Noise Figure and Conversion Loss of the Schottky Barrier Mixer Diode

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Abstract—**The theory of mixer operation is briefly reviewed and the results presented in graphical form convenient to the designer. In particular, the minimum noise figure, conversion loss, and the source and output impedances are plotted as functions of the pulse duty ratio of the diode current. Emphasis is placed on the pulse duty ratio as a more fundamental parameter for defining mixer operation than the magnitude of the diode voltage which is generally nonsinusoidal.**

It is shown that Schottky diode mixers should exhibit single sideband noise figures as low as 3 dB at *X* band when used in conjunction with 1.5 dB noise figure IF amplifiers, provided the diodes have cutoff frequencies higher than 500 GHz.

I. Introduction

CONSIDERABLE interest has recently been shown in the use of the epitaxial Schottky barrier (ESBAR) diode as a microwave frequency down-converter. For the first time it has become possible to surpass the performance of the redoubtable point-contact diodes, through the use of photoresist techniques to achieve small areas and epitaxial material to achieve low series resistance. The new diodes are also more reproducible, have much lower reverse current leakage, lower $1/f$ noise, and can be designed for much higher dynamic range. Fig. 1 compares two Schottky diode I–V characteristics with those of two well-known point contacts. It should be noted that Schottky characteristics can be exponential over a range seven decades of current thus making it feasible to carry out accurate mathematical computations in closed form. Herein it is shown that Schottky diodes should exhibit overall calculated noise figures as low as 3 dB at X band when the image is open circuited and the following IF amplifier has a 1.5 dB noise figure.

II. Mixer Noise and Gain Formulas

The theory for calculating mixer gain has been established for a number of years and has been applied to mixers using point-contact diodes by Herold *et al.*,[1] Torrey and Whitmer,[2] and Strum.[3] The gain and impedance formulas used in this paper are identical to those used by previous workers (except for minor corrections and additions) but the paper differs in that it considers the noise theory and also breaks away from the assumption of a sinusoidal voltage across the nonlinear diode conductance.

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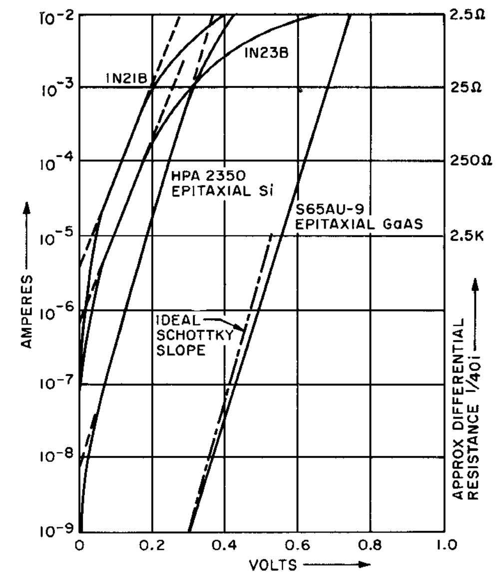

Fig. 1. Point contact and Schottky barrier I–V characteristics. The quantity i_0 in (3) for a particular diode is given by the intercept of the straight dashed line with the y axis. The value of α can be determined from its slope

$$\alpha = \frac{\Delta\ (\log_{10} i)}{\Delta V} \cdot 2.303.$$

The basic noise theory for diode mixers has been developed by Strutt,[4] Messenger and McCoy,[5] and summarized by Kim.[6] This theory neglects thermal noise generated in the diode series resistance and expresses the result in terms of ratios of the Fourier coefficients of conductance g and equivalent shot-noise current I_{eq} where these quantities are time varying due to the application of some arbitrary local oscillator voltage waveform across the diode.

$$g = g_0 + \sum_{n=1}^{\infty} 2g_{cn} \cos n\omega t. \tag{1}$$

$$I_{eq} = I_{e0} + \sum_{n=1}^{\infty} 2I_{en} \cos n\omega t. \tag{2}$$

The theory also neglects the effects of diode parasitic reactances, transit time phenomena, and conversion from frequencies near the higher harmonics of the local oscillator. It should be noted that in the range of voltages where the Schottky diode mixes most effectively the junction capacitance changes by almost 100 percent and only the average component can be tuned out exactly by the source admittance. The fundamental component will cause some reactive mixing and may or may not be a contributor to the conversion loss and/or gain. A general treatment of conversion by means of a nonlinear admittance has been considered by Edwards.[7]

Formulas for conversion loss, minimum noise figure, optimum source impedance, and the corresponding output impedance are presented in the Appendix. Three values of conductance as seen by the diode at the image frequency are considered, namely, a short circuit, an open circuit, and a conductance equal to the signal source. The open-circuited image gives noise and loss figures approximately half a decibel lower than those obtainable with a short circuit.

III. Application of Theory to the Schottky Barrier Diode

At low frequencies the equation relating current and voltage in the Schottky barrier diode is

$$i = i_0(e^{\alpha V} - 1).$$

The effects of diode parasitic reactances which influence the validity of this equation at high frequencies are discussed in Section IV.

The differential conductance is therefore given by

$$\begin{aligned} g &= \frac{di}{dv} = \alpha i_0 e^{\alpha V} \\ &= \alpha(i + i_0) \cong \alpha i \qquad i \gg i_0. \end{aligned} \tag{3}$$

Also, the mean-square shot-noise current is given by

$$\overline{i_{sh}{}^2} = 2qI_{eq}\Delta f$$

where

$$\begin{aligned} I_{eq} &= i + 2i_0 \\ &\cong i \qquad i \gg i_0. \end{aligned} \tag{4}$$

Since g and I_{eq} are simply related through the constant α, it is easy to show using (1) and (2) that

$$\begin{aligned} \frac{g_{c1}}{g_0} &= \frac{I_{e1}}{I_{e0}}, \\ \frac{g_{c2}}{g_0} &= \frac{I_{e2}}{I_{e0}}. \end{aligned} \tag{5}$$

Equations (5) permit a simplification of the noise theory for the Schottky diode, hence, all of the mixer formulas listed in the Appendix can be expressed as functions of the two ratios g_{c1}/g_0 and g_{c2}/g_0 which must approach unity if mixer loss and noise figure are to become small.

Throughout this paper it will be assumed that the peak forward current (i) greatly exceeds the reverse saturation current (i_0) and that the current versus voltage characteristic is purely exponential. From Fig. 1 it can be seen that (3), (4) and therefore (5) will become inaccurate when the maximum forward voltage is less than 0.1 volts thereby introducing errors in the noise calculations; this case is of little interest, however, because of the high noise figure. This conclusion is empirical and will depend on the care with which the junction is made, the material, and the junction area.

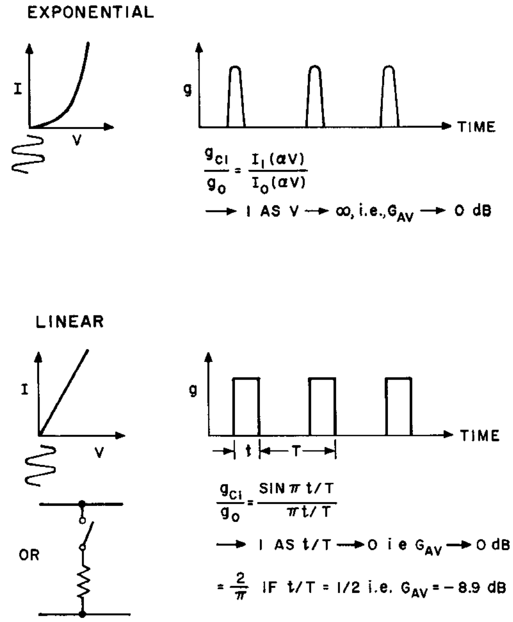

Fig. 2. Exponential and linear mixer conductance waveforms illustrating the definition of an effective pulse duty ratio. (G_{av} = available conversion gain.)

Case I. Sinusoidal Junction Voltage

If a sinusoidal oscillator voltage is assumed, which is certainly not the general case, it is possible to express V as

$$V = V_0 + V_1 \cos \omega t \tag{6}$$

where V_0 is the dc bias voltage. Substitution of this expression in (3) then defines the conductance waveshape

$$\begin{aligned} g(t) &= \alpha i_0 e^{\alpha V_0} e^{\alpha V_1 \cos \omega t} \\ &= \alpha i_0 e^{\alpha V_0}[I_0(\alpha V_1) + 2I_1(\alpha V_1) \cos \omega t \\ &\quad + 2I_2(\alpha V_1) \cos 2\omega t + \cdots] \end{aligned} \tag{7}$$

where I_0, I_1, I_2, etc. are modified Bessel functions with argument (αV_1). Hence from (1) and (7),

$$\begin{aligned} \frac{g_{c1}}{g_0} &= \frac{I_1(\alpha V_1)}{I_0(\alpha V_1)}, \\ \frac{g_{c2}}{g_0} &= \frac{I_2(\alpha V_1)}{I_0(\alpha V_1)}. \end{aligned} \tag{8}$$

The mixer gain and noise formulas can therefore be expressed as functions of the single variable (αV_1). Fig. 2 illustrates the conductance waveform which consists of a series of narrow pulses. The figure also shows the waveform for a conductance in series with an ideal switch. By determining the values of peak sinusoidal voltage and rectangular pulse duty ratio which give identical values of the ratio g_{c1}/g_0 in the two cases, it can be shown that these pairs of parameters will also give values of the ratio g_{c2}/g_0 for the two cases which are equal to within 3 percent or better for pulse duty ratios below 30 percent. It is therefore possible to define an effective rectangular conductance pulse duty ratio PDR

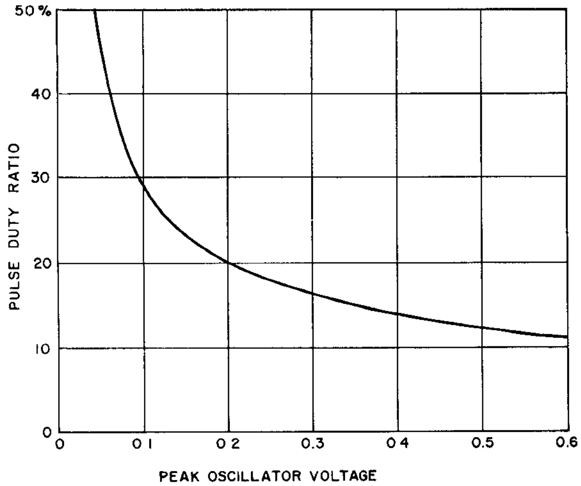

Fig. 3. Effective pulse duty ratio versus peak local-oscillator voltage.

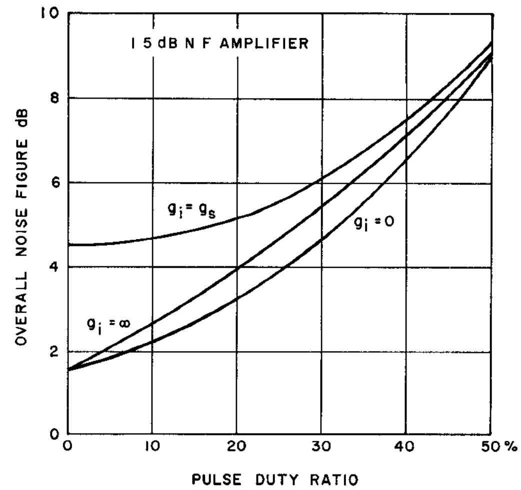

Fig. 6. Computed overall mixer noise figure including a 1.5 dB noise figure IF amplifier. This figure has been derived from Fig. 5 by means of the well-known cascade noise formula.

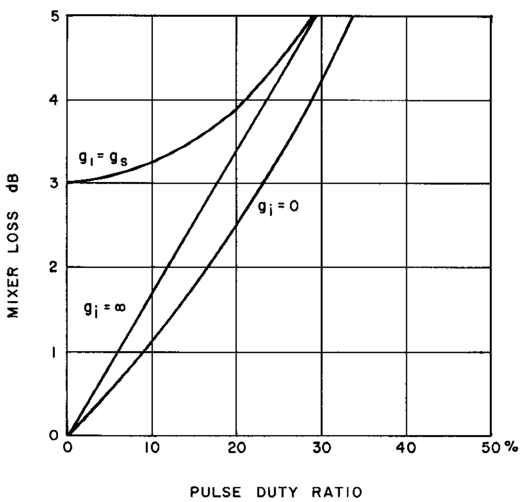

Fig. 4. Computed mixer conversion loss applicable to diodes with no series resistance.

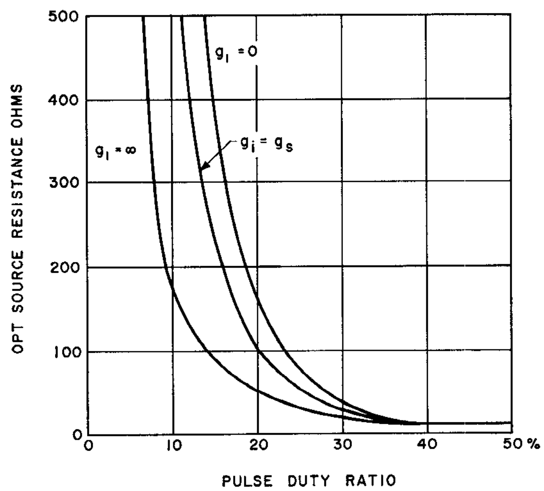

Fig. 7. Computed optimum source resistance for fixed $I_{max} = 10$ mA.

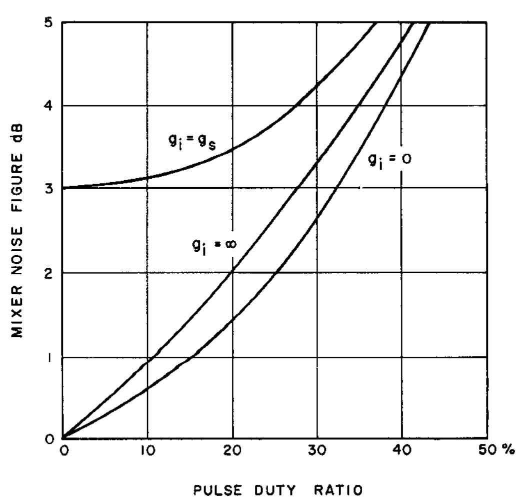

Fig. 5. Computed mixer noise figure applicable to diodes with no series resistance.

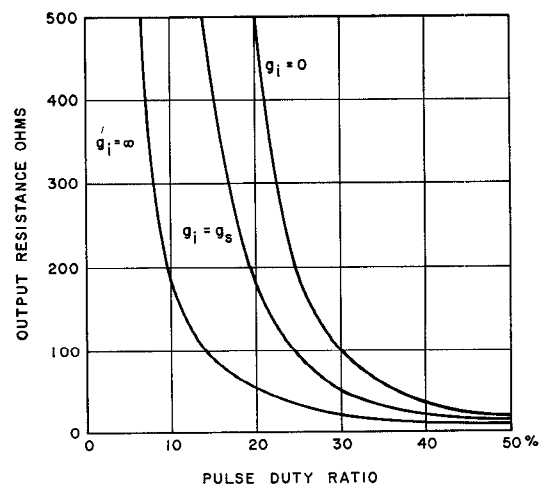

Fig. 8. Computed output resistance for fixed $I_{max} = 10$ mA and optimized source resistance.

$= t/T$ (see Fig. 3) which would result in mixer performance identical to that obtainable from an exponential device pumped by a specified sinusoidal voltage (V_1).

The pulse duty ratio is a very useful single parameter for defining mixer performance and can be shown to be equal to the width of the conductance pulse at 0.22 $g_{\max}$ relative to the interpulse period (T). Fig. 2 shows that the pulse duty ratio should be as small as possible if mixer loss and noise figure are to be reduced, a limit is set, however, by the source and output conductances which ultimately become unreasonably small.

Case II. Nonsinusoidal Junction Voltage

It has been found through low-frequency diode simulation studies, discussed in Section VI, that the conductance versus time waveform is always similar in shape to that occurring when a sinusoidal voltage exists across the Schottky diode junction. Typical conductance and voltage waveforms are shown in Fig. 13 for a case in which the voltage is quite nonsinusoidal. In such cases the effective pulse duty ratio adequately defines the conductance ratios and hence mixer performance. On the other hand, the peak local oscillator voltage or functions derived from that voltage, which were commonly used in the past,[2],[3],[8] have now ceased to be useful parameters.

Computed Results

Figs. 4 through 8 show the computed values of mixer loss, noise figure, and impedances plotted as functions of pulse duty ratio for three values of conductance (g_i) at the image frequency as seen by the diode.[1] In every case the source conductance (g_s) at the signal frequency has been chosen to minimize the noise figure. This condition also results in an input impedance match for all three cases of image loading.[2]

The source and output impedances have been computed assuming that the maximum forward current is 10 mA. It will be clear from Section IV that this is a reasonable assumption, however, for other values of maximum forward current the resistances should be multiplied by $10/I_{\max}$ in milliamperes.

The rectified dc current is a useful quantity for checking mixer operation and can be calculated using (3) and (7):

$$I_{dc} = I_{\max} \cdot e^{-\alpha V_1} \cdot I_0(\alpha V_1).$$

If $e^{-\alpha V_1} \cdot I_0(\alpha V_1)$ is plotted as a function of pulse duty ratio using Fig. 3, the result is found to be a straight line through the origin, hence

$$I_{dc} \cong 0.70 I_{\max} \cdot (\text{PDR}). \tag{9}$$

[1] The curve of loss versus pulse duty ratio for the case of $g_i = \infty$ has been presented previously by Strum.[8]

[2] From Figs. 4 and 5 and the Appendix it can be seen that the mixer noise ratio is less than unity.

The local oscillator power absorbed by the diode can also be calculated if a sinusoidal voltage is assumed and is generally less than half a milliwatt.[9]

IV. Effect of Series Resistance and Junction Capacitance

When the diode has finite series resistance r_s, then both the forward and the reverse impedance at microwave frequencies are affected. In the forward direction, increasing the pump voltage causes the junction conductance to rise until it finally reaches $g_{\max} = 1/r_s$. Beyond this point further drive only widens the effective relative conduction interval t/T (Fig. 9), which reduces the conversion conductance g_{c1}/g_0.

If the forward excursion is arbitrarily restricted to the point where the diode conductance rises to $g_{\max}/5$ ($\cong$10 mA for the GaAs ESBAR diode in Fig. 1) the only way to reduce the loss and noise figure with increased sinusoidal pumping (smaller pulse duty ratio) is to use reverse bias. It should be emphasized, however, that if the voltage swings too far in the negative direction the loss will also increase rapidly due to the series resistance and junction capacitance which lead to a residual diode conductance $g_{\min} = r_s \omega^2 c_j^2$ as shown in Fig. 9. This effect reduces the conversion conductance which now becomes

$$\frac{g_{c1}}{g_0} = \frac{\sin \dfrac{\pi t}{T}}{\pi \left(\dfrac{t}{T} + \dfrac{g_{\min}}{g_{\max}} \right)}$$

and, if the reduction amounts to 1 percent, the conversion loss of a typical mixer will have increased from about 3 to 3.3 dB. This increased loss can be avoided if

$$\text{PDR} > 100 \frac{g_{\min}}{g_{\max}} \cong 500 \left(\frac{f}{f_{c0}} \right)^2 . \tag{10}$$

If a diode is to be operated at 10 GHz with an overall noise figure of 3 dB, including a 1.5 dB IF amplifier, a glance at Fig. 6 shows that the pulse duty ratio must be less than 20 percent and hence from (10) the cutoff frequency must be greater than 500 GHz.

Note that large voltage excursions require the largest possible ratio of reverse to forward diode resistance (f_{co}/f). In general at microwave frequencies, this switching ratio is far lower than the low-frequency switching ratio (r_{rev}/r_s) and it is mainly for this reason that the capacitance of mixer diodes has always been kept very low. From this point of view mixer diodes should have very high cutoff frequencies; values in excess of 500 GHz have been obtained in the best GaAs varactors made by J. C. Irvin of Bell Telephone Labs.

The concept of residual diode conductance used above is very approximate since the junction capacitance is, in actuality, a time-varying quantity and the rules for changing from impedance to admittance are much more complicated than the one used here. This section obviously over-

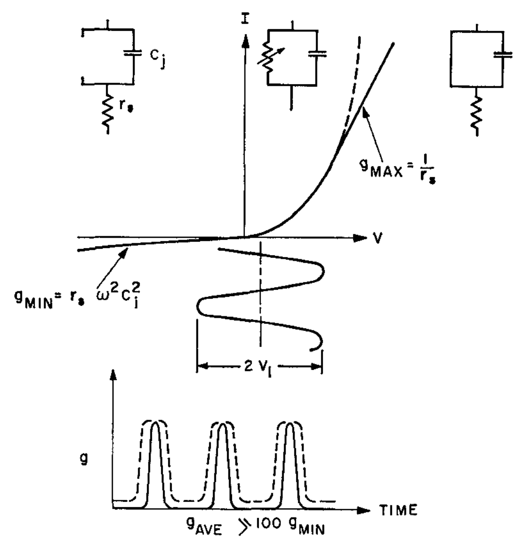

Fig. 9. Effect of diode parasitics on conductance versus time waveform.

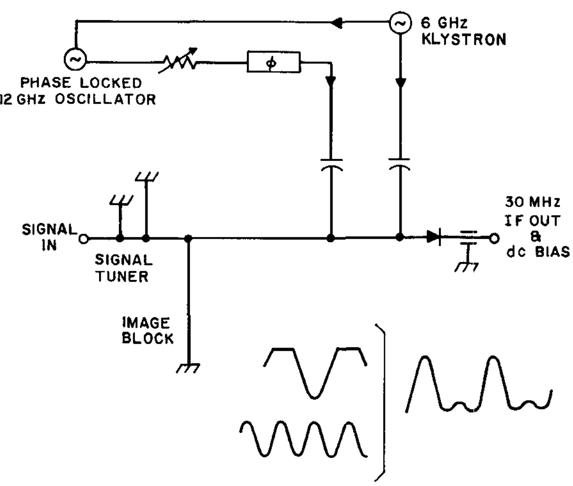

Fig. 10. Mixer circuit with inset illustrating principle of pump waveform improvement. The fundamental voltage will be truncated as shown when the diode resistance falls below that of the pump source.

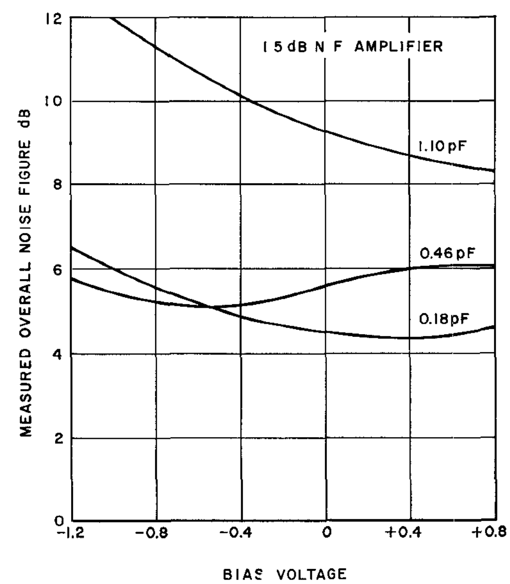

Fig. 11. Measured overall noise figure for conversion from 6 GHz to 30 MHz.

simplifies the actual situation and has been included simply to illustrate some effects on mixer performance which can be caused by parasitic reactances and changes in pulse duty ratio. The thermal noise from the series resistance could also be important, especially at low-noise figures and whenever the junction resistance becomes comparable to r_s.

V. Experimental Results

A series of measurements at 6 GHz have been carried out using the coaxial circuit shown in Fig. 10. The pump waveform could be modified by adding phase locked second harmonic in an attempt to reduce the conductance pulse duty ratio; the principle is illustrated in the Fig. 10 inset and was described originally by Strum.[8] The added complexity of this scheme was found to be unwarranted however, since the overall noise figure improvement was only 0.2 dB. The image block was positioned so that the reactive susceptance seen by the diode junction at the image frequency was intermediate between a short and an open circuit.

Some typical results without pump waveform improvement are plotted in Fig. 11 using GaAs Schottky barrier diodes having series inductances of the order of 0.5 nH and a variety of junction capacitances. It was found that the overall noise figure could be minimized by successive adjustments of bias, pump power, source impedance, and a π network between the mixer output and the 30 MHz IF amplifier; regardless of the order in which the adjustments were carried out, the same set of final conditions was always obtained.

For each diode it was found that the measured values of minimum noise figure, source impedance, output impedance, and the dc current corresponded to the same pulse duty ratio on each of the theoretical curves of Figs. 6 through 8. For example a diode with 0.46 pF junction capacitance exhibited a minimum noise figure of 5.1 dB and thus from Fig. 6 must have operated with a pulse duty ratio of 30 percent. The measured 35 ohms source impedance, 80 ohms output impedance, and 1.4 mA rectified current are in good agreement with Figs. 7 and 8 and equation (9) assuming an image susceptance intermediate between a short and an open circuit. In addition it was found that noise figures could not be lowered when attempts were made to reduce pulse duty ratios below 23 percent by increasing the reverse bias. This may have been due to the limitation imposed by (10) or to the effects of diode parasitic reactances described in Section VI.

The variations in noise figure behavior with forward bias for different diodes can be seen in Fig. 11. This behavior is attributed to the effects of diode parasitic reactances and is discussed in Section VI.

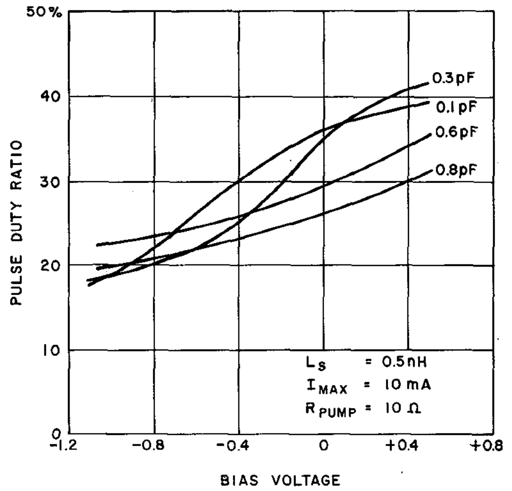

Fig. 12. Pulse duty ratio at 6 GHz measured using a simulated diode at 730 kHz.

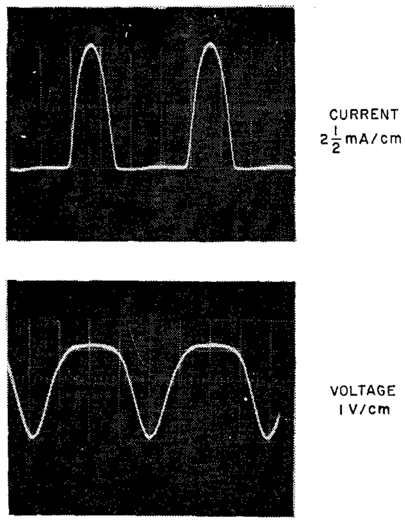

Fig. 13. Current and voltage waveforms at 6 GHz measured using a simulated diode at 730 kHz. $L_S=0.5$ nH, $C_j=0.3$ pF, $R_{\text{pump}}=10$ ohms, $I_{\max}=10$ mA, PDR = 35 percent, and $V_0=0$ volts.

VI. Low-Frequency Diode Simulation Studies

The 6 GHz voltage and current waveforms occurring at the Schottky junction can be studied at 730 kHz if the fixed series inductance and variable junction capacitance are scaled up 8200 times. The variation of junction capacitance with voltage obeys an inverse square-root law and can be simulated using one or more 12 voH breakdown Zener diodes, e.g., the Western Electric 426P whose capacitance is ~800 pF at −0.7 volts and becomes infinite at +0.3 volts. Fortunately the Zener diode capacitance increases more rapidly than its forward conductance which can therefore be neglected. The Zener and Schottky diodes must be separately biased with a constant difference (~0.7 volts) to ensure that the Zener capacitance becomes infinite at the appropriate bias on the Schottky diode which is ~+1.0 volts for the epitaxial GaAs units. Fig. 12 shows measured pulse duty ratios versus bias voltage for a diode with 0.5 nH of inductance when the maximum forward current reaches 10 mA. As expected from Fig. 3 the pulse duty ratio generally falls with reverse bias and increased voltage swing but significant differences occur with forward bias; in particular the simulated 0.1 and 0.3 pF diodes behave in a manner similar to the 0.18 and 0.46 pF diodes at 6 GHz shown in Fig. 11 when the limitations due to finite cutoff frequency [equation (10)] are excluded.

It was found that a reduction of inductance to 0.1 nH produced smaller pulse duty ratios similar to those for the 0.8 pF diode, for a wide range of diode capacitance. Tuning out the inductance with a series capacitor provided a negligible one percent reduction in pulse duty ratio.

The pulse duty ratio also varied with the resistance of the pump source and reached a maximum at 25 ohms which was 5 percent higher than for the 10 ohm source results shown in Fig. 12. Lowest pulse duty ratios occurred with low and high values of source resistance, e.g, 10 and 100 ohms; the use of a low resistance value has been advocated previously by Strum.[8]

It should be emphasized at this point that the external embedding impedance for the diode in the low-frequency simulation is probably very different in the microwave experimental setup. It is unlikely, for example, that the changes in pulse duty ratio produced by changes in diode inductance are real. Actually, the inductance becomes part of the external embedding circuit for the junction and can be tuned out at any number of single microwave frequencies, for example, all of the harmonics of the local oscillator.

When the Schottky diode was driven to forward currents in excess of 10 mA the pulse duty ratio increased at a rate of 1 percent per milliamp probably due to the effects of series resistance as discussed in connection with Fig. 9. Fig. 13 shows typical current and voltage waveforms measured in the simulated model. It should be reiterated that the current pulse retains its typical shape even when the voltage waveform becomes highly nonsinusoidal.

VII. Conclusions

It appears that the pulse duty ratio of the Schottky diode current waveform is a more fundamental parameter for defining mixer operation than the magnitude of an erroneously assumed sinusoidal junction voltage.

Future efforts should be directed at lowering the present 23 percent pulse duty ratio limit which would require diodes with cutoff frequencies in excess of 500 GHz. In addition, diodes should have series inductances smaller than 0.5 nH which would make the choice of junction capacitance between 0.2 and 0.6 pF relatively unimportant.

It may also be desirable to use pump sources with resistances appreciably greater or less than 25 ohms and to break away from the common procedure of matching the pump to the diode.

Following these criteria and designing the mixers with open circuit image impedances, it should be possible to obtain X-band SSB noise figures of 3 dB with reasonable output resistances (250 ohms) driving 1.5 dB IF amplifiers.

Appendix

The following equations were used to compute Figs. 4 through 8. The origins of the equations were discussed in Section II. In the figures, pulse duty ratio has been used as the independent variable in place of g_{c1}/g_0 and g_{c2}/g_0 in the equations. This new variable can be determined from (8) and Fig. 3.

Image Conductance Infinite

$$L = \left(\frac{g_0}{g_{c1}}\right)^2 \left[1 + \left\{1 - \left(\frac{g_{c1}}{g_0}\right)^2\right\}^{1/2}\right]^2$$

$$F = \frac{1}{2} + \frac{L}{2}$$

$$\left(\frac{g_{\text{source}}}{g_0}\right)^2_{\text{opt}} = \left(\frac{g_{\text{out}}}{g_0}\right)^2 = 1 - \left(\frac{g_{c1}}{g_0}\right)^2$$

Image Conductance Equal to Zero

$$L = \left[1 + \left\{\frac{1 + \frac{g_{c2}}{g_0} - 2\left(\frac{g_{c1}}{g_0}\right)^2}{\left(1 - \left(\frac{g_{c1}}{g_0}\right)^2\right)\left(1 + \frac{g_{c2}}{g_0}\right)}\right\}^{1/2}\right]^2 \cdot \frac{\left(1 - \left(\frac{g_{c1}}{g_0}\right)^2\right)\left(1 + \frac{g_{c2}}{g_0}\right)}{\left(\frac{g_{c1}}{g_0}\right)^2\left(1 - \frac{g_{c2}}{g_0}\right)}$$

$$F = \frac{1}{2} + \frac{L}{2}$$

$$\left(\frac{g_{\text{source}}}{g_0}\right)^2_{\text{opt}} = \frac{\left(1 - \frac{g_{c2}}{g_0}\right)^2}{1 - \left(\frac{g_{c1}}{g_0}\right)^2} \cdot \left(1 + \frac{g_{c2}}{g_0}\right)\left(1 - 2\left(\frac{g_{c1}}{g_0}\right)^2 + \frac{g_{c2}}{g_0}\right)$$

$$\left(\frac{g_{\text{out}}}{g_0}\right)^2 = \frac{1 - \left(\frac{g_{c1}}{g_0}\right)^2}{1 + \frac{g_{c2}}{g_0}} \cdot \left[1 - 2\left(\frac{g_{c1}}{g_0}\right)^2 + \frac{g_{c2}}{g_0}\right]$$

Image and Source Conductances Equal

$$L = \left[1 + \left\{\frac{1 + \frac{g_{c2}}{g_0} - 2\left(\frac{g_{c1}}{g_0}\right)^2}{1 + \frac{g_{c2}}{g_0}}\right\}^{1/2}\right]^2 \cdot \frac{1 + \frac{g_{c2}}{g_0}}{\left(\frac{g_{c1}}{g_0}\right)^2}$$

$$F = 1 + \frac{L}{2}$$

$$\left(\frac{g_{\text{source}}}{g_0}\right)^2_{\text{opt}} = \left(\frac{g_{c1}}{g_0}\right)^2 \cdot \left(1 + \frac{g_{c2}}{g_0}\right) \cdot \left[\left(\frac{g_0}{g_{c1}}\right)^2 + \frac{g_0 g_{c2}}{g_{c1}^2} - 2\right]$$

$$\left(\frac{g_{\text{out}}}{g_0}\right)^2 = 1 - \frac{2\left(\frac{g_{c1}}{g_0}\right)^2}{1 + \frac{g_{c2}}{g_0}}$$

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