94-GHz Beam-Lead Balanced Mixer

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***Abstract*—Using a newly developed GaAs beam-lead diode, we have developed and evaluated a balanced mixer at 94 GHz. The various components of the mixer were separately optimized using carefully designed low-frequency model studies as our primary design aid. These studies included the determination of guide impedance and guide wavelength for suspended stripline, and optimization of a waveguide to suspended stripline transition, low-pass filters, and diode location. This 94-GHz mixer exhibits an average single sideband (SSB) conversion loss of 6.2 dB over a 6-GHz RF bandwidth. Together with a bipolar IF amplifier, the system exhibits a 4.5–5.1-dB double sideband (DSB) noise figure over a 50–700-MHz IF bandpass. LO-to-RF isolation was greater than 27 dB over this range of operating frequencies. Finally, severe environmental tests were successfully performed on this mixer between sucessive electrical characterization.**

I. Introduction

RECENTLY, beam-lead diodes with specifications suitable for short millimeter frequencies have been developed [1]–[3]. Prior to this, fabricating mixers for this portion of the spectrum required the use of whisker-contacted diode arrays. Although satisfactory performance has been obtained using this technology [4]–[8], there are several drawbacks to general high volume use.

First and foremost, the assembly of these mixers is a multistep labor-intensive process with a low success rate (~15 percent). Second, the diode contact, which is a mechanical one, often degrades with time. Third, these mixers generally do not withstand military standard vibration, shock and RF power tests unless special "ruggedized" contacts are used which substantially degrade the mixer noise figure. The mixer design reported here works generally as well as the best whisker-contacted mixers but shares none of the abovementioned drawbacks.

In addition to the primary design constraint of employing a beam-lead diode, there were several other highly desirable design constraints. A singly balanced mixer configuration was chosen for three reasons: 1) it reduces the additional noise contribution of typical millimeter oscillators; 2) a nearly constant 50-Ω IF output impedance results, providing an optimum match to available IF amplifiers; 3) significant LO-to-RF isolation is easily realized.

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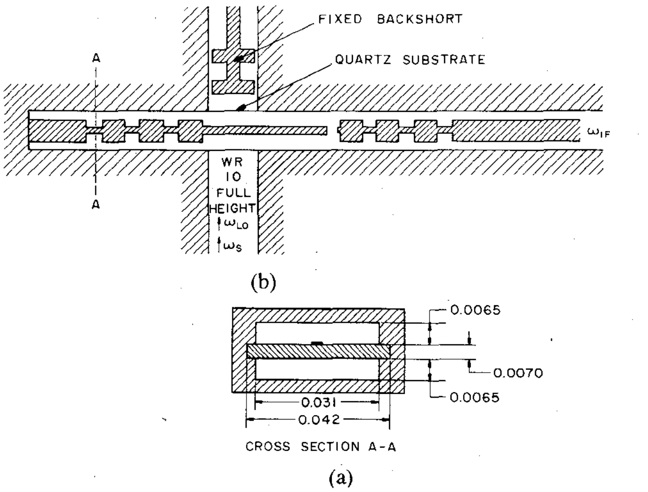


Fig. 1. (a) Cross section of suspended stripline channel for the WR-10 waveguide band. (b) Channel and waveguide together with backshort. Electric field for TE_{10} mode in waveguide lies in the plane of the figure. All dimensions are in inches.

In prototype versions [9], typical conversion loss was 6–6.5-dB single sideband (SSB) with an associated receiver noise temperature of 1520 K (SSB) including a 170 K IF contribution. This noise temperature was achieved over an instantaneous IF bandwidth of 900 MHz. In production versions, conversion loss was 5.5–6.0-dB (SSB) with an associated receiver temperature of 1060 K (SSB) [530 K double sideband (DSB)] or a noise figure of 6.7-dB (SSB) [4.5-dB (SSB)]. This includes the IF amplifier $T_{IF}=170$ K.

II. Design Criteria

Since a diode lies at the heart of every mixer, it is important to characterize it adequately before overall mixer design is to be contemplated. The Alpha diode is a multiple metal layer on GaAs Schottky beam-lead device with a 2-μm diameter anode. The total capacitance and junction capacitance at zero bias average 25 and 15 fF, respectively, and the dc series resistance averages 5 Ω. The anode and cathode beam leads are 125 by 12 μm gold. More information on this diode is contained in [2]. Standard thermocompression bonding techniques are used to attach the diode to thin-film circuits.

A suspended stripline geometry, shown in Fig. 1 was chosen because of the low-loss well-defined embedding network that could be realized. Using high-purity fused silica, impedance levels on the order of 100 Ω are easily

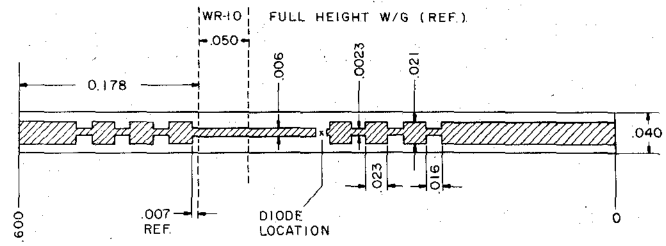


Fig. 2. Detailed dimensions of the quartz substrate metallization.

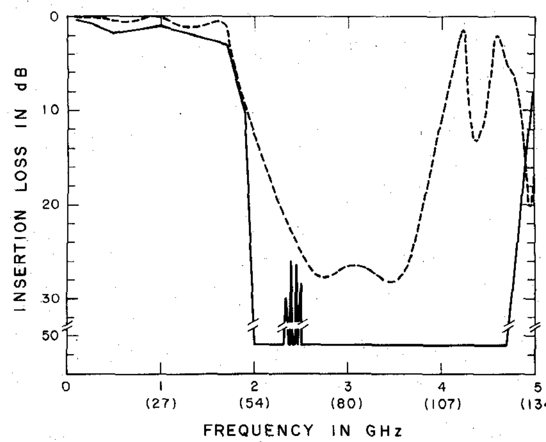


Fig. 3. Insertion $|S_{21}|^2$ for the cascaded five section (dashed curve) low-pass filter and semilumped filter (solid curve).

realized with an accompanying low effective dielectric constant. Higher order modes are not excited below 164 GHz and produce no evidence of playing a significant role in the operation of the mixer [9]–[11].

For the prototype mixer, a ring hybrid was used to realize the balanced configuration, while for the later, production mixers, a folded tee is used. Each diode could be individually dc biased and the IF outputs of each diode were directly combined before IF amplification.

The following design goals were set for the mixer-IF amplifier at 94 GHz:

1) the balanced mixer would exhibit a SSB conversion loss of 6 dB and a SSB receiver noise temperature of 1500 K;
2) an instantaneous IF bandwidth of 100–1000 MHz;
3) an RF bandwidth of ±3 GHz, preferably without mechanical retuning;
4) a local oscillator requirement of 5–10 mW;
5) mechanical and electrical ruggedness;
6) reproduction in system quantities at a low price.

III. Low-Frequency Model Studies

It was decided that a 2-4-GHz low-frequency scaled model would be built and, together with computer-aided design and empirical refinements, the various components of the mixer would be optimized. The diode, substrate, suspended stripline transmission line and waveguide were scaled by a factor of approximately $m=27$. In addition to the obvious requirements that 1) the physical dimensions scale, up by m, and the frequency, down by m, and 2) impedance levels remain constant, care was taken to scale the cutoff frequency of the diode and the diode and metal resistivities [10]–[12]. In these latter two cases, *exact* scaling was not achieved but an estimation of the error involved was made and taken into account.

Two distinct elements were designed separately: a waveguide to suspended stripline transition and a low-pass filter based on these model studies. Fig. 1(a) shows the final 94-GHz suspended stripline channel cross section. Fig. 1(b) shows the substrate passing through both broadwalls of standard (full-height) waveguide (the electric field lies in the plane of the figure). A short distance beyond the left broadwall, a low-pass filter provides an RF short circuit. An (initially) variable backshort tunes the transition to launch a nearly TEM wave onto the transmission line which continues to the right in the channel. To the left of the first low-pass filter, an IF and dc termination is made to the block with a gold ribbon. Fig. 2 shows the detailed dimensions of the substrate metallizations.

Fig. 3 shows the measured insertion loss $|S_{21}|^2$ for two scale model low-pass filter designs, referred to a 100-Ω system. In this and subsequent figures a dual frequency scale is used, representing the actual low-frequency scale model results and the corresponding implied high-frequency performance. One low-pass filter, similar to those in Fig. 2, has a total of 5 cascaded high and low impedance sections, each approximately a quarter wavelength long. The 1-dB cutoff frequency is 1.68 (45) GHz and the loss reaches 20

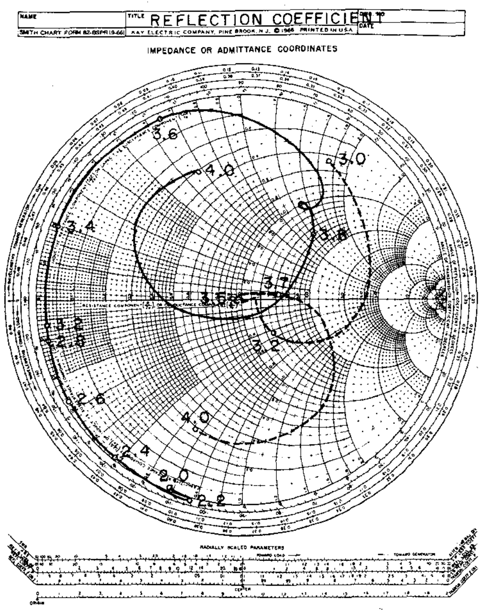


Fig. 4. The solid curve is the complex reflection coefficient S_{11} of the cascaded low-pass filter. The zero reactance frequency is 3.24 (87) GHz. The filter used in subsequent mixer tests had a zero reactance point at 3.5 (94) GHz. The dashed curve is S_{11} for the complete waveguide to stripline transition.

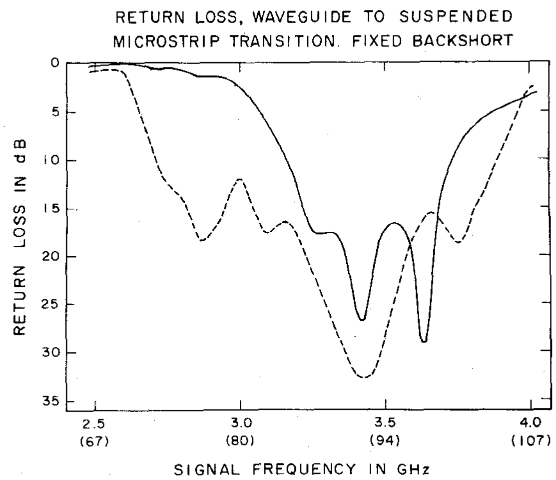


Fig. 5. Return loss for a waveguide to suspended stripline transition. The suspended stripline was terminated in its characteristic impedance. The full and dashed curves refer to cascaded five section and semilumped designs, respectively.

dB at 2.29 (62) GHz. Although this filter shows a passband in the neighborhood of the second harmonic, the insertion loss was greater than 6 dB up to 2.5 times the cutoff frequency. The second filter is a semilumped element design. The insertion loss in the 2–4-GHz region is vastly superior to the cascaded high–low impedance design. This semilumped design has yet to be incorporated into a high-frequency mixer. In Fig. 4, the complex reflection coefficient S_{11}, is shown for the first filter. S_{11} is measured with respect to the physical leading edge of the filter. Also shown in Fig. 4 is S_{11} for the complete waveguide to suspended stripline transition. In this case the 100-Ω line is terminated in the channel with a reflectionless load, and the backshort is optimized and fixed. The VSWR is less than 2:1 over a 0.57 (15.3)-GHz bandwidth. S-parameter data beyond 5 GHz is not shown due to the fact that excitation of modes other than the quasi-TEM mode in the suspended stripline channel complicate the interpretation of the data, beyond this frequency.

Fig. 5 shows the return loss for the same scale model transition. The two traces represent the two different low-pass filter designs, and again the bandwidth shown represents a fixed backshort position. Choosing the arbitrary value of 12 dB for the minimum return loss, the dashed

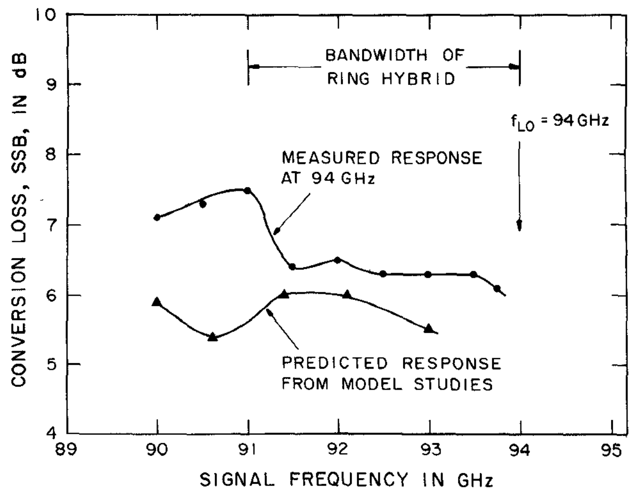


Fig. 6. SSB conversion loss data for both the scaled model and millimeter-wave mixers.

figure represents a 1.1 (30)-GHz fixed-tuned bandwidth.

The next step was to add a diode in series in the suspended stripline transmission line. A second low-pass filter is located beyond the diode followed by the IF port. The exact position of the leading edge of this second low-pass filter relative to the diode location was optimized with respect to the conversion loss in the low-frequency model. The diode chosen for this scale model is a high-barrier silicon Schottky (Alpha DMJ 6784), with the following parameters: $C_{j0} \leq 0.3$ pF, $R_S \leq 10\ \Omega$, $B_v(+10\ \mu\text{A}) \geq 5$ V. Fig. 6 shows the results of the conversion loss measurement on this single-diode low-frequency model. The conversion loss varies between 5.2 and 6.0 dB over a 0.11 (3.0)-GHz IF bandwidth. A small correction (+0.6 dB) has been added to this data to properly scale metal resistivities and diode cutoff frequency.

IV. Millimeter-Wave Mixer Performance

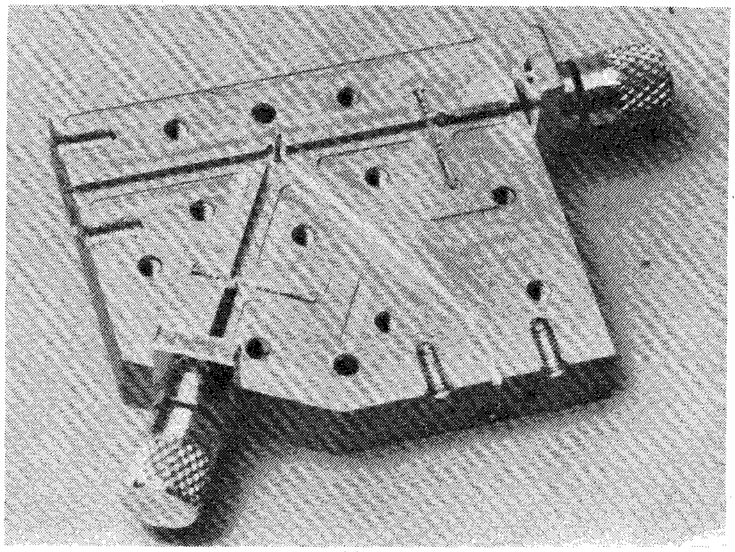
A prototype balanced mixer was fabricated in the WR-10 waveguide band, using a four-port ring hybrid. One half of the split block, including substrates and backshorts, is shown in Fig. 7(a). The hybrid operates as an in-phase/out-of-phase 3-dB power divider, covering a 6-percent bandwidth. A small section of coaxial line connects the output of each of the individual filters to a microstrip circuit on the backside of the split block. There the circuits were combined using dc blocks so that the diodes could be individually biased. No attempt was made to do any IF matching. The production mixer (Alpha/TRG-W9600), shown in Fig. 7(b), employs a folded-tee hybrid which results in lower RF and IF losses.

The substrates were fabricated from fused silica (99.9+ percent), 0.007 in thick, polished to a 1-mil (rms) surface finish. The metallization was all-sputtered Cr/Au, of 0.025 and 2.5-μm thickness, respectively.

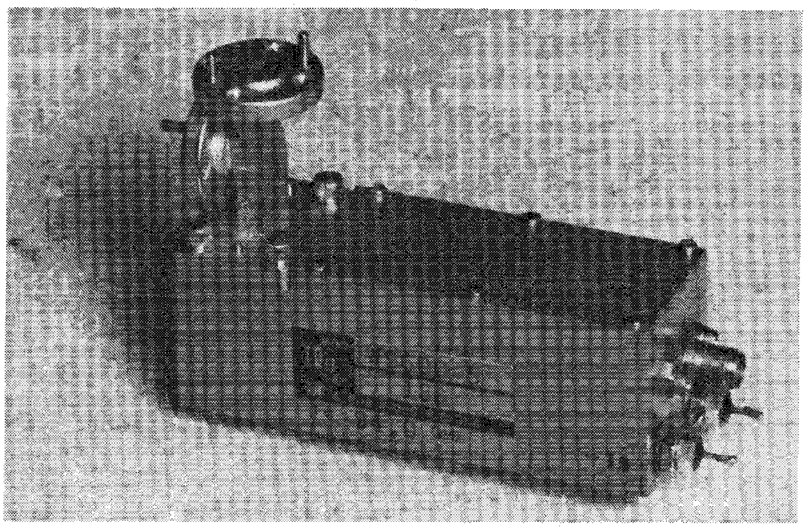
Two types of RF tests were made: SSB conversion loss measurements and DSB noise figure measurements. SSB conversion loss measurements were made using the prototype mixer and the experimental setup shown in Fig. 8(a). Local oscillator frequency, power, and backshort position were optimized for minimum conversion loss. The signal frequency was then varied and the triple stub tuner was readjusted for each signal frequency evaluated. The results are plotted in Fig. 6 in conjunction with the scaled model results. Typical errors for any single data point are ± 0.25 dB. For signal frequencies close to the local oscillator the agreement between the scale model and the millimeter-wave mixers is quite good considering that the extra losses of a balanced millimeter-wave design were not corrected for. At IF frequencies greater than 2.5 GHz the finite bandwidth of the hybrid degrades the balanced mixer design. By readjusting the backshort for each signal frequency a fairly uniform 6.2-dB conversion loss was obtained to the limits of the hybrid bandwidth.

DSB noise figure measurements were made using the production mixer, with fixed backshorts, and the setup shown in Fig. 8(b). It is noteworthy that a free-running Gunn oscillator without isolation or filtering was used for these measurements. The signal in this case was a piece of lossy dielectric either at room temperature or liquid nitrogen temperature.[1] Y-factor measurements were taken (see Appendix). Since the IF system had also been calibrated, conversion loss and mixer noise temperature were also determined. The results of the receiver noise temperature are shown in Fig. 9. The DSB noise figure varies from 4.5 to 6.5 dB depending on IF frequency. At 50-MHz excess noise from the local oscillator and increased IF amplifier noise degrade the mixer performance, while at higher frequencies the IF mismatch, and increased IF amplifier noise contribute to degraded performance. The SSB mixer noise temperature and SSB conversion loss inferred from these measurements range from a best of 620 K and 5.6 dB to 660 K and 6.3 dB near the band edges. No measurements were made at other local oscillator frequencies although

[1] Eccorsorb CV, Emerson and Cuming, Inc., Canton, MA 02021.

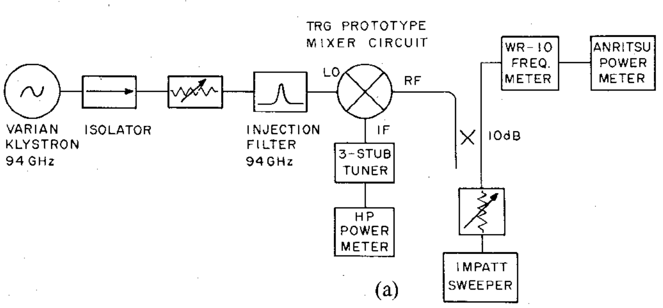


(a)

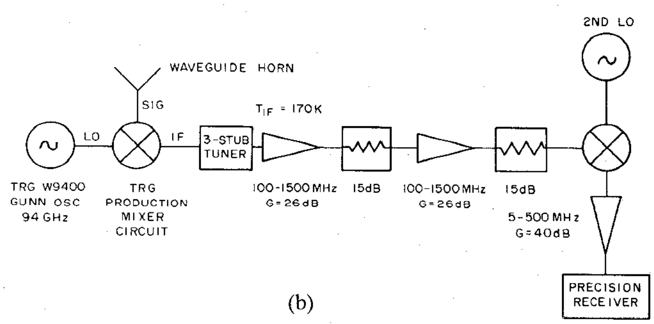


(b)

Fig. 7. (a) Prototype balanced mixer with top half removed, including backshorts and substrates. (b) Production mixer showing folded-tee hybrid and IF amplifier housing.



(a)



(b)

Fig. 8. (a) Experimental arrangement for millimeter-wave SSB conversion loss measurements. (b) Experimental arrangement for DSB noise figure measurements.

this folded-tee hybrid should have an RF bandwidth of about ± 4 GHz for a WR-10 system.

The LO requirements varied from +7 to +9 dBm for unbiased diodes and +3 dBm with dc bias. The optimum value depends whether minimum mixer noise or minimum conversion loss is desired. Typical RF-to-LO isolation was 27 dB at 94 GHz.

Table I summarizes the performance of the production

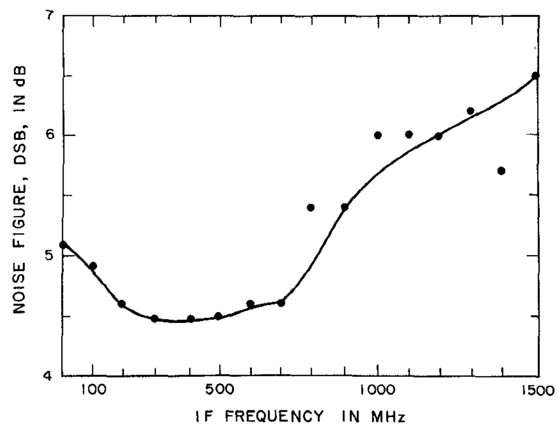


Fig. 9. DSB noise figure data for millimeter-wave production mixer.

TABLE I
RECEIVER PERFORMANCE IN THE 70–120-GHz FREQUENCY RANGE

Reference	T_{SIG}(GHz)	T_M(K)	L_c(dB)	T_{IF}(K)	T_R(K)	Reference	Comments
1. This Design	94	620±50	5.6	170	1240	This Paper	1,2
2. NRAO (1973)	85	800	6.8	—	—	[13]	5
3. NRAO (1973)	85	1030	5.3	—	—	[13]	5
4. NRAO (1975)	85	420	4.6	—	—	[14]	5
5. NRAO (1975)	115	500	5.5	—	—	[14]	5
6. BTL (1977)	72	700	6.2	—	—	[17]	
7. BTL (1978)	94	710±60	6.5	108	1199	[17]	4
8. BTL (1978)	98	390±250	7.4	330	2405	[17]	3,4
9. BTL (1978)	101	720±60	6.5	—	—	[17]	4
10. BTL (1977)	117	840	6.5	—	—	[17]	
11. MPIfR (1977)	107	602	6.2	250	2425	[16]	5
12. MPIfR (1979)	115	660	6.3	—	—	[7]	5
13. MPIfR (1980)	115	740	6.7	—	—	[8]	5
14. AEROSPACE (1977)	90	600	5.4	49	1000	[18]	5
15. AEROSPACE (1977)	115	700	6.2	49	1200	[18]	5
16. AIL (1979)	94	—	6.0	—	—	[3]	
17. MEUDON (1980)	90	850	7.5	—	—	[8]	
18. COLUMBIA-GISS (1979)	115	440	5.3	53	860	[15]	5

1. Beam lead diode
2. Two diode, balanced mixer
3. Image rejection tuning
4. Two-diode subharmonically pumped mixer
5. Corrections to T_M and L_c may include RF and IF losses

mixer and compares it with published data on other mixers in the same frequency range. Several factors complicate an accurate comparison of one receiver with another. First, some mixers are two-diode balanced mixers; others are two-diode subharmonically pumped mixers and still others are single-diode mixers. Second, some authors subtract the deleterious effects of RF losses and IF mismatch, when computing the mixer noise temperature. This is especially true in the case of single-diode mixers which usually contain LO filters and/or diplexers before the mixer. Finally, IF amplifier noise temperatures vary considerably. Nonthеless, a reasonably accurate comparison can be made if a full knowledge of the correction procedure is published.

For most of the entries in Table I, T_M varies between 600 and 850 K. Notable exceptions with lower values are entires 4, 5, 8, and 18. Entries 4 and 5 contain significant corrections for RF losses, IF mismatch, and thermal reradiation. Details of such subtractions are not given and a valid comparison with uncorrected data is not possible. Entry 8 represents a mixer operating in a single sideband mode and additionally T_M possesses a high degree of uncertainty. Entry 18 contains significant subtractions as well but is well documented enough to reconstruct the unmodified T_M for comparison. If input RF and IF transformer losses are taken into account we calculate $T_M = 613$ K and $L = 5.8$ dB. These two numbers can be directly compared with our results, and are seen to be identical within experimental uncertainty. The conclusion to be drawn from this comparison is that this design exhibits an uncorrected T_M and L_C as good or better than any design reported in the literature to date.

V. ENVIRONMENTAL TESTS

Table II summarizes a series of environmental and reliability tests which were performed on these mixers. Among

TABLE II
SUMMARY OF ENVIRONMENTAL AND RELIABILITY TESTS PERFORMED ON THE MILLIMETER-WAVE MIXERS

ENVIRONMENTAL TESTS

TEST	STANDARD
VIBRATION	– MIL-E-5400R FIGURE 2 CURVE Ia 3 AXES
SHOCK	– MIL-E-5400R 3 AXES 400g
TEMPERATURE	– CYCLING FROM 77K TO 400K TEN TIMES
	– SHOCK COOLING IN LN_2
CW RF POWER	– 150mW CONTINUOUS RF POWER 168 HOURS CONTINUOUS
PULSED RF POWER	– 0 5 W, 50 nS PULSE AT 40 kHz REP RATE, 168 HOURS CONTINUOUS

the tests are temperature cycling, vibration, and power standards. RF characterization was performed both before and after these tests and the mixers exhibited no measureable degradation.

VI. SUMMARY AND CONCLUSION

A balanced mixer using beam-lead diodes and suspended stripline technology has been developed for 94-GHz applications. Exhibiting SSB conversion losses of 5.6–6.3 dB and SSB mixer noise temperatures of 620–660 K, these mixers demonstrate a high degree of electrical and mechanical ruggedness. Finally, they are designed for mass production, avoiding the costly and labor intensive procedures of previous, whisker-contacted, designs.

APPENDIX

Y-factor measurements [15], [17] were made on both the production mixer and the IF amplifier, in order to determine the noise temperature of the mixer and IF amplifier and the conversion loss of the mixer. Matched loads at $T_H = 299$ K and $T_C = 77$ K were used in both cases. The Y-factor measurements yield a DSB receiver noise temperature. If the SSB receiver temperature is desired, it is given by

$$T_R(\mathrm{SSB}) = (1 + L_s/L_i)T_R(\mathrm{DSB})$$

where L_s and L_i are the signal and image conversion losses, respectively. The DSB and SSB receiver noise figures F_R(DSB) and F_R(SSB), are given by

$$F_R(\mathrm{DSB}) = 1 + \frac{T_R(\mathrm{DSB})}{290}$$

$$F_R(\mathrm{SSB}) = 1 + \frac{T_R(\mathrm{SSB})}{290}.$$

These last two quantities can also be written in decibels, $NF_R = 10\log_{10} F_R$. In all measurements we found that $L_s = L_i$ to within our measurement uncertainty.

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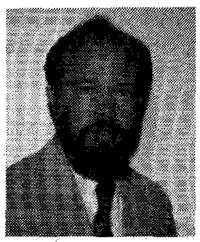

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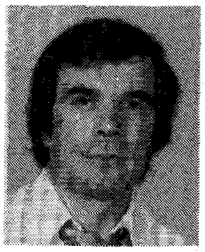

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From 1971 to 1976 he was with the Israeli Ministry of Defence, developing microwave passive and active components. During six months in 1976, he was a Research Fellow at University of Eindhoven, Eindhoven, The Netherlands, working on BARITT diodes. As a Research Assistant at the University of Massachusetts, Amherst, from 1977 to 1978, he developed 94-GHz mixers. From 1978 to 1980 he was employed at F.E.L., Farmingdale, NJ, designing microwave components. Since 1980 he has been with Aerojet ElectroSystems Company, Azusa, CA, developing millimeter-wave front-end receiver components up to 220 GHz. His main interest resides in developing very low noise, spacequalified millimeter-wave mixers for remote sensing sounders.