Measurements on a 215-GHz Subharmonically Pumped Waveguide Mixer Using Planar Back-to-Back Air-Bridge Schottky Diodes

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***Abstract*—This paper presents design and performance data for a 215-GHz subharmonically pumped waveguide mixer using an antiparallel-pair of planar air-bridge-type GaAs Schottky-barrier diodes. The waveguide design is a prototype for a 640-GHz system and uses split-block rectangular waveguide with a 2 : 1 width-to-height ratio throughout. The measured mixer noise and conversion loss are below that of the best reported whisker contacted or planar-diode mixers using the subharmonic-pump configuration at this frequency. In addition, the required local oscillator power is as low as 3 mW for the unbiased diode pair, and greater than 34 dB of LO noise suppression is observed. Separate sideband calibration, using a Fabry–Perot filter, indicates that the mixer can be tuned for true double sideband response at an intermediate frequency of 1.5 GHz. Microwave scale model measurements of the waveguide mount impedances are combined with a mixer nonlinear analysis computer program to predict the mixer performance as a function of anode diameter, anode finger inductance, and pad-to-pad fringing capacitance. The computed results are in qualitative agreement with measurements, and indicate that careful optimization of all three diode parameters is necessary to significantly improve the mixer performance.**

I. Introduction

THE availability of high-quality low-capacitance planar-integrated GaAs Schottky-barrier diodes [1]–[3] has stimulated a reexamination of current millimeter- and submillimeter-wave heterodyne receiver technology. Many high-frequency waveguide mixer designs, which previously were considered to be mechanically impractical, can now be implemented using integrated circuit techniques. As a prototype for a 640-GHz waveguide Schottky-diode system, we have taken a closer look at the subharmonically pumped antiparallel-diode pair arrangement described by [4] and [5], analyzed by [6], and very successfully demonstrated at millimeter wavelengths by [7] and [8]. The antiparallel-diode pair has the advantages of subharmonic pumping and inherent local oscillator (LO) noise suppression [5], [9], both useful for submillimeter-wave operation, where sources are difficult to come by. Two disadvantages of the anti-parallel-diode pair configuration are: increased LO power in the case of commonly or unbiased diodes and, at least for the whisker-contacted configuration, a mechanical structure which is much more difficult to implement, As pointed out in [6], there are also constraints on the obtainable mixer performance due to the practical realization of the diode circuitry (e.g, loop inductance between the diode pair).

Since the antiparallel-diode pair mixer concept was demonstrated at millimeter wavelengths [7], [8], several groups have reported on similar configurations but always with poorer performance [10]–[14] than that of the original Bell Laboratories group. In this paper, we report on a 215-GHz subharmonically pumped two-diode mixer design using planar air-bridge Schottky-barrier diodes which performs slightly better than the whisker contacted devices presented in [8]. Section II describes the mixer mount design and includes measured impedance data taken with a microwave scale model of the mixer block. Section III gives performance data for the 215-GHz mixer mount using a variety of planar diodes and diode substrate configurations. LO noise suppression and sideband ratio measurements are also described. Section IV presents computed mixer performance data, using a version of the program developed by [6], which looks at the effect of diode capacitance, anode finger inductance, diode pad-to-pad fringing capacitance, and mount impedance on the mixer performance.

II. Mixer Mount Configuration

A. Millimeter-Wave Mixer Mount Description

The mixer mount uses a traditional crossed-guide configuration [15] with the local oscillator waveguide perpendicular to the signal guide and electrically coupled with a shielded quartz microstrip line (Fig. 1). The microstrip contains the planar diodes and provides LO, signal and intermediate frequency (IF) isolation through low-pass hammerhead filters [7], [16]. The filters and coupling probes are fabricated lithographically using standard chrome gold liftoff techniques with subsequent buildup of electroplated gold to improve solderability. In an added step, the backside of the quartz is metallized with gaps left under the LO and signal waveguide coupling probes. This has the advantage of allowing the microstrip to be soldered,

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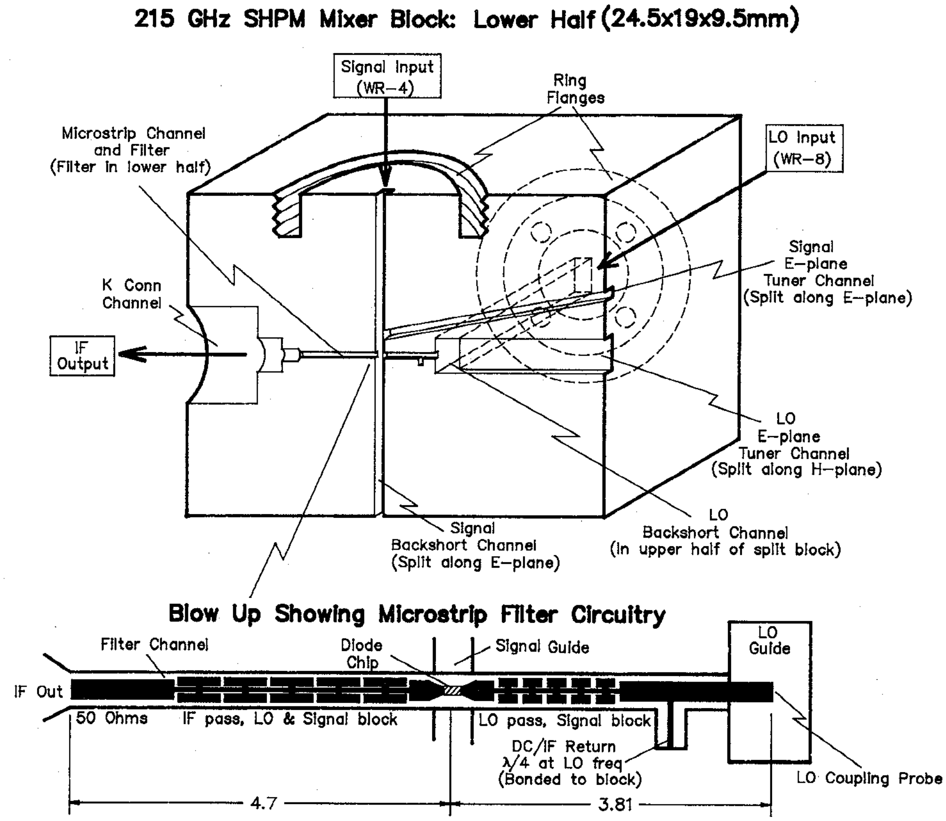

Fig. 1. Schematic of the 215-GHz subharmonically pumped crossed waveguide mixer block and microstrip filter circuitry. Only the lower half of the split block is shown.

rather than glued, into the machined channel, and reduces the effects of surface roughness. A dc/IF return for the diodes is provided by a separate shorting wire on the LO side of the microstrip filter. The LO waveguide is formed in one piece by wire electrodischarge machining (an electroformed insert or split block structure could also be used), and the remaining waveguides are produced with carbide slitting saws in the two halves of the mixer block. All waveguide is full height (2 : 1 aspect ratio), and the microstrip cavity is sized to prevent waveguide mode propagation in the LO and signal filter stop bands. The diode pair is suspended across the center of the broad wall of the signal waveguide with the plane of the filter metallization along the direction of propagation. Tapered probes couple both signal and LO power to the diodes. The LO pass hammerhead filter is designed both to present a short at the waveguide wall for the signal, and to maximize coupling of the LO into the diodes. Local oscillator power is coupled into the microstrip filter through a probe which protrudes from the center of the broad wall, extending halfway across the LO guide. Both the signal and LO guides have E-plane tuners and contacting backshorts behind the microstrip probes. The E-plane tuner for the LO is directly opposite the E-plane coupling probe. In the case of the signal guide, the E-plane tuner is approximately half of a guide wavelength in front of the diodes at the band edge of 200 GHz, and tapers out across the mixer block at an angle of 10° to avoid cutting across the face of the LO guide. IF removal is performed through a K-connector whose center pin is ribboned bonded to the end of the microstrip filter. The mixer mount characteristics were optimized on a 25× frequency scaled model using coaxial probes [18] and capacitively scaled antiparallel pair beam lead diodes [19].

The planar diodes used in the 215-GHz mixer are GaAs Schottky surface channel devices similar to those described in [1]–[3], [14]. All the measured diodes were formed as integrated antiparallel pairs. To facilitate assembly, the diodes are indium bump bonded, using spherical preforms, to pads formed on the ends of the signal coupling probes. Various diode package sizes and substrate configurations were tested (Table I). Since the diodes have a common contact pad on each end, separate dc biasing, to lower the LO power consumption, was not possible. The common dc return was used during RF measurements to slightly unbalance the diode pair in order to optimize the LO coupling by peaking the current in a single device.

B. Microwave Scale Model Measurements

Mount characterization and component optimization were facilitated by the use of an 8.5-GHz scale model which was configured so as to allow impedance probing using the well-known techniques of [18], direct signal and LO reflectometry measurements using the waveguide calibration procedures on an HP8510 network analyzer, or actual mixer conversion loss measurements using capacitively scaled antiparallel-pair beam-lead diodes [19]. The two low-pass hammerhead filters were each empirically designed and optimized in a separate

TABLE I
DIODES USED FOR THE MEASUREMENTS SHOWN IN FIG. 4

Designation diode	mixer	filter	Description	C_{j0} fF	ϕ_b V	η	R_s Ω	I_{sat} 10^{-16}	Anode Diam.	Finger lgth.	Bond Pad w × 1 (μm)
SC1T4/A4	200B4	1A	Small area all GaAs	3.0	1.09	1.28	11.9	2.5	1.4 μm	20 μm	30 × 50
			80w × 2001 × 50t (μm)	3.0	1.07	1.25	12.6	1.4			
SD1T1/A9	200B3	2A	Large area all GaAs	3.0	1.09	1.28	10.7	2.5	1.4 μm	50 μm	120 × 100
			120w × 2601 × 80t (μm)	3.0	1.08	1.25	12.0	1.2			
SR2T1/H4	200B4	2A	Thinned GaAs/Quartz	3.0	1.08	1.26	6.8	2.2	1.2 μm	50 μm	130 × 50
			130w × 2401 × 100t (μm)	3.0	1.06	1.22	7.7	1.3			
SR2T1/G5	200B4	2A	Large area thinned	3.0	1.09	1.25	8.1	1.5	1.2 μm	50 μm	130 × 50
			GaAs no substrate	3.0	1.07	1.22	7.3	0.9			

Notes: All diodes were fabricated at the University of Virginia, Semiconductor Device Laboratory (see [14] for details). Diode parameters (except C_{j0}) are derived from the measured dc I–V curves after mounting. Parameters for both diodes are given. Values of C_{j0} and anode diameter are the nominal for the processed diode wafer. Filters 2A and 1A differ in having diode bonding pads of 100 and 30 μm wide and gaps of 100 and 40 μm, respectively. Mixers 200B3 and 200B4 are nominally identical.

microstrip cavity before being combined in the full mount model.

The hammerhead filter response and conductor dimensions, as measured on the 8.5-GHz scale model, are given in Fig. 2. The sharp stopband peaks around 8 GHz on the IF pass filter [Fig. 2(c)] are real and are also predicted in a finite difference time domain computational analysis of this filter [20].

Several useful measurements were made on the mixer model by employing capacitively scaled antiparallel-air beam-lead diodes obtained form MA/COM [19]. The diodes have a nominal capacitance of 0.15 pF, and are soldered in position at the center of the signal waveguide on the scaled microstrip filter. No attempt was made to scale the diode-package parasitics. Calibrated LO and CW signal power were input to the mixer via coax-to-waveguide transitions. The single-sideband conversion loss into a 50 Ω IF line was measured using a spectrum analyzer over a wide range of IF frequencies, LO power levels, and RF tuning conditions. LO and signal port return losses, fixed-tuned bandwidth, and filter isolation were also measured. Tuner positions which gave the highest conversion efficiency were noted and later used to produce the impedance plots shown in Fig. 3 using the impedance probe technique of [18].

III. MILLIMETER-WAVE MIXER PERFORMANCE

A. Noise, Loss, and Required LO Power

One of the aims of this program is to show that planar diodes can perform as well as whisker-contacted honeycomb structures at 200 GHz and beyond. As the frequency of operation is increased, the diode package size becomes large compared to a wavelength and can have a significant effect on RF performance. As a first step in integrating the packaged planar diode [14] and surrounding mixer filter circuitry, we were anxious to measure the effect of the diode substrate on the mixer conversion loss. Using the etching techniques described in [2], it was possible to measure a variety of diodes in various stages of substrate integration: 1) an all GaAs package, mounted substrate-side-up across the signal coupling probes; 2) a similar sized diode which has had most of the semi-insulating GaAs substrate etched away and substituted with fuzed quartz, also mounted substrate-side up in the signal waveguide; and 3) a diode whose semi-insulating GaAs substrate has been etched away and not replaced by any carrier substrate. This third configuration most closely approximates a fully integrated diode/filter structure. Only a thin 5 μm layer of GaAs remains on the top side of the diode after etching.

The performance of the subharmonically pumped mixer from 195 to 230 GHz is shown in Fig. 4 for the diode packages listed in Table I. Two diode package sizes and three diode substrate configurations were tested. The larger packaged devices were measured with the usual GaAs substrate (UVa type SD1T1), with the substrate chemically thinned to 5 μm, and the device glued to a quartz carrier [2] (UVa type SR2T1), and with the thinned diodes soldered directly to the mixer filter circuitry (glue dissolved and quartz carrier lifted away [2]).The smaller packaged devices (UVa type SC1T4) were measured with the all GaAs substrate in two slightly different mixer mounts. All measurements were performed using a computer controlled noise test system similar to one described in [21].

Fig. 4 indicates a general improvement in mixer performance with decreasing package size and dielectric constant. The best performance obtained so far has been with the small area all GaAs diode package (T_m(SSB) $=$ 1600 K, L_{dB} $=$ 8.7 at 205 GHz), and the worst performance was obtained with the large area all-GaAs package. Although scale model measurements indicate that the larger package all-GaAs devices have a more limited accessible RF signal matching range in this particular mixer mount than the smaller package and lower substrate dielectric constant diodes, the difference in performance might as easily be due to varying diode electrical characteristics. A larger sample set is required before definite conclusions can be made. The required LO power for optimal mixer performance is similar for the smaller package and the thinned GaAs diodes. It is a factor of two larger for the large area all GaAs device and in the mixer block where the signal E-plane tuner has been shifted to within $<\lambda_g/2$ of the diodes (not shown in Fig. 4). The lower-than-usual LO power required for these unbiased diode pairs is attributed to the E-H tuner in the LO waveguide and careful optimization of the signal coupling probe shape and the first LO and IF pass filter section lengths. Table II is included to show comparisons with the better room temperature results from other groups

TABLE II
REPORTED ROOM TEMPERATURE WAVEGUIDE SCHOTTKY DIODE MIXER PERFORMANCE AROUND 200 GHz

Approx. Signal Frequency	Whisker Contacted Diode: Single Diode Fundamental Mixer		Whisker Contacted Diode: Single Diode Harmonic Mixer		Whisker Contacted Diode: Two Diode Subharmonic Mixer			Planar Diode: Two Diode Subharmonic Mixer		
F_{GHz}	T_m	L_{dB}	T_m	L_{dB}	T_m	L_{dB}	P_{LO}	T_m	L_{dB}	P_{LO}
180	750[a]	5.7	2600[e]	10.0	2400[f]	10.5	6.5	2750[h] $T_{Receiver}$		10
205	1250[b]	7.1	2400[e]	9.0	1800[g]	9.6	10	1590[i]	8.7	5.7
								1715[i]	8.7	4.0
								1990[i]	9.3	3.0
230	800[c]	6.2			2400[g]	10.9	10			
	800[d]	6.6								

All results are at room temperature and for an IF frequency between 1 and 2 GHz unless otherwise indicated. T_m = single sideband mixer noise temperature in K, L_{dB} = SSB conversion loss in dB, P_{LO} = required LO power in mW. The authors could not find data for planar diode fundamental and harmonic mixers at these frequencies in the literature.

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using whisker-contacted and planar diodes in fundamental, harmonic, and subharmonically pumped mixer configurations near the same frequency.

B. LO Noise Suppression

The subharmonically pumped antiparallel-pair mixer configuration has inherent LO noise cancellation due to the confinement of the LO noise sideband mixing products in the diode loop [5], [9]. A quantitative measurement of the LO noise suppression level in the 215-GHz mixer was made by using a very noisy local oscillator source, a 75–120 GHz backward wave tube (BWO), and a narrow-band 106-GHz GUNN diode oscillator in the following manner. Signal noise power was injected into the mixer's local oscillator port via a feed horn attached at the end of the through arm of a WR8 directional coupler. LO power at 106 GHz, supplied alternately by the BWO and GUNN, was input into the same mixer port using the −10 dB arm of the coupler. At the same time, dc bias was applied to the antiparallel diode pair so as to produce significant forward current flow in one diode only. Fundamental mixing at 106 GHz could then take place the (212-GHz signal waveguide is cutoff at 106 GHz) and a direct comparison of the mixer noise temperature with each LO source could be made as a function of output IF. To assure that the mixer performance was not being affected by LO power differences between the two sources, the same dc current was maintained in the forward biased mixer diode in each case, and no tuning adjustments were made on the mixer block between measurements. If the assumption is made that the GUNN oscillator adds an insignificant amount of LO noise to the measured mixer temperature, then the difference in performance between the GUNN and BWO can be attributed to the LO noise wings of the backward wave tube at the upper and lower sideband frequencies. Under this assumption, the equivalent noise temperature of the BWO at the signal sideband frequencies can be determined, and numbers in the range of 1.8×10^6–3×10^5 K were obtained over an IF range of 1.1–2.0 GHz, respectively. This leads to a noise power spectral density (2kTB) of 5×10^{-17}–9×10^{-18} W/Hz for the BWO.

The LO noise rejection of the subharmonically pumped mixer can now be inferred by noting the difference in measured mixer noise when the BWO and GUNN are alternately used as pump sources for the antiparallel-diode pair at 212 GHz. The observed change in mixer input noise temperature between the GUNN and the BWO fell between 20 and 60 K out of 2500 K over an IF of 1.2–1.8 GHz. The far ends of the IF band were too noisy to get an accurate difference measurement. By taking the ratio of the difference in measured noise temperature (BWO-GUNN) between the fundamental mixer and the subharmonically pumped mixer, an LO noise suppression of 33–36 dB was obtained over the 1.2–1.8 GHz IF band. At 1.5 GHz, the measured LO noise suppression ratio was 33.7 dB. If we use the approximate equations in [9] to calculate the expected LO noise suppression ratio for a 1.5 GHz IF and an LO frequency of 212 GHz, and assume a perfectly balanced diode pair, we get 34.5 dB!

C. Relative Sideband Ratio

The test set used to derive the data presented in Fig. 4 uses chopped broadband thermal loads and, as such, measures double sideband performance only. To determine the relative sideband response at a given IF frequency, a wire grid Fabry–Perot filter was placed between the mixer signal feedhorn and the loads. To reduce standing waves, no collimating lens was used, and therefore the beam divergence

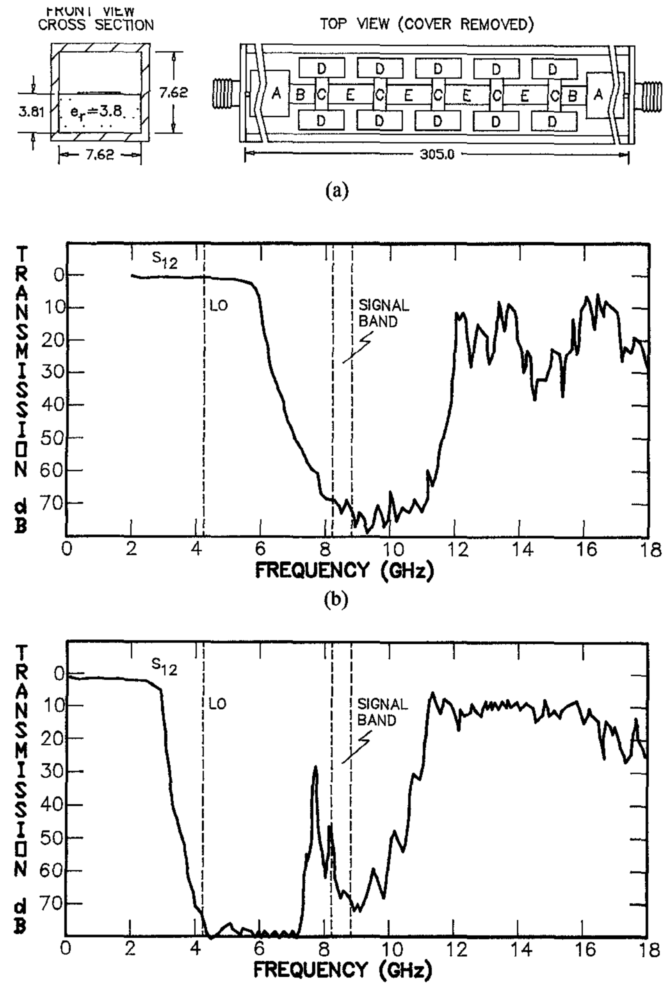

Fig. 2. (a) Hammerhead low-pass filters and microwave scale model test fixture for measuring individual filter frequency response (all dimension in mm). Dimension for the LO pass filter (l × w) are: A = 135 × 4.93, B = 3.43 × 0.76, C = 1.32 × 2.29, D = 4.83 × 2.08, E = 5.33 × 0.76 mm. Dimensions for the IF pass, LO and signal block filter are: A = 120 × 4.93, B = 6.35 × 0.76, C = 1.32 × 2.29, D = 11.43 × 2.08, E = 11.43 × 0.76 mm. The extra long 50 Ω line sections, A, on each side of the hammerhead sections were included to allow time gating out the effect of the coax to microstrip launchers on the input of the test box. (b). Transmission response magnitude of the LO pass, signal block filter as measured in the test box shown in (A). (c). Transmission response magnitude of the IF pass, LO, and RF block filter.

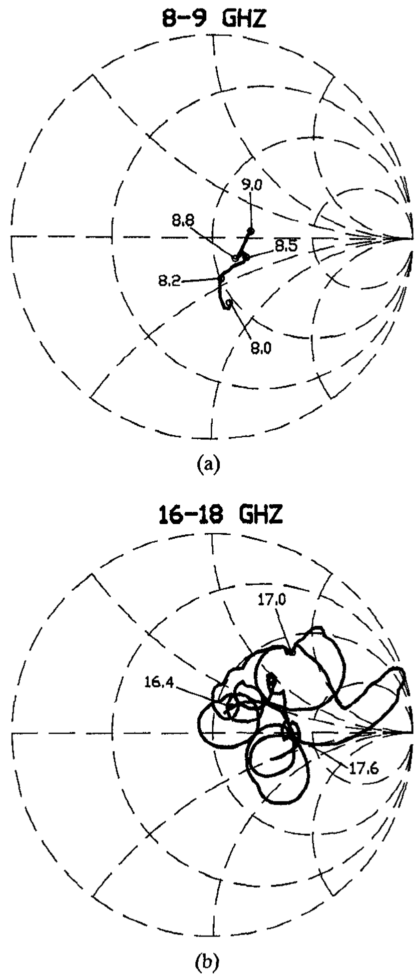

Fig. 3. Measured mixer model mount impedances under optimum tuning conditions at the signal and second harmonic sideband frequencies. No corrections have been made for coaxial line fringing fields or center pin inductance. A stycast block, simulating the GaAs planar diode package, was present for these measurements. Reducing the size of the diode chip most strongly affects the second harmonic impedances, generally increasing the real part.

through the filter resulted in substantial walkoff loss (≈ 3 dB). The Fabry–Perot was initially set to an order (14) which fully separated the two sidebands, and the grid spacing was stepped through several half-wavelengths while the mixer noise and conversion loss were recorded at each position. The resulting curve was then fitted to the expected response curve of the Fabry–Perot using the formulas in [22]. The grid reflectivity and sideband ratio were adjusted as free parameters, the former to partially account for the walkoff losses. A typical plot is shown in Fig. 5. This technique has been used quite successfully to calibrate the response of a similar heterodyne system now flying on the Upper Atmospheric Research Satellite Microwave Limb Sounder [23].

In general, it was found that the mixer blocks operated at or near equal sideband response, although it was possible to optimize one sideband, usually the lower, at the expense of the other. When tuned for optimum performance, as measured with the double sideband IF test set, the actual sideband ratio was generally less than 0.5 dB, making the extrapolated single sideband loss and noise shown in Fig. 4 and Table I accurate to about 11 percent. Some variation in sideband ratio with RF signal frequency was observed, with the response becoming more unequal as the frequency was increased. At the higher end of the RF band (above 220 GHz), the mixer back shorts had little effect on the overall sideband ratio, and true double sideband operation was not possible.

IV. Computed Mixer Performance

The mixer scale model measurements yield mount embedding impedances at the sideband and LO harmonic frequencies as a function of tuner position. These impedances can be used in a nonlinear mixer analysis program [6], [18], along with the planar diode current–voltage (I–V) and capacitance–voltage (C–V) characteristics to compute the performance of the

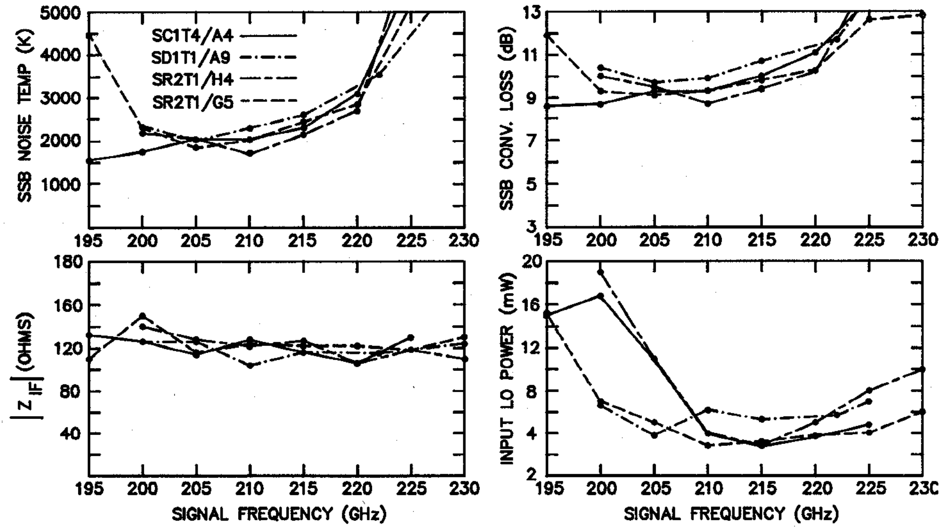

Fig. 4. Measured room temperature mixer noise, conversion loss, magnitude of the IF output impedance, and required LO power between 195 and 230 GHz for the five diodes whose characteristics are given in Table I. Results are given for an IF frequency of 1.4 GHz. The mixer block was optimized for lowest noise temperature at each frequency using the two backshorts and two E-plane tuners. The noise and conversion loss are measured between the input RF signal feed horn and the IF output K-connector. LO power is referenced to the WR-8 input flange. In all cases, SSB values were obtained by doubling the measured DSB results.

subharmonically pumped mixer. For the simulations which are presented here, the fundamental and second harmonic mount embedding impedances, at the tuning position which gave the lowest mixer conversion loss on the scale model (Fig. 3), were used in a version [18] of the mixer analysis program of [6]. Higher harmonics were short circuited outside the diode. The diode pair was assumed to have an I–V relationship which follows the standard thermionic emission equations. Parameter values (R_s, I_s, η, and ϕ) were obtained from the measured dc I–V curve, and are reproduced in the figure captions. The C–V relationship is derived from Poisson's equation and has the usual inverse square root dependence for a uniform doping profile. As in [6], the diode loop inductance is plotted as a variable for a given embedding impedance set and frequency. As a second variable, a parallel fringing capacitance has been added across the diode pair to represent pad-to-pad capacitance. Both of these parasitic elements were too small to be accounted for in the scale model measurements as the MA/COM diode pair was not physically scaled from its millimeter-wave counterpart.

The results of two sets of simulations are shown in Figs. 6 and 7. The differences between the two figures are the zero bias junction capacitance and diode series resistance which are, respectively, 3 fF and 12 Ω in Fig. 6, and 6 fF and 6 Ω in Fig. 7. The computed single sideband mixer noise, conversion loss, and the real part of the IF output impedance are shown as a function of series inductance (anode finger length) and pad-to-pad fringing capacitance. It is not clear where the diodes in Table I fall along the curves in Figs. 6 and 7, as we have not yet derived an accurate model of the anode finger inductance for these devices. The pad-to-pad capacitance has been measured approximately using wafer probing techniques and by scale modeling [24] in a nonwaveguide environment. Values for the diodes in Table I fall somewhere between 3 and 6 fF depending on anode finger length and substrate dielectric constant, but the precise value is not certain.

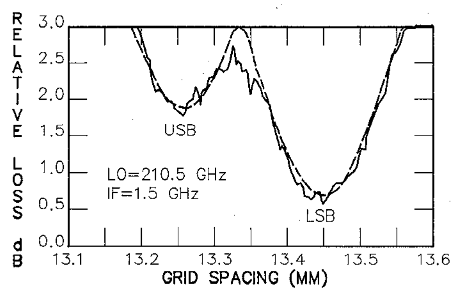

Fig. 5. Typical plot showing the measured (solid) and computed (dashed) mixer sideband response through a plane parallel Fabry–Perot filter over one grid order. The sideband ratio varies with LO frequency and tuning and can generally be adjusted to get equal response within approximately 1.5 dB at 1.5 GHz IF.

Looking at Figs. 6 and 7, it is interesting to compare the measured and computed performance of the 215-GHz mixers. There is qualitative agreement over a reasonable set of parasitic parameter space, but closer matching has not yet been possible. Clearly, the variation of computed mixer performance with diode parasitic parameters is considerable. Much better mixer performance than has been obtained seems possible for the subharmonic pair configuration if the diode anode size, the pad-to-pad capacitance, and the anode finger length are optimized in unison. Single-ended fundamental mixers do not seem to have as much variation in performance with these parasitics [18], and therefore may be much more forgiving when designed without optimization of specific junction parameters.

V. SUMMARY

The design and performance of a 215-GHz subharmonically pumped waveguide mixer using a planar integrated antipar-

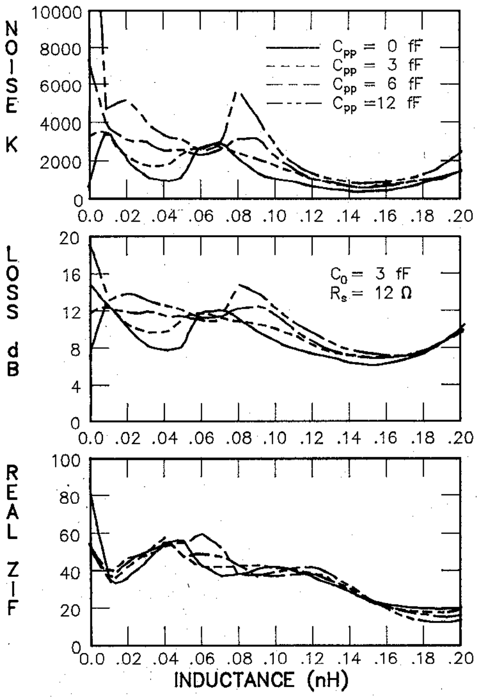

Fig. 6. Computed subharmonically pumped single sideband mixer noise, loss, and real part of the IF output impedance versus series (anode finger) inductance and parallel (pad-to-pad) capacitance. The LO and IF frequencies are 215 and 1.5 GHz, respectively. The mount impedances at the LO, upper, and lower sideband were set to $70 - j12\ \Omega$. The second harmonic sideband impedances were set to $65 + j80\ \Omega$. All higher harmonics were short circuited by the embedding network. Diode parameters for the computation are similar to those listed for the SC1T4 diode in Table I.: $\eta = 1.25$, $I_s = 2.5 \times 10^{-16}$, $\phi = 1.10$, $C_{j0} = 3$ fF, and $R_s = 12\ \Omega$. The LO power was adjusted to maintain a diode dc current of 1 mA and varied from 3.5 to 4.5 mW for the data shown.

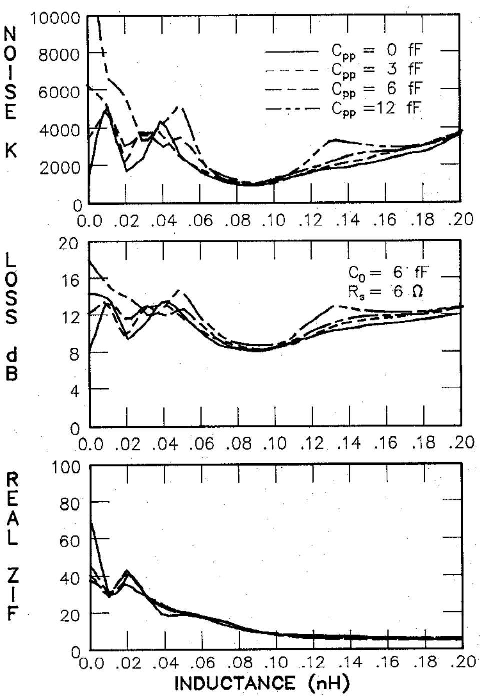

Fig. 7. Computed subharmonically pumped mixer performance versus series inductance and parallel capacitance under the same conditions as given in Fig. 6, but with $C_{j0} = 6$ fF and $R_s = 6\ \Omega$.

allel diode pair are presented. The mixer mount design is intended for use at frequencies up to 650 GHz, and utilizes full height waveguide in a split block configuration for ease of fabrication. The diode and mixer circuitry are formed in an integratable package to minimize assembly and handling. A fabrication procedure to integrate the GaAs diodes with the quartz microstrip structure is under development and will be used in subsequent receivers. The performance of the planar diode package exceeds that of the best whiskered pair at 200 GHz, in spite of the fact that not all circuit parameters have been optimized. In addition, considerably less LO power is required than that reported by other groups with similar mixer configurations, and excellent LO noise suppression is obtained with no external filtering.

Finally, computational analysis based on the program developed by [6] gives reasonable agreement with measured mixer performance when the expected inductive and capacitive parasitic elements are added to the mixer embedding network, and indicates considerable improvement is possible with careful control over these parameters.

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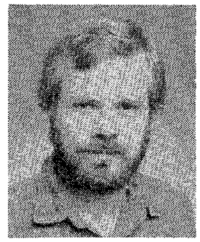

Robert J. Dengler was born in Whittier, CA, in July 1963. He received the Bachelor of Science degree in electrical and computer Engineering from Cal Poly Pomona University in March 1989.

He began his work with Dr. Siegel at the Jet Propulsion Laboratory as an intern in 1988, developing beam pattern acquisition and analysis software. He is now involved in the design and construction of submillimeter-wave receivers, components, and test instrumentation.

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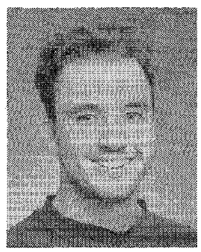
At present he is a Member of the Technical Staff at the Jet Propulsion Laboratory, California Institute of Technology, where his work is largely concerned with the development of millimeter- and submillimeter-wave sold-state devices for space-borne radiometers.



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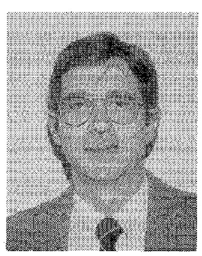
Since graduation he has worked in the Microwave Observational Systems Section of the Jet Propulsion Lab where his primary responsibility has been the numerical analysis of waveguide and microstrip circuits.



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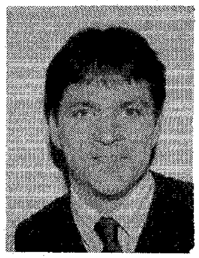
Mr. Bishop is a member of Sigma Xi and the International Union of Radio Scientists.



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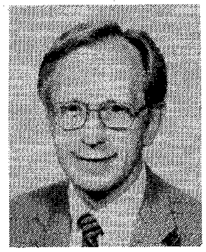

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