An 86–106 GHz Quasi-Integrated Low Noise Schottky Receiver

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***Abstract*—An integrated planar receiver has been developed and tested over the 82–112 GHz bandwidth. The quasi-integrated antenna used in the receiver has a high gain, a high Gaussian coupling efficiency and a wide bandwidth. The novel mixer design consists of a planar GaAs Schottky diode placed at the feed of a dipole-probe suspended inside an integrated horn antenna. The diode uses an etched surface channel and a planar air bridge for reduced parasitic capacitance. At 92 GHz, the room temperature antenna-mixer exhibits a double sideband conversion loss and noise temperature of 5.5 ± 0.5 dB and 770 K ± 50 K, respectively. The measured DSB conversion loss and noise temperature over a 20 GHz bandwidth (86 GHz–106 GHz) remain less than 6.2 dB ± 0.5 dB and 1000 K ± 50 K, respectively. The low cost of fabrication and simplicity of the design makes it ideal for millimeter- and submillimeter-wave receivers.**

I. Introduction

FUNDAMENTAL SCHOTTKY-DIODE mixers are currently used in most millimeter-wave receivers above 100 GHz. The mixers use either a whisker-contacted diode or a planar Schottky diode suspended in a machined waveguide with an appropriate RF matching network. Whisker-contacted Schottky barrier diodes have been the preferred room temperature non-linear elements for submillimeter-wave receivers for the last two decades. Although whisker-contacted Schottky diodes have the lowest shunt capacitance, they require waveguide mounts or corner-cube antennas which are very expensive to machine for frequencies above 200 GHz. Also, whiskered diodes are not generally compatible with integrated planar antennas. Integrated receivers are easier to manufacture, more reliable, smaller, lighter, and much less expensive than waveguide mixers when produced in large quantities. They are an attractive solution for the millimeter-wave region (30–300 GHz) and offer a practical solution for the submillimeter-wave region (300 GHz–3 THz). Integrated receivers can also be arrayed for linear or two-dimensional imaging without a large increase in cost and weight of the system.

Planar antennas have traditionally showed low coupling efficiencies to reflector systems due to their poor radiation patterns and power loss to substrate modes [1]. This has limited their widespread use in millimeter-wave and submillimeter-wave receivers. Research on integrated circuit antennas in the last decade has solved many of the problems associated with planar antennas on electrically thick dielectric substrates [2]. Double-slot [3], double-dipole [4], and spiral antennas [5] on dielectric lenses have been successfully used for planar SIS receivers at millimeter-wave lengths. However, these antennas exhibit a Gaussian coupling efficiency of 70–80% resulting in a 1 to 1.5 dB increase in the receiver conversion loss. Another limitation of planar room temperature receivers was the unavailibility of high quality/low parasitic planar Schottky diodes. This problem has been solved recently with the advent of the University of Virginia (UVa) etched surface channel diode. This structure results in a lower parasitic capacitance and a higher associated RF impedance [6]–[8].

This paper describes a novel wide band planar mixer using a UVa surface-channel Schottky diode. The RF coupling structure is a modification of the integrated horn antenna which consists of a pyramidal cavity with a 70° flare angle etched anisotropically in silicon. The cavity focuses the incoming energy on a dipole-probe suspended on a membrane inside the horn [9], [10]. The antenna has a Gaussian coupling efficiency of 97% which is a considerable improvement over standard integrated-circuit antennas. The mixer circuit, along with the feed dipole, are both integrated on the membrane wafer. The IF signal is taken out through a simple CPS line on the silicon substrate and through an IF matching network. The mixer requires no tunning and no additional RF matching network, and shows excellent performance over a 20 GHz bandwidth. The low cost of fabrication and the simplicity of the design makes it ideal for use in millimeter- and submillimeter-wave receivers.

II. Antenna Design

A quasi-integrated horn antenna has been designed by Eleftheriades *et al.* [11] to overcome the limitations of the large flare angle of the integrated horn antenna introduced in [10], [12]. In this design, a flared machined section is attached to the front part of the integrated horn antenna to result in higher gain patterns (Fig. 1(a)). The abrupt change of flare-angle at the junction of the integrated section and the machined section of the horn acts as a mode converter that excites mainly the TE_{10}, TE_{12}/TM_{12} and TE_{30} modes. These modes are properly phased on the radiating aperture by appropriately selecting the length and the flare-angle of the machined section. The minimum dimension of the machined

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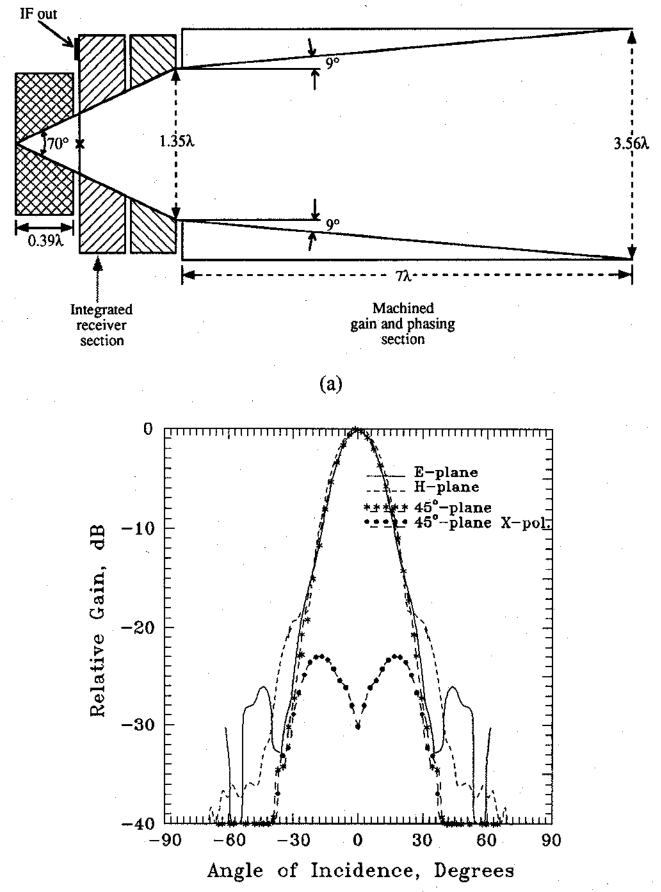

Fig. 1. (a) The 20 dB quasi-integrated horn antenna. (b) The measured E-, H-, 45°-, and x-pol. 45°-plane patterns of the quasi-integrated.

TABLE I

Frequency (GHz)	80	85	90	95	100	105	110
Gaussian coupling efficiency %	86	95	97	97	92	88	86
Gain (dB)	18.7	19.4	20.0	20.6	21.3	21.9	22.7

section is about 1.35λ, which should allow the fabrication of the quasi-integrated horn up to 1.5 THz. The measured patterns at 91.4 GHz of a 20 dB gain quasi-integrated horn antenna (Fig. 1(b)) show low sidelobe-levels, a -22 dB cross-polarization level in the 45° plane, a 10-dB beamwidth of 34°, and a calculated 97% Gaussian coupling efficiency [11]. The aperture dimensions are 3.56λ-square and this results in a 62.5% aperture efficiency (coupling to a plane wave) at 91.4 GHz. Table I shows the calculated gain and coupling efficiency of the 20 dB gain quasi-integrated horn antenna from 80 to 110 GHz. It is seen that the Gaussian coupling efficiency varies between 0.0 dB and -0.5 dB over a 30 GHz bandwidth.

III. Mixer Design

The quasi-integrated receiver consists of an integrated section (Fig. 2), and a machined section, which is attached to the front of the integrated horn antenna and is not shown in

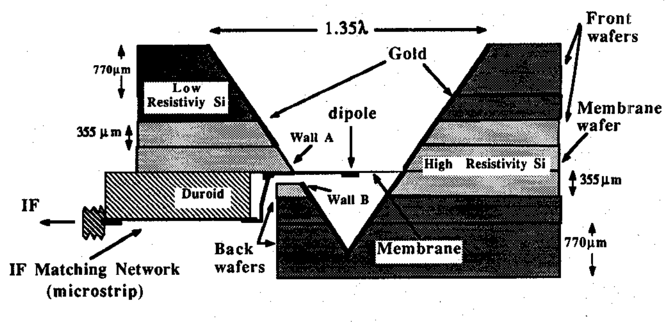

Fig. 2. The integrated horn antenna receiver structure. Wall A is coated with gold, and a V-groove is etched at the side of wall B.

the figure. The receiver structure consists of low and high resistivity silicon wafers etched anisotropically and stacked together to form a pyramidal cavity with a 70° flare angle. The feed-dipole with the CPS lines is integrated on the membrane wafer as shown in Fig. 3. The horn sidewalls of each silicon wafer, except for the wafer containing the membrane, are coated with gold by flood evaporation before assembling the horn atennna. For the membrane wafer, only three sidewalls are coated with gold and the sidewall in front of the coplanar stripline (wall A, Fig. 2) is left uncoated so as not to short the IF and RF signals on the CPS line. A microwave model of the receiver structure of Fig. 2 was built at 2.55 GHz and a 1.1–1.2 dB power loss was measured due to power leakage through the uncoated horn sidewall (wall A). The uncoated sidewall (wall A) does not, however, have any effect on the far-field patterns. For the same reason, the sidewall behind the coplanar stripline (wall B, Fig. 2) must also be far enough from the CPS line. To facilitate the fabrication of wall B, a V-shaped groove is etched in the silicon wafer directly behind the membrane and wall B is coated with gold. The height of the V-shaped groove is 350 μm which is ten times larger than the separation between the two strips forming the CPS line. The cavity is small enough so that no higher-order modes are present at 92 GHz. The effective quasi-TEM dielectric constant is therefore equal to the mean dielectric constant between air and silicon ($\epsilon_e = (1 + \epsilon_r)/2$).

The length of the feed-dipole and its position inside the integrated horn antenna are designed to present an approximate conjugate match to the RF diode impedance [12]. The planar diode is therefore epoxied right at the dipole apex without any additional RF matching network (Fig. 3). An RF choke is found along the CPS line by using two integrated lumped capacitors. The circuit is integrated on highly resistive silicon in order to minimize any losses of the IF signal on the surrounding dielectric substrate. The first capacitor on the membrane itself is $\lambda_o/4$ away from the dipole apex and the second capacitor on the silicon is $\lambda_e/2$ away from the first one, where λ_e is the effective wavelength of the CPS line on silicon covered on top by the V-shaped groove. The capacitors are fabricated by evaporating a 70 μm $\times$ 240 μm gold patch over the CPS line and using a 1.1 μm insulating polyimide dielectric. The separation between the two strips of the CPS line on the silicon wafer is reduced to 30 μm rather than the 60 μm on the membrane in order to minimize radiation losses of the CPS line. The CPS line impedance is 240 Ω on the

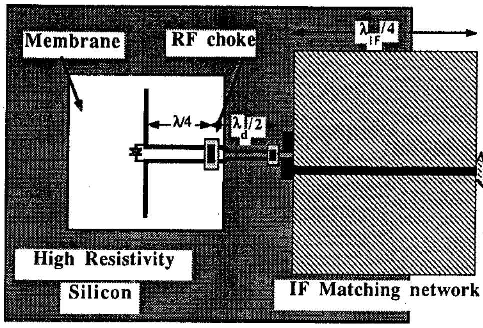

Fig. 3. The mixer design consisting of the Schottky diode epoxied at the feed-dipole apex, the two lumped capacitors forming the RF choke, and the microstrip line IF matching network.

membrane and 80 Ω on the silicon substrate. The capacitor integrated on the membrane is 90 fF approximately and the one integrated on silicon is 50 fF. This results in a LPF with a −3 dB corner frequency of 35 GHz and a rejection of −19 dB at 92 GHz. A 75 Ω microstrip quarter-wave transformer is fabricated on a Duroid 5870 substrate [13] and used to match the 1.4 GHz IF diode output impedance to 50 Ω. The IF matching network is specifically not integrated on the high resistivity silicon substrate to facilitate the use of different IF matching networks. The transition from the integrated coplanar stripline to the Duroid substrate, obtained by either using silver epoxy or bonding wires, is not important due to the low IF frequency used. For higher IF frequencies, the IF matching network can be integrated directly on the wafer by using a transition from coplanar stripline to coplanar waveguide. For array applications with severe space constraints on the matching network, a transformer can be integrated using a lumped capacitor and a lumped inductor [14].

The diode of choice to be used in the mixer design is the UVA SC2T3 planar Schottky diode with 2.5 μm anode diameter, a 5–6 fF zero-bias junction capacitance, an 8–9 fF parasitic capacitance, an ideality factor of 1.12 and a 2.5–3 Ω dc series resistance. The diode chip dimensions are $400\,\mu m \times 140\,\mu m \times 55\,\mu m$ yielding approximately a 6–7 fF parasitic capacitance (C_{p1}) between the contact pads and a 2 fF parasitic capacitance (C_{p2}) between the anode finger tip and the cathode. This diode has shown good performance at millimeter wave frequencies with RF embedding impedances around $50 + j50\,\Omega$ [15]. This impedance could be approximately achieved at 91.4 GHz by using a 0.39λ-long dipole positioned 0.38λ from the apex in a fully conducting horn cavity. The measured input impedance on our scale model was $85 + j11\,\Omega$ at the design frequency due to the effect of the uncoated sidewall (wall A) and the V-shaped groove (wall B). This input impedance drops to $75 - j7\,\Omega$ when a stycast block ($e_r = 12$) is put at the apex of the feed-dipole to model the parasitic capacitance (C_{p1}) of the GaAs diode (Fig. 4). Both measurements were sensitive to the feed geometry and were repeatable to $\pm 5 \pm j5\,\Omega$. Therefore, an input impedance of $80 - j1\,\Omega$ is taken at 91.4 GHz for the dipole antenna (without C_{p1}), and fits well with video responsivity measurements. Table II shows the mixer theoretical performance for the UVA SC2T3 diode for an LO frequency of 90 GHz, an RF frequency of 91.4 GHz, a dc bias of 0.65 V and an available LO power at the dipole terminals of 1.6 mW. The analysis was done using the reflection algorithm developed by Kerr *et al.* [16]. It is seen that a minimum SSB conversion loss of 5.4 dB is achievable without a matching network, and the SSB conversion loss remains under 7.0 dB over a 20% bandwidth.

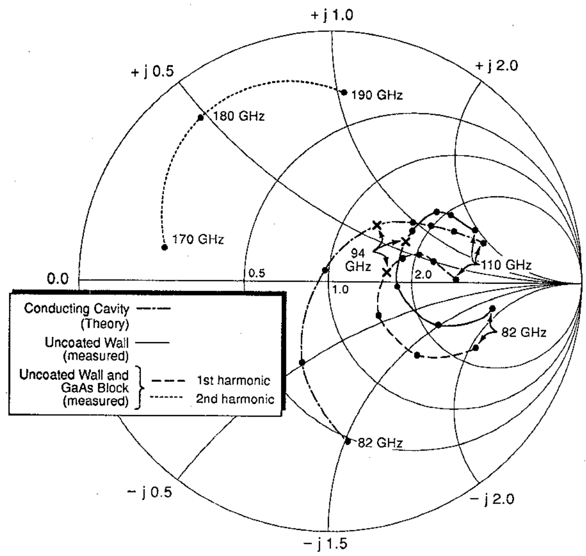

Fig. 4. The measured feed-dipole impedance over the 82–110 GHz and 170–190 GHz range using the 2.55 GHz microwave model of the integrated horn antenna receiver structure. The data points are 4 GHz apart.

TABLE II

f_{LO}(GHz)	90
f_{RF}(GHz)	91.4
$Z^{measured}_{dipole,RF}(\Omega)$ (without C_{p1})	80-j1
$Z^{measured}_{dipole,2RF}(\Omega)$	6+j28
$Z^{in}_{diode,RF}(\Omega)$	69-j28
$Z^{in}_{diode,LO}(\Omega)$	64-j51
$Z^{out}_{diode,IF}(\Omega)$	110+j8
Diode SC2T3 SSB Conversion loss(dB)	5.4
Diode SSB Conversion loss(dB) over 20% BW	5.0-7.0

IV. Receiver Measurements

A quasi-integrated horn antenna receiver was built at 90 GHz and a UVA SC2T3 planar Schottky diode was epoxied at the dipole apex (Fig. 5). Video detection measurements were done at 91.4 GHz by shining a plane wave with known power density on the quasi-integrated horn antenna and measuring the output detected diode voltage in a 120 kΩ load using a lock-in amplifier. The RF plane wave is calibrated to ±5% using an Anritsu power meter and a standard gain horn. The video responsivity is defined here as the ratio of the detected low-frequency voltage across a 120 kΩ load over the total RF plane wave power incident on the antenna aperture. This definition includes a 2.0 dB loss resulting from

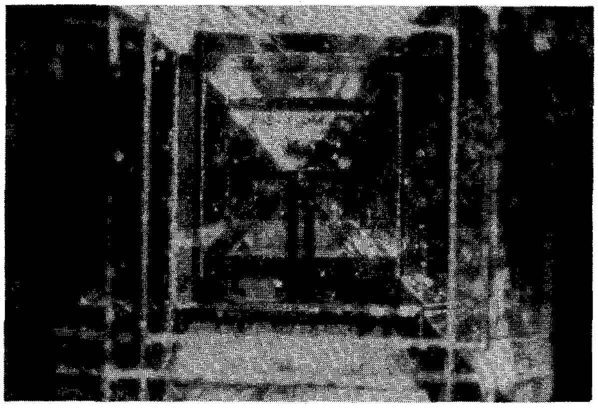

Fig. 5. The integrated horn antenna built at 90 GHz where the feed-dipole, the CPS line, and the first lumped capacitor close to the horn wall can be seen.

the 62.5% aperture efficiency (coupling to a plane wave) of the quasi-integrated antenna. Fig. 6 shows the theoretical and measured video responsivity versus bias current. The theoretical video responsivity is calculated using the measured dipole input impedance and horn sidewall loss obtained from the microwave model, and the diode parameters supplied by the University of Virigina (Table III). The series resistance has been increased to 7 Ω to account for skin-depth losses. It is seen that a very good agreement exists between theory and measurement over four decades of bias current. The peak video responsivity is 1140 V/W which is equivalent to 1.62 V/(mW/cm^2) when referring it to the incident power density on the antenna aperture.

The responsivity can also be defined as the detected low frequency voltage into a 120 KΩ load divided by the power available at the dipole terminals (80 Ω RF source). This definition excludes the aperture efficiency loss (2 dB) and the horn side wall loss (1.2 dB), and results in a peak responsivity of 2380 V/W which is competitive with whisker diodes at these frequencies.

The double-sideband conversion loss and noise temperature of the mixer were measured using a hot/cold load method. A Mach–Zender interferometer is used to combine the RF signal from a hot/cold load and the LO signal from a Gunn oscillator (Fig. 7). The lenses are 4 in diameter an the Gaussian beam waist at the lens interface is designed so that the lens edges are illuminated by the −35 dB power level in the Gaussian beam. A grooved teflon lens with an f-number of 0.75 is used to match the large Gaussian beam waist at the interferometer to the small Gaussian beam waist of the quasi-integrated horn antenna ($w_{o(ANT)} = 1.11\lambda$). The IF chain consists of a 10 dB coupler, a bias-T, a circulator, a low-noise amplifier chain, a 100 MHz band pass filter centered at 1.4 GHz, and a calibrated IF power meter. The 10 dB coupler is used for measuring the IF power reflection coefficient $|\Gamma_{IF}|^2$ at the mixer output port [17]. The IF chain, including the 10 dB coupler, has a 91.9 dB gain (G_{IF}) and a 129 K noise temperature (T_{IF}). The gain increases to 92.9 dB and the IF noise temperature drops to 78 K when the 10 dB coupler is removed. The receiver DSB

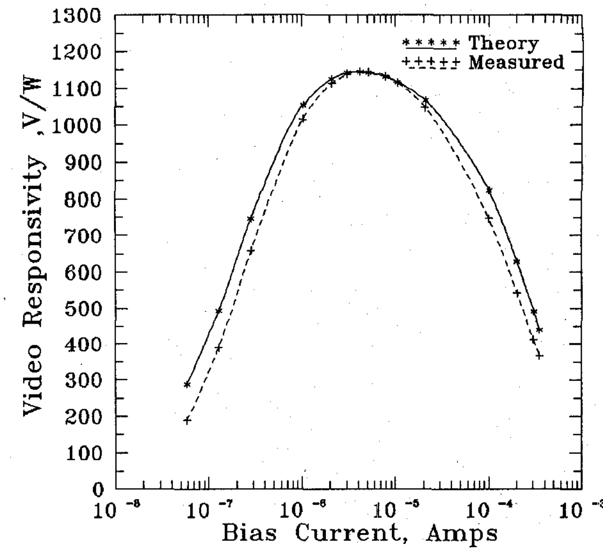

Fig. 6. The theoretical and measured video responsivity of the antenna-mixer versus bias current at 91.4 GHz.

TABLE III

Antenna			Diode				
$\epsilon_{aperture}$	$\epsilon_{losssinwalls}$	Z_{dipole}	R_s	C_j	C_p	ϕ_{bi}	η
-2.0dB	-1.2dB	80-j1Ω	7Ω	5.5fF	8.5fF	0.89V	1.12

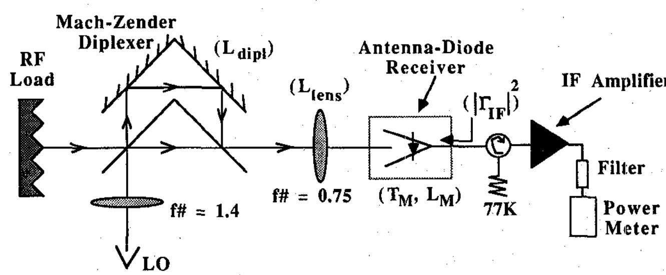

Fig. 7. The receiver DSB measurement set-up.

conversion loss is given by

$$L_{rcvr} = \left(\frac{P_H - P_C}{K(T_H - T_C)B} \right) \Big/ G_{IF} \tag{1}$$

and the receiver DSB noise temperature is given by

$$T_{rcvr} = \frac{T_H - YT_C}{Y - 1}, Y = P_H/P_C \tag{2}$$

with $T_H = 295$ K and $T_C = 85$ K. The T_C is an estimate for Eccosorb AN-72 temperature when dipped in liquid nitrogen at millimeter-wave frequencies [18]. L_{rcvr} includes the antenna-mixer conversion loss (L_M), the diffraction loss in the diplexer (L_{dipl}), the loss in the 0.75 f-number lens due to reflection and absorption (L_{lens}), and the loss due to the non-zero IF reflection coefficient at the mixer output port (L_{IF}). L_M includes the antenna Gaussian coupling loss, the power loss in the uncoated sidewall, the diode intrinsic conversion loss, and the RF mismatch between the diode impedance and the feed-dipole impedance.

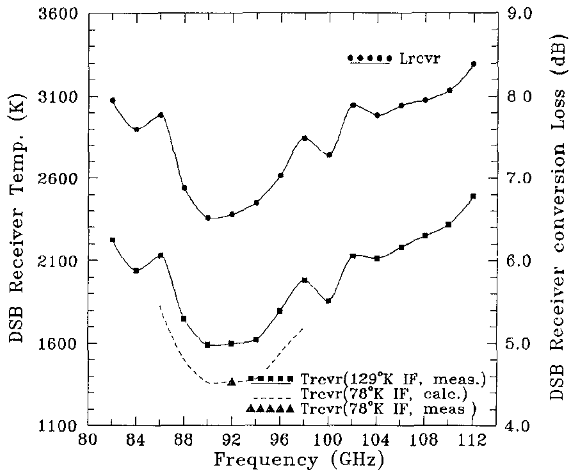

Fig. 8. The measured receiver DSB noise temperature and conversion loss over the 82–112 GHz range.

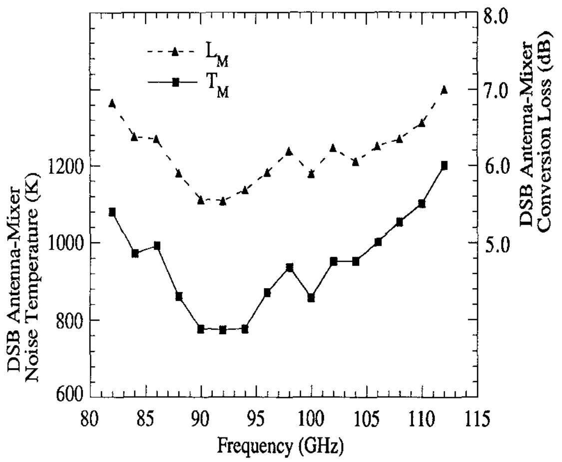

Fig. 9. The measured antenna-mixer DSB noise temperature and conversion loss. The measured values remain less than 6.2 dB ± 0.5 dB and 1000 K $\pm$ 50 K, respectively, over a 20 GHz bandwidth (86–106 GHz).

Hence, L_{rcvr} is given by

$$L_{rcvr} = L_M L_{dipl} L_{lens} L_{IF}$$

with

$$L_{IF} = 1/(1 - |\Gamma_{IF}|^2) \tag{3}$$

The additional losses in the system increase the DSB noise temperature of the receiver, and T_{rcvr} is given by

$$\begin{aligned} T_{rcvr} &= (T_o(L_{dipl}L_{lens} - 1) + T_M L_{dipl} L_{lens}) \\ &\quad + L_{rcvr}(T_{isolator}^{effective}|\Gamma_{IF}|^2 + T_{IF}) \qquad (T_o = 295\,\mathrm{K}) \end{aligned} \tag{4}$$

where T_M is the DSB noise temperature of the antenna-mixer. L_{rcvr} and T_{rcvr} are determined from (1) and (2), and knowing L_{rcvr}, L_{lens}, and $|\Gamma_{IF}|^2$, one can determine L_M and T_M from (3) and (4), respectively.

The receiver DSB temperature and conversion loss were measured from 82 GHz to 112 GHz (Fig. 8). At each frequency, the distances between the quasi-optical components are adjusted to maximize the signal at the receiver. In each case, the LO power and dc bias voltage were adjusted for minimum overall noise temperature T_{rcvr}. All the best data points are obtained for a 1.5–2 mW LO power available at the dipole terminals and a 0.91 V–0.92 V dc bias voltage corresponding to a 1.2 mA–1.5 mA bias current. The LO power level needs to be increased by 30% if it is referred to the aperture of the quasi-integrated horn antenna. A minimum receiver conversion loss of 6.5 dB $\pm$0.3 dB and noise temperature of 1600 K $\pm$ 100K are obtained at 90 GHz and 92 GHz. This noise temperature includes the IF-chain noise temperature of 129 K. A receiver noise temperature of 1370 K was measured at 92 GHz without the 10 dB coupler (Fig. 8).

The loss in the lens L_{lens} was estimated empirically to be 0.5 ± 0.1 dB over the frequency range of measurements. The diffraction loss in the diplexer L_{dipl} varied between 0.3 dB and 0.16 dB over the 82 GHz to 112 GHz range [19]. The IF power reflection coefficient was measured at each frequency and was lower than 0.2 dB at the optimum bias points and LO power. The DSB antenna-mixer conversion loss L_M and noise temperature T_M are calculated using (3) and (4), and a minimum antenna-mixer DSB conversion loss of 5.5 dB $\pm$ 0.5 dB is obtained at 90 GHz and 92 GHz, and a mimimum antenna-mixer DSB noise temperature of 770 K $\pm$50 K is obtained over the 90 GHz to 94 GHz range (Fig. 9). The measured DSB conversion loss and noise temperature over a 20 GHz bandwidth (86 GHz–106 GHz) remain less than 6.2 dB $\pm$ 0.5 dB and 1000 K $\pm$ 50 K, respectively. This includes the drop in the Gaussian coupling efficiency of the quasi-integrated horn antenna at the low and high frequencies. With the 100 MHz band pass filter in the IF chain was replaced by a 50 MHz band pass filter, the DSB measurements at 92 GHz were virtually the same.

SSB measurements for an LO frequency of 92 GHz and an RF frequency of 93.4 GHz were done by halving the interferometer path-length difference and thereby tuning the interferometer as a single-sideband filter [19]. Fig. 10 shows the measured SSB antenna-mixer conversion loss (L_M) at 93.4 GHz versus LO power available at the dipole terminals. The dc bias voltge was adjusted to minimize the IF power reflection coefficient and was measured to be less than 0.15 dB at optimal LO powers. A minimum SSB conversion loss of 7.3 dB ± 0.5 dB was obtained for an LO power between 1 and 2.5 mW. The measured SSB conversion loss of 7.5 dB agrees well with theory which predicts a 6.5 dB conversion loss (1.2 dB + 5.3 dB antenna-mixer conversion loss; 1.2 dB is horn sidewall loss).

V. COMPARISON WITH OTHER MIXER DIODES

The quasi-optical mixer presented in this paper shows excellent performance from 86–106 GHz. The mixer requires no tuning whatsoever, is planar and compatible with monolithic techniques. The measured DSB conversion loss over a 20 GHz bandwidth is 1–2 dB higher than wideband tunerless waveguide mixers [20], [21]. In a narrow band range, the measured data at 92 and 94 GHz compares favorably with waveguide mixers using a planar diode and is within ~1.8 dB of the best published SSB conversion loss measurement, ~3 dB of the best DSB published conversion loss measurement,

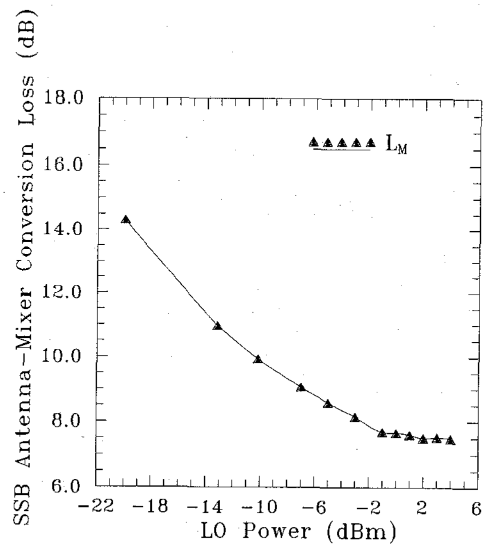

Fig. 10. The measured antenna-mixer SSB conversion loss with $f_{LO} = 92$ GHz and $f_{RF} = 93.4$ GHz.

and ~2.8 dB of the best DSB noise figure at room temperature [8]. The reader is referred to the paper by Garfield *et al.* [8] for a chronological listing of measured SSB conversion loss and noise temperature of whisker-contacted and planar Schottky diode mixers at room temperature. The performance of the quasi-integrated receiver can be improved by 1.5–2 dB by eliminating the 1.2 dB RF power loss in the uncoated sidewall and choosing better dipole impedance. A series of gold coated via-holes in the high resistivity silicon membrane wafer can be etched just next to the uncoated sidewall to synthesize an RF short. Currently, it is easy and inexpensive to array four of these receivers on a single chip. Together, the four planar receivers have a sensitivity similar to that of one of the best tuned waveguide mixers.

VI. Conclusions

An 86–106 GHz quasi-integrated horn antenna receiver has been designed and tested. The measurements show that this new receiver is a very good candidate for millimeter-wave applications. The mixer performance compares favorably with the best waveguide mixers which also use a surface-channel UVa planar diode over the same frequency range. Work is continuing on developing a 250 GHz quasi-integrated receiver using planar Schottky diodes with smaller anode diameter.

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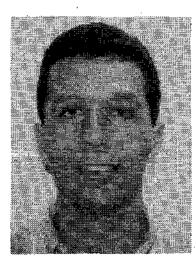
[21] Millitech Corp., *Millimeter and Submillimeter Components Catalog*, Southdeerfield, MA, 1989.



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