Design and Performance Analysis of an Octave Bandwidth Waveguide Mixer

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Abstract—**A new broad-band mixer capable of operating over two full adjacent waveguide bands (18 to 26.5 GHz and 26.5 to 40 GHz) is described. Within the octave bandwidth from 20 to 40 GHz, the maximum conversion loss is 6.5 dB with a corresponding average DSB noise figure of 5.7 dB. A theoretical analysis is given to treat quantitatively the performance of the octave bandwidth waveguide mixer.**

I. Introduction

RECENT TRENDS in millimeter-wave activities have shown increasing interest in broad-band receivers for use in communications, meteorology, and electronic warfare systems. One of the major requirements for a broad-band receiver design is that its front-end mixer be sufficiently broad-band. The need for an ultra-broad-band mixer design becomes more obvious when used in an ECM environment where maximum probability of intercept is required. Full waveguide bandwidth mixers with respectable performance have been reported recently [1], [2], but their operating frequency band is limited to that of the waveguide band. This paper describes a new approach for the design of a waveguide mixer which has an operating frequency band exceeding an octave bandwidth. The novel approach[1] utilizing a crossbar mixer configuration in a double-ridge waveguide mount, hereafter referred to as the double-ridge crossbar mixer, provides both low-noise and broad-band characteristics [3].

This paper describes the design and performance of an octave bandwidth mixer covering the frequency range from 18 to 40 GHz. Following a general discussion on the requirements for achieving more than full waveguide bandwidth performance, a theoretical analysis is given to treat quantitatively the performance of the double-ridge crossbar mixer over a broad range of operating frequencies. Finally, the detailed design and performance results of the octave bandwidth mixer are presented.

II. General Discussion

In a conventional waveguide mixer design, the usable bandwidth is generally restricted by the bandwidth-limiting elements, such as the mixer diodes, RF choke sections, impedance matching sections, etc., but ultimately it is limited by the operating frequency range of the waveguide used. Ordinarily, mixer performance degrades considerably when operating with a bandwidth exceeding 10 to 15 percent. It becomes totally unusable when operating near the waveguide cutoff frequency because of unacceptably high losses. In order to achieve a low-loss octave bandwidth mixer design, a new configuration utilizing a crossbar mixer together with a double-ridge waveguide mount was evolved. This mixer design eliminates practically the critical bandwidth-limiting elements. Fig. 1 shows the pictorial representation of the mixer design. As shown, the two mixer diodes are connected across the ridges of the waveguide mount and the IF signal is extracted via a center crossbar through the sidewall of the ridged waveguide mount where the RF fields are at a minimum. This virtually eliminates the requirements of RF chokes. In addition, the double-ridge waveguide mount provides a relatively low waveguide impedance to match the diode impedance. No bandwidth-limiting quarter-wave impedance transformers are required, thus further enhancing the broad-band performance of the mixer design.

For the design of a broad-band mixer with a bandwidth exceeding the operating frequency range of a standard waveguide, it is necessary that the waveguide cutoff limitation be removed. As an example, for an octave bandwidth (e.g., 18 to 40 GHz) mixer design, the operating frequency range covers two adjacent waveguide bands; i.e., K-band (18 to 26.5 GHz) and K_a-band (26.5 to 40 GHz). It is not possible to use the standard K_a-band waveguide (WR28) for the mixer design since the theoretical cutoff of the WR28 waveguide is at 21.08 GHz. The use of a double-ridge waveguide mount extends the cutoff frequency for the TE_{10} mode to below 18 GHz. Specifically, for the standard double-ridge waveguide (WRD 180 C24), the cutoff frequency is lowered to 15.25 GHz. The crossbar mixer design approach utilizing a standard WR28 waveguide has demonstrated excellent RF performance covering the full K_a-band waveguide bandwidth [4]. A conversion loss as low as 4 dB and an instantaneous RF bandwidth of over 13 GHz (from 26.5 to 40 GHz) was achieved. The present design, combining the advantages of the crossbar mixer and that of a double-ridge waveguide mount, exhibits both low loss and extremely broad-band characteristics.

III. Theoretical Analysis

In a mixer design, the performance parameters of primary concern are the operating bandwidth and the conversion loss. Basically, the requirement for a broad-band mixer design is that its frequency response be uniform and, more particularly, its conversion loss be uniformly low over a wide range of frequencies. For the purpose of analysis, the conversion loss of a mixer is considered to consist of three losses, namely:

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[1] Patent pending.

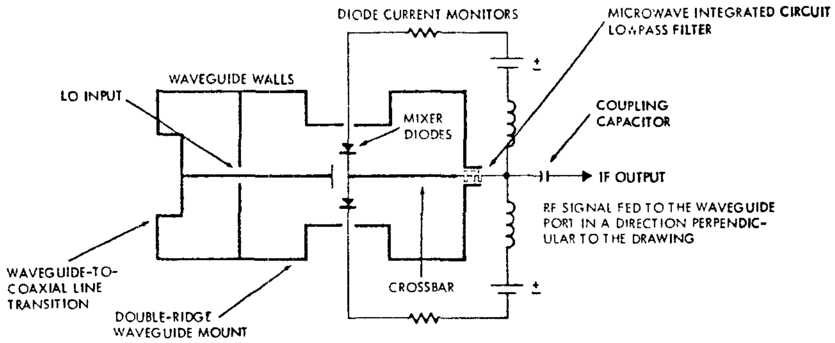


Fig. 1. Schematic diagram of the crossbar mixer.

L_1 mismatch loss due to impedance mismatches at the RF and IF ports

L_2 loss due to the diode junction capacitance and its series resistance

L_3 junction loss—the "intrinsic" loss at the diode junction which is dependent on the characteristic of the mixer diode and its terminating conditions.

Mathematically, these losses can be expressed as

$$L_1 = 10 \log (1 - |\Gamma_{RF}|^2)(1 - |\Gamma_{IF}|^2) \text{ dB} \qquad (1)$$

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

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

where

R_s diode series resistance

C_j diode junction capacitance

R_j diode junction resistance

ω operating frequency.

The "intrinsic" junction loss of a mixer depends on the terminating conditions of the image frequency and is not dependent on frequency. As an example, for the image matched condition, the junction loss is given as [5]

$$L_3 = \left[1 + \left\{\frac{1 + \frac{g_2}{g_0} - 2\left(\frac{g_1}{g_0}\right)^2}{1 + \frac{g_2}{g_0}}\right\}^{1/2}\right]^2 \left(1 + \frac{g_2}{g_0}\right)\left(\frac{g_0}{g_1}\right)^2 \qquad (3)$$

where g_0, g_1, and g_2 are the Fourier coefficients of the diode conductance. As seen from (1)–(3), the frequency-dependent losses are L_1 and L_2, while L_3 is frequency-independent. To achieve broad-band performance, i.e., low and uniform conversion loss over a wide range of frequencies, it is necessary that losses L_1 and L_2 be minimized and be uniformly low over the frequency range of interest.

For the purpose of analysis, a simplified equivalent circuit for the double-ridge crossbar mixer is developed, as shown in Fig. 2, assuming that the crossbar structure is transparent to the RF signals. The crossbar mixer is treated as two single-diode mixers connected in series with respect to the RF signals, except biased under different conditions.

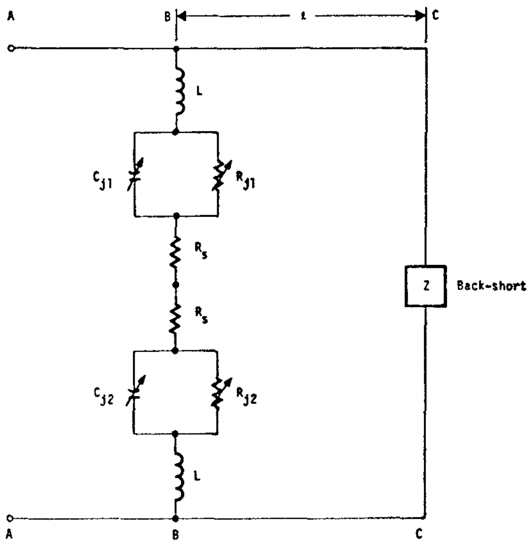


Fig. 2. Simplified equivalent circuit of the crossbar mixer.

Included in the equivalent circuit is a backshort section which is considered as an ideal contacting short located at a distance l from the diode pair.

The following abbreviations are used in Fig. 2.

Z backshort section impedance $= jZ_0 \tan (2\pi l/\lambda_g)$

Z_0 double-ridge waveguide characteristic impedance

λ_g wavelength in the double-ridge waveguide

l distance from diode pair to the backshort section

L whisker inductance

R_s diode series resistance

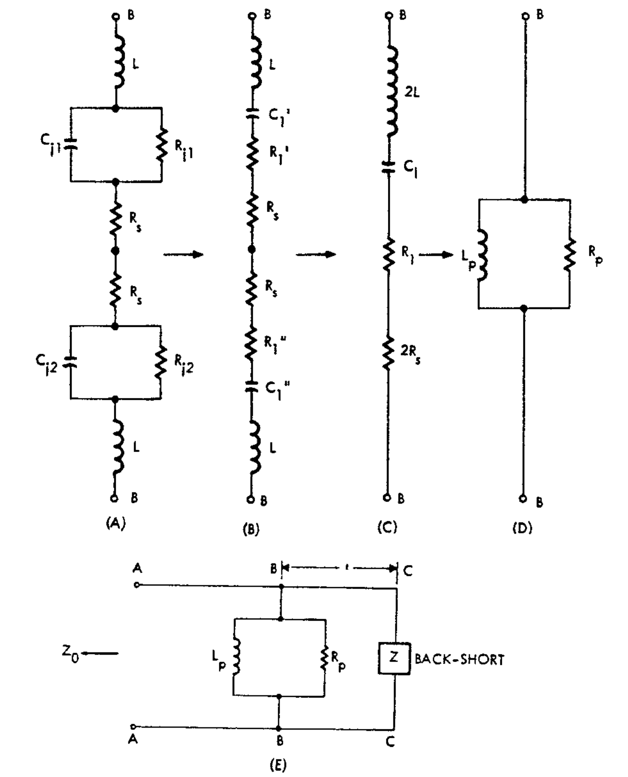
C_{j1},C_{j2} junction capacitance for diode 1 and diode 2 under different bias conditions

R_{j1},R_{j2} junction resistance for diode 1 and diode 2 under different bias conditions

For convenience, the whisker inductance and series resistance of the two diodes are considered to be equal.

The equivalent circuit of the crossbar mixer at plane B–B can be simplified by the impedance transformation as shown in Fig. 3.

By combining the equivalent circuit of Fig. 3(d) with the



where

$$R_1' = \frac{R_{j1}}{1 + \left(\omega C_{j1} R_{j1}\right)^2}; \quad \frac{1}{\omega C_1'} = \frac{1}{\omega C_{j1}\left[1 + \frac{1}{\left(\omega C_{j1} R_{j1}\right)^2}\right]},$$

$$R_1'' = \frac{R_{j2}}{1 + \left(\omega C_{j2} R_{j2}\right)^2}; \quad \frac{1}{\omega C_1''} = \frac{1}{\omega C_{j2}\left[1 + \frac{1}{\left(\omega C_{j2} R_{j2}\right)^2}\right]},$$

$$R_p = (2R_s + R_1' + R_1'') + \frac{\left[2\omega L - \frac{1}{\omega}\left(\frac{1}{C_1'} + \frac{1}{C_1''}\right)\right]^2}{2R_s + R_1' + R_1''}$$

$$\omega L_p = 2\omega L - \frac{1}{\omega}\left(\frac{1}{C_1'} + \frac{1}{C_1''}\right) + \frac{(2R_s + R_1' + R_1'')^2}{\left[2\omega L - \frac{1}{\omega}\left(\frac{1}{C_1'} + \frac{1}{C_1''}\right)\right]};$$

$$\text{for } \omega L > \frac{1}{\omega}\left(\frac{1}{C_1'} + \frac{1}{C_1''}\right)$$

Fig. 3. Equivalent circuit transformation.

backshort section, a new equivalent circuit for the crossbar mixer is obtained as shown in Fig. 3(e). It is seen that for impedance matching, the following conditions should hold.

$$R_p = Z_0 \tag{4}$$

and

$$\omega L_p = -Z_0 \tan \frac{2\pi l}{\lambda_g}. \tag{5}$$

It is recalled that the frequency-dependent losses are the impedance mismatch loss L_1, and the diode parasitic loss L_2, while the junction loss L_3 is frequency-independent. For the calculation of bandwidth limitation of a mixer design, only the frequency-dependent losses L_1 and L_2 will be considered.

In order to simplify calculations, we assume that the impedance mismatch at the IF port is a negligibly small constant. Thus L_1 becomes

$$L_1 = -10 \log (1 - |\Gamma_{RF}|^2). \tag{6}$$

Using the equivalent circuit of Fig. 4, it can be shown that the reflection coefficient is given as

$$|\Gamma_{RF}| = \left[\frac{\left(1 - \frac{Z_0}{R_p}\right)^2 + \left(\frac{Z_0}{\omega L_p} + \frac{1}{\tan 2\pi l/\lambda_g}\right)^2}{\left(1 + \frac{Z_0}{R_p}\right)^2 + \left(\frac{Z_0}{\omega L_p} + \frac{1}{\tan 2\pi l/\lambda_g}\right)^2}\right]^{1/2} \tag{7}$$

The mismatch loss L_1 is given as

$$L_1 = -10 \log \frac{4Z_0}{R_p\left[\left(1 + \frac{Z_0}{R_p}\right)^2 + \left(\frac{Z_0}{\omega L_p} + \frac{1}{\tan 2\pi l/\lambda_g}\right)^2\right]}. \tag{8}$$

The diode parasitic loss L_2 not only is frequency-dependent but also depends on the values of R_s, R_j, and C_j. For minimum losses, the following condition exists

$$\frac{\partial L_2}{\partial R_j} = 0 \quad \text{or} \quad R_j = \frac{1}{\omega C_j}. \tag{9}$$

By substituting (9) into (2), we have

$$L_{2\,\min} = 10 \log \left[1 + 2\frac{f}{f_c}\right] \tag{10}$$

where

$$f_c = \frac{1}{2C_j R_s}$$

where f is the operating frequency.

Based on the equivalent circuit model developed for the crossbar mixer, the mismatch and parasitic losses are calculated over an octave bandwidth of operating frequency range, i.e., 18 to 40 GHz as shown in the Appendix. The calculated results are tabulated in Table I.

For the fully conducting diode having the parameters as described in the Appendix, the average "intrinsic" junction loss L_3 for a mixer under an image matched condition is approximately 3.8 dB over the frequency range of interest [6]. Assuming that the IF mismatch loss is 0.5 dB, the total conversion loss L of the mixer (i.e., $L = L_1 + L_2 + L_3 + 0.5$ dB) is plotted against frequency as shown in Fig. 7(a). It is seen that the frequency response of the mixer is remarkably uniform over an octave bandwidth frequency range, clearly indicating the broad-band characteristic of the mixer design.

IV. Octave Bandwidth Mixer Design

In the design of a low-noise broad-band mixer, careful consideration must be given to reducing parasitic losses in the circuit components and eliminating critical bandwidth-limiting elements. The key components of primary concern in a broad-band mixer design generally include the mixer

TABLE I
MIXER MISMATCH AND PARASITIC LOSS

Frequency (GHz)	Mismatch Loss, L_1 (dB)	Parasitic Loss, L_2 (dB)		Total Loss (dB) $L_1 + L_2$
		Diode 1	Diode 2	
18	1.2414	.6946	.0811	2.0171
19	.8478	.6950	.0860	1.6288
20	.5718	.6955	.0913	1.3586
21	.3767	.6960	.0968	1.1695
22	.2398	.6965	.1025	1.0388
23	.1455	.6970	.1086	.9511
24	.0828	.6975	.1148	.8951
25	.0431	.6981	.1214	.8626
26	.0200	.6987	.1281	.8468
27	.0083	.6993	.1352	.8428
28	.0039	.7000	.1425	.8464
29	.0037	.7006	.1500	.8543
30	.0050	.7013	.1578	.8641
31	.0062	.7020	.1659	.8741
32	.0061	.7028	.1742	.8831
33	.0043	.7035	.1827	.8905
34	.0016	.7043	.1915	.8974
35	.0000	.7051	.2006	.9057
36	.0033	.7060	.2099	.9192
37	.0183	.7068	.2194	.9445
38	.0556	.7077	.2291	.9924
39	.1317	.7086	.2391	1.0794
40	.2710	.7095	.2494	1.2299

diodes, the impedance matching elements, and the RF choke section for the extraction of the IF signals. The performance of these components is frequency-limited—particularly the stepped impedance transformer and the RF choke section ordinarily used in the conventional mixer design, which seriously degrades the bandwidth performance of the mixer. In the present mixer design approach, a crossbar mixer together with a double-ridge waveguide mount is utilized. This eliminates the bandwidth-limiting circuit elements, such as the stepped impedance transformer and the RF choke sections, and provides an inherent impedance-matching condition for both the RF and the IF signals, resulting in a mixer design with octave bandwidth performance.

The basic requirements for designing a mixer with an operating frequency range over an octave bandwidth are as follows.

1) The waveguide components used for the fabrication of the mixer must have an effective usable bandwidth well exceeding an octave bandwidth.

2) The circuit elements, such as the mixer diode and its embedding network, must be sufficiently broad-band and must be properly matched to the waveguide impedance over the octave bandwidth frequency range.

Based on these requirements, an octave bandwidth mixer, utilizing a crossbar configuration in a double-ridge waveguide mount, was designed, fabricated, and tested. This mixer is capable of operating over the frequency range from 18 to 40 GHz. Fig. 4 shows the mechanical configuration of the mixer design where the mixer mount is built from the standard double-ridge waveguide mount (WRD 180 C24).

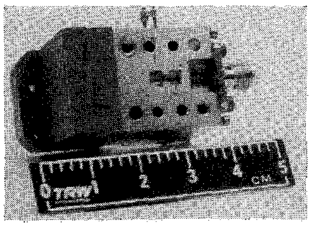


Fig. 4. Mechanical configuration of the crossbar mixer.

The octave bandwidth crossbar mixer consists of a double-ridge waveguide mount, a ridged waveguide-to-coaxial line transition, two Schottky-barrier diodes, and a low-pass filter, as shown in Fig. 1. The double-ridge waveguide mount also serves as the backshort housing of the mixer mount.

The heart of the mixer is a pair of low-noise GaAs Schottky-barrier diodes connected in series across the broad walls of the ridged waveguide. As shown in Fig. 1, one of the electrodes of each diode is connected to a metal crossbar which serves as a mechanical support for the diodes and also as a transmission line for the incoming LO power and the IF output signals. In actual operation, the RF signal is fed directly to the ridged waveguide port. On one end of the crossbar, LO power is fed via a ridged waveguide-to-coaxial line transition and capacitively coupled to the diodes. The IF output is extracted from the opposite end of the crossbar via a microwave integrated-circuit low-pass filter (LPF). The LPF was fabricated on sapphire substrate with a cutoff frequency exceeding 12 GHz. The use of the LPF at the IF port is to prevent LO power leaking to the output. Electrically, the two mixer diodes are connected in series with respect to the RF signal and in parallel with the IF output. This provides a higher impedance level to the RF signal and a lower impedance level to the IF signal than a single diode mixer. Therefore, an inherent impedance match condition for both the RF and IF signals is achieved for broad-band performance. In addition, the relatively low impedance of the ridged waveguide further improves impedance matching of the mixer diodes. This results in an extremely broad-band mixer design with uniformly low losses over an operating frequency range exceeding an octave bandwidth.

The other microwave circuit components essential for the octave bandwidth mixer design are the double-ridge waveguide mount and the ridged waveguide-to-coaxial line transition. These components must provide sufficient bandwidth to cover the operating frequency range, e.g., 18 to 40 GHz.

The double-ridge waveguide used for the 18 to 40-GHz mixer design has the dimensions shown schematically in Fig. 5.

The key parameters essential for a ridged waveguide design are:

1) the effective usable bandwidth defined by the cutoff frequencies of the TE_{10} and TE_{20} modes

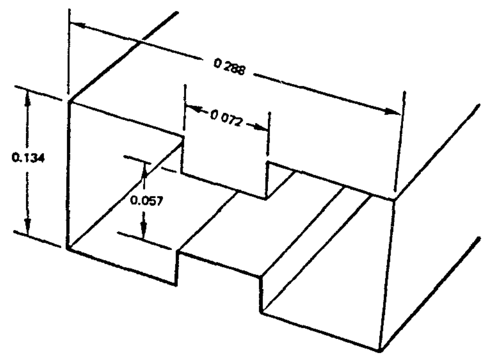


Fig. 5. Double-ridge waveguide (all dimensions in inches).

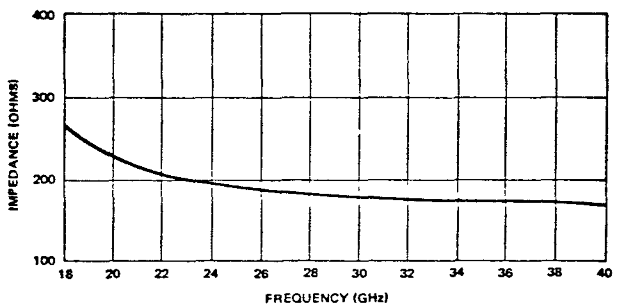


Fig. 6. Double-ridge waveguide impedance versus frequency.

2) the waveguide attenuation
3) the waveguide impedance.

According to Hopfer [7], with the dimensions shown in Fig. 5, the double-ridge waveguide has a usable bandwidth from 15.25 to 44.8 GHz, which is more than adequate for the 18 to 40-GHz operating bandwidth. The attenuation of the double-ridge aluminum waveguide is approximately 0.6 dB/ft at 18 GHz and 0.2 dB/ft at 40 GHz, clearly insignificant for this application.

The characteristic impedance for a double-ridge waveguide is given by Chen [8] as

$$Z_0 = \frac{Z_{0\infty}}{\sqrt{1 - \left(\frac{f_c'}{f}\right)^2}} \tag{11}$$

where

$Z_{0\infty}$ the characteristic impedance at infinite frequency for the TE_{10} mode
f_c' the TE_{10} mode cutoff frequency of the double-ridge waveguide
f the frequency of operation.

Fig. 6 shows a plot of the double-ridge waveguide impedance versus frequency. It is seen that the impedance level of a double-ridged waveguide is substantially lower than that of a standard K_a-band (26.5 to 40 GHz) waveguide (WR28), which varies from 443 to 623 Ω.

The ridged waveguide-to-coaxial line transition used in the 18 to 40-GHz mixer is a double-ridge aluminum waveguide construction (Model R45, MRC). This transition provides acceptable performance for the present design. Over the frequency range of 18 to 36 GHz, the typical VSWR is 1.3 : 1, with a maximum insertion loss of 1 dB within the 18 to 36 GHz range and 2 dB from 36 to 40 GHz. The higher insertion loss at the high end of the frequency band is not important for the present application because it is at the LO port. It will not affect the conversion loss of the mixer except that higher LO power is required to drive the diodes.

V. Mixer Diode Characterization

The Schottky-barrier diode plays the key role in the performance of the mixer. In fact, the ultimate bandwidth and noise performance of a mixer is determined by the quality of the diodes. It is, therefore, essential that the Schottky-barrier diodes be fully evaluated prior to their use in the mixer circuits. Evaluation is generally performed by measuring the "n-factor," R_s and f_c, of the mixer diode. The n-factor and R_s are defined by the I–V characteristics of a forward-biased diode as follows:

$$I = I_0 \exp \frac{q(V - IR_s)}{nkT} \tag{12}$$

where

I diode current
V applied voltage
I_0 diode saturation current
q electron charge
k Boltzmann's constant
T absolute temperature
n diode ideality factor
R_s diode series resistance.

It is seen from (12) that the n-factor and R_s can be determined by the I–V characteristic of the diode. The values of the n-factor and R_s are used to determine the dc quality of the mixer diode. Generally, for a good quality mixer diode, the n-factor and R_s should be less than 1.1 and 10 Ω, respectively.

The cutoff frequency f_c of a mixer diode is defined as

$$f_c = \frac{1}{2\pi C_j R_s} \tag{13}$$

where C_j is the diode junction capacitance. The cutoff frequency sets the ultimate limit of the mixer for use at high frequencies. It is desirable that the cutoff frequency be high for high-frequency applications. As can be seen from (13), for high f_c, both the C_j and R_s must be minimized. Minimization of C_j requires a small diode junction area as well as the use of low carrier concentration semiconductor materials. However, both of these (small junction area and low concentration material) increase the series resistance of the diode, which in turn lowers the cutoff frequency [see (13)]. To reach a compromise, an extremely thin epitaxial layer of moderately low carrier concentration is used. As an example, GaAs epitaxial layers with a carrier concentration of 6×10^{16} cm^{-3} and a thickness of less than 0.3 μm have been used successfully for the fabrication of Schottky-barrier diodes of 4-μm diameter. These mixer diodes have typical C_j's and R_s's of 0.01 pF and 5 Ω, respectively, under zero-bias conditions. This type of mixer diode has been used over the frequency range from 20 to 75 GHz with excellent performance.

VI. Performance Evaluation and Results

The bandwidth performance of a mixer is primarily restricted by the frequency-dependent elements associated with the mixer as well as its embedding network. In order to achieve ultra-broad-band and low-loss performance, considerations must first be given to minimizing the parasitic inductances, capacitances, and loss elements. These include the whisker inductance, the junction and package (if any) capacitances, and the series resistance. However, for a given mixer diode (i.e., C_j and R_s fixed), the ultimate broad-band performance of the mixer can be achieved by proper selection and implementation of the following critical bandwidth-limiting elements:

contact whiskers
adjustable backshort section
crossbar geometry.

An octave bandwidth mixer (Fig. 4) was designed and fabricated using GaAs Schottky-barrier diodes having $C_j = 0.01$ pF and $R_s = 5\ \Omega$. In the mixer design, performance optimization was achieved by precision design and fine tuning of every single element in the mixer circuit. The length of the contact whisker must be properly selected so that sufficient inductance will be provided for tuning out the parasitic capacitances of the mixer. A whisker length of approximately 0.02 in using a 0.001-in diameter wire was used in the present design. In addition, a contacting short of the plunger type was used to provide external tuning for the mixer. Ideally, the metal crossbar embedded in the waveguide should be transparent to the microwave signal and be lossless. In fact, for a properly designed crossbar, it can be considered as a microwave circuit with reactive elements only. It is, therefore, essential that the geometry and dimensions of the crossbar be properly designed such that its resonances do not occur within the operating frequency band of interest. In the present design, a 0.025-in diameter cylindrical crossbar of full waveguide width is used.

Evaluation of the octave bandwidth mixer was carried out using K-band (18 to 26.5 GHz) and K_a-band (26.5 to 40 GHz) test equipment with appropriate ridged waveguide-to-standard waveguide transitions. Fig. 7 shows the mixer conversion loss and noise figure performance over the 18 to 40-GHz range, measured with the mixer fixed tuned at 35 GHz. Also shown in Fig. 7 is the calculated conversion loss of the mixer obtained by the method and the equivalent circuit shown in Section III. It is seen that the measured conversion loss is in good agreement with the calculated value over the operating frequency range. The uniform distribution of the conversion loss over the frequency band fully demonstrates the broad-band characteristics of the mixer design. The double-sideband noise figure of the mixer was measured using an IF amplifier with a 1.5-dB noise figure. Over the octave bandwidth from 20 to 40 GHz, the measured minimum noise figure was 4.8 dB and the maximum 7 dB. It should be noted that from 22 to 40 GHz, the minimum and maximum noise figures are 4.8 and 5.7 dB. A differential of only 0.9 dB over a bandwidth of 18 GHz clearly indicated the excellent broad-band performance of the mixer design.

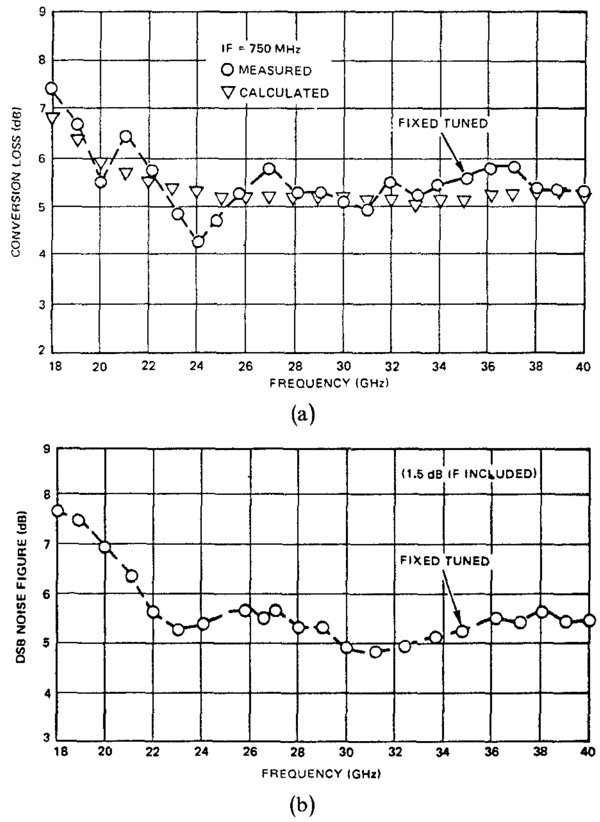


Fig. 7. (a) Octave bandwidth mixer conversion loss versus frequency. (b) Noise figure versus frequency.

VII. Conclusion

The performance of the octave bandwidth mixer is extremely encouraging, particularly, the bandwidth performance which exceeds an octave bandwidth. It was found that the most critical bandwidth-limiting parameter is the junction capacitance of the mixer diode which ultimately limits the operating bandwidth of the mixer design. An equivalent circuit model for the double-ridge crossbar mixer was developed, which treats quantitatively the bandwidth performance of the mixer over a wide range of frequencies. This circuit model provides a guideline for selecting the essential circuit parameters of the mixer design to achieve broad-band performance.

The design principle of the octave bandwidth mixer can definitely be extended to other frequency ranges by proper scaling. It is possible that the same design approach may be used for the design of mixers operating from 40 to 80-GHz and 80 to 160-GHz regions. This offers the technical feasibility of covering the 20 to 160-GHz frequency range with just three mixers.

Appendix
Calculation of Mismatch and Parasitic Losses of the Crossbar Mixer

To calculate the mismatch and parasitic losses of the mixer, we need to know the parameters, R_s's, R_j's, and C_j's of the mixer diodes. The exact values of these parameters cannot be determined without knowing the bias and local-oscillator drive conditions on the mixer diode. Since the local-oscillator drive level on the diode is not precisely known, in order to get an estimate of the value of these parameters, we make the following assumptions.

1) At each half cycle of the LO frequency, one of the diodes is fully conducting while the other one is partially conducting and being biased with a dc bias only.

2) Both of the diodes are operating under a small-signal condition such that

$$R_j = \left(\frac{\partial I}{\partial V}\right)^{-1} = \frac{nkT}{qI}, \quad \text{for } V > IR_s$$
$$= \frac{29}{I(ma)}, \quad \text{for } n = 1.1, \quad T = 298 \text{ K}. \quad \text{(A-1)}$$

Equation (A-1) is derived from the I–V characteristic of a mixer diode [see (12)].

The junction capacitance of the diode can be determined as follows [9]

$$C_j = C_{j0}\sqrt{\frac{\Phi}{(\Phi - V)}} \quad \text{(A-2)}$$

where

C_{j0} zero bias junction capacitance $= A\left[\frac{\varepsilon q N_d}{2\Phi}\right]^{1/2}$

A junction area

ε semiconductor permittivity $= 1.11 \times 10^{-12}$ F/cm for GaAs with $\varepsilon = 12.5$

q electron charge

N_d donor concentration

assuming that GaAs Schottky-barrier diodes are used for the fabrication of the crossbar mixer operating over the frequency range from 18 to 40 GHz. The Schottky barrier diode has the following characteristics:

carrier concentration	$N_d = 6 \times 10^{16}$ cm^{-3}
barrier potential	$\Phi = 0.9$ V
diode area (4 μm diameter)	$A = 1.25 \times 10^{-7}$ cm^2
series resistance	$R_s = 5\ \Omega$.

Under the condition of minimum conversion loss, the measured bias currents were typically 1 mA for the fully conducting diode and 50 μA for the partially conducting diode. Based on the I–V characteristic of the GaAs Schottky-barrier diode used, this corresponds to a bias voltage of 0.75 V for the fully conducting diode and 0.6 V for the partially conducting diode. Using the preceding bias conditions, the parameters of the two diodes are calculated as follows:

$$C_{j0} = A\left[\frac{\varepsilon q N_d}{2\Phi}\right]^{1/2} = 9.6 \times 10^{-15} \text{ F}.$$

For the fully conducting diode,

$$C_{j1} = C_{j0}\sqrt{\frac{\Phi}{(\Phi - V)}} = 9.6 \times 10^{-15}\sqrt{\frac{0.9}{0.9 - 0.75}}$$
$$= 2.35 \times 10^{-14} \text{ F}$$
$$R_{j1} = \frac{29}{1} = 29\ \Omega.$$

For the partially conducting diode,

$$C_{j2} = 9.6 \times 10^{-15}\sqrt{\frac{0.9}{0.9 - 0.6}} = 1.66 \times 10^{-14} \text{ F}$$
$$R_{j2} = \frac{29}{50 \times 10^{-3}} = 580\ \Omega.$$

By substituting R_s, R_{j1}, R_{j2}, C_{j1}, and C_{j2} into the equation shown in Fig. 3 at a fixed tuned frequency of 35 GHz, we have

$$R_1' = 28\ \Omega,\ C_1' = 1.07 \times 10^{-12} \text{ F}$$
$$R_1'' = 106\ \Omega,\ C_1'' = 2.03 \times 10^{-14} \text{ F}.$$

Using the impedance match condition, $R_p = Z_0$, (4), where Z_0 is the double-ridge waveguide impedance at 35 GHz, and solving for L, we have $L = 6.66 \times 10^{-10}$ h. Then, we calculate ωL_p using the equation shown in Fig. 3, i.e., $\omega L_p = 386$ Ω. The backshort distance l can be obtained by solving (5) for l as follows:

$$\omega L_p = -Z_0 \tan \frac{2\pi l}{\lambda_g}$$

where

$$\lambda_g = \frac{\lambda_0}{\sqrt{1 - \left(\frac{f_c'}{f}\right)^2}} = 0.94 \text{ cm at 35 GHz}.$$

Thus $l = 0.298$ cm.

Now with the mixer fixed tuned at 35 GHz, i.e., the backshort position is fixed, the mismatch and parasitic losses from 18 to 40 GHz are calculated. The calculated results are tabulated as shown in Table I. The parasitic loss is the sum of the losses contributed from the two diodes under different biased conditions.

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