Design and Performance of *W*-Band Broad-Band Integrated Circuit Mixers

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Abstract —**Broad-band integrated circuit mixers using a crossbar suspended stripline configuration and a finline configuration were developed with GaAs beamlead diodes. For the crossbar suspended stripline balanced mixer, less than 7.5-dB conversion loss for 15-GHz instantaneous IF bandwidth was achieved with the LO at 75 GHz and the RF swept from 76 to 91 GHz. With the LO at 90 GHz, a conversion loss of less than 7.8 dB was achieved over a 14-GHz instantaneous bandwidth as the RF is swept from 92 to 105 GHz. For the finline balanced mixer, a conversion loss of 8 to 12 dB over a 32-GHz instantaneous IF bandwidth was achieved as the RF is swept from 76 to 108 GHz. Integrated circuit building blocks, such as filters, broadside couplers, matching circuits, and various transitions, were also developed.**

I. Introduction

RAPIDLY EXPANDING activities in millimeter-wave hardware developments have created an urgent need for broad-band mixers for the receivers used in electronic warfare, surveillance, meterology, radiometer, and communication systems. During the past decade, significant improvements have been made in millimeter-wave mixers. Most of these, however, have been limited to narrow-band applications. Recently, several broad-band mixers have been built in waveguide circuits [1–3]. Among these is a crossbar configuration reported by Yuan [3]. This circuit was later realized at *W*-band using a stripline structure and pill type mixer diodes [4], and using a waveguide configuration and honeycomb mixer diodes [5].

Integrated circuit technologies provide the advantages of low cost, light weight, and small size. These also have the potential of direct translation into monolithic circuits and even large-scale integration.

The use of beamlead diodes in integrated circuit mixers avoids a mechanical diode contact in whisker-contacted mixers and therefore has better reliability and needs less assembly time. A 94-GHz mixer using beamlead diodes has recently been reported with a single sideband conversion loss of 6.2 dB [6] and 5 to 9 dB [7] over a 3-GHz RF bandwidth.

This paper describes the performance of a *W*-band wide-band integrated circuit mixer using beamlead diodes with a wide instantaneous IF bandwidth [8]. This mixer is especially important for millimeter-wave receivers which, in addition to accepting a wide range of RF frequencies, should provide wide instantaneous IF bandwidth with a fixed LO frequency.

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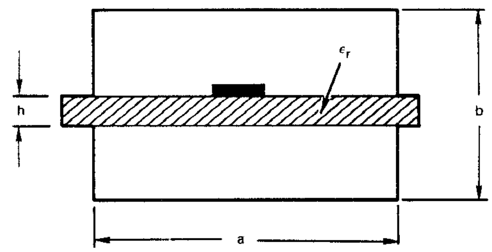

Fig. 1. Suspended stripline configuration.

A crossbar stripline mixer was developed to achieve less than 7.5-dB conversion loss for a 15-GHz instantaneous IF bandwidth with the LO at 75 GHz. A conversion loss of less than 7.8 dB with 14-GHz IF instantaneous bandwidth was achieved with the RF swept from 92 to 105 GHz. For narrow-band operation at 94 GHz, less than 5.5-dB conversion loss was achieved for a 500-MHz bandwidth and less than 6 dB for a 2-GHz bandwidth. A finline balanced mixer was also developed. It operates over a 32-GHz IF instantaneous bandwidth with a conversion loss of 8 to 12 dB, with the LO at 74 GHz as the RF is swept from 76 to 108 GHz. With the LO at 82 GHz, a conversion loss of less than 10 dB has also been achieved with the RF swept from 77 to 107 GHz. These results represent state-of-the-art performance in integrated circuit mixers.

To facilitate the mixer development, filters, broadside couplers, matching circuits, and various transitions were developed in integrated circuit form with low insertion loss. The performance of these components will also be discussed.

II. Suspended Stripline-to-Waveguide Transition

A suspended stripline-to-waveguide transition is essential for individual component testing before final integration. The transition should have low loss and a bandwidth sufficient for the application. Two types of suspended stripline-to-waveguide transitions used in mixers were investigated: 1) electric probe transition for crossbar suspended stripline mixer, and 2) waveguide-to-finline-to-stripline transition for finline mixer.

Theoretical studies have considered the case of an *H*-plane slab of dielectric, centrally located within the waveguide as shown in Fig. 1. The dominant waveguide mode

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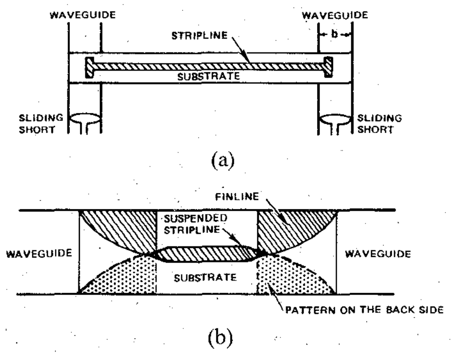


Fig. 2. Suspended stripline-to-waveguide transitions including two transitions for testing purposes. (a) Electric probe type transition. (b) Waveguide-to-finline-to-suspended stripline transition.

can be either the first longitudinal section magnetic (LSM_{11} or quasi TE_{10}) mode or the distorted TE_{01} mode, depending on the dielectric permittivity and guide dimensions. For this application, we are interested in the case where $h/b \leqslant 0.25$ and $\epsilon_r \leqslant 10$. The cutoff frequency [9] of the dominant waveguide mode LSM_{11} can be approximated by

$$f_c = \frac{c}{2a}\sqrt{1-\frac{h(\epsilon_r-1)}{b\epsilon_r}} \qquad (1)$$

where

- a width of channel;
- b height of channel;
- h thickness of the substrate;
- ϵ_r relative dielectric constant;
- c velocity of light in a vacuum.

The a dimension of suspended stripline is chosen so that the first higher order mode is much higher than the frequency band of interest. In the case of a channel cross section of 0.050×0.025 in^2 and a substrate thickness of 0.005 in, the cutoff frequency for duroid is 112 GHz, which guarantees a simple quasi-TEM mode of operation for *W*-band.

The electric probe type transition consists of an electric probe inserted into the waveguide formed by an extension of the suspended stripline or microstrip line beyond the ground plane (Fig. 2). The idea is very similar to the conventional waveguide-to-coaxial line transition. The transition was originally developed for narrow-band operation at *Ka*-band [10]. The advantage of this transition is that the probe can be fabricated as an integral part of the stripline and the difficulty of making reliable electrical contacts is avoided. By optimizing the probe shape, broad bandwidth has been accomplished with low insertion loss. As shown in Fig. 3, excellent results have been achieved at *W*-band using a rectangular probe shape. The total insertion loss for two transitions and a 1-in line is typically 1 dB. The insertion loss of each transition is about 0.25 dB over an 18-GHz range. The upper operating frequency can be easily adjusted by shaping the probe geometry. The assembly of this type of transition is relatively simple, and a sliding short can be used to optimize the performance at a specified frequency.

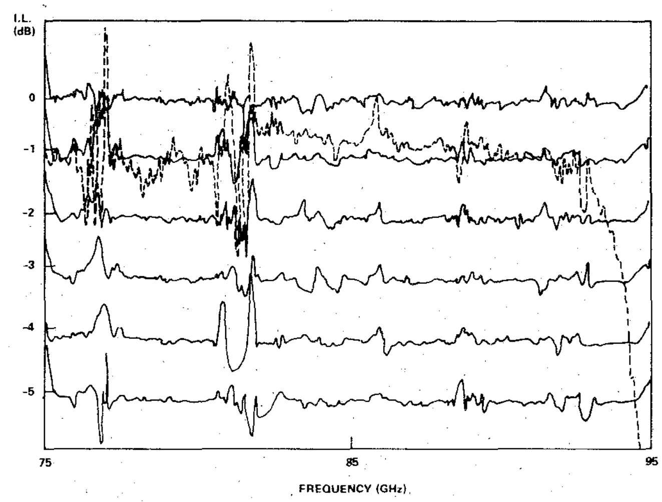


Fig. 3. Performance of *W*-band electric probe type transition (loss including two transitions and a 1-in line). The spikes shown are due to the sweeper.

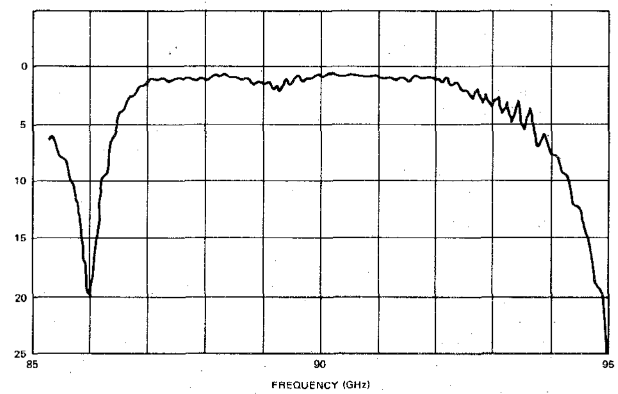


Fig. 4. Performance of waveguide-to-finline-to-stripline transition (including two transitions and a 0.5 in stripline).

The second type transition [11] is designed to couple the power from waveguide-to-finline and then to stripline. As shown in Fig. 2, the finline is tapered down to a small gap. The performance of this transition is shown in Fig. 4. The total insertion loss for two transitions and a half-inch stripline is about 1 dB over a 5-GHz bandwidth. The insertion loss per transition is therefore about 0.25 dB. The narrow bandwidth is believed due to the discontinuities introduced at the fin and stripline interface.

Because of the low loss, wide bandwidth, and easy assembly, the electric probe type transitions were used for most of our component testing.

III. Filter Design

In the mixer, a low-pass filter is required to pass the IF frequency and reject the LO and RF signals. The RF and LO ports are isolated in the crossbar mixer because the input waveguides are orthogonal to each other. In the finline mixer, however, a bandpass filter is needed to increase the LO/RF isolation. This section discusses the development of these filters.

A Chebyshev IF filter was selected. The element values for a Chebyshev filter with a 0.2-dB ripple are computed

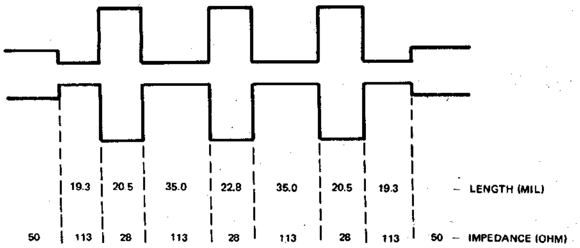


Fig. 5. Low-pass filter circuit layout.

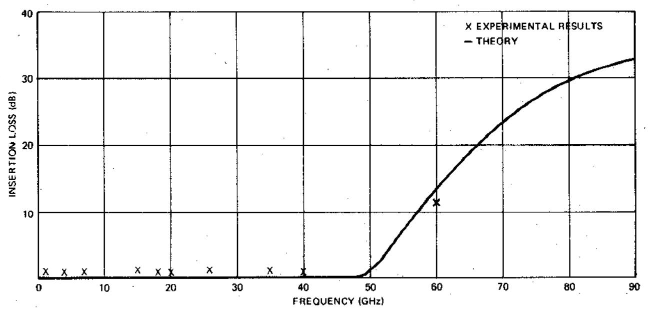


Fig. 6. Predicted and measured performance of a low-pass filter.

Fig. 7. Direct-coupled resonator filter.

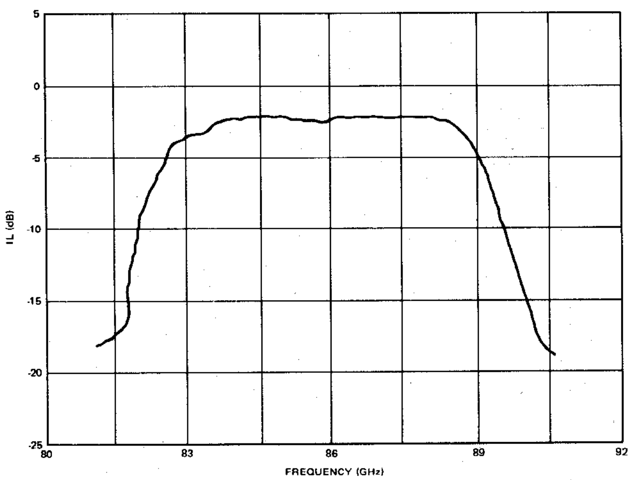


Fig. 8. Performance of a *W*-band bandpass filter (including two transitions and a 0.5 in extra length).

from tables given by Matthaei *et al.* [12]. The filter performance is optimized through the COMPACT computer program. The cutoff frequency is designed to be 50 GHz to provide an attenuation greater than 20 dB at the RF and LO frequencies.

A seven-element semi-lumped structure consisting of three capacitive and four inductive sections is shown in Fig. 5. The predicted and measured performances are given in Fig. 6.

The bandpass filter consists of multiple-coupled resonators and can be realized from end-coupled half-wavelength strips in a suspended stripline configuration, as shown in Fig. 7. The minimum gap is set at 3 mils for ease of fabrication.

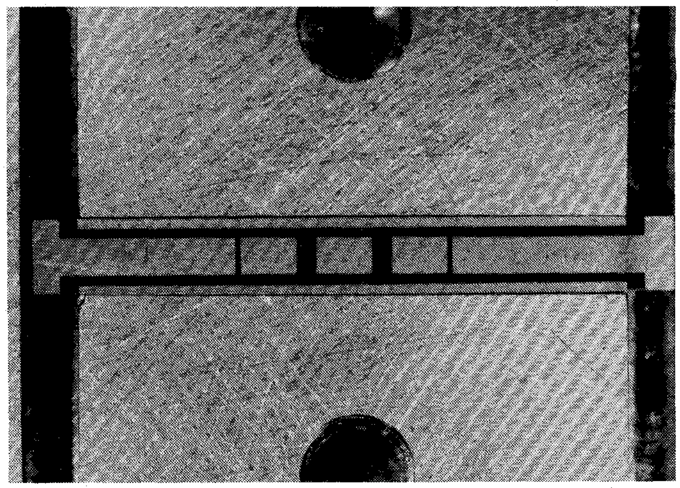
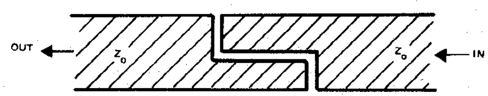
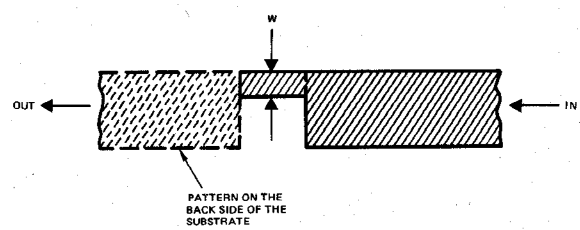


Fig. 9. *W*-band bandpass filter with two transitions.



(a)



(b)

Fig. 10. Two different configurations of broadside coupler. (a) Edge coupler with lines on the same side of substrate. (b) Broadside coupler with lines separated by the substrate.

A bandpass filter has been designed using the COMPACT program to pass the LO and reject the RF frequency. This computer program calculates the required capacitances from which the gap dimensions can be determined [13], [14].

Fig. 8 shows typical performance of a *W*-band bandpass filter centered at 86 GHz. The insertion loss includes two transitions and a 0.5 in extra length of stripline. The insertion loss of the filter is thus about 1.5 dB over a 5-GHz bandwidth after the correction of transition loss. Fig. 9 is a photo of the filter with two transitions made for testing purposes.

IV. Broadside Coupler Development

The RF power propagating down a suspended stripline couples to the stripline on the other side of the substrate through a broadside coupler. The broadside coupler could serve as a dc block or present an open circuit to the IF frequencies.

At lower frequencies, the conventional edge coupler shown in Fig. 10(a) has been used quite successfully. At *W*-band, to achieve minimum insertion loss, it is necessary to have very tight coupling with a gap too small to realize practically. In this case, thin lines are superimposed and

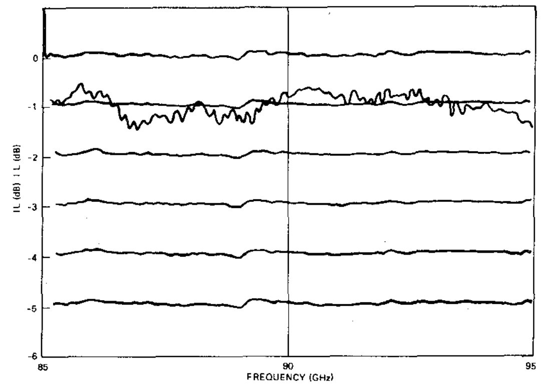


Fig. 11. Performance of a broadside coupler (including a two transitions and a 0.5-in section).

separated by a substrate material of low dielectric constant, with thickness on the order of 0.005 in, as shown in Fig. 10(b). The design of this coupler is straightforward and can be found in Matthaei [12]. The characteristic impedance of the coupled lines can be specified in terms of Z_{oe} and Z_{oo} as even- and odd-mode impedances such that

$$Z_{oe} = Z_o\sqrt{\frac{1+K}{1-K}} \tag{2}$$

$$Z_{oo} = Z_o\sqrt{\frac{1-K}{1+K}} \tag{3}$$

$$Z_o = \sqrt{Z_{oe}Z_{oo}} \tag{4}$$

where Z_o is the impedance of the terminating transmission lines and K is the coupling coefficient.

The width of the coupling lines is determined by Z_{oe} and Z_{oo}, and the substrate thickness as well as the distance between the ground planes above and below the thin substrate [12]. The length of the coupling section is about a quarter of a wavelength corrected for the compensation due to discontinuities. Typical performance is shown in Fig. 11 for the frequency range of interest. The insertion loss shown includes two transitions for testing purposes and a 0.5-in extra length of transmission line. The insertion loss of the coupler is thus less than 0.2 dB.

V. Crossbar Suspended Stripline Balanced Mixer

The circuit configuration of our crossbar stripline is shown in Fig. 12. The RF signal is applied to mixer diodes from a waveguide perpendicular to the circuit board. The crossbar configuration is formed by two mixer diodes with opposite polarity connected in series across the broadwall of the waveguide. The mixer diodes are thus in series with respect to the RF signal and in parallel with respect to the IF circuit. The IF signal is extracted via a low-pass filter and the LO signal is injected from the other side through a broadside coupler and an electric probe type transition.

A. Design Considerations

In a mixer design, the performance parameters of primary concern are the operating bandwidth and the conversion loss. To treat this analytically, we have developed a circuit

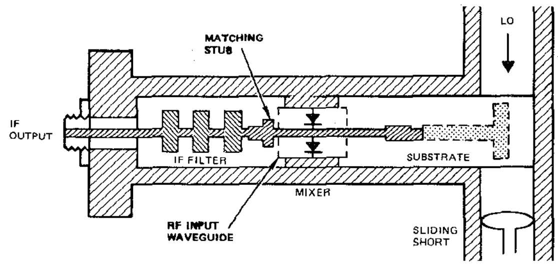


Fig. 12. Crossbar stripline mixer.

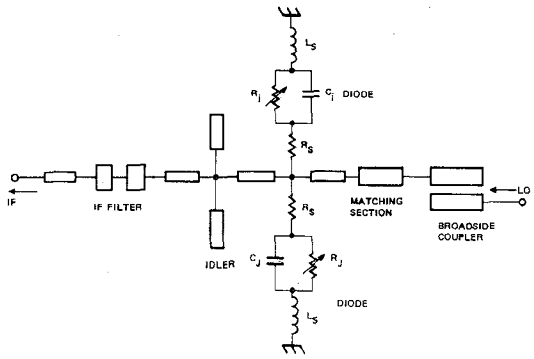


Fig. 13. Crossbar stripline mixer equivalent circuit.

model (Fig. 13) which deals quantitatively with the mixer performance.

In general, the conversion loss of a mixer is considered to consist of three parts:

- L_1 mismatch loss due to impedance mismatch at RF and IF ports
- L_2 diode parasitic loss due to its junction capacitance and series resistance
- L_3 intrinsic junction loss of the ideal diode.

Mathematically, these losses can be expressed as [3]

$$L_1 = 10\log\left[\left(1-|\Gamma_{\mathrm{RF}}|^2\right)\left(1-|\Gamma_{\mathrm{IF}}|^2\right)\right]\,\mathrm{dB} \tag{5}$$

where Γ_{RF} and Γ_{IF} are the reflection coefficients at the RF and IF ports and

$$L_2 = 10\log\left[1+\frac{R_s}{R_j}+(\omega C_j)^2 R_s R_j\right]\mathrm{dB} \tag{6}$$

where

- R_s diode series resistance;
- C_j diode junction capacitance;
- R_j diode junction resistance;
- ω operating frequency in radians.

The intrinsic junction loss of a mixer depends on the terminating conditions of the image frequency and nonlinearity of the diode conductance and is not dependent on frequency. As an example of the image matched condition, the junction loss is given as

$$L_3 = \left\{1+\left[\frac{1+\frac{g_2}{g_0}-2\left(\frac{g_1}{g_0}\right)^2}{1+\frac{g_2}{g_0}}\right]^{1/2}\right\}^2\left[1+\left(\frac{g_2}{g_0}\right)\right]\left(\frac{g_0}{g_1}\right)^2 \tag{7}$$

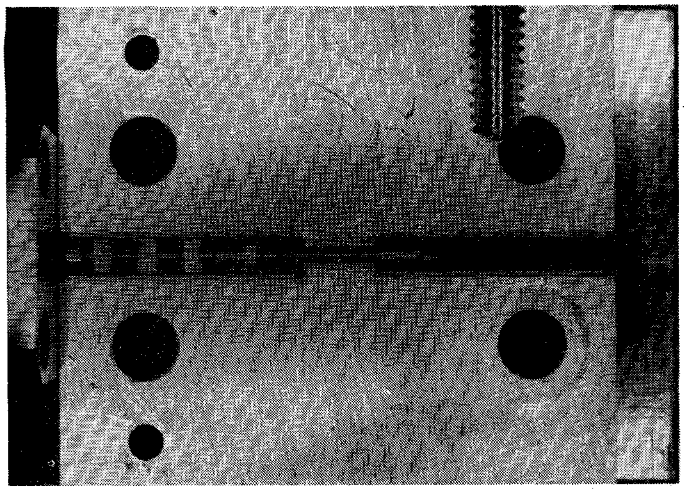


Fig. 14. Circuit layout of *W*-band crossbar stripline mixer.

where g_0, g_1, and g_2 are the Fourier coefficients of the diode conductance [15].

The diode parasitic loss L_2 depends on the values of the R_s, R_j, and C_j of the diode. For specified diodes, L_2 is fixed and cannot be improved. Consequently, circuit optimization was concentrated on the reduction of mismatch loss (L_1) and intrinsic junction loss (L_3).

L_1 is minimized by optimizing the RF and IF circuit matching. The junction resistance R_j of the mixer diode is varied with LO voltage and its value can be as low as 100 to 150 Ω under fully turned-on conditions. Waveguide impedance is in the range of 400 to 600 Ω and can be matched to the diode impedance by a reduced-height taper transformer. The sliding short on the opposite side of the RF port will tune out the reactance part of the diode.

IF and LO matching are aided by a computer analysis of the equivalent circuit shown in Fig. 13. An IF filter passes the IF frequency band and rejects the LO and RF signals. The connecting transmission line between the IF and LO ports can be optimized to provide matched conditions at the LO and IF ports.

A broadside coupler was designed to present an open circuit to the IF frequencies to prevent dissipation of IF power in the LO port. The coupler also serves as a dc block as the mixer is integrated with an MIC local oscillator. It can be designed to present an open circuit at the image frequency and thus reflect back the power for remixing with the LO to generate power at the IF frequency. A double open stub (Fig. 12) was used to facilitate the LO matching.

B. Mixer Performance

A photograph of a *W*-band crossbar stripline circuit layout is shown in Fig. 14. With the LO at 75 GHz, a conversion loss of less than 7.5 dB for 15-GHz instantaneous IF bandwidth was achieved with two beamlead diodes as the RF is swept from 76 to 91 GHz (Fig. 15(a)). The beamlead diodes are commercially available diodes with a C_j of approximately 0.04 to 0.05 pF and R_s of 5 to 7 Ω. With the LO at 90 GHz, a conversion loss of less than 7.8 dB was achieved over a 14-GHz instantaneous IF bandwidth as the RF is swept from 92 to 105 GHz (Fig. 15(b)). The mixer can be tuned for narrow-band operation at 94 GHz. Less than 5.5-dB conversion loss was achieved for 500-MHz bandwidth and less than 6 dB for 2-GHz bandwidth as shown in Fig. 15(c). RF/LO isolation of over 20 dB was achieved. These results represent state-of-the-art performance in this frequency range for an integrated circuit mixer.

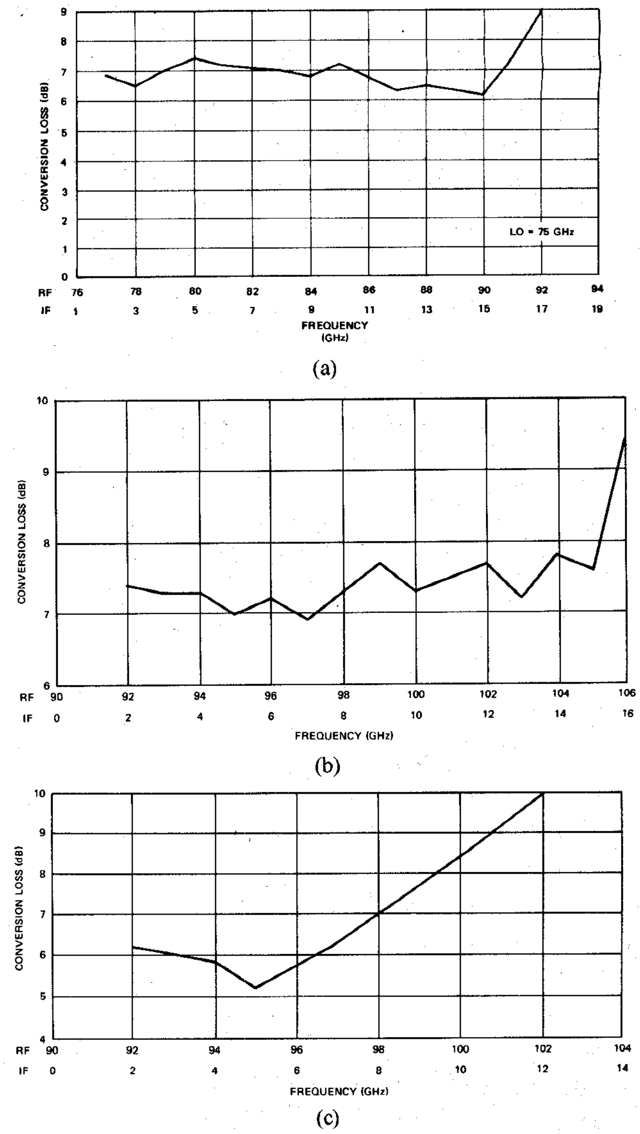


Fig. 15. Crossbar stripline mixer performance. (a) LO = 75 GHz, RF = 76–90 GHz. (b) LO = 90 GHz, RF = 92–105 GHz. (c) Narrow-band operation.

VI. Finline Balanced Mixer

In conjunction with the crossbar stripline balanced mixer development, a finline balanced mixer was also developed. Finline has a simple geometry and is compatible with metal waveguides. A finline *W*-band balanced mixer has been reported by Meier for narrow-band operation [16] and recently realized for medium bandwidth operation by others [17], [18]. This paper presents a design which results in wide-band operation.

A circuit layout is given in Fig. 16 showing the basic configuration of this mixer. The RF is fed from the right

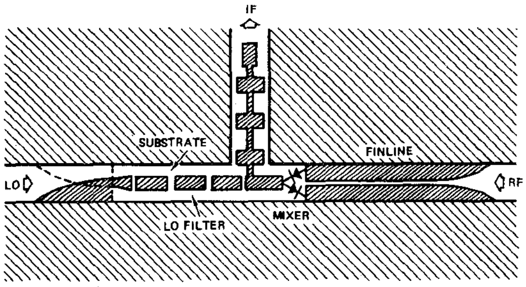


Fig. 16. Finline mixer circuit layout.

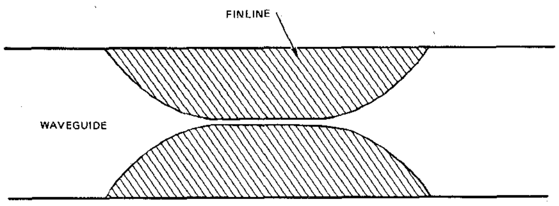


Fig. 17. Waveguide-to-finline-to-waveguide transition for testing.

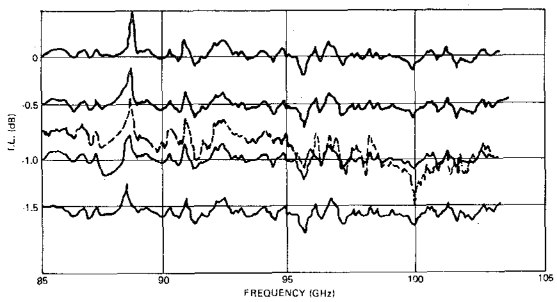


Fig. 18. Insertion loss measurement (including two transitions and a 0.5 in finline section).

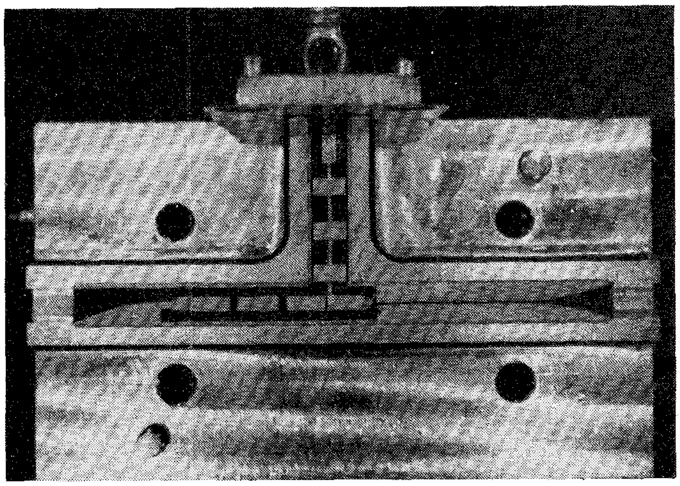


Fig. 19. Photo of finline mixer.

through a tapered finline and LO power is coupled to the mixer diodes through a waveguide-to-finline-to-suspended stripline transition and a suspended stripline bandpass filter. The LO bandpass filter built on suspended stripline is implemented to achieve good RF to LO isolation. IF output is taken out via a low-pass filter.

To achieve good RF matching, a cosine taper finline transition was incorporated into the design. The transition

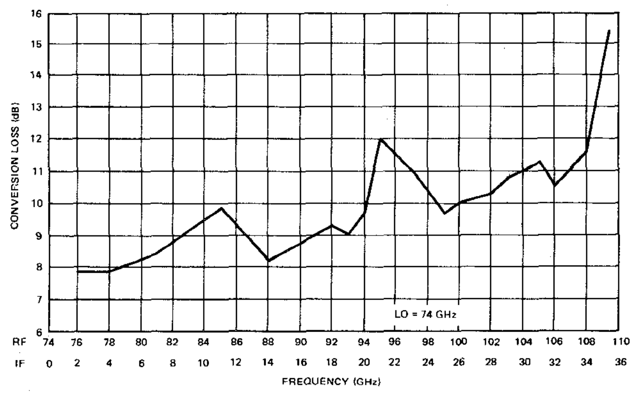


Fig. 20. W-band finline mixer performance (LO at 74 GHz).

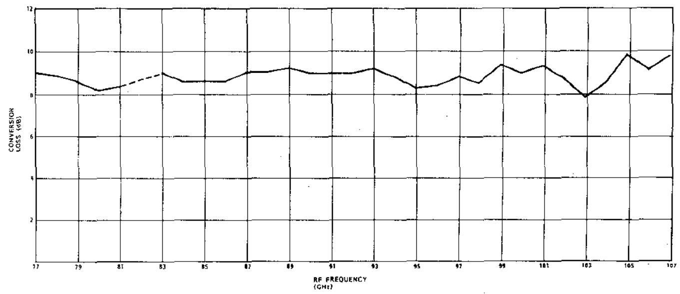


Fig. 21. W-band finline mixer performance (LO at 82 GHz).

was first built, tested, and optimized as an individual component (Fig. 17). Fig. 18 shows the characterization of this transition with typical insertion loss per transition of about 0.2 to 0.3 dB. It is believed that this transition has a much wider bandwidth than that shown in Fig. 18. The gap size between the tapered finlines is varied to achieve the optimum impedance matching to the mixer diodes. To match the low diode impedance (100 to 150 Ω), the gap size was designed to be less than 0.004 in, as optimized by a computer program.

A photograph of this mixer is shown in Fig. 19. A conversion loss of 8 to 12 dB was achieved over 32-GHz IF instantaneous bandwidth with the LO at 74 GHz as the RF is swept from 76 to 108 GHz (Fig. 20). Compared to the crossbar stripline mixer, the finline mixer operates at a wider bandwidth with slightly higher conversion loss. With the LO at 82 GHz, a conversion loss of less than 10 dB has been achieved with the RF swept from 77 to 107 GHz (Fig. 21).

VII. Conclusions

W-band broad-band integrated circuit mixers were developed using beamlead diodes. The mixer configurations are crossbar suspended stripline and finline. For crossbar suspended stripline mixers, a conversion loss of less than 7.5 dB over a 15-GHz instantaneous IF bandwidth was achieved with a fixed LO at either 75 or 90 GHz. For a finline mixer, a conversion loss of 8 to 12 dB was achieved over a 32-GHz instantaneous IF bandwidth with the LO at 74 GHz and the RF swept from 76 to 108 GHz. With the LO at 82 GHz, a conversion loss of less than 10 dB has also been achieved with the RF swept from 77 to 107 GHz.

All these results represent state-of-the-art performance in integrated circuit mixers.

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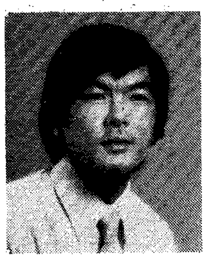
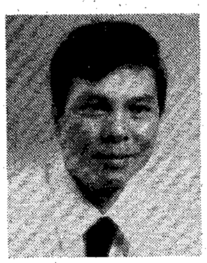
He joined the University of Bradford, England, in 1968, where he became engaged in microwave engineering and was awarded the M.S. degree for work on image cancelling mixers in 1969. Since 1970 he has has been engaged in research and development work in microwave circuits, and has contributed to the design and development of a wide range of microwave components and subsystems. From 1977 to 1981 he was with Hughes Aircraft Company where he was responsible for the development of low cost broad-band millimeter-wave mixers and frequency doublers. He also participated in the development of low-noise Gunn VCO's, step recovery diodes, and varactor multiplier chains. Currently, he is with the Millimeter Wave Technology Department of TRW Defense and Space Systems, Redondo Beach, CA, where he is involved in millimeter-wave integrated circuit development.



George M. Hayashibara was born in Long Beach, CA, on October 26, 1955. He received the B.S. degree from California State Polytechnic University, Pomona, in 1978. Currently, he is pursuing the M.S. degree in microwave systems engineering at the University of Southern California, Los Angeles.

From 1978 to 1981 he worked at Hughes Aircraft Company, Electron Dynamics Division, where he was involved in the development of millimeter-wave solid-state oscillators and power combiners. In August 1981 he joined TRW Defense and Space Systems Sector, Redondo Beach, CA, as a Member of the Technical Staff in the Millimeter Wave Technology Department. His current interests are concerned with millimeter-wave integrated circuits.



Kai Chang (S'75–M'76) was born in Canton, China, on April 27, 1948. He received the B.S.E.E. degree from National Taiwan University, Taipei, Taiwan, the M.S. degree from the State University of New York at Stony Brook, and the Ph.D. degree from the University of Michigan, Ann Arbor, in 1970, 1972, and 1976, respectively.

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