for GaAs SAW applications. If temperature compensation is required, it must be accomplished by other means such as thin-film overlays [8] or digital compensation [9].

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Planar Millimeter-Wave Diode Mixer

N. J. CRONIN AND V. J. LAW

***Abstract*— A new mixer has been built, using a planar GaAs Schottky-barrier diode, for operation at frequencies around 100 GHz. The mixer has low noise temperature and conversion loss and low local oscillator power requirement. The design is such that construction of scaled versions should be possible for operation up to 200 GHz.**

I. Introduction

For frequencies above 100 GHz, most high-performance mixers still utilize whisker-contacted Schottky-barrier diodes [1]. Satellite borne applications in the short millimeter region are of increasing importance, and there is thus a growing demand for systems capable of withstanding the rigors of space flight. Whisker-contacted diodes have been used in space [2] but more rugged and reliable mixers are being sought. Excellent beam-lead diodes have been developed in a number of laboratories around the world [3], [4] but parasitics associated with the contact leads have limited their use to around and below 100 GHz. There is a need, therefore, to develop new mixers which combine the ruggedness of the beam-lead structures with the high-frequency capability of whiskered devices [5].

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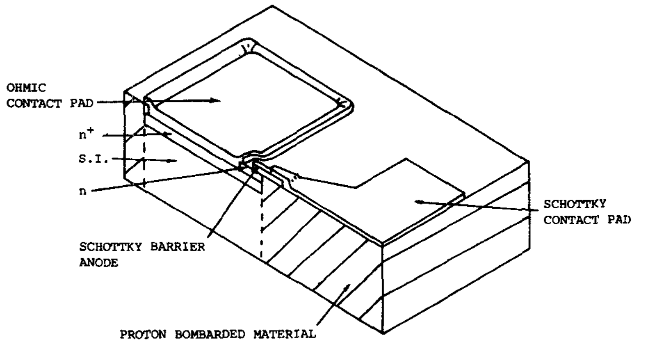

Fig. 1. Planar GaAs Schottky-barrier diode.

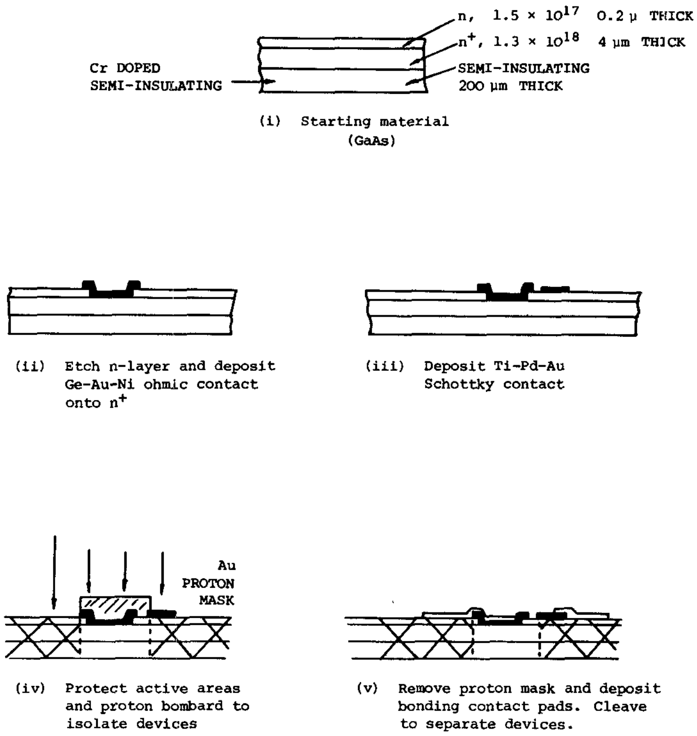

Fig. 2. Schematic representation of the planar diode process.

We report a new mixer, operating in the band 90 to 110 GHz. The design utilizes a custom-built, planar, GaAs Schottky-barrier diode soldered directly into a suspended-substrate stripline circuit without the use of bonding leads. This configuration exhibits low parasitics and should not be subject to the frequency limitations of conventional beam-lead designs.

II. The Diode

Fig. 1 is a sketch of the overall configuration of the planar diodes used. The devices were fabricated by a process which is summarized in Fig. 2. Two different types of diode have been assessed. Type-A were fabricated by the authors at the Plessey, Allen Clark Research Centre, Towcester, England; type-B were kindly loaned to us by Dr. B. J. Clifton of M.I.T. Lincoln Laboratories, MA. Fig. 3 shows the dimensions of the type-A

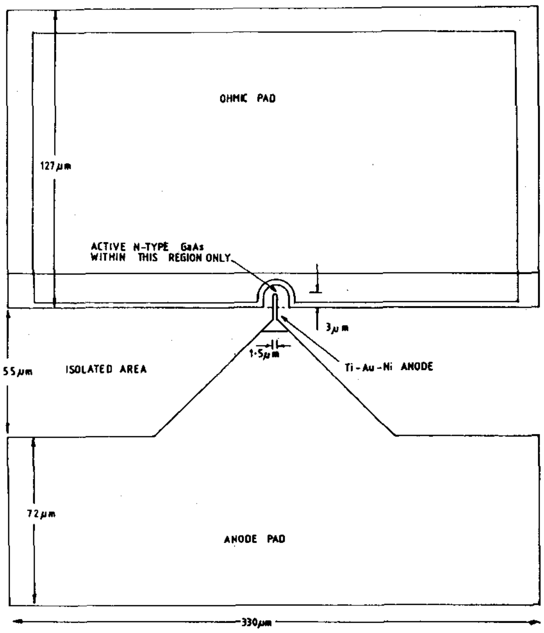

Fig. 3. Type-A diode.

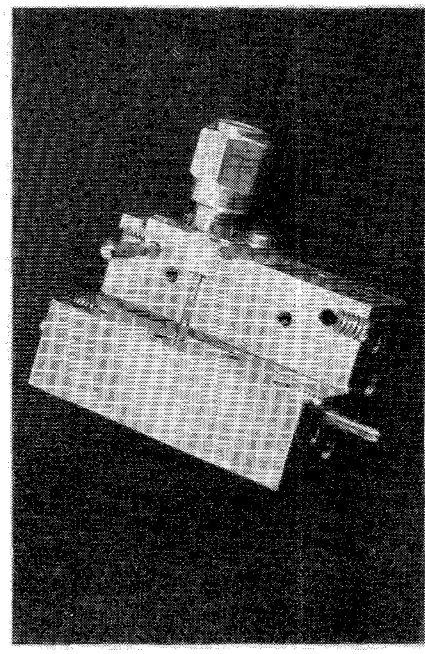

Fig. 4. Partially disassembled mixer block.

devices. The characteristics of the diodies are as follows:

	Type-A	Type-B
Series resistance R_s	$<10\Omega$	$<10\Omega$
Capacitance (Total, zero bias)	20–30 ff	20ff
Ideality	≈ 1.5	<1.1
Stray capacitance	<5 ff	<5ff

The diodes are designed as "flip chips," that is, in operation they are solder bonded, face down, into a circuit. The bonding pads have been designed to match the circuit into which the diodes are soldered in order to minimize their effect upon the operation of the mixer.

III. Mixer Design and Assembly

The mixer developed is of the split-block-type using suspended-substrate stripline and reduced-height waveguide. Fig. 4 is a photograph of a partially disassembled mixer, Fig. 5 is a sketch of the overall layout of the design, and Fig. 6 is a detailed drawing of the suspended-substrate circuit.

The signal enters in WG 28 (WR 8), which tapers to half-height. Power is coupled to the 50-Ω line by a simple capacitive probe coupler [6]. The optimum length of the probe was found to be half of the height of the waveguide (i.e., one-quarter full height). The diode is shunt-mounted to ground in front of a low-pass IF filter [7] as shown in the figure. The filter is based upon a design by Lidholm [8], and is scaled to produce a cutoff frequency of around 80 GHz. The IF output and dc bias are through a 3-mm SMA connector.

A full analysis of the mixer circuit awaits the outcome of scale-model measurements currently under way. However, from our millimeter-wave measurements, an approxiamte semi-empirical formula may be given which enables a mixer to be optimized for any required frequency in the range 90 GHz to 120 GHz. The critical dimension is that marked as X in Fig. 6. To optimize the mixer for operation at a frequency of ν GHz, the required value is given by

$$X = -0.032\nu + 4.35 \text{ millimeter.}$$

The design of the mixer is such that assembly is very straightforward and requires no sophisticated bonding or whiskering

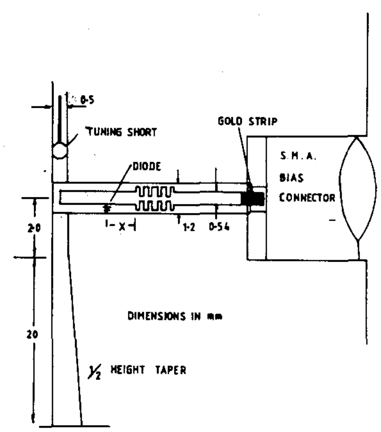

Fig. 5. Layout of the mixer.

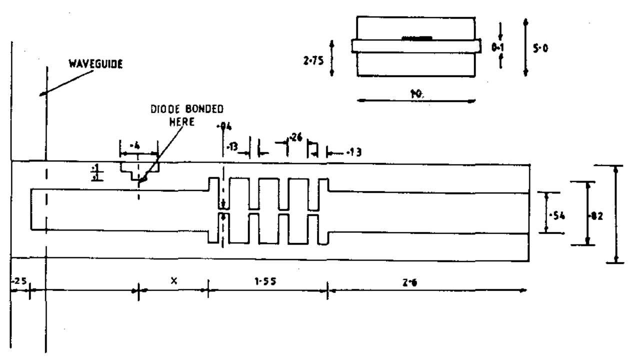

Fig. 6. The stripline filter circuit.

equipment. Simple soldering using low melting point indium solder is used both for diode bonding and coupling of the stripline to the SMA output connector.

IV. Mixer Performance

The double-sideband conversion loss L and noise temperature $T_{MXR'}$ of the mixer have been measured. The method adopted was basically that of Weinreb and Kerr [9] in which a radiometer

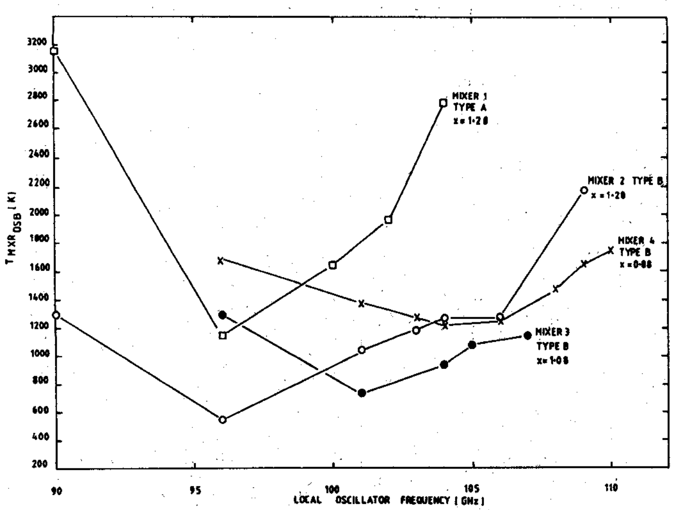

Fig. 7. Measured double sideband noise temperature against local oscillator frequency for four different mixers.

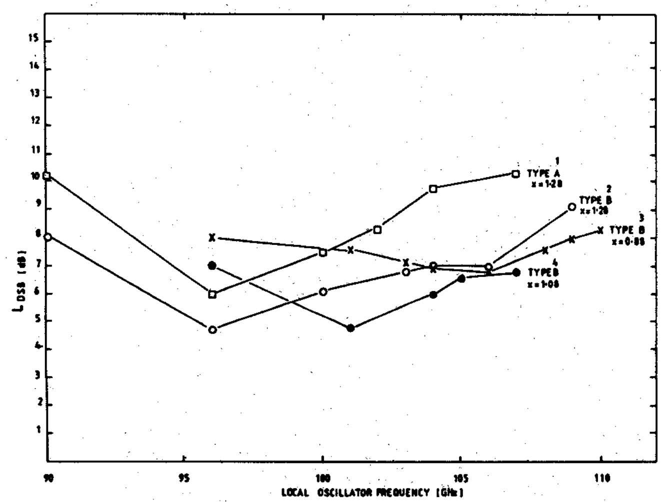

Fig. 8. Double-sideband conversion loss against local oscillator frequency.

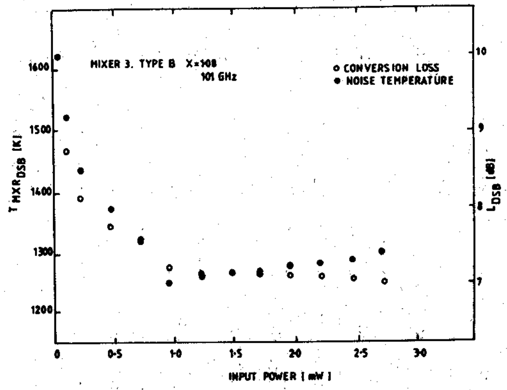

Fig. 9. Mixer noise temperature and conversion loss as a function of local oscillator power level.

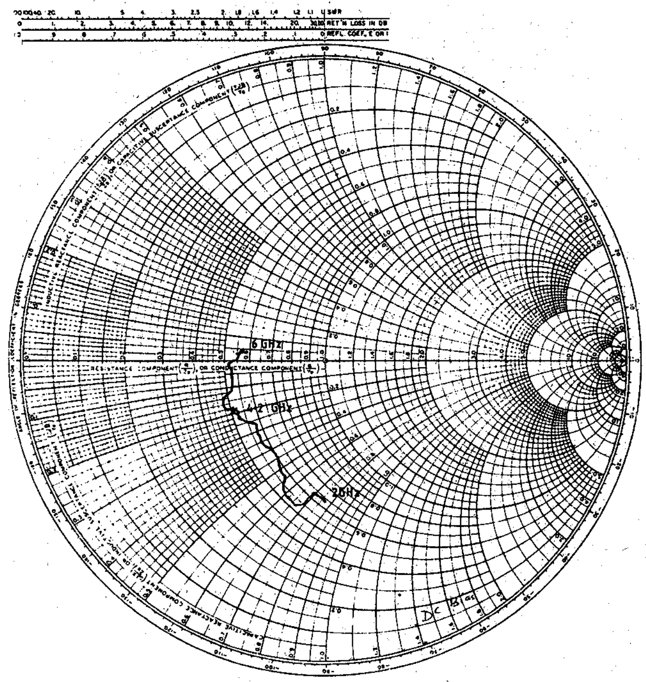

Fig. 10. Mixer IF impedance.

incorporating the device under test is used to observe microwave absorber at ambient and liquid nitrogen temperatures. Measurements were made with the local oscillator frequency varying in the range 90 to 110 GHz. The number of spot frequencies which could be measured was limited by the availability of suitable sources. Figs. 7 and 8 show the results obtained with four mixers, one of type-A and three of type-B. Conversion loss is quoted with respect to the input waveguide flange and the IF output connector. No correction has been made for IF mismatch. The IF frequency used was 3.9 GHz with an instantaneous IF bandwidth of 80 MHz.

Fig. 9 shows the conversion loss of mixer 3 as a function of LO power. As can be seen from the figure, less than 1 mW of LO drive is required for optimum mixer noise temperature.

Fig. 10 shows the mixer IF impedance as a function of IF frequency. Here, the mixer was LO pumped at 95 GHz to a dc current of 3 mA with a dc bias voltage of 0.55 V.

V. Discussion and Conclusions

The measured performance of our new mixer is comparable to that of the current state-of-the-art beam-lead and planar diode mixers. For comparison, Table I shows our results together with those of a recent commercially available mixer from Mullard.

TABLE I
Comparison of Bath Type-A and Mulland 760CL5A Performance

Mixer	Bath Type A	Mullard 760CL5A
Frequency Range	90 - 110 GHz	75 - 110 GHz
Noise Figure	8.7 dB*	7.5 dB**
L.O. Power	1 mW	10 mW

* Measured at 94 GHz including 1 dB i.f. contribution and 3 dB D.S.B. → S.S.B. allowance

** Measured at 85 GHz including 1 dB i.f. contribution

As may be seen from the table, our noise figure is marginally worse than the commercial component; however, we have allowed 3 dB to convert our double-sideband measurements to single-sideband for comparison. This may well be too large an increase since we are thereby assuming the mixer to respond equally in both sidebands, which is unlikely given the resonant nature of the diode–stripline–IF filter combination.

The local oscillator power requirement of our mixer is seen to be relatively low; this becomes important if the design is to be used at higher frequencies [10]. Fig. 10 shows that there is some residual IF mismatch at 3.9 GHz which could be removed by an appropriate IF impedance matching transformer leading to a further small improvement in performance.

The most significant advantage of the new design lies in the simplicity of the structure. Given the availability of diodes, a mixer with good performance can be assembled without sophisticated bonding or whiskering equipment. Furthermore, the reduction of all linear dimensions (including those of the Schottky barrier) by up to a factor of two would appear to present no difficulty either in diode processing or mixer circuit construction and assembly. We, therefore, anticipate that the same basic design can be used for operation up to 200 GHz.

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They are especially indebted to B. J. Clifton of M.I.T. Lincoln Laboratories for his advice and support throughout and for the provision of devices used in the program, and also to Prof. D. H. Martin, Queen Mary College, London, for his encouragement of their work.

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Scattering at an *N*-Furcated Parallel-Plate Waveguide Junction

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***Abstract*—Using the conservation of complex power technique (CCPT), this paper presents a solution to the problem of EM scattering at the junction of a parallel-plate waveguide and an N-furcated parallel-plate waveguide with arbitrarily spaced thick septa. Although this junction can be regarded as an $(N+1)$-port configuration, the problem is formulated so that it is viewed mathematically as a generalized 2-port. This leads to very simple expressions for the scattering parameters of the junction. Convergent numerical results are presented for bifurcated, trifurcated, and 4-furcated structures, and the effects of varying the thickness of the septa are investigated. The formulation is directly applicable to N-furcated rectangular waveguide junctions having TE_{n0} excitation, with application in the design of E-plane filters.**

I. Introduction

Electromagnetic scattering at the junction of a parallel-plate waveguide and a bifurcated parallel-plate waveguide with a septum of vanishing thickness has been studied by Mittra and Lee [1], who provided analytical solutions using the residue calculus method and the Wiener–Hopf technique. Moreover, a quasi-static solution using the singular integral equation method has been given by Lewin [2] for the case of a centrally located infinitely thin septum.

Trifurcated waveguide junctions were treated by Pace and Mittra [3], who considered the structure to be two bifurcated junctions in tandem; the overall solution was deduced with the help of the generalized scattering matrix technique [1].

The N-furcated junction has also been considered, in early papers, by Heins [4] and Igarashi [5]; however, their methods apply only to equally spaced thin septa.

In regard to bifurcated guides with thick septa, one may use the generalized scattering matrix technique, considering the junction as a bifurcated junction with a thin septum followed by a step discontinuity [6]. However, it would be very laborious to apply this technique repeatedly for the problem of an N-furcated waveguide junction with $N-1$ arbitrarily spaced thick septa.

In some recent papers [7]–[9], the conservation of the complex power technique (CCPT) has been used to obtain theoretically exact solutions with numerically convergent results to the problem of scattering at certain waveguide junctions. In this paper, the CCPT is applied to the specific case of the junction of a parallel-plate waveguide and an N-furcated parallel-plate waveguide, as shown in Fig. 1. The thicknesses $t_1, t_2, \cdots, t_{N-1}$ of the $N-1$ septa are not necessarily equal, nor are the separations between plates $L_1, L_2, L_3, \cdots, L_N$ of the N waveguides; the sole constraint is that $t_1 + t_2 + \cdots + t_{N-1} + L_1 + L_2 + \cdots + L_N = L$, where L is the separation between plates of the guide which forms the junction at $z = 0$ with the N-furcated guide. Note also that the dielectric constant ϵ_{rn} in each waveguide is arbitrary.

Although in this contribution we only consider the problem of N-furcated parallel-plate waveguide junctions for TE_n and TM_n excitation, the formulation is also directly applicable to the

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