An Integrated-Circuit Balanced Mixer, Image and Sum Enhanced

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***Abstract*—GaAs Schottky-barrier diodes with a zero-bias cutoff frequency of 800 GHz have been used in an integrated-circuit balanced diode mixer operating with a signal frequency centered at 9.3 GHz and a local-oscillator (LO) frequency at 7.8 GHz. For an instantaneous bandwidth of 1.0 GHz, the conversion loss (including all circuits and connector losses) was under 3.15 dB. Over the center 0.5 GHz of the band, the conversion loss was less than or equal to 2.8 dB. The conversion loss at the image-band edges was greater than 25 dB; the loss at the center of the image band was greater than 35 dB.**

I. INTRODUCTION

THIS PAPER reports generally on a microwave mixer for operation at X band, and more specifically on an integrated mixer circuit having very low conversion loss. Microwave diode mixers have been used for many years to obtain conversion of a signal at microwave frequencies to one at a much lower frequency. Such mixers have been the subject of much study and development. However, there is a continuing need for improvements which can result in better electrical performance, higher reliability, improved reproducibility, and lower production costs.

Well known [1]–[4] are the techniques for the enhancement of mixer operation by the proper control of the impedances at each of the mixer terminals and at each of the frequencies of importance. The frequencies of importance are the modulation products which exist according to the heterodyne principle by which the mixer operates. The received signal (RF), together with a higher level signal from a local oscillator (LO), are applied to a nonlinear element. The signal is mixed with the LO producing the sum frequency LO + RF, the difference (or intermediate) frequency (IF), LO − RF, and the image frequency 2LO − RF.

It has been known for some time that this loss in converting an RF signal to an IF signal can be minimized by properly terminating the sum and image frequencies. However, the realization of the proper termination can represent a severe problem. Prior integrated-circuit forms [3], [5] of image-enhanced mixers have generally been single-ended (unbalanced) mixers as opposed to balanced mixers, and have suffered the limitation of narrow-band operation imposed by the use of a narrow-band filter for image termination control.

Image-enhanced mixers can yield substantial improvement in performance only if high-performance diodes are available. The measure of potential performance is indicated by the frequency cutoff of the diode. Very high frequency cutoff (f_{co}) is required for low conversion loss. Until the advent of the Schottky-barrier diode, and in particular, the GaAs Schottky barrier, sufficiently high f_{co} diodes were not available and image-enhanced mixers were only an academic curiosity. Now such Schottky barriers are readily available and low-conversion-loss mixers are a reality.

II. SCHOTTKY-BARRIER JUNCTION PROPERTIES

The Schottky barrier used for mixers primarily requires the variable resistance property of a junction, thus it is commonly called a varistor. The variation of resistance of a varistor as a function of applied voltage is dramatic. The reverse-bias resistance is on the order of many megohms. The resistance decreases rapidly with increasing forward bias until the forward-bias series resistance R_s dominates over the effect of the junction resistance.

The junction resistance is in parallel with a junction capacitance C_j which is also a voltage variable component. The varistor must be designed such that the junction capacitance is minimized for a given series-limiting resistance. To compare varistors of differing R_s and C_j values, it is useful to define a cutoff frequency, $f_{co} = (2\pi C_j R_s)^{-1}$. For this comparison the zero-bias value for f_{co} is useful. It has been found that his value correlates well with measured results.

Figs. 1 and 2 show the f_{co} as computed for silicon and GaAs Schottky barriers. The value for R_s is made up of two parts. The first is the resistance of the epitaxial region X_e, and the second is the parasitic resistance due to the spreading of the current from the epitaxial layer into a substrate of finite conductivity ρ_s. Two values of junction diameter D_j were assumed. Three values of X_e were assumed. The first curve shows the limiting value of f_{co} due to the junction alone (no substrate resistance), and the epi-layer thickness is taken to be $X_e = W_B$. The remaining curves assumed a substrate spreading resistance. The second curve was calculated assuming the epitaxial-layer thickness X_e to be just enough to accommodate the space charge region at breakdown W_B. The third and

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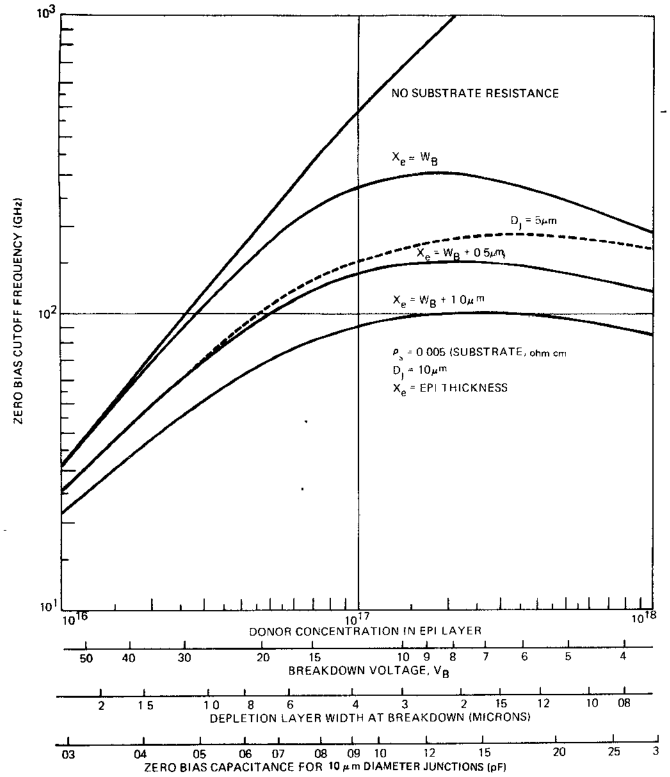

Fig. 1. Theoretical parameters of epitaxial silicon Schottky barriers.

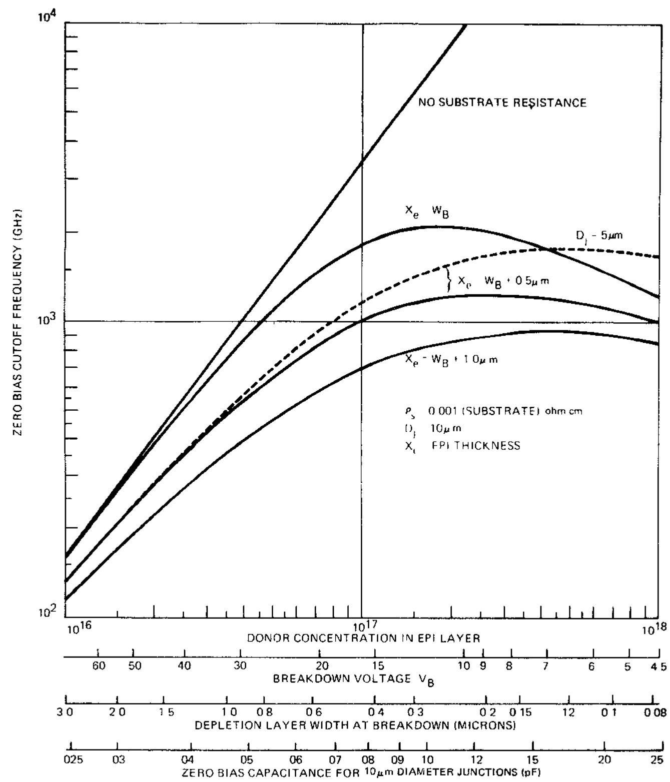

Fig. 2. Theoretical parameters of epitaxial GaAs Schottky barriers.

fourth curves make an allowance in the epi-layer thickness of 0.5 and 1.0 μm, respectively.

The effect of the parasitic substrate resistance is clearly apparent in the figures, both in the drastically reduced f_{co} and in the occurrence of a maximum in the curves [3]. The breakdown-voltage–impurity-density relationship of Sze and Gibbons [6] for an abrupt junction has been plotted also in the figures. These breakdown voltages represent bulk breakdown characteristics. Note that the f_{co} calculation has been made assuming at least an epi-thickness sufficient to be fully depleted at breakdown. This allows maximum breakdown voltage for a given doping level. If a much reduced epi-thickness is used, then an improvement in f_{co} can be obtained. Assume a GaAs epi-layer doping of 10^{16} and a thickness of 0.5 μm. Then one can calculate an f_{co} (for the 10-μm-diam junction on GaAs) of 2283 GHz, a breakdown voltage of 22 V, and a punch-through voltage of 2.0 V. The zero-bias capacitance is 0.024 pF. Now suppose that an excess of 0.5 μm were left on this epi-layer (due to processing variables). Now the value for f_{co} drops to 604 GHz, a drastic change in the value for R_s. Thus the epi-layer thickness for this diode becomes an extremely critical control parameter. However, such a thin-layer diode is attractive, especially for high-burnout diodes, because, for a given application, a junction diameter of 20 μm on a 0.5-μm epi-layer of 10^{16} will give about the same impedance level and f_{co} as a junction dimaeter of 10 μm on a 0.5-μm epi-layer of 2×10^{17}. Such a diode would have a two-to-one thermal impedance improvement over the smaller diode. The case for which the barrier depletion layer at zero bias extends through, or nearly through, the entire lightly doped epi-layer represents a special form of metal-semiconductor barrier known as a Mott barrier [3].

Note in Figs. 1 and 2 that there is an order of magnitude difference in the f_{co} scales between Si and GaAs. Realistic substrate resistances have been assumed. It can be seen that if one assumes a nominal 0.5-μm excess of epi-layer material for both Si and GaAs, that for 10-μm junctions the Si devices will yield an f_{co} of no more than 150 GHz; the GaAs devices can be expected with f_{co} on the order of 1000 GHz. These numbers are realistic and have been readily approximated in practice. Thus GaAs is the natural choice for very low-conversion-loss mixers.

III. DESIGN CONSIDERATIONS

Barber [4] has presented an analysis of microwave mixers and has shown that the pulse-duty ratio of the Schottky-diode current waveform is the most fundamental parameter for defining mixer operation because the diode current pulse retains its typical (switched) shape even when the voltage waveform becomes highly nonsinusoidal.

It can be shown that most microwave mixer diodes (adjusted for lowest conversion loss) behave as though the barrier itself were switched on and off at the LO rate, and that the resistance in the ON state is just that of the limiting series resistance (R_s), and the impedance in the OFF state is just that expected of the series resistance R_s in series with the barrier capacitance C_j. Of course the barrier capacitance is a function of voltage and time, but good correlation with measured results are obtained if the zero-bias capacitance value is used. Thus the frequency cutoff is $f_{co} = (2\pi R_s C_j)^{-1}$.

Using these considerations, an extension of Barber's analysis [7] has allowed the calculation of the conversion loss as a function of the pulse-duty ratio and as limited by the operating frequency to cutoff-frequency ratio (f/f_{co}). Fig. 3 shows the expected mixer conversion loss that would be obtained for the broad-band case (wherein the image termination equals the signal termination). Fig. 4 shows the computed mixer conversion loss for the case wherein the image is short circuited.

Figs. 5, 6, and 7 show the computed values of mixer terminal impedances plotted as functions of the pulse-duty ratio (t). In each case the RF-signal impedance R_{RF} has been chosen to minimize the mixer noise figure.

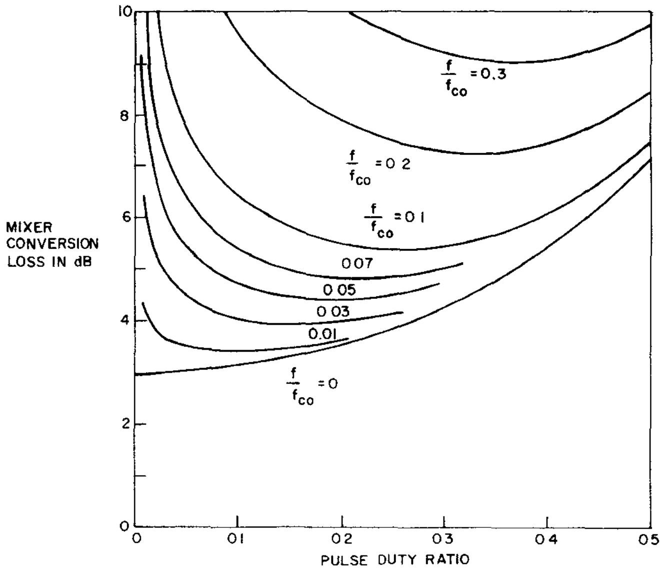

Fig. 3. Computed mixer conversion loss for the broad-band case. (Image termination equals signal termination.)

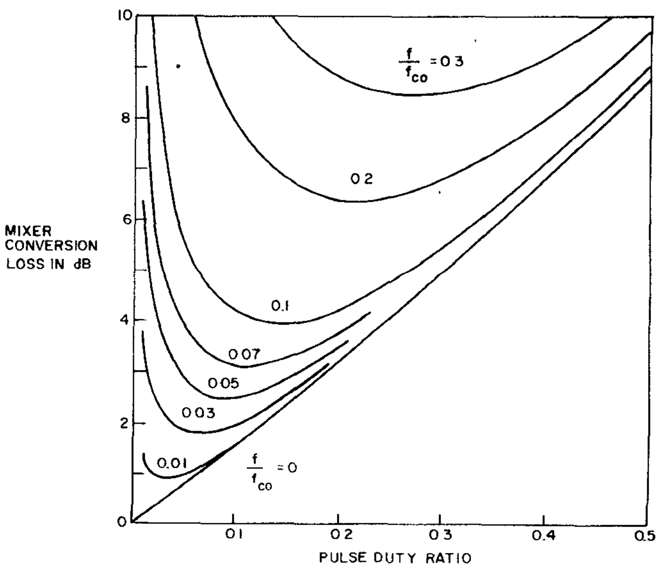

Fig. 4. Computed mixer conversion loss for the short-circuited image case.

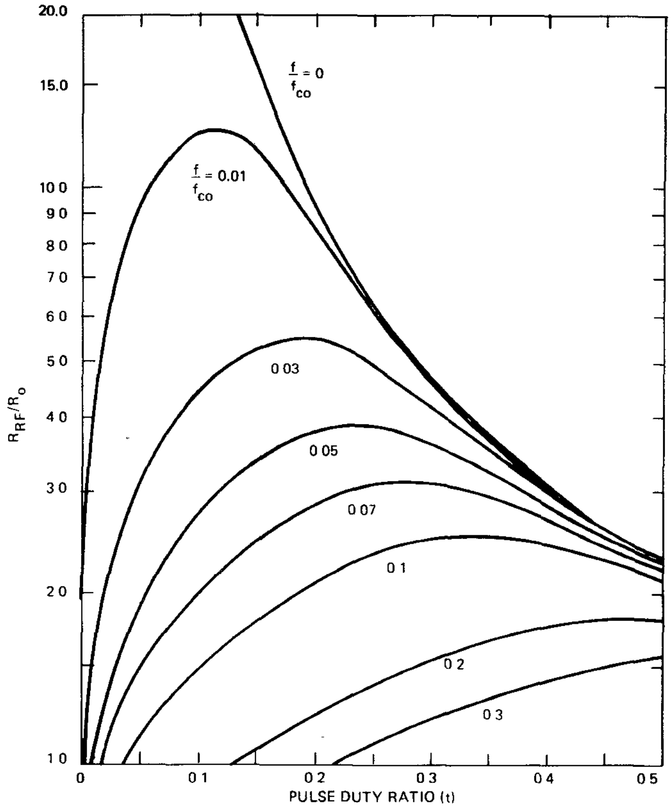

Fig. 5. Computed RF impedance for the broad-band case.

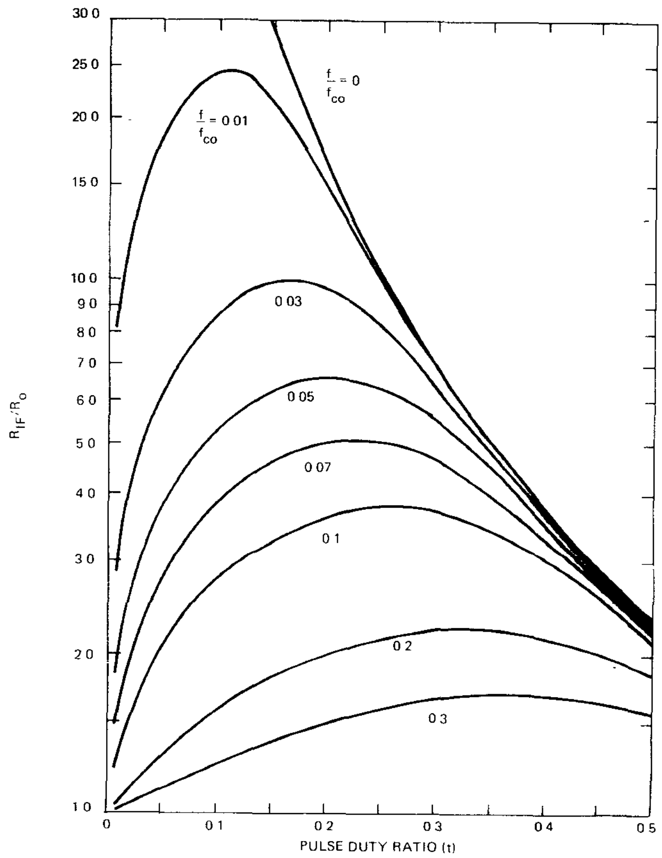

Fig. 6. Computed IF impedance for the broad-band case.

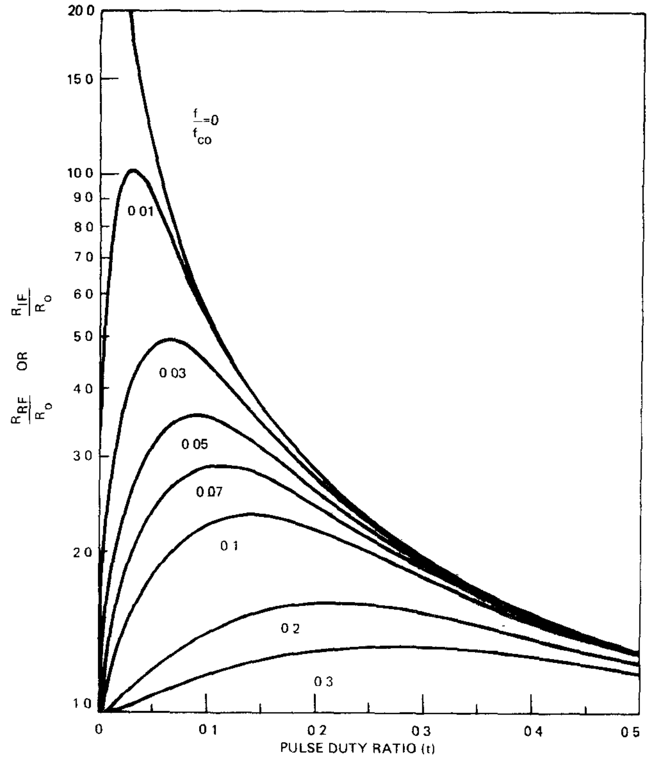

Fig. 7. Computed terminal impedance for the short-circuited image case.

This condition also results in an input impedance match. The RF and IF impedances have been computed as quantities normalized to R_0. R_0 is the average diode impedance and is well approximated by the simple expression

$$R_0 = R_s/t. \tag{1}$$

This is the time-averaged diode impedance and thus is the impedance presented to the LO.

The rectified dc current is a useful quantity for checking mixer operation and can be calculated by using (2).

$$I_{dc} \cong \left(\frac{Pt}{R_s}\right)^{1/2} \left[\frac{0.9 \sin(\pi t) - \sqrt{2}t \cos(\pi t)}{1 + t}\right] \tag{2}$$

where P is the LO power, t is the pulse-duty ratio, and R_s is the diode-limiting resistance.

The LO power can be estimated by (3).

$$P = \frac{t(V_0 - V_b)^2}{R_s[1 + \cos(2\pi t)]} \tag{3}$$

where V_0 is the forward potential drop of the Schottky barrier and V_b is the bias voltage. Typical values for V_0 are 0.75 V (GaAs), 0.5 V (Si-Schottky barrier), and 0.15 V (Si-point contact).

IV. RESULTS

Fig. 8 shows the implemented design which allows complete realization of the desired image-enchanced balanced mixer using GaAs Schottky-barrier diodes. This is a plan view of the microwave-integrated-circuit (MIC) mixer as viewed from the ground-plane side of the alumina substrate. The RF signal of frequency 9.3 GHz enters the substrate on the right edge via the microstrip (signal-input) port. The RF is coupled to the diodes via a broadband microstrip-to-slot-line transition and through a bandpass image-reject impedance-matching filter consisting of microstrip lines coupled to the slot line. The pair of mixer diodes terminate this filter.

The LO at a frequency of 7.8 GHz is injected via the LO-input microstrip terminal. The LO power then passes through the directional filter and to the mixer diodes by way of a microstrip-to-coplanar-line transition (pin through the substrate). Slot-line stubs at the end of the

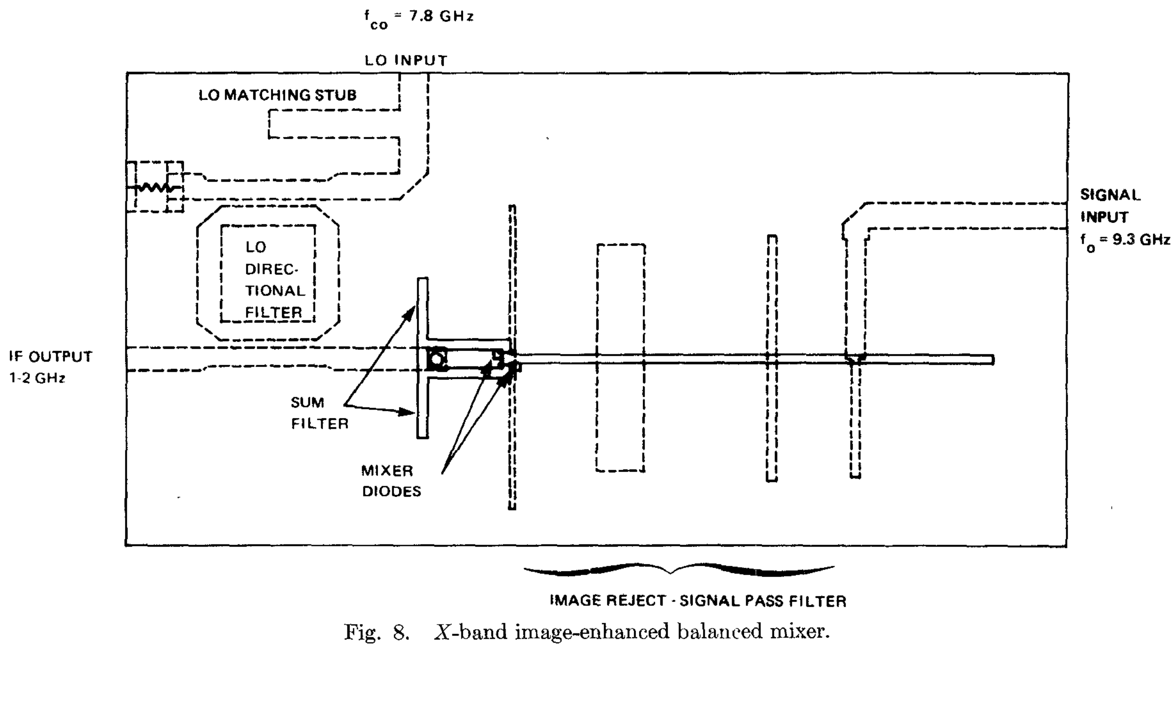

Fig. 8. X-band image-enhanced balanced mixer.

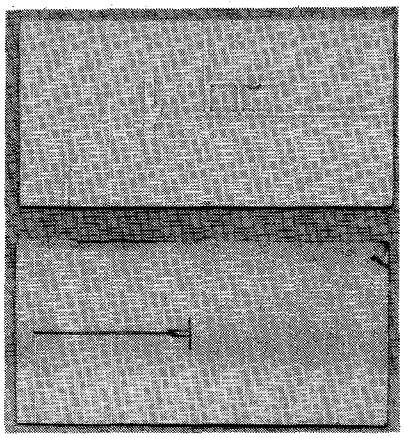

Fig. 9. Photograph of the X-band balanced mixer.

coplanar-line section present a short circuit to the mixer diodes at the sum frequency. The diodes are arranged such that the IF-output line is common with the LO input. The wide frequency separation of the IF and LO, and the directional- and frequency-selective properties of the directional filter allow very simple diplexing of the two signals with essentially zero bandwidth limiting of the IF port.

The two diodes are in parallel to the IF and LO ports, but in series to the signal port. As the conventional IF amplifier input impedance is nominally 50 Ω, the microstrip-to-coplanar-line characteristic impedance is set to 50 Ω. In a short-circuited image mixer the signal impedance is equal to the IF impedance. Thus the diode impedance must be 100 Ω, the two in parallel then matching the 50-Ω line. But the two in series will present 200 Ω to the signal slot line. The slot line can readily be made with a characteristic impedance of nominally 100 Ω. The 50-Ω microstrip to 100-Ω slot-line transition has well in excess of 10-percent bandwidth. The signal filter then is designed to supply the impedance transformation required to match the 200-Ω diode impedance to the 100-Ω slot-line impedance.

Reference to Fig. 4 will show that a conversion loss of under 2.0 dB can be achieved with diodes of $f_{co} \cong 800$ GHz and with PDR $\cong 0.10$. The signal-load impedance R_{RF} can be obtained from Fig. 7, and is found to be $R_{RF} = 5.5\, R_0$. The impedance to the LO port is given by (1), so that if $R_L = 100$ Ω and $t = 0.10$, then $R_{LO} = 18.0$ Ω. The LO line will then have a VSWR $\cong 5.5{:}1$. As the directional filter is narrow-band and transparent to the LO power, the LO matching stub can be placed as shown in Fig. 8 without loss of performance.

Using (1), and the fact that $R_L = 100$ Ω, one finds $R_s = 1.8$ Ω for $t = 0.10$. Assuming a diode frequency cutoff of 800 GHz, the junction capacitance is found to be $C_j = 0.11$ pF. A junction diameter of about 10 μm will yield this value of C_j with an attendant $f_{co} = 800$ GHz.

A computer program has been written which allows the analysis and optimization of microwave circuits with embedded mixer diodes. The diode model used was that which assumes that the Schottky diode can be represented as a junction switched at the LO rate. In the ON condition the resistance is that of the limiting series resistance; and in the OFF state it is the series resistance in series with the barrier capacitance. The zero-bias capacitance was used for the analysis and appears to give good correlation with measured results. In this computer routine a three-frequency analysis is used. That is, the diodes are represented by black boxes with terminals at the RF, IF, and image frequencies with couplings set by the 3×3 conductance matrix. All other frequencies generated are assumed to be short circuited. A nodal analysis routine is used to model the external circuitry at the three frequencies of interest and the diodes are added in parallel. An optimization program is then used to vary circuit values until the desired results are obtained. The computer results tracked the measured values within 0.4 dB over the band.

The signal filter loss was estimated to be ≈0.85 dB. The conversion loss was $L_c \cong 1.6$ dB. The microstrip-to-slot-line transition is no more than 0.05 dB so that an overall conversion loss for the complete mixer should be about 2.5 dB. An additional loss of about 0.2 dB must be added for the 3-mm coaxial connectors used to bring in the RF-input signal and to remove the IF in the test fixture. Thus the total expected conversion loss (band center) is expected to be 2.7 dB.

An estimate for the LO power is obtained from (3). It is obvious by inspection of Fig. 8 that the diodes are dc short circuited and thus $V_b = 0$. For the GaAs devices being used, $V_0 \cong 0.75$ V. Thus for the diode with $R_s = 1.8$ Ω, an LO power of about 17 mW is required to attain the PDR = 0.1. As two diodes are being used, an available LO power of ≈35 mW is required for full modulation and attainment of reasonable impedance levels.

A mixer embodying the previously described design has been built. A photograph of the substrate with diodes mounted (top and bottom views of the substrate) is shown in Fig. 9. The following characteristics indicate the advancement of the state-of-the-art performance obtained. The measured curve of conversion loss versus frequency is shown in Fig. 10. These data (Table I) represent all circuit losses including connector losses.

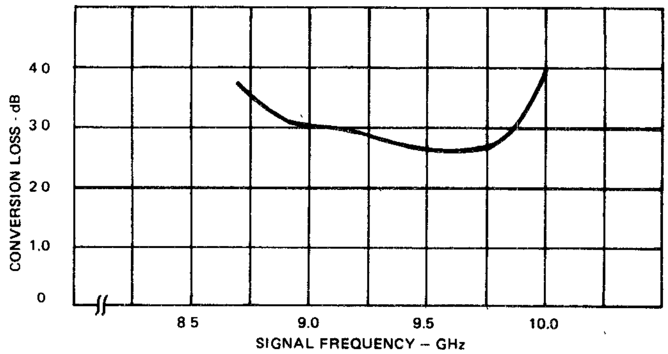

Fig. 10. Curve of conversion loss versus frequency.

TABLE I
BALANCED MIXER PERFORMANCE

Signal frequency (center)	9.4 GHz
LO frequency	7.8 GHz
Signal bandwidth	1.0 GHz
Conversion loss (1.0-GHz band)	≤3.15 dB
Conversion loss (0.5-GHz band)	≤ 2.8 dB
Conversion loss (best point)	2.6 dB
LO power	≈ 40 mW
Dynamic range (input signal for 1.0-dB compression)	+ 13 dBm
Image band isolation	> 25 dB
VSWR (across signal band)	< 1.4:1

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