f_{\mathrm{IF}}$, $n > 1$, it is consistent with observed mixer performance, and is a useful aid to design.

III. THE DIODE MOUNT

A. Mount Configuration

The choice of a physical configuration for the diode mount is governed by the following considerations.

1) The mount must be easily tunable, preferably by means of a control such as a waveguide short circuit behind the diode. A broad-band RF choke structure is required in the IF and bias connection to the diode to ensure that the impedance seen by the diode will vary as little as possible over the tuning range.

2) The IF circuit must operate with wide bandwidth at a frequency of several gigahertz where low-noise cryogenic paramps are available for use as IF amplifiers. An RF choke which is highly reactive at the IF frequency should therefore be avoided.

3) The diode mounting structure should not introduce excessive parasitic capacitance around the diode thereby reducing its effective cutoff frequency. For a diode whose capacitance is $\sim$0.01 pF this effectively precludes the use of ribbon-contacted or beam-lead diodes in their present forms, and strongly points to the use of a whisker-contacted diode.

These requirements can be fulfilled by a waveguide mount similar in some respects to the wafer mount introduced by Sharpless [1], but using very much reduced-height waveguide, and a different RF choke structure.

B. Mount Equivalent Circuit

Eisenhart and Khan [10] and Eisenhart [11] have made a detailed investigation of the driving-point impedance seen by a small device connected across the gap G in a waveguide mount as shown in Fig. 3(a). The approximate equivalent circuit of the mount is shown in Fig. 3(b), where

$$Z_g = 2\left(\frac{\mu}{\epsilon}\right)^{1/2} \frac{b}{a} \frac{\lambda_g}{\lambda} \tag{8}$$

is the TE_{10}-mode guide impedance, L_s is the post inductance due to the evanescent TE_{m0} modes ($m > 1$), C is the gap capacitance due to the evanescent TE_{mn} and TM_{mn} modes, $n > 1$, and C_1 and L_1 are the capacitance and inductance due to the TE_{m1} and TM_{m1} modes. This equivalent circuit characterizes the mount in the normal operating range of the waveguide for which $f_c < f < 2f_c$, where f_c is the cutoff frequency for the TE_{10} mode.

The gap impedance, Z_{gap} in Fig. 3(b) is strongly affected by elements C_1 and L_1 which are series resonant at the frequency f_1 for which the waveguide height $b = \lambda/2$. For full-height waveguide $b \simeq a/2$, and the resonance f_1 occurs close to $2f_c$. Over most of the useful waveguide band L_1 and C_1 cause a rapid variation of Z_{gap}, both real and imaginary parts, which is clearly undesirable for a mixer in which broad tunability must be simply achieved. By reducing the waveguide height, however, it is possible to raise the resonant frequency f_1 until L_1 and C_1 are equivalent to a small capacitance C', which is independent of frequency for $b^2 \ll \lambda^2/4$.

The element C of Fig. 3(b) is independent of frequency for $b^2 \ll \lambda^2$, and can be considered together with C' as a single capacitance C'', provided $b^2 \ll \lambda^2/4$. In the case of a mixer, the gap of Fig. 3(a) is the depletion region of the diode. C'' is then the junction capacitance C_d of the diode, and can conveniently be measured by a capacitance bridge connected to the IF port of the mixer while the diode is being contacted.

C. Mount Analysis

We now investigate the reduced-height waveguide mount of Fig. 4(a) whose equivalent circuit is shown in

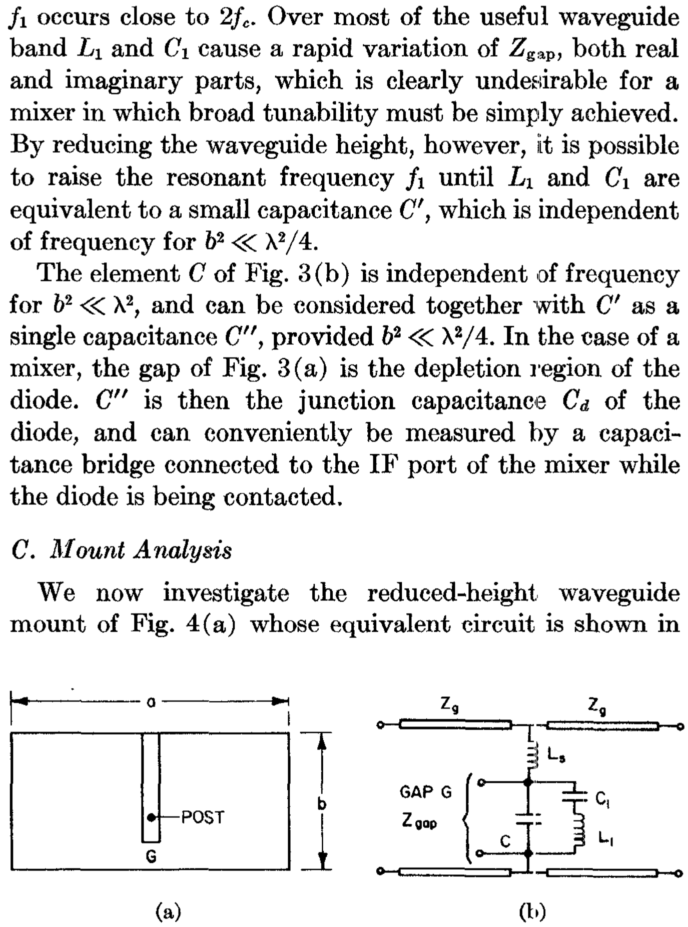

Fig. 3. (a) The simple waveguide mount. The diode is mounted across the gap G. (b) The equivalent circuit of the mount for frequencies $f_c < f < 2f_c$.

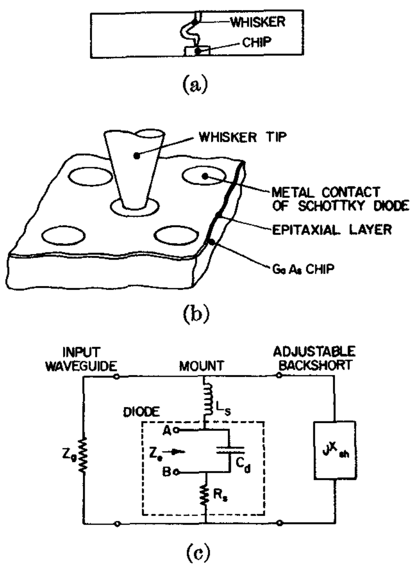

Fig. 4. (a) Reduced-height waveguide mount with a whisker-contacted diode. (b) Diode contact details. (c) Equivalent circuit of the mount and diode parasitics as seen by the junction resistance of the diode.

Fig. 4(c). The impedance Z_e is the embedding impedance seen by the junction resistance R_j of the diode. For efficient mixing, Z_e must be real and equal to some optimum source impedance. It is of interest to examine the real values of Z_e that are possible for this circuit. In particular we shall determine the values of X_{sh} (i.e., the backshort settings) for which Z_e is real, and the effect of L_s, C_d, R_s, Z_g, and frequency on these real values.

The equivalent circuit of Fig. 4(c) was analyzed by computer to determine the values of X_{sh} for which Z_e is real. Fig. 5 shows the real values of Z_e and corresponding values of X_{sh} as functions of $\omega L_s/Z_g$. The main curves are for $R_s = 0$, and typical points are indicated for $R_s = 0.05Z_g$. It is seen that there are, in general, two values of X_{sh} for which Z_e is real, and those real values may differ by a factor of 10 or more.

IV. MOUNT DESIGN FOR 80–120 GHz

A. Diodes

The Schottky-barrier diodes used in this work [15], [16] were formed by electroplating a platinum anode, followed by gold, on epitaxial gallium arsenide. Typical characteristics are shown in Table I. The parameters η and R_s are defined by the diode equations

$$i = i_0 \left\{ \exp\left(\frac{qv'}{\eta kT}\right) - 1 \right\} \tag{9a}$$

$$v = v' + iR_s. \tag{9b}$$

The diodes were supplied by Dr. R. J. Mattauch of the University of Virginia.

B. Electrical Design

The first step in the mount design is to use the loss curves of Fig. 2 to determine the optimum value of R_d,

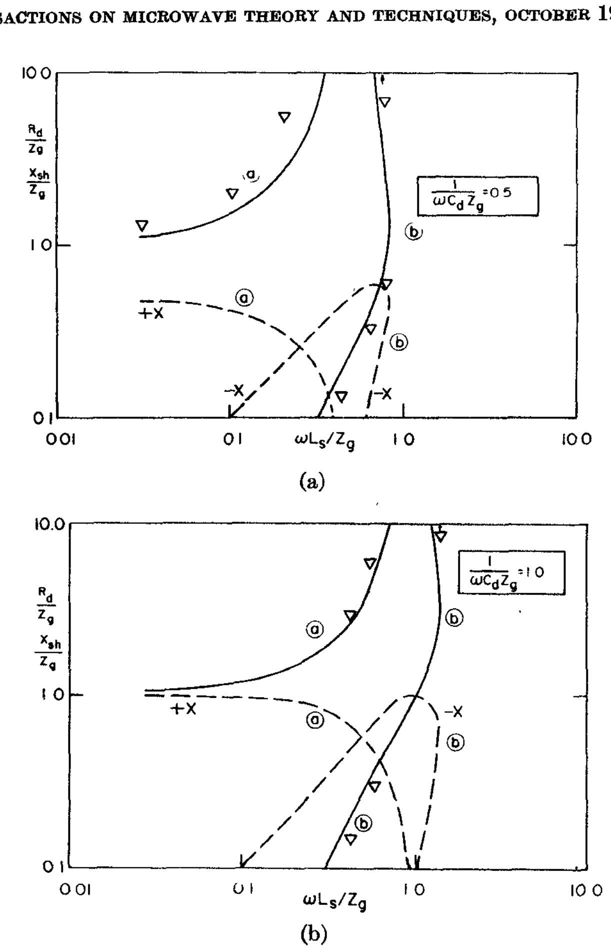

Fig. 5. Mount-matching curves showing values of diode impedance R_d which can be matched (solid curves), and the corresponding backshort reactance X_{sh} (broken curves), both as functions of whisker reactance ωL_s. Normalized reactance of diode capacitance, $1/(\omega C_d Z_g)$ = (a) 0.5, (b) 1.0. Diode series resistance is assumed zero; points (▽) are for $R_s = 0.05\ Z_g$.

TABLE I
CHARACTERISTICS OF THE GALLIUM ARSENIDE SCHOTTKY-BARRIER DIODES AT ROOM TEMPERATURE

EPITAXIAL LAYER	Doping	3×10^{17} cm^{-3}
	Thickness	0.5 ± 0.25 μ
SUBSTRATE	Orientation	(1 0 0)
	Type	n
	Doping	$2\text{-}3 \times 10^{8}$ cm^{-3}
DIODE DIAMETER	2.5 μ	3.5 μ
MEASURED PARAMETERS		
η	1.11	1.10
R_s (measured at DC)	8.0 Ω	3.6 Ω
C_d (at 0.0V, 1 MHz)	0.007 pF	0.012 pF
V_b (at −0.1 μA)	−8 V	−8 V
CALCULATED PARAMETERS		
C_d at V_{bias}	~0.011 pF	~0.020 pF
$\frac{1}{\omega C_d}$ at 100 GHz	145 Ω	80 Ω
R_s at 100 GHz[a]	10 Ω	6 Ω
f_c at V_{bias} and 100 GHz	1450 GHz	1330 GHz
$\frac{f_c}{f_{sig}} = \frac{\omega_c}{\omega}$	14.5	13.3
FROM FIG. 2		
Optimum $\omega C_d R_d$	1 - 2	1 - 2
R_d for minimum δ	145 - 290 Ω	80 - 160 Ω
δ	0.7 - 1.1 dB	0.8 - 1.2 dB

[a] The values of R_s at 100 GHz include contributions from skin effect in the whisker and diode substrate material.

the RF impedance of the diode, for which the power loss in R_s is minimized. The value of C_d used in this calculation is assumed to be the value at the bias voltage. Experience has shown that for gallium arsenide diodes a forward bias of 0.4–0.7 V is required. Table I gives the values of R_d and δ for the two diode types available. Since the IF impedance is known only within the limits set in Section II-C, R_d and δ can only be determined to lie within corresponding limits.

The next step in the mount design is to use the matching curves of Fig. 5 to determine the value(s) of diode impedance R_d which can be matched in the mount shown in Fig. 4. Dimensions assumed are as follows: waveguide width[2] $a = 0.100$ in, diode chip thickness $t = 0.006$ in, contact whisker length $l = b - 0.006$ in, and whisker radius $r = 0.00025$ in. An approximate formula for the inductance of a thin wire across the center of a reduced-height waveguide of height b is given by Sharpless [1]

$$L_s = 2 \times 10^{-7}\, l \log_e \left(\frac{2a}{\pi r}\right) \text{ (MKSA units).} \tag{10}$$

Table II gives the salient calculations in determining the matchable values of R_d for three mounts with different waveguide heights and for two different diodes. Predicted values of the conversion loss and IF impedance are also given.

During the initial part of this work only the 3.5-μm diodes were available, and for these the one-quarter-height mount provides the best match. This mount was used for all the mixers described in this paper. For the 2.5-μm diodes the impedance level of the one-quarter-height mount is somewhat lower than the optimum value; however, Fig. 2 indicates a degradation in conversion loss of less than 0.1 dB.

TABLE II
CALCULATION OF MATCHABLE R_d VALUES AND CORRESPONDING CONVERSION LOSS AND IF IMPEDANCE FOR VARIOUS WAVEGUIDE HEIGHTS AND DIODES

DIODE DIAMETER	2.5 μ			3.5 μ		
Waveguide Height as a Fraction of Full Height	1/2	1/3	1/4	1/2	1/3	1/4
Z_g at 100 GHz, eq. 8	233 Ω	156 Ω	117 Ω	233 Ω	156 Ω	117 Ω
$\frac{1}{\omega C_d Z_g}$ using Table I	0.6	0.9	1.2	0.3	0.5	0.7
$\frac{\omega L_s}{Z_g}$ using eq. 10	1.4	1.2	0.9	1.4	1.2	0.9
$\frac{R_d}{Z_g}$ from Fig. 5	no match	1.5	0.9	no match	no match	1.0 (or~10.0)
R_d	---	230 Ω	110 Ω	---	---	117 Ω
Loss δ dB when diode is matched -- from Fig. 2	---	0.7 - 1.0 dB	0.7 - 1.3 dB	---	---	0.7 - 1.1 dB
$L_{SSB} = 3 + \delta$ dB	---	3.7 - 4.0 dB	3.7 - 4.3 dB	---	---	3.7 - 4.1 dB
Expected IF impedance -- from Section II-C	---	115 - 560 Ω	55 - 220 Ω	---	---	58 - 234 Ω

[2] The choice of $a = 0.100$ in allows the possibility of TE_{20}-mode propagation above 118 GHz. For a centrally mounted diode, however, there is no asymmetry to excite this mode. Our measurements have indicated no higher mode problems.

C. Mechanical Design

Room-Temperature Mixer: The room-temperature mixer, shown in Fig. 6, consists of two main parts, a waveguide transformer and the main body. The transformer is electroformed copper, shrunk into a brass block, and is designed to have a VSWR < 1.06 from 80 to 120 GHz [17]. The main body of the mixer is a brass block, split across the narrow walls of the waveguide. The upper part contains the RF choke supported in Stycast 36-DD dielectric,[3] and the lower part accepts an accurately machined copper post supporting the contact whisker. The diode chip is soldered in place on the end of the RF choke before the two halves of the block are finally assembled. The aluminum insert shown around the choke in Fig. 6 became necessary when it was found that during curing the Stycast reacted chemically with any copper-bearing metal. The positioning of the contact whisker was monitored with a capacitance bridge connected between the diode and the body of the mixer. This ensured that excessive capacitance was not introduced in parallel with the diode due to deformation of the whisker tip after contacting the diode. The whisker position was controlled to a fraction of a micron by a differential micrometer. The backshort is of the contacting finger type, milled from a single piece of beryllium-copper shim stock. Contact between the SMA connector and the RF choke is made by a small bellows spring.

The RF choke was designed to give low loss over 80–120 GHz while having low capacitance as seen at the IF. It consists of four coaxial sections of, alternately, 12- and 70-Ω characteristic impedance, inside an outer conductor of 0.027-in diameter. The cutoff frequency of the TE_{11} mode on the high impedance sections of the choke is ~170 GHz. Calculation of the choke impedance Z_c as seen from inside the waveguide gives $\mathrm{Re}[Z_c] < 0.2\ \Omega$ and $\mathrm{Im}[Z_c] < 5\ \Omega$ in the frequency range 80–120 GHz.

Cryogenic Mixer: The room-temperature mixer described in the preceding was found to be unstable when cooled because of movement between the diode and contact whisker. This was caused by differential contraction of the Stycast dielectric with respect to the metal body of the mount. To eliminate differential contraction poses a

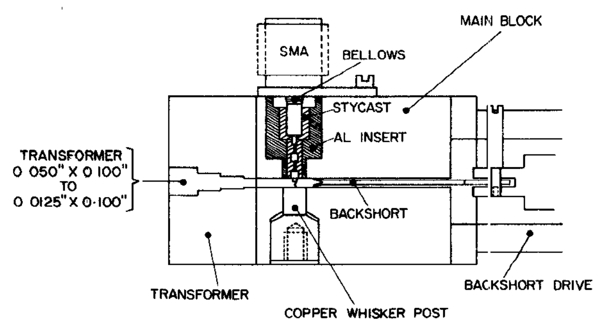

Fig. 6. Cross section of the room-temperature mixer.

[3] Emerson Cuming Company. $\epsilon_r = 1.7$.

difficult materials problem, but its effect can be controlled by using the quartz diode package shown in Fig. 7(b). Differential contraction between the contact whisker and the quartz is small enough to be taken up by the spring of the wisker. Fused quartz was chosen as the structural material because it has high mechanical strength and rigidity, relatively low dielectric constant and loss tangent, is easily cut by scribing and breaking, and is easily metallized with gold over a thin chromium adhesion layer.

It was desired to keep the electrical properties of the mount as close as possible to those of the room-temperature design, and for this reason the mount configuration shown in Fig. 7(a) was used. The main electrical difference between this and the room-temperature design is the quartz member across the waveguide adjacent to the diode. The additional shunt susceptance of this member can be tuned out by adjustment of the backshort.

The quartz diode mount is constructed from three strips of 0.006 × 0.015-in quartz as shown in Fig. 7(b). Two strips are metallized with the RF choke pattern, and the longer unmetallized third strip forms the mechanical support between the choke strips. On one choke strip two 0.001-in gold brackets are ultrasonically bonded, one to contact the IF connector, the other to support the diode which is soldered to it. The contact whisker is soldered to one end of the second choke strip. The three strips are assembled using Eastman 910 adhesive: first the strip carrying the diode is glued to the long support strip, and then the strip carrying the whisker is slid into

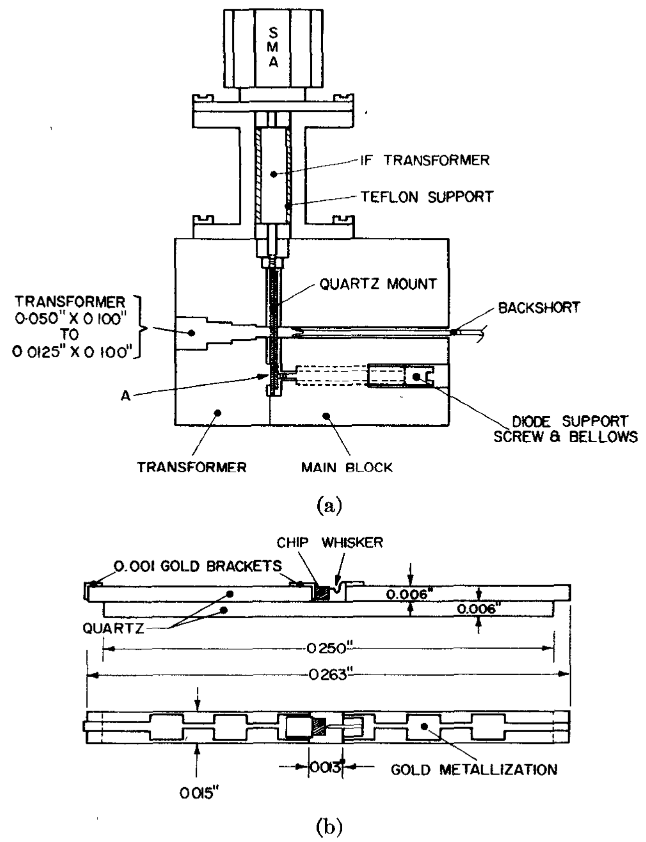

Fig. 7. The cryogenic mixer. (a) Cross section of the mixer. (b) Details of the quartz diode mount. Not to scale.

contact with the diode and glued. The positioning of the whisker point on the chip is observed through a high-power microscope and monitored with an I–V curve tracer. A differential screw is used to control the position of the whisker strip within a fraction of a micron.

The quartz diode assembly is supported across the waveguide, as shown in Fig. 7(a), by the pressure of two springs. One spring holds the assembly against a raised part (A) of the block, ensuring a dc return path, and RF and IF grounds. The second spring, on the end of the IF transformer, contacts the gold bracket at the end of the quartz structure. The diode structure is then free to expand relative to the brass housing.

V. PERFORMANCE

The noise and conversion loss measurements given in the following were made using the IF noise radiometer/reflectometer described by Weinreb and Kerr [12]. This instrument enables the mixer performance to be determined without matching the IF port, which is expedient when a large number of measurements are to be made under conditions of varying IF port impedance. Results obtained in this way have been in good agreement with measurements made by the Y-factor method with the IF port matched using an appropriate transformer.

Typical performance figures for the room-temperature mixers are shown in Table III. The considerable superiority of the smaller diode is believed to be due to its smaller capacitance, enabling it to behave more nearly as an ideal switching mixer.

Table IV gives typical figures for the cooled mixers. These mixers were all constructed with 2.5-μm diodes. The cooled measurements made at 77 K were found to be close to those at 18 K; laboratory measurements were therefore generally made at 77 K for convenience. The mixers had 0.2–0.5-dB greater conversion loss when operating at 4.75-GHz IF than at 1.4-GHz IF. This was probably due to the following: 1) higher IF transformer losses at 4.75 GHz, and 2) the wider spacing (9.5 GHz)

TABLE III
MEASURED CHARACTERISTICS OF THE ROOM-TEMPERATURE MIXERS (f_{IF} = 1.4 GHz)

LO Frequency	85 GHz		115 GHz	
Diode	2.5 μ	3.5 μ	2.5 μ	3.5 μ
L_{SSB}	4.6 dB	6.2 dB	5.5 dB	6.7 dB
$T_{M_{SSB}}$	420°K	700°K	500°K	1400°K
Bias	0.4 v	0.6 v	0.4 v	0.4 v
	2.0 mA	4.0 mA	2.0 mA	4.0 mA

TABLE IV
MEASURED CHARACTERISTICS OF THE CRYOGENIC MIXERS (f_{LO} = 115 GHz)

TEMP.	IF FREQ.	L_{SSB}	$T_{M_{SSB}}$
298°K	1.4 GHz	5.4 dB	740°K
77°K[a]	4.75 GHz	5.8 dB	300°K

[a] Similar results were obtained at 18 K.

between the signal and image bands resulting in a poorer RF match.

The measured IF impedance levels all lie within the limits predicted in Section II-C.

VI. CONCLUSION

An approach to mixer design has been presented for cases where the diode is driven hard by the LO and can be approximated by a switch whose duty cycle depends on the basis voltage and LO level. The ideal diode is connected through a parasitic network, containing the diode's series resistance and capacitance, to the embedding network (mount). The optimum impedance of the embedding network is shown to depend primarily on the parasitic resistance and capacitance. For the particular diodes used in this work it was necessary to reduce the height of the waveguide in the mount to $\sim\frac{1}{4}$ of the standard height.

Two mixers have been described. One is for room-temperature operation, and the other, a modification of the first with a quartz diode mounting structure, is suitable for cryogenic cooling. Typical values of the SSB conversion loss and SSB mixer noise temperature [defined in (1)], measured at 115 GHz, are 5.5 dB and 500 K operating at room temperature with a 1.4-GHz IF, and 5.8 dB and 300 K when cryogenically cooled to 77 or 18 K with a 4.75-GHz IF. The difference between the measured conversion loss and the predicted value is due to nonideal switching behavior of the diode, and to dissipation of signal power converted to higher order sidebands, $nf_{LO} \pm f_{IF}$, $n \geq 2$, which were assumed to be reactively terminated.

The mixers described in this paper are currently in use on the National Radio Astronomy Observatory's 36-ft radio telescope at Kitt Peak, Ariz.

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