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Super-Schottky Mixer Performance at 92 GHz

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***Abstract*—As part of a program to explore the behavior of superconducting Schottky mixers at high frequencies ($\nu_{RF} \geqslant 90$ GHz), the mixing and video performance of several super-Schottky diodes have been tested at 92 GHz. The diodes used ($\sim$3-μm active diameter, doping concentration $\sim 2\times10^{19}$ cm^{-3}) were identical to those recently developed at Aerospace for use in a 31-GHz mixer. The WR-10 mixer mount, designed specifically for this experiment, utilizes a quartz stripline assembly for the diode, whisker, and IF choke, suspended across quarter-height RF waveguide.**

At 92 GHz, video responsivities were typically $\sim$80 A/W (corrected for RF mismatch). Conversion loss (corrected for both RF and IF mismatches) was typically measured to be $\gtrsim$18 dB. As expected, T_{diode} was small (<5 K). Video responsivity and conversion loss were also measured at an RF frequency of 3.95 GHz. These data were used with the measured I–V characteristics of the diodes to compare theoretical predictions of diode performance at 92 GHz in both the video and mixing modes, with the high-frequency data.

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I. INTRODUCTION

THE DEVELOPMENT of increasingly sensitive receivers at millimeter wavelengths imposes stringent mixer performance requirements. The super-Schottky di-

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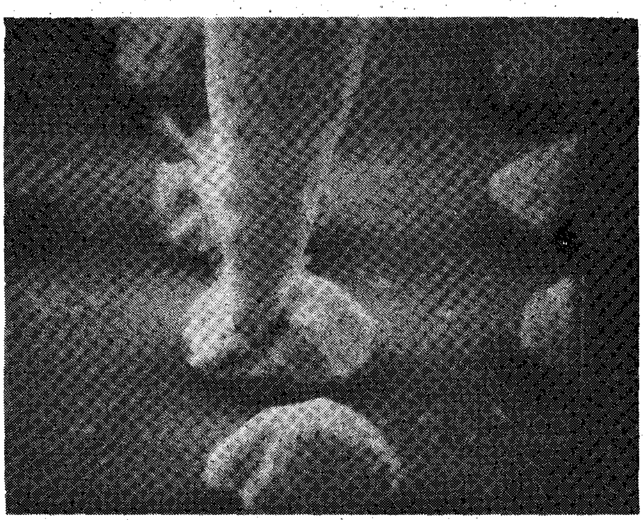

Fig. 1. Scanning-electron microscope (SEM) photograph of a super-Schottky diode array and whisker contact. The diameter of the diode mushrooms is 5 μm.

ode, a superconductor–semiconductor tunneling junction, has performed well as a mixer at both 9 and 31 GHz. At 9 GHz, the diode output temperature was measured to be 1.2 K with a single-sideband conversion loss of 7 dB [1]. At 31 GHz, use of an improved diode [2] also yielded an output temperature of 1.2 K; however, conversion loss increased slightly to 9 dB [3]. Given its low output noise temperature, the super-Schottky mixer appeared promising for low-noise receiver applications at short millimeter wavelengths. Uncertainties persisted, however, regarding the extent to which high-frequency (i.e., 90 GHz and above) performance would be degraded by junction parasitics and possible quantum-loss mechanisms [4]. Consequently, a program to study the performance of a super-Schottky mixer at ~90 GHz was undertaken. Results of the first stage of this program, that of testing available diodes originally fabricated [2] for use at 31 GHz, are reported here.

II. Diode and Mixer Mount

The super-Schottky diodes used in this experiment were provided by the Solid State Electronics Department of the Aerospace Corporation's Electronics Research Laboratory and were similar to those tested recently at 31 GHz [2], [3]. Arrays of 3-μm-diameter Pb dots were laid down by high-field pulse plating over 2×10^{19} cm^{-3} p-type gallium arsenide (GaAs) prepared as described in [2]. The Pb contacts were formed into mushroom shapes with $\approx$5-μm diameter to eliminate mechanical stress at the Pb/GaAs Schottky interface, which could destroy the extreme nonlinearity of the diode. A thin gold overplating was then applied to the contacts to inhibit oxide growth. Once fabricated, the large chip was sawed into 0.005×0.010-in squares for mounting. All diodes tested in this experiment were derived from the same parent chip. A scanning-electron microscope (SEM) photograph of a typical diode array (with a whiskered contact) is shown in Fig. 1.

The mixer mount design was based on previous uncooled and cryogenic millimeter-wave mixers [5]. A cross section of the mount is shown in Fig. 2. The signal enters via WR-10 waveguide, which is then stepped down to lower impedance quarter-height guide. A contacting short behind the mixer is used for tuning. The diode, whisker, and IF choke are mounted on a quartz stripline assembly across the reduced-height waveguide. Further details of the mount's construction and assembly are given in [6].

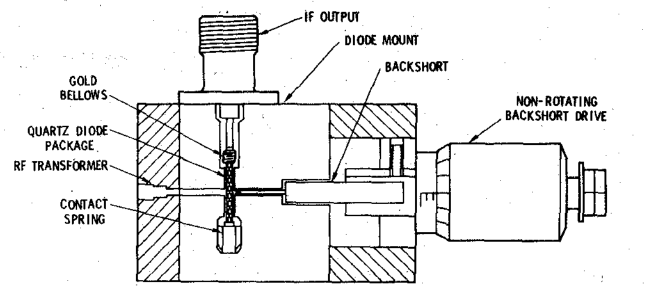

Fig. 2. Cross-sectional view of the 92-GHz super-Schottky mixer mount used in these measurements. The input waveguide is WR-10 (0.100×0.050 in).

III. Measurement System and Test Procedures

A. Measurement System

A number of similar measurement system configurations were used in the course of this experiment. For clarity, we focus on the configuration used in the last series of tests, as shown in Fig. 3(a)–(c) in block diagram form. A convenient division of the test system into three subsystems—cold assembly, RF system, and IF system—facilitates discussion. Each is described below.

1) Cold Assembly: This subsystem incorporates the mixer, RF waveguide components, and IF coaxial components which reside in the dewar. The RF assembly consists of two WR-12 silver waveguide lines, with a section of thin-wall stainless-steel waveguide inserted in each run to increase thermal isolation. The signal in one line (usually the local oscillator (LO) signal) passes through a 20-dB metallized mica attenuator. The other waveguide run (used for calibration and tuning signals) couples into this line below the attenuator via a high-directivity ($\gtrsim$40 dB) 10-dB directional coupler.

Three identical 0.141-in-diameter semirigid coaxial lines carry output signals from the mixer, a 50-Ω termination, and a short circuit up through the header plate to a coaxial switch. These lines have a stainless outer jacket (for thermal isolation) and a beryllium–copper inner conductor (for relatively low IF loss). A fourth stainless coaxial line with a 50-Ω load can be used to terminate the IF receiver input circulator. The super-Schottky mixer and the IF lines mounted in the cold assembly are shown in Fig. 4.

Careful waveguide and coaxial-loss measurements were essential to derive desired mixer parameters. These loss measurements were repeated several times at 295 and 1 K in the course of the experiment.

2) RF System: As shown in Fig. 3(a), this subsystem incorporates the mixer, LO assembly, and klystron, calibration gas tube, and signal and RF reflection assembly with an auxiliary 92-GHz receiver. The LO power incident at the mixer was determined by measuring the klystron power

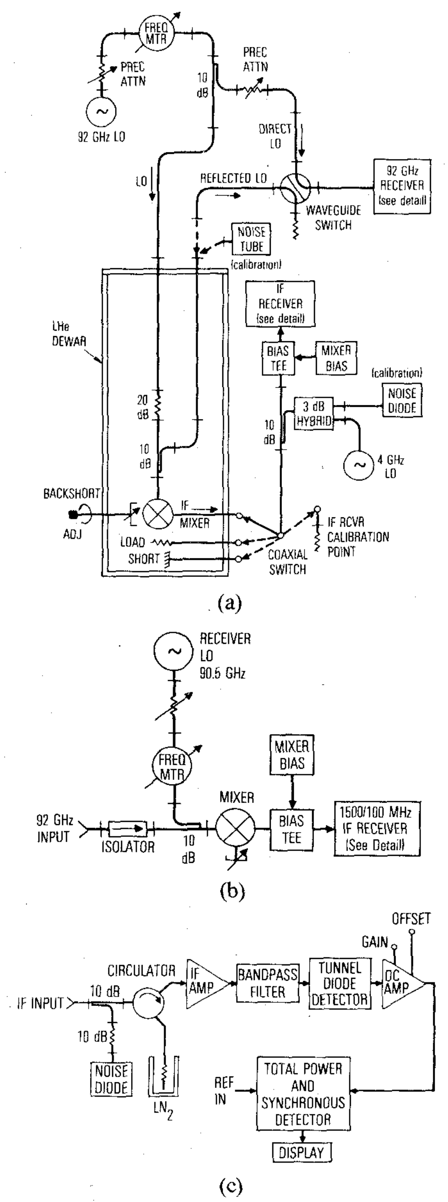

Fig. 3. (a) Block diagram of the super-Schottky measurement system. (b) Detail of RF reflectometer receiver. The IF receiver used is functionally identical to that shown in Fig. 3(c). (c) Block diagram of IF receivers used in this experiment. The two receivers used had center frequencies/bandwidth (MHz) of 1500/100 and 350/100.

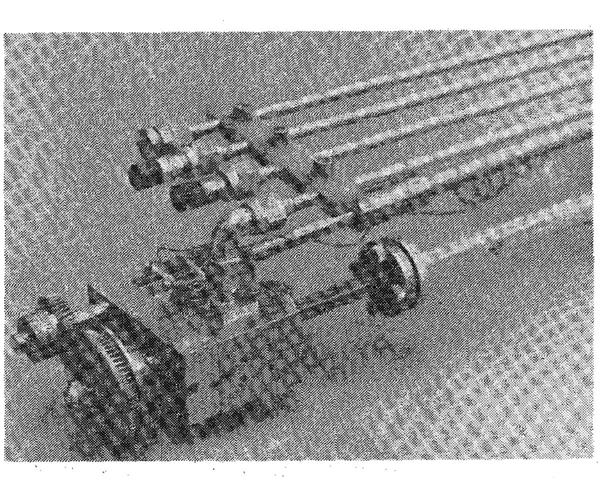

Fig. 4. Photograph of the super-Schottky mixer and IF coaxial lines mounted in the dewar cold assembly. A carbon resistor temperature sensor is mounted on top of the mixer.

at the dewar input flange and accounting for the waveguide losses in the cold assembly. In most of our measurements, the mixer was tuned for optimum performance by connecting a gas tube noise source and synchronously modulated ferrite switch directly on the signal line at the dewar header. After tuning, the ferrite switch was removed and the noise tube, whose radiometric output temperature was measured to be $(12.7 \pm 0.6) \times 10^3$ K at 90 GHz, was placed directly on the signal flange to measure mixer conversion loss.

For RF reflection measurements, the 92-GHz auxiliary was used to compare the reflected LO signal from the mixer (up through what is normally the calibration line) with the attenuated LO signal coupled into the receiver via the waveguide switch shown in Fig. 3(a). Knowing the round-trip losses in the dewar assembly and the attenuation of the direct LO signal then allowed an estimate of the high-level mixer return loss to be made. In subsequent discussions we assume this to be nearly identical to the small-signal RF return loss.[1]

3) IF System: Two IF receivers at 350 and 1500 MHz were used during the course of these measurements. The design of both is basically the same; a block diagram is shown in Fig. 3(c). The receivers incorporate a noise diode injected into a directional coupler (for IF reflection measurements), an input circulator with a cooled termination, low-noise amplifiers, a filter to define the bandpass, and a tunnel diode detector followed by a low noise dc amplifier to provide an output signal proportional to input power. Gain and offset controls are provided on the dc amplifier to scale the output to read directly in degrees Kelvin. A synchronous detector was also incorporated in all receivers and its output signal was used to optimize mixer performance when the input RF signal was synchronously modulated. The bulk of the measurements reported here were made with the 350-MHz (100-MHz bandwidth) radiometer because it generally gave the best mixer performance; possibly this was due to the lower IF loss in the mixer, less sideband gain imbalance, or lower calibration errors (since line losses were lowest at 350 MHz).

B. 4-GHz RF System

Because all diodes tested here are taken from the same chip, it was necessary to verify that the quality of the junctions tested was similar to those previously measured at lower frequencies. This was done by determining mixer output temperature, conversion loss, and video responsivity using an RF of 4 GHz, with a 350-MHz IF. Even without a backshort to assist in RF matching, it was expected that

[1]In general, the high- and low-level return losses may be different. However, at 3.95 GHz the corrections for RF mismatch applied to our data are small; moreover, a comparison of high- and low-level return losses for a super-Schottky mixer at 9 GHz [1] showed the two to differ only slightly (R. J. Pederson, private communication). At 92 GHz, the small-signal return loss was calculated from the impedance mismatch between the 1/4-height waveguide and the mixer, modeled following Lidholm [15]. The diode parameters were taken from the present set of measurements, whisker inductance was calculated from the whisker dimensions [6], and a purely reactive backshort was assumed. A minimum value (at optimum backshort setting) of $|\Gamma^2| \sim 0.6$ was obtained, very close to that measured for the high-level LO.

similarity of these results to those obtained previously at X-band [2] would be a good indication of unimpaired junction quality (see Section IV-A). Further, as will be seen, the 4-GHz measurements assisted in analyzing the observed mixer and video behavior at 92 GHz.

The 4-GHz measurements were made by using a high-directivity coupler mounted at the IF input to inject LO and RF calibration signals along the common coaxial line to the mixer IF port (see Fig. 3(a)). A 4-GHz signal generator and a calibrated noise diode were combined in a 3-dB hybrid to provide the signals. Line losses to the dewar header were measured and added to the dewar coaxial losses at the respective RF and IF frequencies to calibrate the measurements.

C. *Video Responsivity Measurements*

Video responsivity was measured by frequency modulating the mixer LO ($\sim$100 Hz) at various incident power levels and synchronously detecting the video output voltage with a lock-in amplifier. At 90 GHz, the LO klystron was injected through the LO waveguide, while at 4 GHz, the signal was injected through the mixer IF port. The mixer was biased for optimum output response in these measurements using a constant-current bias supply.

D. *Mixer Measurements*

Mixer conversion loss, output-noise temperature, and IF reflection coefficient can be determined from mixer and short outputs when the calibration and IF noise diode signals are cycled [5]—provided that mixer output voltages are calibrated in degrees Kelvin at the IF port. In practice, IF radiometer output voltages were calibrated in degrees Kelvin at the coaxial switch using ambient and liquid-nitrogen-cooled terminations. Consequently, corrections were applied to the data, as detailed in [6], to account for the loss and emission of the dewar IF line, and IF receiver noise and mismatches when the noise diode was fired. IF data from the mixer, load, and short allow these contributions to be self-consistently determined assuming that coaxial-line mismatches occur within the IF receiver. This assumption is conservative, in the sense that mixer output-noise temperatures so derived are in fact upper limits.

IV. Results and Discussion

A. *Results*

Experimental results for the super-Schottky diodes are presented in Table I. Tabulated quantities are typical values selected from the larger set of many measurements. Estimated worst case standard errors are also listed in the table and were derived assuming all measurement uncertainties to be uncorrelated [6]. The tabulated uncertainties are believed to be conservative.

Several trends are evident in the data. Despite very low diode output-noise temperatures (expected for a super-Schottky junction [1]), mixer conversion efficiency at 92 GHz is poor; likewise, video responsivity is observed to be roughly two orders of magnitude worse at 92 GHz than at either 9 or 31 GHz [1], [3]. As noted in Section III-B, the mixer and video responsivity performance observed at 3.95

TABLE I
Summary of Super-Schottky Results

			MIXER				DETECTOR	
ν_{LO} (GHz)	ν_{IF}/BW (MHz)	P_{LO} (nW)	SSB[a] L_{conv} (dB)	T_{diode} (K)	$\lvert\Gamma_{IF}\rvert^2$	$\lvert\Gamma_{RF}\rvert^2$	R_d (Ω)	$\boldsymbol{R}$ (A/W)
92	1500/100	100	25[b]	≤ 3[c]	0.20[d]	—	—	—
92	350/100	500	22[b]	≤ 3[c]	0.14[d]	0.63[g]	140	30
3.95	350/100	16	8.3[e]	≤ 5[f]	0.06	0.15[g]	156	2800

Notes:
(a) Single-sideband conversion loss, corrected for IF mismatch
(b) Estimated worst-case error $\leq$ 2 dB (Ref. 6)
(c) Estimated worst-case error $\leq$ 80% (Ref. 6)
(d) Estimated worst-case error $\leq$ 10% (Ref. 6)
(e) Estimated worst-case error $\leq$ 1 dB (Ref. 6)
(f) Estimated worst-case error $\leq$ 4 K (Ref. 6)
(g) See § IIIA.2.

GHz suggests that the diode chip fabricated for this experiment is comparable to those previously tested at X-band [1]. Although the mixer RF reflection coefficient was found to be rather large (roughly 50 percent of the incident RF signal being lost by reflection), the conversion loss would only improve by $\sim$3 dB if a perfect match could be attained. Moreover, sideband imbalances can neither satisfactorily account for the generally poor conversion efficiency at 350-MHz IF (a sideband separation of 700 MHz), nor is it required to explain the marginally worse performance (just below our estimated uncertainty level) at the 1500-MHz IF.

B. *Discussion*

Previously developed theoretical models [1], [2] for super-Schottky diodes can be used to discuss the observed video and mixing performance of the junctions at both 4 and 92 GHz. Within the limitations of the theories, a comparison of data at both frequencies affords an important check of the experiment's internal self-consistency. Moreover, a good agreement between the experimental results and theoretical predictions would strongly argue for the basic validity of the theoretical models.

1) Video Responsivity: Consider first the video data at 4 GHz. Correcting the observed responsivity for RF mismatch yields

$$\boldsymbol{R}_{\text{cor}}(4) = \frac{\boldsymbol{R}_{\text{obs}}(4)}{\left(1 - |\Gamma_{RF}|^2\right)} \sim 3294 \text{ A/W}. \quad (1)$$

Theoretically, $\boldsymbol{R}_{\text{cor}}$ is related to the diode's S-parameter (a measure of I–V nonlinearity—cf., below) and physical temperature T, by [7], [8]

$$\boldsymbol{R}_{\text{cor}} = \frac{S}{2L_p'} \left[\frac{\tanh(h\nu/2kT)}{(h\nu/2kT)} \right] \quad (2)$$

where L_p' is the diode's parasitic loss in the video mode, h and k are Planck's and Boltzmann's constants, respectively, and ν is the RF frequency. The factor in square brackets bridges the high- and low-frequency regimes of video operation. In the absence of parasitics, $\boldsymbol{R}_{\text{cor}}$ [1] corresponds to the energy per unit charge of electrons undergoing photon-assisted tunneling. At low frequencies, an energy $\sim(q/2kT)^{-1}$ is required to raise an electron above the

Fermi sea for tunneling; this corresponds to the absorption of many LO photons. At high frequencies, however, each electron can only absorb a single photon—hence, $\boldsymbol{R}_{\text{cor}}$ must have the asymptotic limit $\sim q/h\nu$. These limits are, therefore, in general applicable to all video detectors. The explicit $\tanh(x)/x$ form in (2) is a consequence of the (assumed) exponential I–V characteristic of the super-Schottky diodes around the bias voltage point. At $T=1$ K, the bracketed factor has the values 1.00 and 0.44 at 3.95 and 92 GHz, respectively.

The parasitic loss of the diode appearing in (2) can be expressed as [1]

$$L_p' = 1 + R_s/R_d + \omega^2 C^2 R_s R_d. \tag{3}$$

Here

- ω RF angular frequency, rad·s^{-1}
- R_s spreading resistance, Ω
- R_d dV/di dynamical diode resistance (Ω), from I–V curve (Table I)
- C junction capacitance, F.

At dc, the super-Schottky diodes tested here (with a doping concentration $N=2\times10^9$ cm^{-3}) have a spreading resistance [9] $R_s \cong 8.3$ Ω. At 92 GHz, skin-depth effects [11] can be expected to increase this value; for example, Held and Kerr [12] have measured an increase of ~ 2 Ω at 115 GHz in R_s for a 2.5-μm GaAs diode. In what follows, therefore, we shall assume $R_s \cong 8.3$ Ω at 4 GHz, and $R_s \cong 10$ Ω at 92 GHz, respectively.

The junction capacitance can be found from [10]

$$C = A\left[\frac{q^2 \epsilon \epsilon_0 N}{2V_B}\right]^{1/2} \text{ F} \tag{4}$$

where

- A junction area, m^{-2} $=\pi D^2/4$
- q electron charge, 1.6×10^{-19} C
- ϵ dielectric constant of GaAs at 1 K, ~ 12.5
- ϵ_0 $=8.85\times10^{-12}$ F·m^{-1}
- N doping concentration $\cong 2\times10^{25}$ m^{-3}
- V_B barrier height near zero bias $\cong 0.4$ eV$=6.4\times10^{-20}$ J.

Hence $C \cong 149\times10^{-15}$ F.

Since $R_d = 156$ Ω was measured at 4 GHz, one predicts $L_p' = 1.07$ (0.3 dB) at that frequency. Relations (1) and (2) then imply $S\sim 7058$ V^{-1}. Now assuming a perfectly exponential I–V characteristic with no leakage to describe the I–V curve measured, one has $I = I_0[\exp(SV)-1]$, where I_0 is the saturation current. Using the I–V curve measured when the above data were taken, we obtain $S\sim 5600$ V^{-1}, a difference of only ~ 25 percent.

Using the measured value of S, the predicted video responsivity at 92 GHz using (2) is

$$\boldsymbol{R}_{\text{pred}} = \frac{5600}{2L_p'(92)} \times 0.44 = \frac{1232}{L_p'(92)} \text{ A/W}. \tag{5}$$

Evaluating $L_p'(92)$ using (3) and the measured value $R_d =$ 140 Ω then yields

$$L_p'(92) = 11.42 \text{ (10.6 dB)} \tag{6}$$

and hence

$$\boldsymbol{R}_{\text{pred}}(92) \cong 108 \text{ A/W}. \tag{7}$$

In fact, correcting $\boldsymbol{R}_{\text{obs}}(92)\sim 30$ A/W for the observed RF mismatch yields $\boldsymbol{R}_{\text{cor}}(92)\sim 81$ A/W. This is within about 30 percent of the predicted value and therefore in reasonable agreement.

TABLE II
PREDICTED VERSUS OBSERVED SUPER-SCHOTTKY CONVERSION LOSSES AT 4 AND 92 GHz

	CALCULATED			OBSERVED	
ν_{LO} (GHz)	L_p [1] (dB)	L_o [2] (dB)	L_{pred} [3] (dB)	L_{obs} [4] (dB)	$(L_{pred} - L_{obs})$ (dB)
3.95	0.5	6.7	7.2	7.6	-0.4
92	13.7	6.7	20.4	17.7	2.67

NOTES: 1 Parasitic loss (see text).
2 Intrinsic conversion loss from DC resistance ratio (see text).
3 $L_p + L_o$
4 Corrected for RF and IF mismatch; ν_{IF} = 350 MHz.

2) Conversion Loss: A comparison of mixer conversion loss at the two RF frequencies, while more uncertain than the foregoing, is also of interest. Neglecting ohmic losses, the single-sideband conversion loss of a broad-band mixer (corrected for RF and IF mismatches) is conventionally represented as a product of two terms

$$L = L_0 L_p. \tag{8}$$

L_p is the parasitic loss of the mixer and is functionally similar to (3) [2]

$$L_p = (1 + R_s/R_{\text{IF}})(1 + R_s/R + \omega^2 C^2 R_s R). \tag{9}$$

Here R is the RF resistance of the nonlinear junction, R_{IF} its IF resistance, and the other terms have their previous meanings. Since R and R_{IF} were not directly measured, a straightforward evaluation of L_p is impossible. However, R_{IF} can be estimated from the slope of mixer I–V curve at optimum bias (with the LO applied). Since the bias point and LO drive levels for optimum mixing were similar to their best video response values, the approximation $R_{\text{IF}} = R_d$ may be used. This leads to a contribution ~ 0.3 dB to L_p from the first parenthetical term in (9). Further, to within a factor of ~ 2, the approximation $R = 2R_{\text{IF}}$ is conventionally employed [11]. Noting that the uncertainty in this last assumption translates into <0.1 and ~ 3-dB errors in L_p at 3.95 and 92 GHz, respectively, we obtain the results given in the Table II.

L_0 is the intrinsic loss associated with the frequency conversion process. It can be estimated from the following relations [1] using the measured dc resistance ratio $(R_{\max}/R_{\min})$ for the diode (≈ 140). Let

$$x = \tfrac{1}{2}\ln(R_{\max}/R_{\min}). \tag{10}$$

Then

$$L_0=(2/\eta)\left[1+\sqrt{1-\eta}\right]^2 \quad (11)$$

where

$$\eta=\frac{2I_1(x)^2}{I_0(x)\left[I_0(x)+I_2(x)\right]} \quad (12)$$

and $I_j(x)$ is a modified Bessel function of the first kind of order j.

Several uncertainties are associated with this estimate. First, it implicitly assumes a Y-connected embedding network for the mixer (i.e., all harmonics of the RF and LO short-circuited), probably an adequate assumption given the large capacitance of the diode. Second, (9) assumes the LO drive level to encompass a full sweep of the diode conductance characteristic, a situation not likely to be fully attained. In both cases, deviations from the ideal will raise L_0. A final assumption implicit in the (classical) relations above is the neglect of collective quantum effects. These might either further degrade L_0 or improve it in a manner similar to the conversion gain predicted and observed in SIS mixers [13], [14]. Detailed theoretical modeling will be required to settle this issue.

Bearing these limitations in mind, a comparison of observed conversion losses at 4 and 92 GHz with those anticipated from the above relations (subject to the assumptions stated) was made. Results are given in Table II. The close agreement between measured and predicted losses may be fortuitous, given both the experimental and modeling uncertainties. However, the agreement between the experimental results at two widely differing RF frequencies and the theoretical models invoked here does suggest that these models are adequate for evaluating the potential of the super-Schottky diode at frequencies $\gtrsim$100 GHz. This being the case, it is clear that the major factor contributing to the poor diode performance observed at 92 GHz is junction parasitics. Therefore, an effort to reduce junction capacitance while maintaining spreading resistance values comparable to those here will be required, if substantially improved mixing and video performance at high frequencies is to be attained.

V. Summary

Both the video responsivity and mixing performance of heavily doped ($N=2\times10^{19}$ cm^{-3}) 3-μm active-diameter super-Schottky diodes have been measured at 92 GHz. A typical single-sideband conversion loss of $\sim$18 dB (corrected for RF and IF mismatches) was measured; a video responsivity of $\sim$80 A/W (likewise corrected for RF mismatch) was also observed. Conversion loss and responsivity were also measured at an RF frequency of 3.95 GHz, enabling the data at both frequencies to be compared using theoretical models. Agreement between the observations and model predictions is surprisingly good considering the various uncertainties (both theoretical and experimental) which enter into the comparison. This argues for both the internal consistency of the experiment performed here, as well as the basic validity of the theories developed to describe the super-Schottky diode. A major effort to reduce junction parasitics in the device will be required before mixing performance comparable to that observed at 10 and 31 GHz can be attained at frequencies $\gtrsim$100 GHz.

Acknowledgment

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