The present method can also be applied to other three-dimensional problems by using all merits of the method [9]–[11].

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A Broad-Band, Ultra-Low-Noise Schottky Diode Mixer Receiver from 80 to 115 GHz

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Abstract—**A cryogenic 3-mm receiver has been developed which fully utilizes the low-noise potential of Schottky diodes by approaching the shot-noise limit within 10 percent. With a broad-band mixer design which properly terminates the input sidebands and reactively terminates the second harmonic of the local oscillator and its sidebands, the double sideband (DSB) mixer noise temperature is 35 K in the best case. This design has given an average DSB receiver noise temperature of 75 K over the 80 to 115-GHz band with a best noise temperature of 62 K.**

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I. Introduction

Studies of cooled mixer diodes began in 1956 with Messenger's [1] cooling of *X*-band 1N26 diodes to lower their noise temperature. However, no further work was published on cooling mixers until 1973, when Weinreb and Kerr [2] investigated the noise mechanisms in millimeter mixers and predicted that a 40 K double sideband (DSB) mixer temperature was possible by cooling Schottky diodes to 20 K. Only now has that sensitivity been achieved and surpassed with a minimum DSB mixer noise temperature of 35 K and a DSB receiver noise temperature of 62 K at 100 GHz.

This sensitivity is the result of several years of mixer development at the Five College Radio Astronomy Observatory (FCRAO) which has required considerable

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effort in every aspect of the receiver design to achieve this low noise over a broad RF bandwidth. At the FCRAO, the 3-mm receiver is used on the 13.7-m-diam radio telescope for both spectroscopic studies of molecules and continuum radiometry of quasars and active galaxies.

A broad-band mixer and receiver design was necessary to fully utilize the excellent Schottky diodes now available. Especially important to the low noise of this receiver was the development of Schottky diodes with low doping for cryogenic operation by several laboratories. The receiver improvements include a broad-band vacuum window with an average loss less than 0.05 dB, scalar feed horns to launch a Gaussian beam, improved mixer design, and an IF noise temperature of 10 K at 1.4 GHz. The mixer required the most design effort. It incorporates a broad-band RF filter and noncontacting backshort which not only short circuit the local oscillator (LO), signal, and image bands, but also are reactive at the second harmonic of the LO and its sidebands. The RF impedance of the contact whisker has been optimized to achieve a broad-band response from 80 to 115 GHz with an average receiver temperature of 150 K SSB. An integral IF matching circuit on alumina microstrip has given low loss and broad-band operation.

The following sections discuss in detail the receiver design, the mixer design, and the results as a double sideband system. Then these results are compared with previous work.

II. Receiver Design

The receiver block diagram is shown in Fig. 1. All of the RF components except the vacuum window are cooled to 20 K with a closed-cycle helium refrigerator [3]. The fused silica vacuum window has $\lambda/4$ layers of teflon epoxied on each side as antireflection coatings. The result is an average loss of 0.05 dB over the 75 to 115-GHz band [4].

The scalar feed is designed to launch a Gaussian beam into the quasi-optics system. Its loss is estimated at 0.1 dB. The ring filter is a tunable filter which is used to couple in the LO signal with, while maintaining low loss in the signal sidebands 1.4 or 4.75 GHz away. It is manufactured by Custom Microwave after a design by Davis [5] with a race-track-shape coupling ring [6] to minimize the loss in the signal path. The ring filter has a signal loss of 0.4 to 0.5 dB with a LO coupling loss of 5 to 6 dB. The LO waveguide from 300 to 20 K is gold-plated stainless steel to limit the loss to less than 3 dB and heat loading on the 20-K station to less than 0.1 W. The ring-filter and mixer backshort have mechanical tuning drives.

The first IF amplifier with about 25 dB of gain is also cooled to 20 K and located close to the mixer. The IF was originally over the 4.5 to 5.0-GHz band using a parametric amplifier and a cooled FET as a second stage for a net IF noise temperature of 24 K. Subsequently, an improved IF system, centered at 1.4 GHz, with a noise temperature of only 10 K [7] has been used for the best results.

The receiver temperatures were measured using Emerson-Cuming CV-3 absorber at room temperature and 77 K in front of the dewar window with the mixer-tuned DSB. No input filtering was done to separate the upper

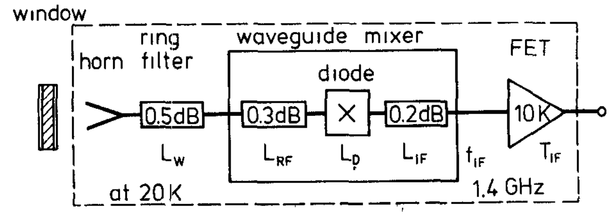

Fig. 1. Receiver block diagram. The fused-silica vacuum window has $\lambda/4$ matching layers. The scalar feed horn, LO injection filter, mixer, and initial IF amplifiers are cooled to 20 K.

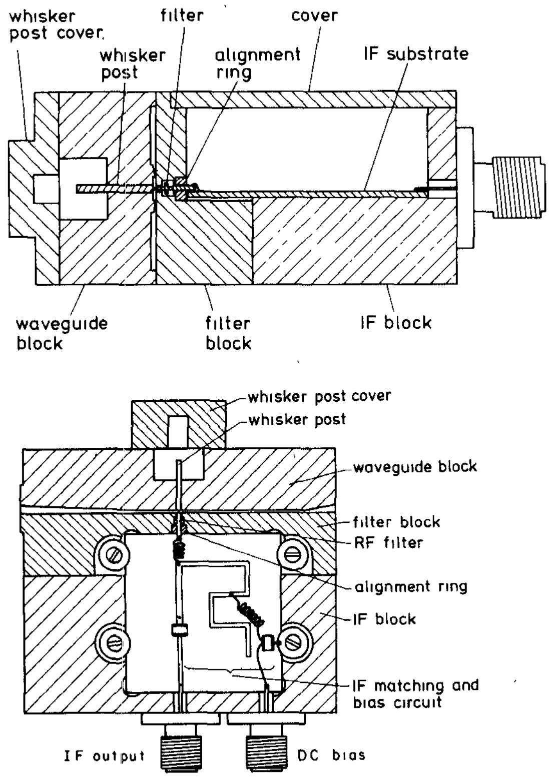

Fig. 2. Mixer waveguide mount. (a) A cross section through the reduced-height waveguide, whisker post, RF filter, and microstrip IF matching circuit, and (b) shows the waveguide taper from full to 1/4 height, and the 1.4-GHz matching circuit.

and lower sidebands and the backshort was within one wavelength of the diode so that the conversion loss was the same in the two sidebands. The linearity of the system extends to an input temperature of at least 6000 K. The second harmonic response of the receiver was measured to be 45 ± 5 dB below the fundamental. This eliminates any apparent lowering of the receiver noise temperature due to inputs at the second harmonic.

III. Mixer Design

The mixer was designed to be tunable over the entire 80 to 115-GHz band since the receiver was to be used for radio astronomy observations over this range, and to have a large instantaneous IF bandwidth. A minimum IF bandwidth of 400 MHz was required since our galaxy and external galaxies have differential Doppler velocities of up to 1000 kM/s, corresponding to a frequency range of 385 MHz at 115 GHz.

The mixer is shown in Figs. 2 and 3. Fig. 2(a) is a cross section through the diode showing the 0.8-mm-diam whisker

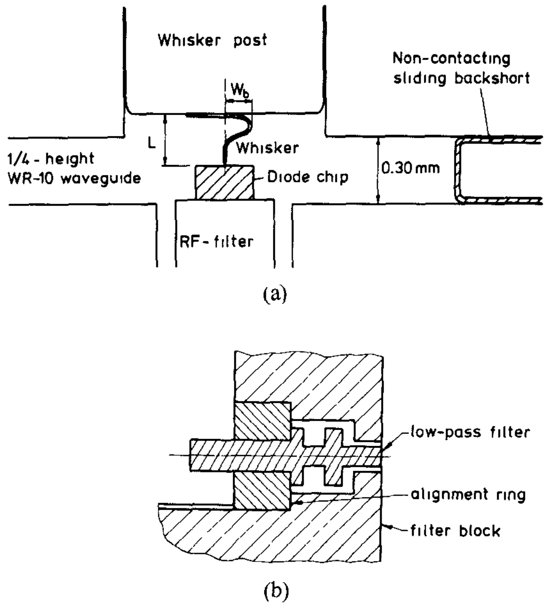

Fig. 3. Details of diode, whisker, and RF filter. (a) shows the diode chip on the RF filter contacted with a 12-μm-diam whisker, and (b) is a cross section of the coaxial RF filter which is held by a Macor dielectric ring.

post, the RF filter, and the alumina substrate where the IF matching circuit is implemented in microstrip. Fig. 2(b) is an orthogonal view which shows the IF matching circuit for 1.4 GHz and the linear taper from full-height to 1/4-height WR-10 waveguide. This reduction lowers the RF impedance seen by the diode and, consequently, the IF impedance. The measured loss is 0.2 to 0.3 dB when the mixer block is machined from OFHC copper. No gold plating was done because the pure copper has a better performance by about 0.1 dB [8]. A linear taper can be readily machined and its loss is estimated to be only 0.1 dB more than a $\lambda/4$ step transformer.

Fig. 3(a) shows the details of the diode with its whisker contact and the noncontacting backshort, while Fig. 3(b) shows the coaxial RF filter. The diode chip is typically 250 μm square by 120 μm thick. The RF filter and whisker inductance are in series with the parallel combination of the waveguide admittance and the backshort. The RF filter was empirically designed to be close to a short circuit in the 80 to 120-GHz band while still being reactive in the second harmonic band. The filter parameters are summarized in Table I. A Smith chart representation of the filter impedance is shown in Fig. 4. For comparison, the analogous plots for the mixers of Linke *et al.* [9] and Kerr [10] are also included. Our filter is reactive over the range 75 to 225 GHz, covering both the fundamental and second harmonic frequencies. These are the most important frequency components for mixer performance, as was shown by the work of Held and Kerr [11], [12]. The mixer designed by Linke *et al.* [9] was for the 60 to 90-GHz band (WR-12) so that their filter is resistive in the lower part of the fundamental band, reactive above 75 GHz and a short circuit for the second harmonic band. The $\lambda/4$ design used by Kerr [10] is excellent as a short for the fundamental 75

TABLE I
RF FILTER PARAMETERS

The coaxial RF filter dimensions, dielectric constant, and impedance for each coaxial section are given. The diode is mounted on section 1 and the IF matching circuit connected to section 6.

Section	ID (mm)	OD (mm)	Length (mm)	ϵ_r	Z_0 (Ω)
1	.45	.61	.67	1.0	19
2	.45	1.59	.31	1.0	76
3	1.32	1.59	.41	1.0	11
4	.53	1.59	.43	1.0	66
5	1.32	1.59	.25	1.0	11
6	.67	2.39	1.25	5.75	32

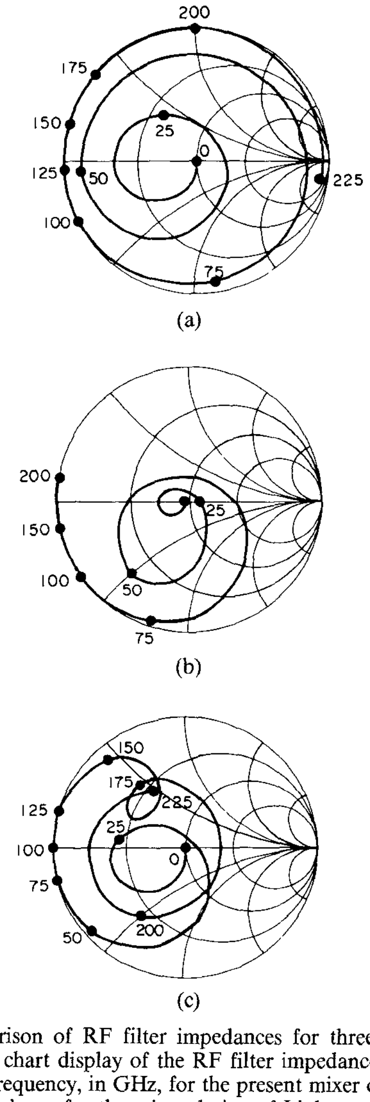

Fig. 4. Comparison of RF filter impedances for three mixer designs. (a) is a Smith chart display of the RF filter impedance, normalized to 50 Ω, versus frequency, in GHz, for the present mixer design, (b) is the RF filter impedance for the mixer design of Linke *et al.* [9], and (c) is the impedance for the design of Kerr [10].

to 115-GHz band, but is not reactive for the second harmonic band. These filter calculations start with a 50-Ω termination impedance at the IF end of the filter and treat each section as a transmission line, taking into account the fringing capacitance [13] at each change in the transmission-line impedance.

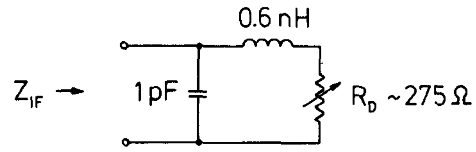

Fig. 5. Equivalent circuit for IF matching. The diode's dynamic impedance is ≈ 275 Ω. The inductive and capacitive elements are due to the RF filter.

Machinability of the filter was taken into account during the design process. As a compromise, only the section nearest the WR-10 waveguide is cut off to the TE_{11} coaxial mode in the 75 to 115-GHz band (see Fig. 3(b)). The dielectric support ring at the IF end of the filter was machined from Macor [14]. This dielectric has a relative permittivity (ϵ_r) of 5.75 and a loss tangent of 0.015 at 100 GHz [15]. For best stability during cycling between 295 and 20 K, no epoxy was used. The center conductor was pressed into the Macor ring and the filter sections machined concentric to the Macor ring. This assembly then was pressed into the mixer block. OFHC copper was used for the inner conductor since the majority of the losses in a coaxial line are due to the inner conductor and the dc electrical conductivity of OFHC copper is 300 times that of brass at cryogenic temperatures [8]. Also an important consideration for cryogenic operation, the thermal conductivity of OFHC copper is up to 20 times that of brass at 20 K [16].

The backshort was designed to be noncontacting for higher reliability since Held and Kerr [12] reported problems in repeating their results with a contacting backshort and Linke *et al.* [9] successfully used a noncontacting backshort. The backshort used in this work has three low and two high impedance sections. The design reported by Brewer and Räisänen [17] gives a VSWR of 90 in the 80 to 120-GHz band and about the same in the second harmonic band of the local oscillator. Thin (0.019 mm) mylar tape [18] is used for the dielectric and has given consistent performance over several years of cryogenic operation.

As was shown in Fig. 2(b), the microstrip matching circuit is built into the mixer block adjacent to the RF filter. This minimizes the electrical length before the filter impedance can be matched to 50 Ω and maximizes the IF bandwidth. As seen from the IF side of the RF filter, the combination of the filter and diode can be modeled by the circuit in Fig. 5. The inductance of 0.6 nH is the sum of the inductances of the individual coaxial sections of the filter, while the 1 pF of capacitance is the sum of the coaxial sections plus the fringing capacitances. When the mixer is optimized for the lowest receiver temperature, the IF impedance is the same as the differential resistance of the pumped diode $I-V$ curve and is measured from the slope of the dc $I-V$ curve. This impedance varies from 200 to 500 Ω over the 75 to 115-GHz band. This experimental result was reported by Räisänen [8] and was found also from computer modeling of the mixer by Lehto and Räisänen [19]. A compromise value of 275 Ω was used in designing the matching circuit. At 1.4 GHz, the net IF

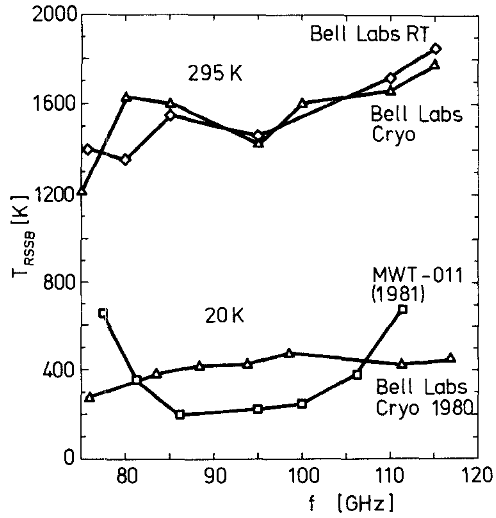

Fig. 6. Single sideband receiver noise temperatures with a 4.75-GHz IF. The two upper curves and the two lower curves are at 295 and 20 K, respectively.

impedance is near 50-Ω resistive plus a capacitive component so a match is obtained with a series inductor whose reactance is about 100 Ω plus an open-circuited $\lambda_{IF}/2$ stub and a transformer for broadbanding. A dc-bias circuit also is included on the alumina substrate. For a 4.5 to 5.0-GHz IF, two open-circuited stubs are used to broaden the response and a $\lambda/4$ section is used to transform the resulting impedance to 50 Ω. The output end of the RF filter is connected to the 4.75-GHz microstrip circuit with a loop of 12-μm-thick copper foil to minimize mechanical strain on the RF filter as the mixer is temperature cycled.

The remaining part of the RF circuit which can be varied is the whisker inductance, as reported by Räisänen *et al.* [20]. As will be discussed in detail in Section IV-B, the whisker length was varied until the receiver response was an optimum over the 80 to 120-GHz band.

IV. Results

A. Comparison of 4.75-GHz and 1.4-GHz IF

The initial receiver tests were with a 4.75-GHz IF with a net temperature of 24 K when cooled to 20 K. Two Bell Telephone Laboratory (BTL) diodes were tested at room temperature. The "room-temperature" diode had a doping of 2×10^{17} cm^{-3} while the "cryogenic" diode had a lower doping of 5×10^{16} cm^{-3} for improved performance at 20 K. As is shown in Fig. 6, both diodes gave essentially the same receiver temperatures at 295 K over the entire 75 to 115-GHz band. The low-doped diode gave a very flat performance of ≈ 400 K SSB when the system was cooled. The total SSB conversion loss, including 0.6 dB for the feed and ring filter, was 7.0 to 7.7 dB over the 75 to 120-GHz range [4], [21], [22]. This system was used for the 1980/81 observing season of the FCRAO 13.7-m radio telescope.

Subsequently, in 1981, diodes made by Millimeter Wave Technology (MWT) were tested in the receiver with a 4.5 to 5.0-GHz IF. As is also shown in Fig. 6, the cooled receiver performance was improved by almost a factor of 2

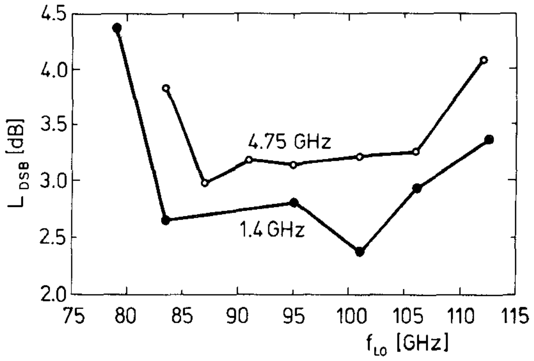

Fig. 7. Total DSB conversion losses at 295 K over the 80 to 115-GHz band for a 4.75 and 1.4-GHz IF. These losses include the feed horn, ring filter, and all mixer losses (waveguide, diode conversion, and IF matching).

over the 85 to 105-GHz range with the MWT diode due to its steeper $I-V$ curve (see (5) and Table IV). Below 85 GHz, the mixer performance with the MWT diode is worse than with the BTL diode because of the increased length of the RF filter. When the MWT diode was mounted, the section of the RF filter nearest the diode was lengthened to move the diode into the waveguide as was shown in Fig. 3(a). This was done to see the effect of different diode positions. This longer RF filter is not as reactive below 80 GHz, which causes the performance to deteriorate. The poorer results above 105 GHz were due to excess inductance from a long whisker as will be discussed in the next section.

When cooled FET amplifiers became available at 1.4 GHz with noise temperatures as low as 10 K [7], the mixers were adapted to this IF frequency by only changing the alumina microstrip circuit in the mixer block. In addition to the lower IF noise temperature when changing from a 4.75 to a 1.4-GHz IF, there is the additional benefit of lower total conversion loss in the mixer. This improvement is shown in Fig. 7 for a mixer with a MWT SD-011C diode. The IF matching circuit was changed between the tests. The DSB conversion loss (L_{DSB}) was measured at room temperature over the 78 to 113-GHz range. Referring to Fig. 1, the total loss L_{DSB} is composed of; L_W, the losses in the window, scalar feed, and ring filter; L_{RF}, the losses in the mixer waveguide and linear taper; L_D, the diode conversion loss; and L_{IF}, the resistive and reflected losses in the IF matching circuit

$$L_{\mathrm{DSB}} = L_W L_{RF} L_D L_{IF}. \tag{1}$$

Window losses are $\simeq 0.05$ dB, the scalar feed is assumed to have a loss of 0.1 dB and the ring filter 0.5 dB. This sum was assumed to decrease slightly to 0.5 dB when cooled. The mixer RF losses at room temperature were 0.5 dB, 0.3 dB from the input taper, and 0.2 dB from the reduced-height waveguide and the backshort. The mixer losses were assumed to improve from 0.5 to 0.3 dB at 20 K because of the increase in the conductivity of OFHC copper. The IF losses are $\simeq 0.2$ and $\simeq 0.6$ dB at 1.4 and 4.75 GHz, respectively. The DSB conversion loss L_D is the same for both the IF's since they are a small fraction (< 5 percent)

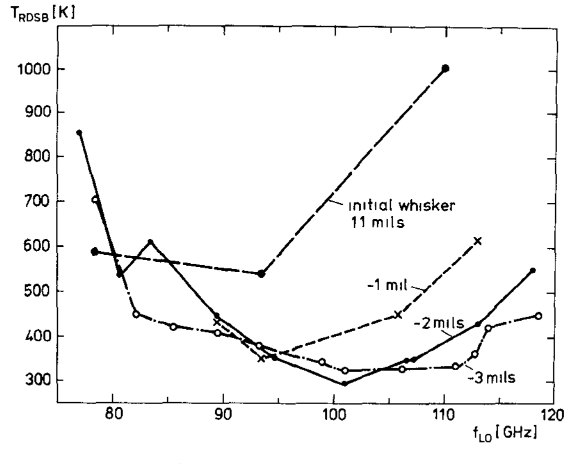

Fig. 8. Optimization of the receiver noise temperature over the 80 to 115-GHz range by varying the whisker inductance. The DSB receiver noise temperature at 295 K with a 1.4-GHz IF is plotted for the initial 0.011-in-long whisker and for whiskers 0.001, 0.002, and 0.003 in shorter. The whisker inductance is directly proportional to length.

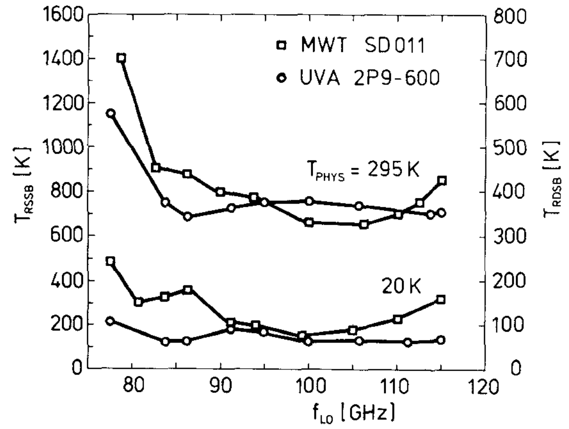

Fig. 9. DSB receiver noise temperatures with a 1.4-GHz IF are plotted versus LO frequency for a MWT and a UVA diode in different mixers at room temperature and 20 K.

of the signal frequency. A similar improvement in conversion loss between 4.75 and 1.4 GHz was noted by Kerr [10]. For comparison, Nussbaum *et al.* [23] have achieved SSB conversion losses of 6 ± 1 dB over the 90 to 120-GHz band with an image-enhanced mixer using beam-lead diodes.

B. Optimization of the RF Impedance

Fig. 8 shows the DSB receiver temperature at 295 K using a MWT SD-011C diode with a 1.4-GHz IF. The first results were somewhat narrow band using a whisker of 12-μm diam and 280-μm (.011 in) length. The total length is approximately $L + W_b$ (see Fig. 3(a)) [10]. To investigate the RF response of the mixer versus whisker length, the same whisker was shortened in 25-μm (.001 in) steps by etching. At each length, the diode was recontacted and its performance was measured as is shown in Fig. 8 [20]. The optimum length was in the range from 200 to 230 μm, corresponding to an inductance of ≈ 0.2 nH [19]. The effect of whisker inductance has been studied by Siegel and Kerr [24]. Their numerical results showed the importance of tuning the whisker inductance for a given diode capacitance. Held [25] also found the importance of whisker length from scale model measurements and experimental results.

Once the importance of the whisker length was established, a University of Virginia (UVA) diode was mounted in another mixer block and the whisker inductance optimized to give even lower noise and broader performance. The results for these two mixer blocks are summarized in Fig. 9. Both the MWT and UVA diodes gave very similar results at 295 K, being < 800 K SSB over much of the band. When cooled to 20 K, the UVA 2P9-600 diode results are only slightly better (~ 10 percent) in the center of the band from 90 to 105 GHz, but this diode maintains its low-noise performance (< 200 K SSB) over the entire 78 to 115-GHz range. The average SSB temperature is 150 K with a best temperature of 124 K, measured with a 50-MHz bandwidth. The DSB receiver temperature only varies from 60 to 80 K over an IF of 1.2 to 1.6 GHz when measured in a 50-MHz bandwidth. This mixer has been used in the FCRAO-cooled receiver since 1981.

C. Contributions to the Receiver Noise Temperature

The total receiver noise temperature is divided into 3 parts: 1) contributions from input losses; 2) mixer contributions including mixer waveguide losses, diode losses, and IF matching circuit losses; and, 3) the IF contribution. With this in mind, the DSB receiver temperature can then be written as

$$T_{\mathrm{RDSB}} = (L_W - 1)T_{\mathrm{phys}} + L_W T_{\mathrm{MXR}} + L_{\mathrm{DSB}} T_{IF} \quad (2)$$

where T_{phys} is the physical temperature of the feed horn, ring filter, and mixer block, and T_{IF} the IF noise temperature, which is 40 K at room temperature and 10 K at cryogenic temperatures for a 1.4-GHz IF. The total DSB mixer noise is

$$T_{\mathrm{MXR}} = (L_{RF} - 1)T_{\mathrm{phys}} + L_{RF}(L_D - 1)T_{eq} + L_{RF}L_D(L_{IF} - 1)T_{\mathrm{phys}}. \quad (3)$$

The equivalent temperature T_{eq} of the diode as a lossy element will approach T_D as a limit when $R_s = 0$, higher harmonics are reactively terminated, parametric effects are negligible, and there is no excess diode noise as a function of current. The short-noise limit for a diode [26] is given by

$$T_D = (qV_o/2k) \quad (4)$$

when the diode current i for a voltage V across the diode and its series resistance R_s is

$$i = i_s e^{[(V - iR_s)/V_o]}. \quad (5)$$

The electron charge is q and Boltzman's constant is k. This would be the case if the noise from the mixing process is dominated by shot noise, which was found to be true for room-temperature mixers by Held and Kerr [12].

Two different diodes in separate mixer blocks were measured at their best operating frequencies with a 1.4-GHz IF. These results are presented in Table II for the MWT SD-011 and UVA 2P9-600 diodes at both 295 and 20 K.

TABLE II
DETAILED MIXER NOISE TEMPERATURE CONTRIBUTIONS WITH A 1.4-GHz IF

Two diodes at physical temperatures of 295 and 20 K are compared. Their dc-series resistance and slope parameter V_o are given. The measured DSB conversion loss L_{DSB} and noise temperature T_{RDSB} are used to derive the equivalent noise temperature of the diode T_{eq}. This is compared to the shot-noise limited diode noise temperature T_D.

Diode	MWT SD-011		UVA 2P9-600	
Physical Temp.	295 K	20 K	295 K	20 K
R_s (Ω) V_o (mV)	4.4 31	8 13	12 28	12 8.0
L_{DSB} (dB) L_D (dB)	2.6 1.3	2.3 1.3	3.4 2.1	3.1 2.1
T_{RDSB} (K)	294	68	344	62
T_{IF} (K)	40	10	40	10
T_{MXR} (DSB) (K)	155	43	185	35
T_{eq} (K)	249	108	178	48
$T_D = qV_o/2k$ (K)	180	77	165	47

The diode slope parameter V_o and series resistance R_s are fit over the 1 μA to 5-mA range. Typical operating conditions are 500 μW for 800 μA at 0.8 V and 200 μW for 300 μA at 0.8 V for the LO power and diode dc current and voltage of the MWT and UVA diodes, respectively. The total loss was measured at room temperature. The conversion loss with the MWT diode is 0.8 dB lower than with the UVA diode, partly due to more capacitance variation and corresponding parametric effects and also due to its lower series resistance. The voltage variable capacitance of the diodes can be written as

$$C(V) = C_{jo}(1 - V/\phi)^{-\gamma} \quad (6)$$

where ϕ is the barrier potential, and C_{jo} is the junction capacitance for zero bias. Both diodes have similar exponents, $\gamma \approx 0.4$, but the MWT diode has a higher C_{jo} (see Table IV). It also requires more LO power because of its higher V_o when cooled [2], which will give a larger voltage swing and consequently more capacitance variation and larger parametric effects.

In Table II, T_{eq} is derived from room-temperature and cryogenic measurements for two diodes in similar mixers. The series resistance R_s and slope parameter V_o were measured at dc. The total conversion loss was measured at room temperature and the diode loss L_D derived using (1). The DSB receiver and the IF noise temperatures give the DSB mixer temperatures from (2). Finally, T_{eq} is derived from (3) and compared to the theoretical shot-noise limit T_D. The derived T_{eq} is with 40 percent and 10 percent of T_D for the MWT and UVA diodes, respectively. The most likely cause for this discrepancy is noise from parametric effects due to increased LO power (500 μW for the MWT diode versus 200 μW for the UVA diode).

Both of these diodes gave excellent results for two reasons. One, the diodes have a low effective noise at 1.4 GHz as a function of dc bias, rising only to 300 K at a current of 5 mA [27]. As was pointed out by Held [28], a good dc

TABLE III
COMPARISON OF 3-mm COOLED SCHOTTKY DIODE MIXERS

Various cryogenically cooled mixers are compared for their: input losses L_W; total SSB conversion loss L_{SSB}; SSB mixer temperature T_{MSSB}; IF noise temperature T_{IF}; and total SSB receiver noise temperature T_{RSSB}. The receiver temperature for [29] is derived by assuming a T_{IF} of 25 K. The results of the last two mixers are from this paper.

DIODE	UVA 3.5µ φ	UVA 2.5µ φ	BTL 280-92	BTL	BTL cryo	BTL 280-92	MWT SD-011	UVA 2P9-600
L_W (dB)	0.6	0.6	-	0.6	0.5	0.4	0.5	0.5
L_{SSB} (dB)	7.2	5.8	6.7	7	6	7.5	5.3	6.1
T_{MSSB} (K)	280	300	209	200	120	91	86	70
T_{IF} (K)	20	20	22	20	24	25	10	10
T_{RSSB} (K)	445	435	312	348	250	240	136	124
Reference	[2]	[10]	[9]	[28]	[22]	[29]		

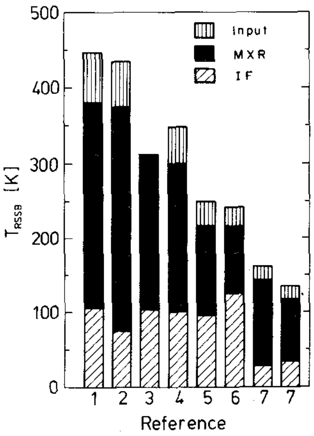

Fig. 10. Comparison of noise contributions to 3-mm cooled mixers. The SSB receiver noise temperature is divided into 3 parts: 1) due to the conversion loss and IF noise temperature (lower section); 2) due to the mixer (center section); and 3) due to input losses before the mixer. References are: 1-[2]-[10], 3-[9], 4-[28], 5-[22], 6-[29], 7-[this paper].

TABLE IV
COMPARISON OF 3-mm SCHOTTKY DIODES

Various whisker-contacted diodes are compared for their: zero-bias capacitance C_{jo}; γ and ϕ are from (6); doping; junction area; physical temperature; dc-series resistance R_{SDC}; and slope parameter V_o; the log of the saturation current I_s; the shot-noise limited diode temperature T_D; and the experimental diode equivalent temperature T_{eq}.

DIODE	UVA 2.5µ φ [10,12]	BTL N280-92 [9]		BTL "room temp" [8]		BTL "cryo" [8]		MWT SD-011		UVA 2P9-600		UVA 2P8-500 [31]	
C_{jo} (fF)	7	14.7		19		13		8		6.5		6.5	
γ	0.4	≈0.2		0.5		≈0.2		0.5		0.4		0.4	
ϕ (V)	0.95	1.0		1.0		1.0		1.1		1.1		1.06	
Doping (cm^{-3})	3×10^{17}	3×10^{16}		2×10^{17}		5×10^{16}		2×10^{16}		3×10^{16}		3×10^{16}	
Junction Area ($\mu m)^2$	4.9	11		14.5		9.7		5		3.1		3.1	
T_{phys} (K)	295	295	18	295	20	295	23	295	18	295	22	295	18
R_{SDC} (Ω)	8.0	4.5	7.4	8	8	6	6	4.4	8	12	12	13	13
V_o (mV)	28	28	12	31	12	29	26	31	13	28	8.0	27	18
$-\log_{10}(i_s)$ (A)	16.1			16	42.5	17	21	15.9	38.2	16.0	57.4	17.6	29.6
$T_D = qV_o/(2k)$	164 K	164	69	180	70	170	154	180	77	165	47	160	105
T_{eq} (K)	323		47	250		150		249	108	178	48		

noise curve is necessary, but not sufficient, to give a low (< 200 K) equivalent temperature for the pumped diode. Depending on the mixer circuit, correlated shot-noise from high harmonics averaged over a cycle of the LO can make the average noise greater than 300 K [26]. However, our mixer circuit is such that these correlated noise components are minimized and a low equivalent diode temperature is realized.

D. Comparison with Previous 3-mm Schottky Receivers

In analyzing the excellent results that have been obtained with the present receiver, it is important to separate out improvements in the Schottky diodes, improvements in the embedding circuit due to the entire mixer design, and reductions in the IF noise temperature. The present results are compared with previous work on cryogenic 3-mm receivers in Table III and Fig. 10 for 7 different mixers. The total SSB conversion loss is given by L_{SSB}, while L_W and T_{IF} are the same as in Section IV-C. The net SSB receiver temperature is

$$T_{RSSB} = 2(L_W - 1)T_{phys} + L_W T_{MSSB} + L_{SSB}T_{IF}. \quad (7)$$

In Fig. 10, the total receiver temperature is separated into 3 parts. The IF contribution is denoted by the lower section of each bar, the mixer contribution by the center part, and the input losses by the top section of each bar. The system noise temperature given by Linke *et al.* [9] includes losses in the Fabry-Perot filter used for sideband rejection and LO injection, so the input losses were not separated out from the total SSB loss.

There has been a steady improvement in the noise temperatures since millimeter cryogenic receivers were first reported in 1973 [2], in both mixer and IF contributions. The IF noise temperature has improved by a factor of two with the introduction of cooled 1.4-GHz FET amplifiers. This, in addition to an improved conversion loss of 0.2 to 1.2 dB, has lowered the IF contribution from 90 ± 10 K to 35 ± 3 K.

Mixer noise temperatures which were 300 K SSB for the first cooled 3-mm receivers were reduced to 200 K and now to 70 to 86 K. While the input losses L_W have remained constant at 0.5 to 0.6 dB, their effect has diminished from 60 to 15 K as the IF and mixer noise have decreased, since the input contribution is 6 K plus 15 percent of the mixer and IF contributions.

The parameters of the various diodes used by Kerr [10], [12], Linke *et al.* [9], the BTL diodes, MWT and UVA diodes used in the present mixer design, and the UVA diodes used by Räisänen *et al.* [31] are summerized in Table IV. On comparing the UVA diode used by Kerr [10] with the MWT or UVA diodes used in this design at room temperature, one notes that the electrical parameters are quite similar. The noise from the dc-biased diode is essentially the same [27], [28], so excess noise at high currents cannot explain the difference. So, the improvement in the equivalent temperature of the mixer as an attenuator T_{eq} from 323 to 178 K is entirely due to the mixer design.

Although the room temperature parameters have been the same since 1975, the slope parameter V_o of the lower doped ($<10^{17}$ cm^{-3}) diodes have, in general, been im-

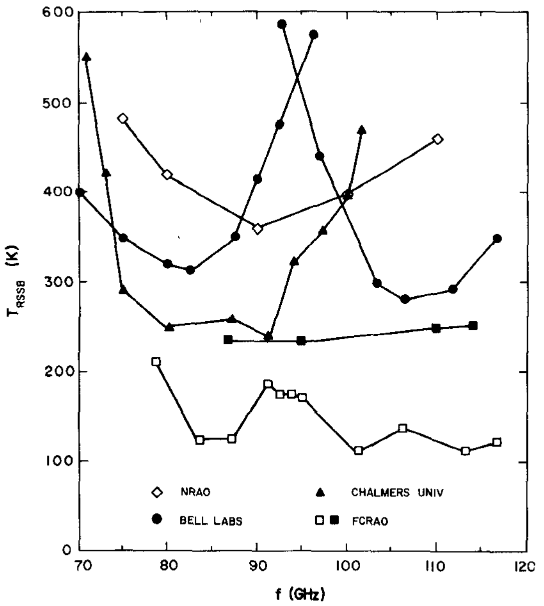

Fig. 11. Frequency response of 3-mm cooled mixers. Single sideband receiver noise temperatures are plotted over the 70 to 120-GHz range for the NRAO mixer [31], the Bell Telephone Laboratory system [9], the Chalmers University system [29], and the present design which is used at the FCRAO. The FCRAO DSB and SSB measurements were made with 50 and 500-MHz bandwidths, respectively. Open symbols are twice DSB results and filled symbols are SSB temperatures.

proved, as recommended by Viola and Mattauch [32], [33]. Occasionally, the $I-V$ curve of the cooled diode is not a single exponential but a combination of exponentials, due to different parts of the diode having different barrier heights [34], [35]. The slope parameter and series resistance of the diodes in Table IV have been derived by a fit of the dc $I-V$ curves with only a single exponential over the 1 μA to 5-mA current range.

While Fig. 10 compared various 3-mm mixers at their best operating points, the frequency response of the SSB receiver temperatures are compared in Fig. 11 for the NRAO receiver [36], the BTL system [9], the Chalmers University system [30], and the FCRAO receiver with a UVA 2P9-600 diode. The FCRAO data is just twice the measured DSB temperature in a 50-MHz bandwidth, since the sideband gains have been measured equal to within 4 percent. When used on the telescope, the FCRAO receiver has a Martin-Puplett [37] type sideband filter which, together with coupling optics, has $\simeq 0.5$-dB loss. The SSB receiver temperature with this filter is also plotted in Fig. 11 for a measurement bandwidth of 500 MHz. The only better receivers in this frequency range are the recent SIS junction mixers reported by Pan *et al.* [38] which have achieved 68 K SSB at 115 GHz.

V. Conclusions

The combination of a broad-band mixer design, excellent Schottky barrier diodes, and low-noise FET amplifiers has given mixer receivers whose noise temperatures are the best ever obtained with Schottky diodes. This has been accomplished over the broadband of 80 to 115 GHz. At a physical temperature of 20 K, the SSB mixer noise temperature is 70 K in the best case, an improvement factor of two over the best previous results.

The equivalent temperature of the diode as an attenuator is within 10 percent of the shot-noise limit due to the improved embedding circuit for the diode. This has been accomplished by an RF filter and backshort design which are near to a short circuit in the fundamental band and reactive in the second harmonic band. The whisker inductance has been optimized for a broad-band performance over the 3-mm band. The use of OFHC copper in the mixer construction with its high thermal and electrical conductivity is important in keeping the circuit losses low.

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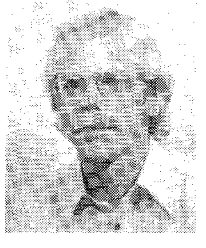

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Imaging Polarimeter Arrays for Near-Millimeter Waves

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Abstract —**An integrated-circuit antenna array has been developed that images both polarization and intensity. The array consists of a row of antennas that lean alternately left and right, creating two interlaced subarrays that respond to different polarizations. The arrays and the bismuth bolometer detectors are made by a photoresist shadowing technique that requires only one photolithographic mask. The array has measured polarization at a wavelength of 800 μm with an absolute accuracy of 0.8° and a relative precision of 7 arc min, and has demonstrated nearly diffraction-limited resolution of a 20° step in polarization.**

I. Introduction

Recently, imaging antenna arrays have been developed that make images at near-millimeter wavelengths [1]–[3]. The idea is that an image is focused on an array of antennas with individual detectors, and the power received by each antenna is plotted to form an image. Fig. 1 shows how these systems work. An objective lens focuses an image onto the array through a lens on the back of the substrate. This substrate lens takes advantage of the fact that antennas are most sensitive to radiation from the substrate side. These arrays have demonstrated diffraction-limited resolution at 1.2 mm [2]. The antennas in these arrays are all linearly polarized, and measure only one component of the electric field so that the polarization angle is not measured directly.

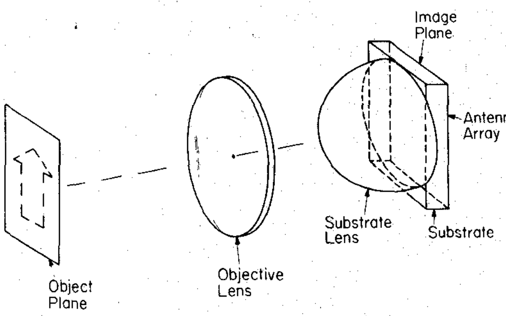

Fig. 1. Substrate-lens coupled optical system.

Polarimeters measure polarization, and can be useful in determining material properties. For example, radars can measure surface roughness by analyzing polarization changes on reflection because rough surfaces depolarize, while smooth surfaces maintain polarization [4]. In biochemistry, the concentration of sugars can be measured by the rotation of polarization of transmitted light [5]. In plasma diagnostics and semiconductors, the polarization change by Faraday rotation is proportional to the magnetic field [6], [7].

A variety of different polarimeter schemes have been implemented. In microwaves, two linearly polarized antennas measuring orthogonal components of an electric field form a polarimeter [8]. In optics, the two orthogonal components can be split by a Wollaston prism, and measured independently [9]. In near-millimeter waves, three other

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