A 30-GHz Monolithic Single Balanced Mixer with Integrated Dipole Receiving Element

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Abstract—This paper will describe a 30-GHz monolithic low-noise balanced mixer which has been developed using an integrated bow-tie antenna to waveguide transition and low parasitic Mott diodes. The diodes and mixer circuit were developed using MBE material and were fabricated using a plated airbridge technology. Measurements on the diode at dc and RF showed that the zero bias junction capacitance was 0.025 pF and the series resistance was 10 Ω. A mixer conversion loss of 6 dB was measured at 30 GHz with an IF of 1 GHz.

I. Introduction

THIS PAPER will describe the design and performance of a monolithic 30-GHz mixer and antenna to waveguide transition on GaAs. The mixer was intended for use in EHF phased-array modules for space applications, and a variety of designs were considered bearing in mind the requirement for high performance and low cost. Although the original requirement was for a communication link at 30 GHz, it was decided at the outset to select an approach that could be easily scaled to higher frequencies, such as 44, 60, and 94 GHz, without any significant compromise in RF performance. In recent years, there have been a number of publications describing monolithic single balanced mixers [1]–[3]. Previous experience at higher frequencies had shown that it was difficult to make a satisfactory transition to an MMIC with repeatable performance both for microstrip and coplanar transmission lines. It was, therefore, decided to investigate an approach where the transition was incorporated onto the chip in the form of a receiving element or antenna which could act as a free-space receiving element or as part of a transition to a rectangular waveguide. A second consideration was that the transmission lines be readily compatible with the coplanar realization of the diodes. Studies were made of different forms of transmission lines including microstrip, coplanar waveguide, and coplanar strips, and a comparison was made between them in terms of impedance range, normalized guide wavelength, and conductor and dielectric losses. Fig. 1 is a comparison between microstrip and coplanar transmission lines and shows conductor loss at 30 GHz as a function of characteristic impedance for a 4-mil-thick GaAs substrate [4]. The impedance of microstrip line is a function of the width W to height H ratio, and since the height of the substrate is fixed, the loss versus impedance characteristic is represented by one curve. However, for a fixed substrate thickness, the impedance of coplanar line is a function of width W to spacing S ratio. Since this ratio can be achieved by a range of different widths and spacings, the loss versus impedance characteristic is represented by an area. In general, it was found that from an electrical standpoint the most significant difference between different transmission-line designs was in the impedance ranges which could be realized with acceptable loss. It was found that the edge-coupled lines such as coplanar waveguide and coplanar strips had practical impedance ranges that were about twice those for broadside-coupled lines, such as microstrip, for given values of loss. In view of the coplanar format of the diode and the coplanar receiving elements, which showed promise, the main part of the mixer circuit was designed using coplanar transmission lines.

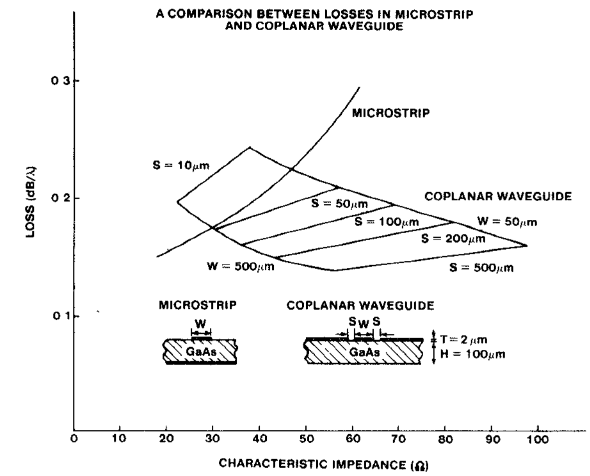

Fig. 1. A comparison between losses in microstrip and coplanar waveguide.

II. Mixer Circuit Design

A planar circuit for the mixer and antenna was designed and this is shown in Fig. 2. Apart from a ground plane for the microstrip elements, all the metallization for critical conductor geometries and the devices was on one side of the substrate. A balanced dipole bow-tie antenna was selected feeding a 100-Ω balanced coplanar strip transmis-

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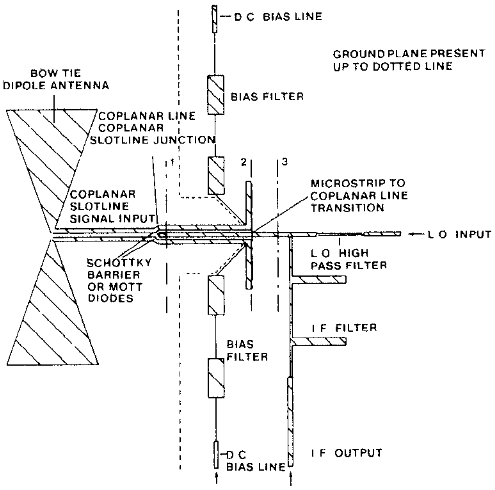


Fig. 2. A 30-GHz monolithic planar single balanced mixer incorporating a bow-tie antenna/WG transition element.

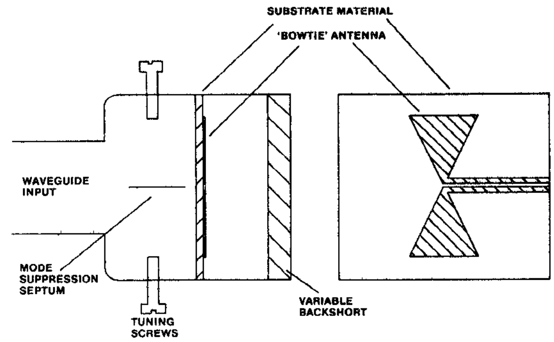


Fig. 3. Key features of WG to balanced line transition using a bow-tie antenna element.

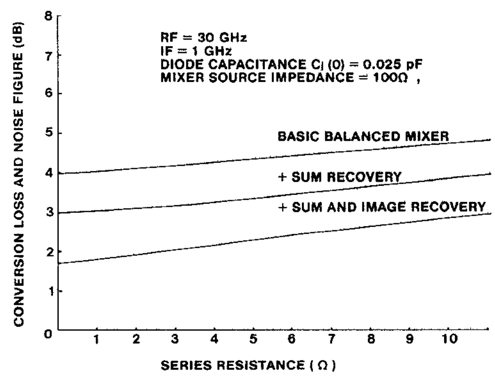


Fig. 4. Conversion loss and noise figure versus diode series resistance.

sion line and the mixer design was assessed with a bow-tie as a free-space receiving element and as a transition to waveguide as shown in Fig. 3. This transition will be described later.

The two most important factors in a mixer design are 1) the nonlinear element and 2) the embedding structure or circuit surrounding the device. The two are interdependent and, therefore, it is usual to specify one and design the other around it. A device was designed which had optimum characteristics for operation in the 100-Ω coplanar strip transmission line at 30 GHz. The two most important parameters in the device are 1) the zero-bias junction capacitance and 2) the series resistance. The tradeoffs between the two were established from previous work [5] and a junction capacitance of 0.025 pF determined.

A plot is given in Fig. 4 showing the change in conversion loss as the device series resistance is varied from 0–12-Ω. It has been shown [5] that the most significant proportions of power are generally lost at the sum and image frequencies. Three curves are given which show the conversion loss for the basic mixer circuit together with the improvement as energy is recovered from the sum and image frequencies. These results do not include losses in the input feed line or IF filter circuits and are based on the high LO drive condition where negligible power is dissipated in the junction.

A Mott diode [6], [7] was chosen for the nonlinear device. The Mott diode is a form of Schottky diode in which the active layer is made very thin so that it remains virtually fully depleted over a wide range of junction voltages yielding a very flat capacitance voltage characteristic. This feature is important because it enables good conversion loss and noise figure to be obtained at lower LO powers than with the standard Schottky diode. The reason for this will now be explained.

In a given mixer, minimum conversion loss is obtained with a particular value of LO duty ratio (commonly referred to as pulse duty ratio for a bilinear diode and defined as the ratio of low impedance or "on" state to high impedance or "off" state of the diode). The pulse duty ratio S is given approximately by the equation [5]:

$$s = \frac{1}{\pi}\cos^{-1}\left(\frac{V_T - V_{\mathrm{dc}}}{V_{\mathrm{LO}}}\right) \tag{1}$$

$$V_T = \frac{1}{\alpha}\ln\left(\frac{1}{\alpha I_s (R_s + r_s) F}\right) \tag{2}$$

$$\alpha = \frac{q}{nkT} \tag{3}$$

where

- V_T effective turn-on voltage of the diode (V),
- V_{dc} dc bias voltage (V),
- V_{LO} magnitude of local oscillator fundamental appearing across diode junction (V),
- α exponential factor (V^{-1}),
- I_s reverse leakage current (A),
- R_s signal source resistance (Ω),
- r_s diode series resistance (Ω),
- F an empirical factor, typically 0.45, which is weakly dependent on the mixer embedding structure,
- q charge on an electron $= 1.6 \times 10^{-19}$ (coulombs),
- n diode ideality factor,
- k Boltzmann's constant $= 1.38 \times 10^{-23}$ (J/K),
- T temperature (K).

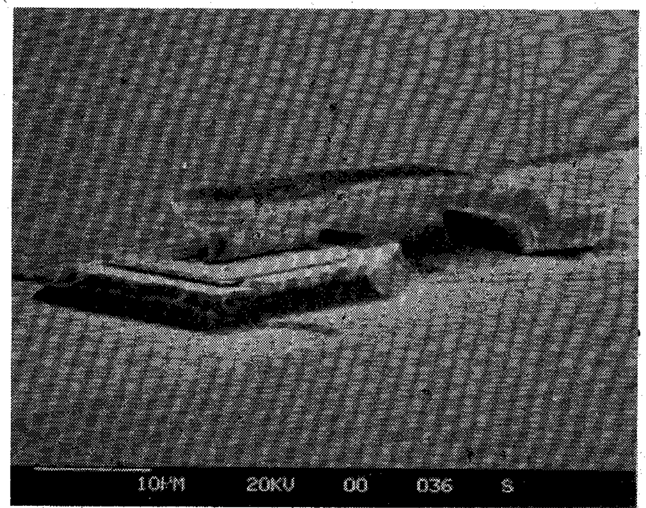
In a practical mixer, a particular value of pulse duty ratio is achieved by a combination of dc bias and LO power as given by (1). As the LO power is decreased, it is necessary to apply increasingly more positive dc bias to maintain the correct operating point to achieve the optimum pulse duty ratio for minimum conversion loss. In the conventional Schottky diode, a significant increase in capacitance occurs when forward bias is applied which is much greater than with the Mott as shown in Fig. 15. This causes a significant shunting on the magnitude of the signal voltage appearing across the junction — an effect which increases as the dc operating point becomes more positive. Since the frequency conversion process occurs primarily in the nonlinear resistance of the junction, any shunting effect from the capacitance degrades the conversion loss. This problem is significantly overcome in the Mott diode due to the flat capacitance voltage characteristic.

The second effect of the nonlinear capacitance is on the noise figure. The nonlinear capacitance causes parametric amplification of the noise voltages of all mixed products appearing across the diode junction. This effect caused an increase in the noise temperature of the device and, hence, the overall mixer noise figure.

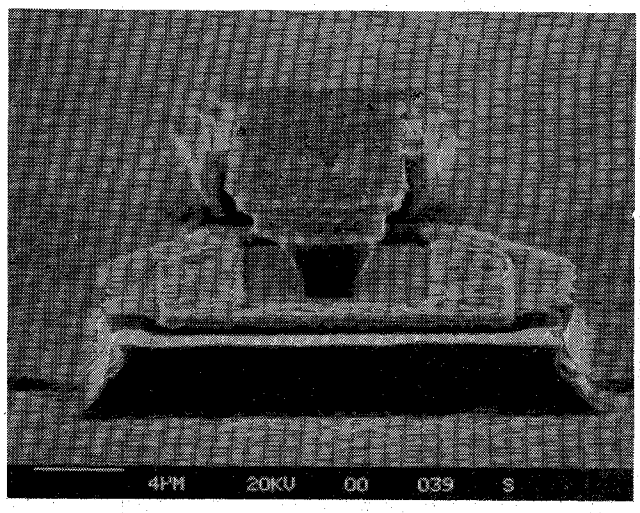
The mixer circuit described in this paper is of single balanced design where the signal and LO are mutually isolated due to the circuit geometry and not due to the use of filters. The signal is fed to the diodes via a balanced line in coplanar strips [8]. The mixer proper is formed at the junction between the coplanar line (with reduced width ground planes) and the coplanar strips. The diodes are mounted at this junction, which operates in a similar manner to the coplanar slotline junction commonly used in hybrid circuits. The slots formed either side of the coplanar line are made to be an integer number of wavelengths long at the signal frequency in order to provide an open-circuit plane at the diodes ($\lambda/2$ in this case). A transition between the microstrip line and the coplanar line is achieved by grounding the finite ground plane of the coplanar lines with quarter-wavelength stubs in microstrip. In addition to overcoming the requirement for via holes, this type of transition is important because it allows a dc voltage to be developed across the diodes. The dc bias is supplied via low-pass hi–lo filters realized in microstrip. The LO is connected to the coplanar line via a transition where the top conductor is common and the finite-width ground planes are ground ed to the underneath ground plane using quarter-wavelength microstrip stubs mounted at the transition point. A high-pass quarter-wavelength coupled line filter provides dc isolation between the diodes and the LO and also prevents the LO from loading the IF signals emerging from the mixer. A simple line and stub filter is used to extract the IF and provides more than 30 dB of LO isolation.

The microstrip lines on the 4-mil GaAs were interfaced to the APC 3.5-mm LO and IF ports via a 10-mil alumina substrate. The dimensions of a specially designed microstrip launcher and GaAs/alumina substrate spacing were chosen to minimize reflections [9].

The bow-tie was investigated as both a receiving element and as a transition to waveguide by making measurements on 3 to 1 scale models at 10 GHz.



(a)



(b)

Fig. 5. SEM micrographs of a typical Mott diode.

III. Fabrication

The monolithic single balanced mixer was fabricated using MBE material. The layer structure consisted of a 3-μm-thick n^+ layer doped at 2×10^{18} cm^{-3} followed by an 0.2-μm-thick layer doped at 5×10^{16} cm^{-3} grown on undoped LEC substrates. The process sequence is as follows. The cathode contact was formed by etching down to the n^+ layer and defining the ohmic contact metal by a liftoff technique. After alloying the AuGeNi contact, the devices were isolated by etching 4-μm-deep mesas. Next, contact areas were defined on the top and at the bottom of the mesa to form the cathode contact pad. At the same time, the anode contact was defined on top of the mesa along with a contact pad on the semi-insulating substrate. The Schottky barrier was defined by sputtering TiW and Au. The Schottky metal formed the base metal layer for the following plating step which defined the airbridge between the Schottky barrier and the pad. Next, the circuit metallization was defined by a conventional plating process. The final step in fabrication was polishing the wafer down to a thickness of 4 mil and completing the backside metalliza-

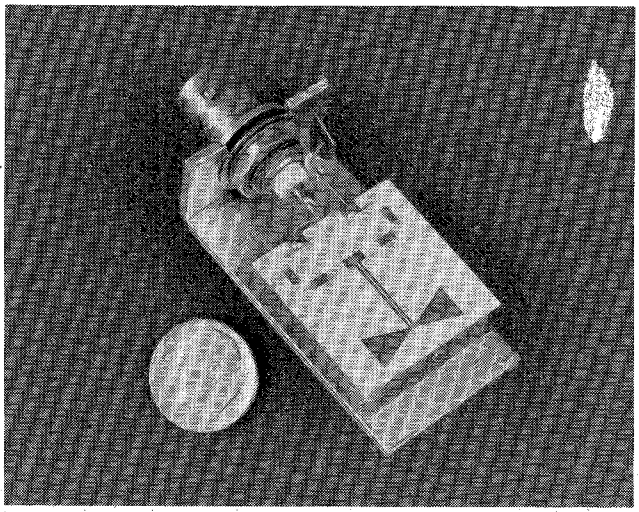


Fig. 6. 10-GHz bow-tie antenna fixture.

tion process, again by a plating technique. This metal provided the ground plane for the microstrip areas and did not exist under the coplanar lines feeding the antenna.

Fig. 5(a) and (b) are SEM micrographs of a typical Mott diode and illustrate the three most notable features of the technology employed.

1) The airbridge interconnect between the active anode and the contact pad at the mesa bottom minimizes the parasitic capacitance commonly observed when a conventional dielectric assisted crossover is used to run the anode finger over the mesa edge (Fig. 5(a)).

2) The thick mushroom-shaped anode finger which is naturally generated by the plating technique employed allows one to have a small active footprint (1 μm) with a thick feed line (see Fig. 5(b)). This enables one to have a small capacitance coupled with a low metal resistance which is necessary for low loss.

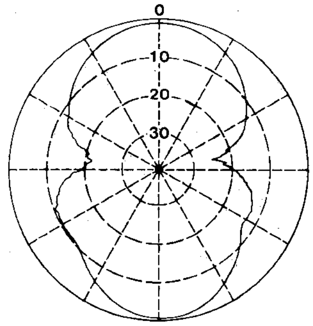
3) A more subtle advantage of the airbridged diode process is that it does not depend on the critical lipped resist profile which is necessary for liftoff. Hence, to a first order, the technology developed is independent of the mesa height employed, which is untrue for a liftoff process.

IV. Measured Results

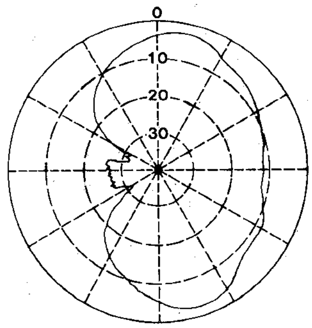
A. Bow-Tie as a Receiving Element

The bow-tie receiving element was required to match to the 100-Ω coplanar strips section of the mixer. Based on a model for a biconical dipole [10] and some previous experimental work conducted at General Electric, an angle of 40° was selected for the small angle of the bow-tie. The length of the antenna was selected as 1.3 times a half-wavelength in free space at the center frequency of 30 GHz.

A 3 to 1 scale model of this antenna without the mixer was built on 12-mil alumina for ease of testing and fabrication. A detector diode was placed across the balanced transmission line to eliminate the need for a balun, as shown in Fig. 6. A low-pass filter section enabled a dc signal proportional to the received RF signal to be extracted directly. Measured E-plane and H-plane patterns of the scaled bow-tie are shown in Fig. 7. The null at 90° in the H-plane patttern is due to blockage from the output connector.



(a)



(b)

Fig. 7. (a) E-plane and (b) H-plane patterns of bow-tie antenna at 10 GHz.

The antenna was investigated for use in a phased-array that requires a unidirectional pattern; therefore, a ground plane was placed on the back side of the antenna spaced such that the reflected energy added in phase with the "front lobe." E-plane and H-plane patterns of this configuration are shown in Fig. 8.

The 30-GHz version of this bow-tie was built on 4-mil alumina, and the H-plane pattern of this antenna with the reflecting ground plane is shown in Fig. 9.

The E-plane and H-plane patterns measured at 10 and 30 GHz were virtually identical, and a plot of relative gain versus frequency is shown in Fig. 10. The figure shows only the relative gain which was determined from the detector output. Calculations from the antenna patterns indicated the maximum gain to be 5 dB. This figure shows a 3-dB bandwidth of 25 percent which is considerably wider than that of other planar receiving elements such as the microstrip patch.

B. Bow-Tie as a Waveguide Transition

Following work on the bow-tie as a free-space receiving element, the structure was examined as a transition to and from rectangular waveguide. This element had the advantage that it was dc isolated from the waveguide which enabled a dc voltage to be developed across the feed point and that the substrate was mounted broadside across the waveguide enabling a very compact transition to be real-

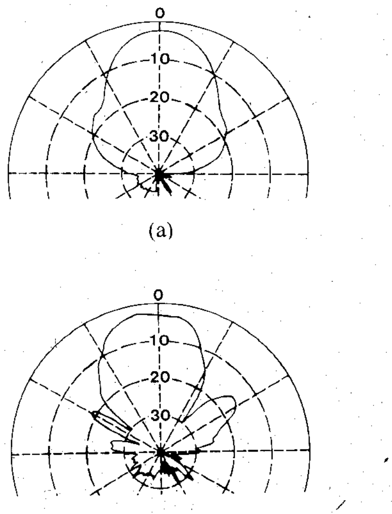


Fig. 8. (a) E-plane and (b) H-plane patterns of bow-tie antenna with ground plane at 10 GHz.

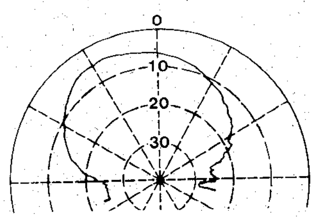


Fig. 9. H-plane pattern of bow-tie antenna with ground plane at 30 GHz.

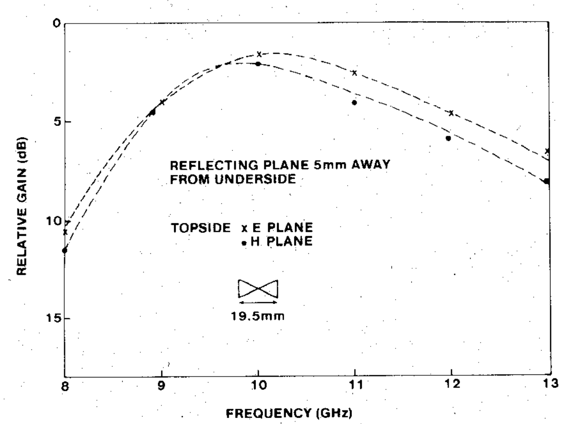


Fig. 10. Bow-tie dipole antenna gain versus frequency.

ized. This transition is essentially a balanced form of an E-plane probe transition to a TEM or quasi-TEM line where a dipole forms the transition to the balanced line. The transition from waveguide is shown in Fig. 3. A standard rectangular waveguide input was stepped to double height a quarter guide wavelength from the plane of the bow-tie, a septum being included to suppress higher order waveguide modes. A variable backshort and capacitive tuning screws enabled the match to the transition to be optimized. The design was investigated and optimized using a 10-GHz scale model shown in Fig. 11. The antenna transition was intended to match into a 100-Ω balanced line; therefore, a series of bow-tie elements of different

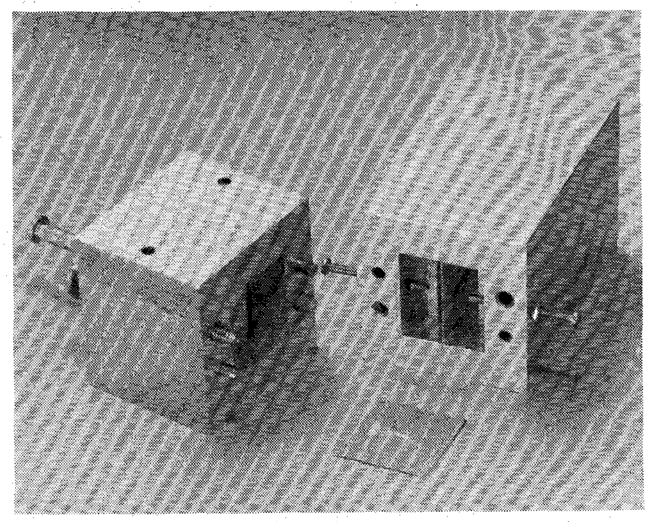


Fig. 11. 10-GHz WG to balanced line test fixture.

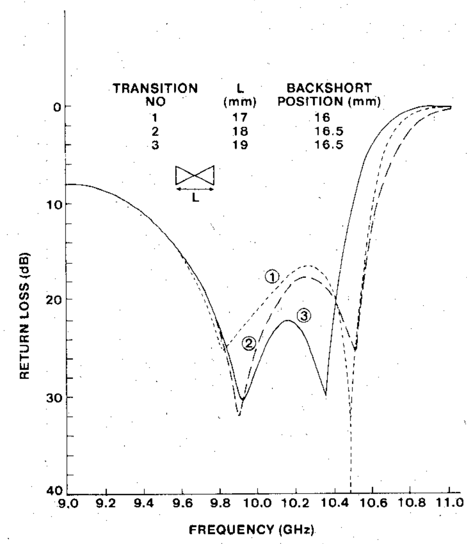


Fig. 12. Return loss of bow-tie waveguide to coplanar balanced line transition.

lengths were made on 12-mil alumina substrate using 100-Ω chip resistors mounted at the center feed point of the elements. The position of the backshort and tuning screws were optimized to give 20-dB return loss over the widest bandwidth centered at 10 GHz. Fig. 12 shows three typical results where it can be seen that the bandwidth is 6.5 percent. The sharp skirt at the higher end of the passband makes this kind of transition suitable for use in an image rejection or recovery mixer circuit. Following these measurements, the 30-GHz transition was developed by scaling the dimensions.

C. DC and RF Test Results of Diode and Mixer

Both dc and RF wafer probing of discrete Mott diodes was performed on approximately 50 devices. DC measurements indicated the diodes had a reverse breakdown of about -7 V, and by accurately measuring the V–I characteristic, the series resistance was determined to be 10 Ω and the ideality factor about 1.08.

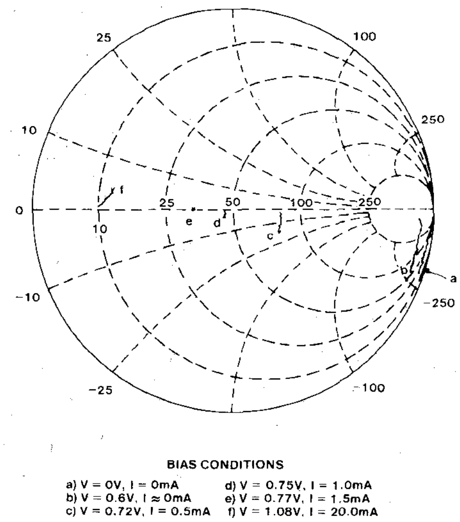


Fig. 13. S_{11} of a shunt mounted Mott diode as a function of dc bias voltage.

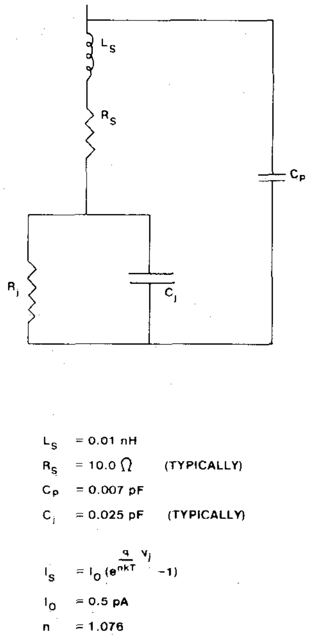


Fig. 14. Diode equivalent circuit.

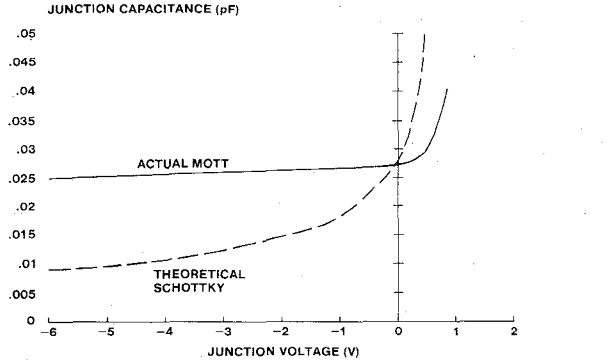


Fig. 15. Capacitance–voltage characteristic of Mott and Schottky diodes.

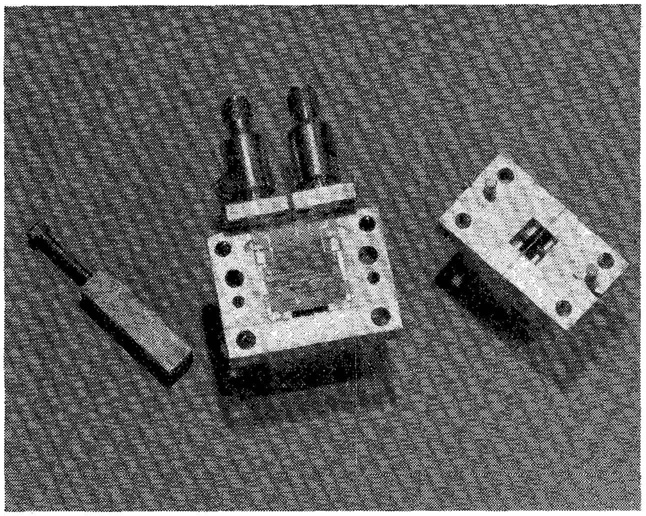


Fig. 16. Disassembled mixer.

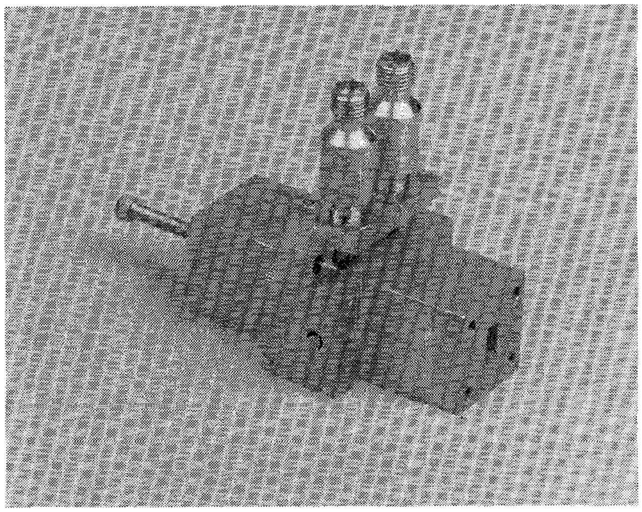


Fig. 17. Complete EHF mixer.

An RF probe station [11] was used to make small-signal one-port S-parameter measurements for shunt-mounted diodes over the frequency range 2–18 GHz.

Fig. 13 shows Smith Chart plots of S_{11} as the forward bias was increased from 0 V to 1.08 V, where the current was 20 mA. The response before the diode starts conducting indicates a small capacitance. As the bias voltage is increased, the resistance becomes more and more dominant and the response is seen to move across the Smith Chart towards a short circuit. When the current through the diode was 20 mA, the plot shows a series resistance of about 10 Ω.

A simple equivalent circuit for the diode was calculated from the one-port S-parameters for different bias conditions, and a zero-bias equivalent circuit is shown in Fig. 14. The element values in the circuit were obtained by computer optimization of the measured versus modeled S-parameters. The starting values for L_s, C_j, and C_p were obtained from the physical geometry of the device and processing parameters, and the series resistance R_s was calculated from dc measurements.

The zero-bias cutoff frequency of the diodes was determined to be 640 GHz. Fig. 15 shows the variation of

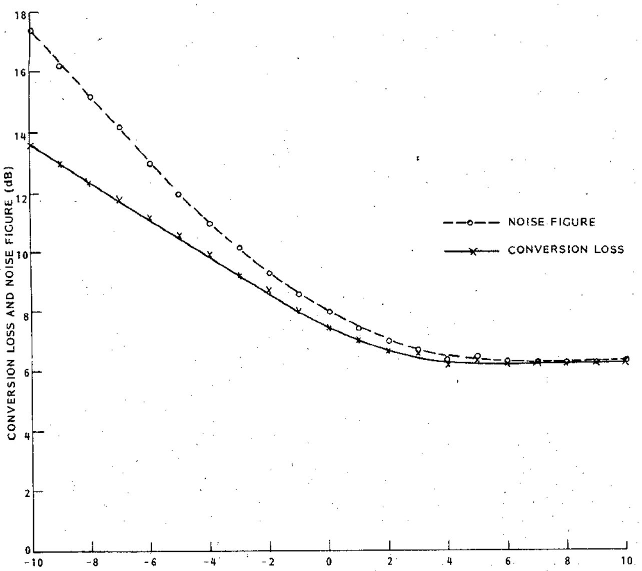


Fig. 18. Conversion loss and noise figure versus LO power of 30-GHz monolithic single balanced mixer.

junction capacitance as a function of junction voltage for both the actual Mott diode and a theoretical Schottky. The Mott diode clearly shows flatter capacitance variation with voltage, being approximately 0.025 pF from -6 V to $+0.5$ V.

Fig. 16 shows the two portions of the waveguide fixture sketched in Fig. 3. The GaAs substrate was mounted on alumina for mechanical strength using a nonconducting cyanoacrylate cement. The complete circuit was mounted in the fixture as shown in Fig. 17.

Measurements were made of conversion loss, and noise figure as a function of LO power with a 30-GHz LO and a 31-GHz signal producing a 1-GHz IF. These are shown in Fig. 18. The variable backshort and inductive tuning screws were adjusted for minimum conversion loss and the dc bias voltage was optimized for each LO setting. A single-sideband conversion loss and noise figure of 6 dB (including fixture losses) was measured with 10 dBm (10 mW) of LO power. This increased by approximately 1 dB when the LO power was decreased to 0 dBm (1 mW).

V. Conclusion

A 30-GHz monolithic low-noise mixer has been described using an integrated bow-tie antenna to waveguide transition and low parasitic Mott diodes. Although the design requires further development to reduce the size and improve the passive circuits, initial measured results for the first design are encouraging. This design incorporates tuning elements in the waveguide mount and can be made compact as the circuit is mounted broadside across the waveguide.

Acknowledgment

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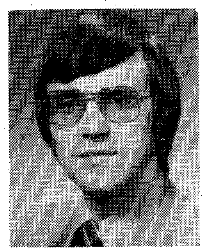

Stephen J. Nightingale (M'82) was born on May 7, 1949, in Redhill, Surrey, England. He joined Philips Research Laboratories, England, in 1967 for a five-year student apprenticeship in mechanical and electrical engineering. He studied at South Bank Polytechnic in London and received a Dip.E.E. in electrical engineering from the Institute of Electrical Engineers (IEE) in 1974 under the joint sponsorship of Philips and the IEE. He received the Ph.D. degree in electronic engineering from Kent University in Canterbury, Kent, in 1980, and became a Chartered Engineer (C.Eng) and a Corporate Member of the IEE in 1981. His work in recent years was concerned with the design and construction of microwave/millimeter-wave components and subsystems for frequencies up to 100 GHz. This work included the design and development of millimeter-wave radiometer systems for target tracking and imaging applications in the 27–40- and 75–110-GHz frequency bands.

From 1977–1979, he was seconded to the Philips Forshungslaboratorium, Hamburg, West Germany, where he worked on the noise characteristics of Schottky diodes and radiometric measurement techniques. He joined the Electronics Laboratory of General Electric, Syracuse, NY, in September 1982, where he is the Technology Manager for millimeter-wave applications. His responsibilities include the management of and technical contributions to several millimeter-wave programs for satellite communication and radar systems.

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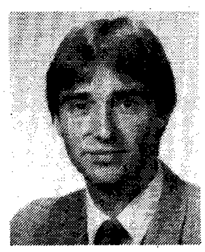


M. Alastair G. Upton (M'84) was born in Bradford, West Yorkshire, England, on May 31, 1959. He received the B.Sc (Hons.) degree in electrical and electronic engineering from the University of Leeds, England, in 1980.

From October 1980 to July 1983, he was employed by Philips Research Laboratories in Redhill, England, where he worked on microwave integrated circuits for instantaneous frequency measurement (IFM) receivers. In August 1983, he joined the Electronics Laboratory at General

Electric in Syracuse, NY, where he has been engaged in monolithic microwave and millimeter-wave circuit design. This work includes mixers, low-noise amplifiers, and characterization of FET's, HEMT's and Mott diodes for EHF circuit applications. He is also involved with the modeling of discontinuities in microwave transitions and junction effects between transmission lines on two different substrate materials.

Mr. Upton is an associate member of the Institute of Electrical Engineers of England.

From 1983 to August 1985, he was a Principal Staff Engineer with the Electronics Laboratory of General Electric in Syracuse, NY, where he directed the internal research and development activity in the area of 0.25-μM gate-length HEMT's for both cryogenic and room-temperature operation. He was also involved with the development of low-loss, monolithically integrable, surface-oriented EHF mixer diodes. Since September 1985, he has been Assistant Professor of Electrical Engineering and Computer Science at the University of Michigan, Ann Arbor, MI.

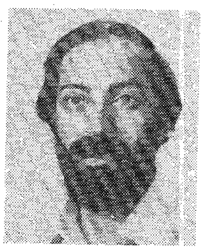
Bruce K. Mitchell (S'76–M'79) received the B.S. degree in electrical engineering in 1979 from Pennsylvania State University.

From 1979 to 1981, he was an antenna design engineer at General Electric's Reentry Systems Division. He transfered to GE's Electronics Laboratory in Syracuse, NY, as a microwave design engineer. He is currently Project Manager on a number of millimeter-wave communication and radar programs.

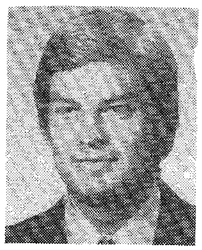
Mr. Mitchell is currently the Vice Chairman of the Syracuse IEEE section and past Chairman of the MTT/AP chapter.

Susan C. Palmateer (M'84) received the B.S. degree in chemistry from Monmouth College, in 1979, and the M.S. (1982) and Ph.D. (1984) degrees from Cornell University, Ithaca, NY, both in electrical engineering.

Prior to graduate school, she worked at Bell Laboratories on molecular-beam epitaxial growth and characterization of indium gallium arsenide. She has been an employee of the General Electric Electronics Laboratory since 1983. She is responsible for the growth and characterization of high-purity molecular-beam epitaxial GaAs AlGaAs material for microwave device applications.



Umesh K. Mishra (M'83) was born in Pune, India, on September 25, 1958. He received the Bachelor of Technology degree from the Indian Institute of Technology, Kanpur, India, in 1979. The subject of his Bachelor's thesis was oxide semiconductor-on-silicon solar cells. He received the M.S. degree Lehigh University, Bethlehem, PA, in 1981, where he worked on metal-tunnel oxide-silicon switching devices. In 1984, he received the Ph.D. degree from Cornell University, Ithaca, NY. His thesis dealt with submicron vertical electron transistors in GaAs with heterojunction (AlGaAs) cathodes.



Phillip M. Smith (S'81–M'82) received the B.S., M.S., and Ph.D. degrees in electrical engineeering in 1976, 1978, and 1982, respectively, from Cornell University, Ithaca, NY. His thesis work consisted of an experimental investigation of the high-field transport properties of semiconductors.

Since joining the General Electric Electronics Laboratory in 1981, he has worked on a variety of microwave devices and associated monolithic microwave IC's, and has established accurate S-parameter, power and noise device characterization capabilities. From 1981 to 1984, he spent a portion of his time as a Visiting Scientist at Cornell. He is currently involved in the design, characterization, and analysis of GaAs power FET's and short gate-length GaAs FET's and HEMT's for applications ranging from 1 to 94 GHz.