Analysis and Optimization of Millimeter- and Submillimeter-Wavelength Mixer Diodes

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Abstract—**An analysis of the noise of millimeter- and submillimeter-wavelength mixers with GaAs Schottky diodes is presented. This analysis accounts for the correlation of the downconverted components of the time-varying hot-electron noise in the series resistance, and is thus accurate even for cryogenically cooled mixers operated at submillimeter wavelengths. This paper shows that the terms of the series-resistance noise correlation matrix are functions of the Fourier coefficients of the squared diode current $F(I^2)$ rather than the square of the Fourier coefficients of the diode current $[F(I)]^2$, as has been previously presented in the literature. The analysis is used to evaluate the optimization of cryogenic mixer diodes. It is shown that minimization of the diode's $I-V$ slope parameter V_o is more critical than reduction of the parasitic elements for millimeter-wavelength operation, while at frequencies above 600 GHz ($\lambda < 0.5$ mm), the junction capacitance is the most crucial parameter. Experimental results from several research groups working with a variety of mixers are presented to substantiate these results.**

I. Introduction

HETERODYNE receivers for submillimeter-wavelength astronomical observations have recently been developed [1]–[3]. The performance of these receivers is dependent on the quality of the nonlinear mixer element, usually a GaAs Schottky barrier diode. Optimization of the diode for use at these high frequencies will lead to an immediate improvement in the performance of the receivers.

A mixer's performance can be improved by increasing the nonlinearity of the diode's $I-V$ characteristic or by reducing the diode's series resistance R_s and junction capacitance C_j. However, these parameters are linked, and improvement of one inevitably leads to degradation of the other. Optimization of the diodes for millimeter- and submillimeter-wavelength applications requires an examination of the tradeoffs of these three parameters.

The basic principles of frequency conversion are well described by Torrey and Whitmer [4]. Following their work, many efforts have been made to predict the performance of real mixers with reasonable accuracy. Using the work of several earlier authors [4]–[7], Held and Kerr [8] developed numerical techniques which yielded quantitative agreement with experimental results [8], [9]. Their approach allows the consideration of arbitrary embedding impedances at the harmonics of the local oscillator (LO) and sideband frequencies, properly takes into account the correlation of the frequency components of the junction shot noise, and allows the use of any diode capacitance law. However, their analysis neglects the correlation of the frequency components of the hot-electron noise generated in the series resistance. This limits the accuracy of the technique in cases where the hot-electron noise substantially affects the mixer performance, namely in cryogenically cooled receivers and at submillimeter wavelengths. The importance of this noise has been realized by the present authors and, independently, by Hegazi *et al.* [10].

The accuracy of the diode equivalent circuit can also affect the mixer analysis. In an analysis of intrinsic conversion loss, McColl [11] assumed that the diode junction voltage and current could not exceed the "flat-band" values due to the large junction capacitance that occurs at flat-band.[1] That assumption led to predictions of conversion loss limitations for cryogenically cooled and small-area diodes which were significantly worse than have been observed in practice. These predictions have, however, been used in subsequent analyses of mixer performance [12], [13]. In a recent paper [14], it was shown that the restriction imposed by McColl, which limits the minimum diode resistance to a value significantly greater than the series resistance, is not realistic. With a model of the diode junction that allowed greater junction currents and a minimum diode resistance equal to R_s, the mixer conversion loss can be predicted more accurately. This paper also showed that the high value of junction capacitance that occurs near flat-band does not severely limit the performance of the mixer.

In the present paper we have two goals. First, we extend the earlier mixer analysis to include the periodically varying hot-electron noise. Our derivation corrects two fundamental errors in the analysis given in [10]. Second, we use the extended analysis and an accurate model of the diode, which allows the junction current to exceed its flat-band value, to investigate the optimization of GaAs Schottky

Manuscript received June 25, 1986; revised September 8, 1986. This work was supported in part by the National Science Foundation under Contract ECS-8412477.

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IEEE Log Number 8611629.

[1] The flat-band voltage is the junction voltage required to shrink the depletion layer width to zero. The flat-band current is the current at this bias.

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barrier mixer diodes for millimeter- and submillimeter-wavelength applications.

We start in Section II-A with the model of the Schottky diode including noise sources. The circuit model and the effect of electron heating are discussed in some detail. The dependence of the series resistance on the current is described with a previously developed model of the electron energy distribution.

In Section II-B, we incorporate the hot-electron noise in the mixer analysis, taking into account its time-varying nature.

In Section III, we use the mixer analysis to investigate the optimization of diodes for use in high-frequency receivers. This analysis demonstrates that at submillimeter wavelengths the R_sC_{jo} product is more important than the diode slope parameter V_o and that at frequencies greater than 600 GHz the most important parameter is the junction capacitance. Conversely, in the case of millimeter-wavelength receivers, V_o is shown to be the most important diode parameter.

Experimental results supporting these conclusions are discussed.

II. Mixer Analysis

A. The Schottky Diode

The equivalent circuit of a Schottky diode and the band diagram of a metal–semiconductor junction are shown in Fig. 1. The incremental diode junction capacitance can be expressed as

$$C_j = \begin{cases} S\left\{\dfrac{q\epsilon N}{2(V_{FB}-V_j)}\right\}^{1/2}(1+3x/d) & \text{if } V_j < V_{FB} \\ 0 & \text{if } V_j = V_{FB} \end{cases} \quad (1)$$

where

- S diode contact area $= \pi d^2/4$,
- q electronic charge,
- ϵ the dielectric constant of the semiconductor,
- N the semiconductor doping density,
- V_{FB} the flat-band potential (defined in Fig. 1(b)),
- V_j the voltage applied to the junction,
- x the depletion layer width,
- d the diode diameter.

This equation is derived from the standard depletion approximation [15] and includes the correction for the fringing capacitance given by Copeland [16]. As the forward voltage increases, the incremental junction capacitance increases substantially due to the decrease in the depletion layer width. When the junction voltage is equal to the flat-band voltage ($V_j = V_{FB}$), the depletion layer width is zero. Further increases in the diode bias cannot result in any additional charge being stored in the junction. Thus, at flat-band the junction capacitance is zero, and the diode equivalent circuit reduces to the series resistance.

The junction conductance current can be expressed as

$$I_j = I_{sat}\left[\exp(V_j/V_o) - 1\right] \quad (2)$$

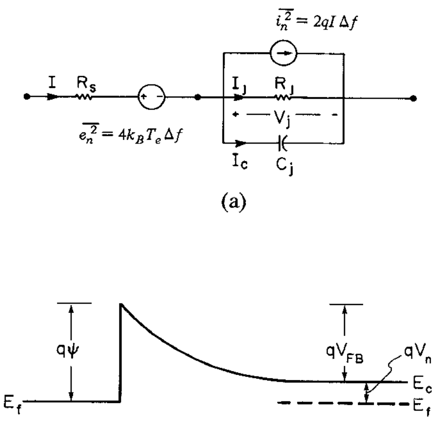

Fig. 1. (a) The circuit model of a GaAs mixer diode with noise sources. T_e is the effective noise temperature of the series resistance (see Section II). (b) The energy-band diagram of a metal–semiconductor junction at zero bias indicating the barrier height ψ, the flat-band voltage V_{FB}, the parameter $V_n = \psi - V_{FB}$, the conduction band edge E_c, and the Fermi level E_f.

where I_{sat} is the saturation current and V_o is the diode slope parameter, which is equal to $\eta k_B T/q$, where η is the diode ideality factor, k_B is Boltzmann's constant, and T is the temperature. For ideal thermionic emission, I_{sat} is given by

$$I_{sat} = A^*ST^2\exp(-\psi/V_o) \quad (3)$$

where A^* is the reduced Richardson constant (8.6 A/cm^2K^2 for GaAs) and ψ is the barrier height.

The small-signal junction resistance is $r_j = V_o/I_j$, and can be expressed as

$$r_j = \frac{V_o\exp\left[(\psi - V_j)/V_o\right]}{A^*ST^2}. \quad (4a)$$

If thermionic emission were the dominant conduction mode for all bias voltages, the junction resistance at flat-band would reduce to

$$r_{j,FB} = \frac{V_o\exp(V_n/V_o)}{A^*ST^2} \quad (4b)$$

where V_n is the potential difference between the conduction band edge and the Fermi level in the undepleted semiconductor (see Fig. 1). It is given by

$$V_n = \frac{k_BT}{q}F_{1/2}^{-1}\left\{\frac{n\sqrt{\pi}}{2N_c}\right\} \quad (5)$$

where $F_{1/2}^{-1}$ represents the inverse Fermi function of order one half. With these results, the minimum junction resistance of an ideal thermionic diode can be investigated. If we assume a temperature of 100 K, a diameter of 1 μm, a doping density of 2×10^{16}, and an ideality factor of unity, we have

$$V_o = 8.6 \text{ mV}$$

$$S = 0.8\times10^{-8} \text{ cm}^2$$

$$V_n = 12.3 \text{ mV}.$$

With (4b), these values yield $r_{j,FB} = 52\ \Omega$. Since this value is significantly larger than the minimum measured junction resistance of real diodes (which typically approaches zero), it is clearly not valid to assume ideal thermionic emission at voltages approaching flat-band, or to limit the diode current to the value that occurs at flat-band. In our example, this assumption has led to an increase of the minimum resistance of the diode of 52 Ω. This error has caused previous authors [11]–[13] to make rather pessimistic predictions of diode conversion loss.

Tunneling and hot-electron effects prevent real diodes from behaving as ideal thermionic emitters. Padovani and Stratton have shown that tunneling increases both V_o and I_{sat} [17], while hot-electron effects have been shown to cause V_o and I_{sat} to increase with forward current [18]. These effects reduce the junction resistance and allow the total resistance of real diodes to approach R_s at flat-band.

Surface effects can also cause variations in V_o. Verlangieri *et al.* [19] and Aydinli *et al.* [20] have shown that the surface preparation of the semiconductor just prior to anode contact formation has a strong effect on V_o. For optimal performance these surface nonidealities must be minimized.

While it is not possible to determine $V_o(I,T)$ and $I_{sat}(I,T)$ accurately from simple $I-V$ measurements, it is possible to develop a model of the diode with sufficient accuracy for mixer analysis. In this paper, we assume values of V_o obtained from curve fitting of measured data to (2), and values of I_{sat} which yield a small value of r_j at flat-band (0.5 Ω). These values of I_{sat} are slightly larger[2] than those obtained from curve fitting over the entire current range, but are most accurate near flat-band. These assumptions preserve the steepness of the $I-V$ characteristic and allow a minimum total diode resistance approximately equal to R_s. Also, this model does not incorporate an abrupt change in the diode resistance at flat-band (since $r_j \ll R_s$ at flat-band) and is thus believed to be more accurate than the model used in our earlier work [14], which assumed lower values of I_{sat}.

The Schottky diode's noise sources, shown in Fig. 1(a), are due to the shot noise in the junction and thermal noise in the series resistance. By Schottky's theorem, the mean-squared shot noise current through the junction is given by [21]

$$\overline{i_n^2} = 2qI_{eq}\Delta f \qquad (6)$$

where I_{eq} is the sum of the magnitudes of the forward and reverse currents in the junction, which, for mixer applications, is effectively the forward current since the reverse current is negligibly small in comparison. This equation assumes that the transit-time effects are negligible in the frequency range of interest.

The mean-squared thermal noise voltage of a resistance R at the temperature T is given by Nyquist's theorem as [22]

$$\overline{e_n^2} = 4k_BTR\Delta f. \qquad (7)$$

This assumes that the electron energies have a Maxwell–Boltzmann distribution at the temperature T, which is generally true for a nondegenerately doped semiconductor at low current densities. However, when the forward current is large, the electron distribution is heated and the noise will exceed that predicted by (7) [23]. For this reason, the equivalent noise temperature of the series resistance of a mixer diode will vary throughout the LO cycle.

To describe the time variation of the series resistance noise, we adopt the model of Hegazi *et al.* [10]. In their work, the effective noise temperature of the series resistance was expressed as

$$T_e = A(f)R_sI^2. \qquad (8)$$

Here, R_s is the resistance of the epilayer, I is the current through the epilayer, and $A(f)$ is a frequency-dependent coefficient.[3] $A(f)$ was determined from empirical data published by Keen and Zirath [24]. Although not all diodes exhibit a current-squared dependence, this model does represent a good first approximation for well-behaved diodes.

A theoretical basis for this model has been presented by Kollberg [25] and recently by Zirath [26]. This model, which is similar to that used by Sze [27] to investigate Gunn devices, makes use of the fact that the average electron in the epilayer must gain energy from the electric field at the same rate it loses energy to the lattice. These rates can be approximated as

$$r_{gain} = qEv_d \qquad (9)$$

$$r_{loss} = 3k_B(T_e - T)/(2\tau_e) \qquad (10)$$

where E is the electric field, v_d is the average drift velocity, τ_e is the average energy relaxation time of the electrons, and T_e is the electron distribution temperature that describes the rate of energy loss to the lattice. For the low-field case, we can express the forward current in terms of the carrier concentration n, the electron mobility μ, and the electric field, yielding

$$I = Sqn\mu E = Sqnv_d. \qquad (11)$$

Equating the two rates and using (11), we obtain

$$T_e = T + KI^2 \qquad (12)$$

where the noise constant K is given by

$$K = \frac{2\tau_e}{3k_Bq\mu n^2S^2}. \qquad (13)$$

If the electron energy distribution is nearly Maxwellian, the noise temperature of the epilayer resistance is ap-

[2] The assumed values of I_{sat} are typically less than a factor of four greater than those obtained from curve fits. This translates to a shift of the $I-V$ curve of about 10 mV (roughly 1.5 percent) for a typical cryogenic diode.

[3] For typical diodes, the epilayer resistance is the major component of R_s. In this example, the other components, such as the spreading resistance in the substrate and the contact resistance, are neglected. In general, these components should be treated as separate terms in the noise calculations.

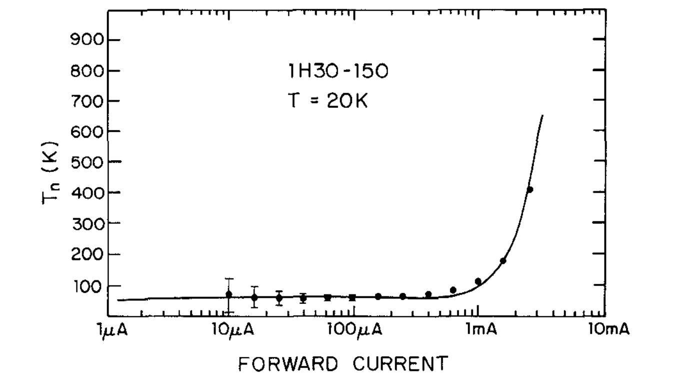

Fig. 2. Typical equivalent noise temperature of a dc-biased diode versus forward current. Experimental data (●) measured at 1.4 GHz and the noise predicted by (14) are shown. Nominal diode parameters ($N_d = 1\times 10^{17}\text{cm}^{-3}$, $\mu = 5000\ \text{cm}^2/\text{Vs}$, $d = 1.2\ \mu\text{m}$, and $K = 7\times 10^7\ \text{K/A}^2$) have been assumed. R_s, r_j and V_o were calculated from best fits of the diode's $I-V$ curve to (2).

proximated by T_e [23], yielding the current-squared noise dependence.

This analysis depends on the validity of (9), (10), and (11) and on the assumption of a near-Maxwellian distribution; thus it should be valid at low field strengths (less than ≈ 2 kV/cm). However, measured data for a well-behaved Schottky diode, shown in Fig. 2, show that this model yields good results throughout the range of currents typically used in mixer applications. In this figure, the noise temperature (measured at 1.4 GHz) of a dc-biased Schottky diode is plotted (●) versus the forward current and compared to the prediction of the diode noise theory (solid curve), given by [28]

$$T_n = \frac{qV_o}{2k_B}\frac{r_j}{R_s + r_j} + T_e\frac{R_s}{R_s + r_j}. \tag{14}$$

The parameters V_o, R_s, and r_j are determined from best fits to the diode $I-V$ data, and T_e is determined from (12). Nominal diode parameters and an energy relaxation time of 1 ps have been assumed. The curve fit is quite good; the error in the 0.5-mA range is most likely due to the uncertainty in R_s and the inaccuracy of the assumption of a Maxwellian distribution.

Unfortunately, not all diodes have well-behaved noise characteristics. Quite often the diodes exhibit excess noise over wide ranges of currents. This effect has recently been linked to stresses at the GaAs–SiO_2 interface [29] and to surface preparation [30]. Recent evidence also indicates that the hot-electron noise is decreased at high frequencies (>100 GHz) [24], [25], [31]. This effect may be explained by the presence of traps at the junction, transit-time effects, or, perhaps, intervalley scattering if the epilayer is thick enough. However, these effects are beyond the scope of this paper and will not be considered further.

B. The General Mixer Circuit

The mixer analysis developed by Held and Kerr [8] takes into account the embedding impedances seen by the diode at any finite number of sidebands and LO harmonics, the

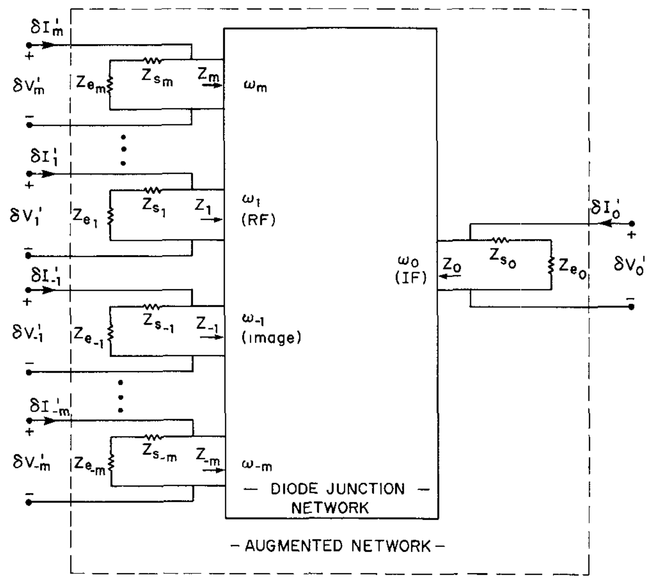

Fig. 3. The small-signal representation of the mixer as a multifrequency linear multiport network. The augmented network includes the embedding impedances and the series resistance, as well as the diode junction (r_j and C_j).

voltage-variable junction capacitance, and the correlation between the various frequency components of the junction shot noise. However, their analysis did not include the correlation of the frequency components of the series resistance noise. Although their assumption of a constant noise temperature in the series resistance has yielded excellent results for room temperature mixers for frequencies as high as 220 GHz [9], it is expected to be less accurate when there is significant heating of the electron distribution in the series resistance. This will occur in cryogenic mixers (due to the low ambient temperatures) and in high-frequency mixers at room temperature (due to the high current densities).

The general analysis, which is clearly developed in [8], will be outlined here to show its extension to include the correlation of the series-resistance noise.

The representation of the small-signal diode circuit as a multifrequency linear multiport network is shown in Fig. 3.[4] The diode series impedance Z_s is treated as a part of the embedding circuit, and the analysis is performed in terms of the "augmented" network enclosed within the dashed line. The small-signal currents at each frequency $\delta I_m'$ can be expressed in terms of the small-signal voltages $\delta V_m'$ by a system of equations of the form

$$\delta I_m' = \sum_{n=-\infty}^{\infty} Y_{mn}'\delta V_n', \qquad -\infty < m < \infty \tag{15}$$

or, in matrix notation, by

$$\underline{\delta I'} = \underline{Y'}\ \underline{\delta V'} \tag{16}$$

where an underline indicates a matrix quantity.

[4] The subscripts on the currents, voltages, impedances, and angular frequencies are consistent with the notation developed by Seleh [7]. In this notation, $\omega_m = \omega_{IF} + m\omega_{LO}$ and I_m, and V_m, and Z_m are the current, voltage, and impedance at the frequency ω_m.

The conversion loss from the RF source to the IF load can be expressed as

$$L = \frac{|Z_{e_0} + Z_{s_0}|^2 |Z_{e_1} + Z_{s_1}|^2}{4|Z'_{01}|^2 \operatorname{Re}(Z_{e_0}) \operatorname{Re}(Z_{e_1})} \tag{17}$$

where Z'_{01} is the 0,1 element of $\underline{Z}'$, the inverse of the $\underline{Y}'$ matrix, and Z_{s_m} and Z_{e_m} are, respectively, the diode's complex series impedance (including skin [8] and transient [33] effects) and embedding impedance at the frequency ω_m. Similarly, the single sideband mixer noise temperature can be expressed as

$$T_m = \frac{\langle |\delta V_{N_0}|^2 \rangle |Z_{e_1} + Z_{s_1}|^2}{4 k_B \Delta f |Z'_{01}|^2 \operatorname{Re}(Z_{e_1})} \tag{18}$$

where the ensemble average $\langle |\delta V_{N_0}|^2 \rangle$ is the equivalent mean-squared noise voltage at the IF port of the intrinsic diode which will deliver the same IF power to the load as all of the sources of noise in the actual diode. It now remains to calculate $\langle |\delta V_{N_0}|^2 \rangle$ from our model of the diode.

The noise generated by the Schottky diode can be separated into two distinct components, the junction shot noise and the noise generated in the series resistance. Because they are generated in physically separate regions of the diode, these noise components are assumed to be uncorrelated.

The noise voltage at the IF port can be calculated using the inverse of (16) provided all of the diode's noise sources can be represented by equivalent current sources in parallel with the diode junction. Since the shot noise sources are already in this position, they are readily considered. From [8], the mean-squared noise voltage at the output port due to the shot noise can be expressed as

$$\langle |\delta V_{S_0}|^2 \rangle = \underline{Z'_0} \langle \underline{\delta I'_S} \ \underline{\delta I'^{\dagger}_S} \rangle \underline{Z'^{\dagger}_0} \tag{19}$$

where $\underline{Z'_0}$ is the middle row of the $\underline{Z}'$ matrix, $Z^{\dagger}$ indicates the complex conjugate transpose of a matrix Z, and $\langle \underline{\delta I'_S} \ \underline{\delta I'^{\dagger}_S} \rangle$ is a square matrix whose (m, n) term represents the correlation of the components of the noise at the frequencies ω_m and ω_n.

The mean-squared noise current in the junction is given by

$$\langle \delta I^2 \rangle = 2 q I_j \Delta f. \tag{20}$$

Dragone [16] and Uhlir [5] have shown that if a local-oscillator voltage is applied to the diode so that the current I_j varies with time, the terms of the correlation matrix are related to the Fourier coefficients of I_j by

$$\langle \delta I_{S_m} \delta I^*_{S_n} \rangle = 2 q I_{j_{m-n}} \Delta f \tag{21}$$

where $I_{j_{m-n}}$ is the $(m-n)$th Fourier coefficient of the time-varying term in (20), in this case the diode current, and the superscript $*$ represents the complex conjugate.

The mean-squared noise current in the series resistance can be expressed as

$$\langle \delta I^2 \rangle = \frac{4 k_B \Delta f}{Z_s} T_e = \frac{4 k_B}{Z_s} (T_o + K I^2) \Delta f. \tag{22}$$

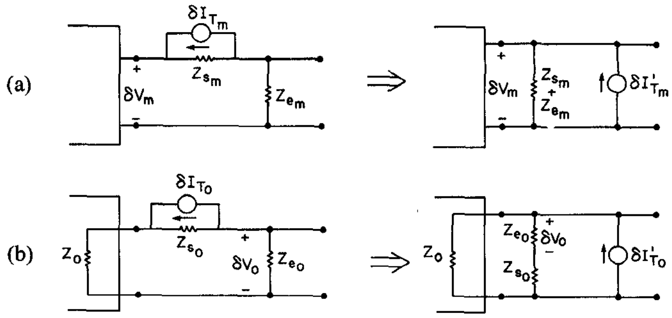

Fig. 4. The transformation of the hot-electron noise sources to equivalent noise sources in parallel with the diode junction. (a) The case where $|m| \neq 0$. (b) The special case of the IF output port $m = 0$. The transformations are such that the quantities δV_m (and hence the noise delivered to the IF load) remain constant for all m. The quantity Z_0 is the output impedance of the junction at the frequency ω_0 (see Fig. 3). The output port, in this case the IF, must be treated differently from the others.

The second term in the parentheses is time-dependent (because of the LO); thus, its frequency components are correlated. This equation can be treated in the same manner as (20), only here the constants are different and the time-dependent term, $(T_o + K I^2)$, is a function of the total diode current squared, rather than the first power of the diode current. The correlation between two frequency components of this noise is thus expressed as

$$\langle \delta I_{T_m} \delta I^*_{T_n} \rangle = \frac{4 k_B \Delta f}{\sqrt{Z_{s_m} Z^*_{s_n}}} \Theta_{m-n} \tag{23}$$

where the temperature Θ_{m-n} is the $(m-n)$th Fourier coefficient of the time-dependent term of (22), given by

$$\Theta_{m-n} = \frac{\omega_{LO}}{2\pi} \int_{-\pi/\omega_{LO}}^{\pi/\omega_{LO}} (T_o + K I^2) \times \exp[-j(m-n)\omega_{LO} t] \, dt. \tag{24}$$

Note that the Fourier integrand is a function of *the diode current squared* (the constant T_o can be taken from the integral) and *not the current*, as was assumed in [10].

The noise sources δI_{T_m} must be transformed into equivalent sources in parallel with the diode junction in such a way that their contributions to the voltage across the IF load (Z_{e_0}) remain unchanged. The transformations are depicted in Fig. 4. Note that the transformation at the IF is different from that at the other sidebands, a subtle point which was considered in [8] and [9], but neglected in [10].

The equivalent noise sources can be expressed in terms of the original sources as

$$\delta I'_{T_m} = \delta I_{T_m} \frac{Z_{s_m}}{Z_{s_m} + Z_{e_m}} \qquad \text{for } m \neq 0 \tag{25}$$

and

$$\delta I'_{T_0} = -\delta I_{T_0} \frac{Z_{s_0}}{Z_0}. \tag{26a}$$

If the IF load is conjugate matched $Z_0 = Z^*_{e_0} - Z_{s_0}$ and

$$\delta I'_{T_0} = \delta I_{T_0} \frac{Z_{s_0}}{Z_{s_0} - Z^*_{e_0}}. \tag{26b}$$

The transformed sources are equal to the original sources multiplied by a complex constant. Thus, the correlation of the sources is not affected (except for the constants), and the terms of the transformed matrix can be expressed as

$$\langle \delta I'_{T_m} \delta I^*_{T_n} \rangle = \frac{4k_B \Delta f}{\sqrt{Z_{s_m} Z^*_{s_n}}} \Theta_{m-n} \beta_m \beta_n^* \tag{27}$$

where Θ_{m-n} is given by (24), and

$$\beta_p = Z_{s_p} / \left(Z_{s_p} + Z_{e_p} \right) \qquad \text{when } p \neq 0 \tag{28a}$$

and

$$\begin{aligned} \beta_0 &= -Z_{s_0}/Z_0 \\ &= Z_{s_0} / \left(Z_{s_0} - Z^*_{e_0} \right) \qquad \text{for a matched load.} \end{aligned} \tag{28b}$$

The equivalent mean-squared noise voltage at the output port due to the noise in the series resistance can be expressed in terms of the correlation matrix, as in (19), and provided the shot and thermal noise are assumed to be uncorrelated, the total mean-squared noise voltage is expressed as

$$\langle |\delta V_{N_0}|^2 \rangle = \underline{Z'_0} \left\{ \langle \underline{\delta I'_S} \; \underline{\delta I'^{\dagger}_S} \rangle + \langle \underline{\delta I'_T} \; \underline{\delta I'^{\dagger}_T} \rangle \right\} \underline{Z'^{\dagger}_0}. \tag{29}$$

With this result, the mixer noise temperature can be found from (18) provided the diode parameters and the embedding impedances are known.

III. Diode Optimization

A. Introduction: The Circuit Description

The analysis given in Section II is valid for arbitrary embedding impedances. In reality, the embedding circuit can be expected to vary widely among the many different mixer mounts in use today. Since the goal of the present investigation is the optimization of the mixer diode, we will assume simplified embedding circuits which present real impedances to the diode at all of the sideband frequencies except for the IF, which is assumed to be conjugate matched. This simplification is justified by the fact that we are searching for major trends in mixer performance as the diode parameters are varied, and not for subtle effects which will depend on a precise knowledge of the mixer mount. The number of sidebands considered is limited by the amount of computer power available. As was pointed out by Held and Kerr [8] "truncation of the Y-matrix . . . implies that all higher sidebands are short circuited. This is likely to be a good approximation for millimeter wave mixers because the mean junction capacitance approaches a short circuit for very high frequencies." Siegel and Kerr [9] have reported that the use of six nonshorted sidebands (excluding the IF) has yielded sufficient accuracy. For the present work, we have also found that sidebands beyond the third harmonic of the LO do not significantly affect the results.

The LO source is assumed to supply a sinusoidal voltage ($V_{LO} \sin \omega_{LO} t$) to the diode (including R_s and C_j), and the matrix elements Y_{mn} are found from the solution of the nonlinear circuit. The method of solution has been described previously [14], and a general method for solving the LO circuit has been described by Siegel and Kerr [9], [32].

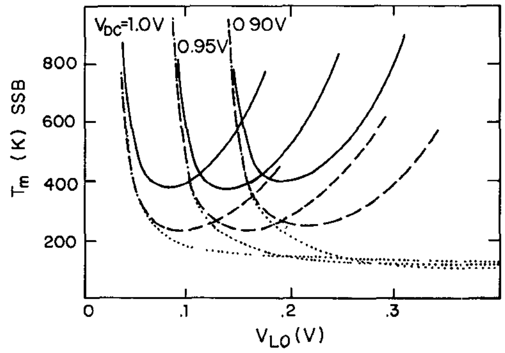

Fig. 5. T_M versus V_{LO} for three values of V_{dc} (as indicated) assuming a 1H30–150 diode at 20 K with an LO frequency of 300 GHz. The solid curves are for the complete noise theory developed here, the dashed curves are for the model developed in [8], and the dotted curves neglect the hot-electron noise entirely.

The diode equivalent circuit has been described in Section II-A. Following [8], [9], and [10], we assume that the series resistance is independent of the diode current and can be included in the embedding circuit. While the analysis of Section II-B can accommodate a frequency-dependent series resistance, in the present work we will assume that the series resistance is real and independent of frequency, thus neglecting skin effects and charge carrier inertia [33].

Although charge carrier inertia has a negligible effect at the frequencies considered here, the skin effect is believed to add 2 to 3 Ω to the series resistance at 100 GHz and up to 10 Ω at 1 THz (see, for example, [8]). We make this simplifying assumption because the skin effect resistance will not vary significantly as the other diode parameters (V_o, C_j, and R_s(epi)) are varied. Incorporation of these effects will yield somewhat larger values of noise temperature; thus, geometrical techniques to minimize these components of R_s are very important [34]. However, this simplification of the analysis will not significantly affect the general trends to be investigated here.

B. Theoretical Results

In this section, we use the mixer analysis to investigate the relative importance of the three diode parameters R_s, C_{jo}, and V_o. These parameters are varied independently to study their effects on the mixer performance. We have chosen to link the junction capacitance to the noise constant K (defined in (12) and (13)), since any effort to minimize C_{jo} will increase K.[5] In reality, all of the diode's electrical parameters are linked through the physical properties of the diode, and cannot really be varied independently. With this in mind, we will examine the very important tradeoffs between the $R_s C_{jo}$ product and V_o and between R_s and C_{jo}.

[5] We assume that C_{jo} is varied by changing the diode diameter. Thus, a reduction of C_{jo} by a factor of two will increase K by a factor of four.

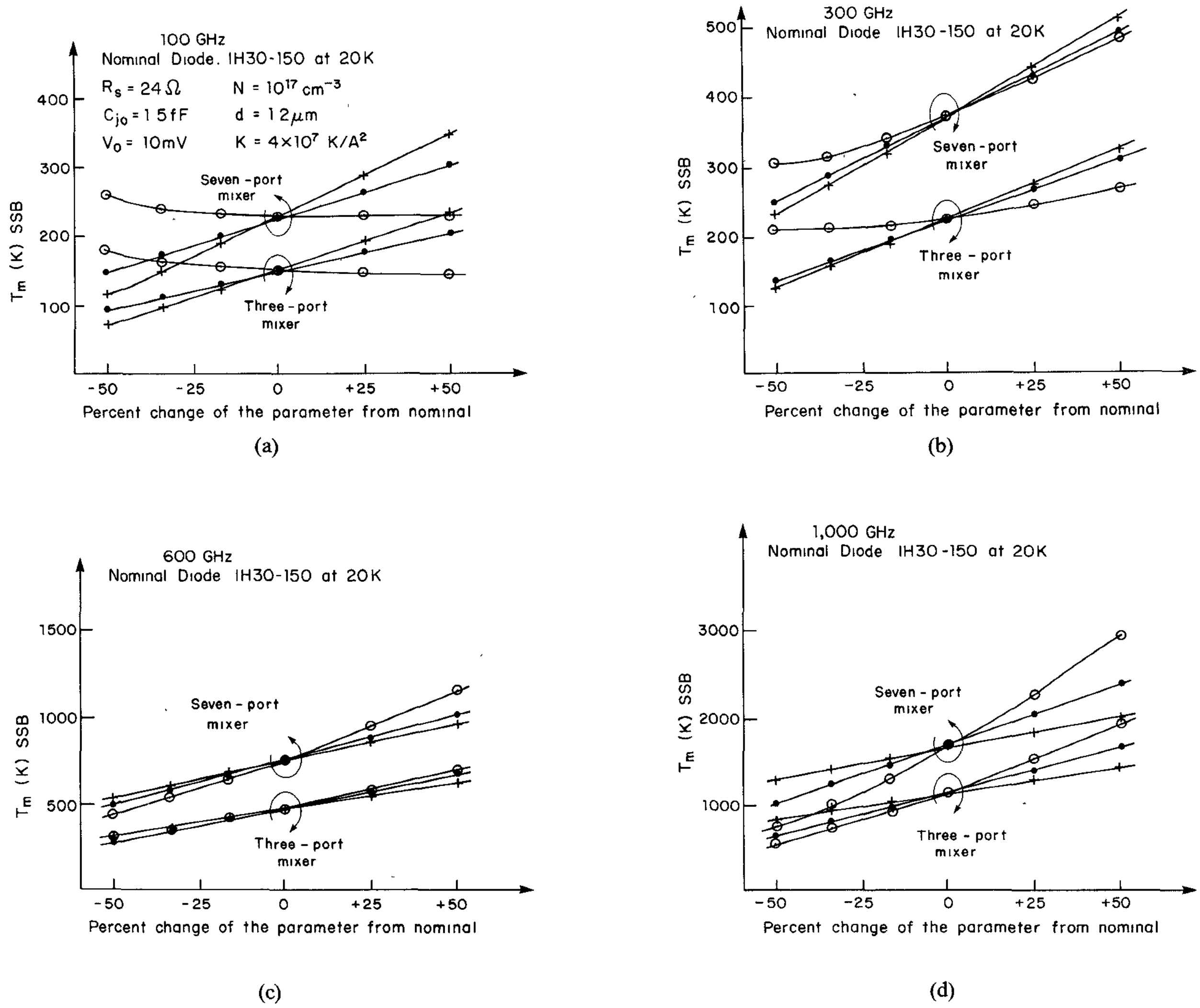

Fig. 6. The minimum predicted mixer noise temperature of a 1H30–150 diode at 20 K as the individual diode parameters are varied from their nominal values (listed in (a)). The LO frequency is 100, 300, 600, and 1000 GHz in parts (a), (b), (c), and (d), respectively. Results for two different circuits (described in the text) are shown. Note that the losses due to the increase in the series resistance at high frequency (skin effect) have not been included.

Fig. 5 shows the predicted mixer noise temperature versus LO voltage amplitude V_{LO} for a 300-GHz mixer at 20 K with $Z_{e_m}=100\ \Omega$ for $0<|m|\leqslant 3$ and a matched IF load. The three curves are for different dc bias voltages V_{dc}, as indicated. The diode parameters are those of a typical GaAs mixer diode (UVa type 1H30–150), and are listed in Fig. 6(a).[6] The solid curves represent the results of the complete noise model, the dotted curves show the noise if the hot-electron effects are neglected (i.e., $T_e=T\neq f(t)$), and the dashed curves assume the noise model of [8] and [9] (i.e., $T_e=T_{\mathrm{avg}}\neq f(t)$). The inclusion of the hot-electron noise with correlation has a significant effect on the minimum T_M and the optimum V_{LO}.

The mixer noise temperature is plotted as a function of the diode parameters for LO frequencies of 100, 300, 600, and 1000 GHz in Fig. 6(a)–(d), respectively. Results for two embedding circuits are shown in each figure. The first circuit is the same as that assumed in Fig. 5, while the second is a three-port mixer with $Z_{e_m}=100\ \Omega$ for $|m|=1$, $Z_{e_m}=0$ for $|m|>1$ and a matched IF load. For each individual curve, one diode parameter is varied independently while the other two are held at the nominal values for a 1H30–150 diode (listed in Fig. 6(a)). These results show that the three-port mixer yields substantially better performance at all frequencies, thus indicating the importance of the embedding circuit to the mixer performance. However, the trends of the mixer performance versus the diode parameters are the same for both circuits. Thus, exact knowledge of the embedding circuit is not necessary for the optimization of the diode.

The results in Fig. 6 show that all three diode parameters are of vital importance, and that their relative importances vary greatly with frequency. At 100 GHz (Fig. 6(a)), V_o is the most critical parameter, followed by R_s. The junction capacitance is virtually insignificant; in fact, reduction of C_{jo} below its nominal value actually degrades the mixer performance. This is caused by the increase in the noise constant K, which becomes important when C_{jo}

[6] Note that the assumed series resistance is significantly greater than the expected skin effect resistance, even at 1 THz.

TABLE I
REPORTED RECEIVER PERFORMANCE IN THE MILLIMETER- AND SUBMILLIMETER-WAVELENGTH RANGES

Freq. (GHz)	Temp. (K)	$T_M(K)$ SSB	L(dB) SSB	$T_{rec}(K)$ SSB	Diode Data*					
					Type	R_s (Ω)	C_{jo} (fF)	R_sC_{jo} (ΩfF)	$V_o(mV)$ at $1\mu A$	Ref.
100	20	70	6.2	124	2P9-300	12	6.5	78	8	[37]1984
110	20	81	5.7	—	2P9-300	14.7	8	120	9.5	[38]1985
230	20	230	6.0	330	1H2	19	1.8	34	12	[39]1985
256†	21	504	8.2	—	2P9-300	16	7	110	9.0	[40]1984
266	20	275	6.4	—	1E2	10	4	40	16.5	[41]1985
350	20	—	—	900	1H2	19	1.8	34	12	[42]1986
460	20	—	—	1,740	1H10-150	(15)	(2.5)	(38)	(11)	[43]1986
693	20	—	—	3,200	1H11-250	(21)	(2.3)	(48)	(11)	[42]1986
693+	300	—	—	4,850	1E7	(12)	(1.3)	(16)	(31)	[44]1985
803	300	—	—	4,200	1E7a	(20)	(0.8)	(16)	(32)	[44]1986
2,500	300	12,500	18	17,000	1E3	(12)	(1.5)	(16)	(31)	[45]1984

*Methods of measuring R_s, C_{jo}, and V_o vary among laboratories. Values shown here are those reported in the references when available or are approximate values measured in our lab for diodes of the same type (in parentheses).
†Best result in this frequency range with a 2P9–300 (high R_sC_{jo} product) diode.
+Best result at 693 GHz with a room-temperature receiver.

is very small. At 300 GHz (Fig. 6(b)), the situation is similar, except that the three parameters are of more nearly equal importance. At 600 GHz (Fig. 6(c)), the situation is reversed, with C_{jo} being the most important parameter and V_o the least. Finally, at 1 THz (Fig. 6(d)), the junction capacitance is the most important parameter by a wide margin and V_o is the least.

With a real Schottky diode, it is not possible to reduce any of the three diode parameters without affecting the others. However, certain physical properties of the diode can be chosen to reduce one parameter at the cost of increasing another, in order to optimize the overall mixer sensitivity. Our theoretical results yield important guidelines for diode optimization. For example, reducing the diode diameter reduces the junction capacitance and increases the series resistance, while maintaining a nearly constant R_sC_{jo} product. At high frequencies (above about 600 GHz), it is beneficial to reduce the capacitance as much as possible. (The lowest capacitance currently achievable in our lab is 0.7 fF, roughly half that of the 1H30–150-type diodes.) At low frequencies (< 300 GHz), the capacitance is not so important and it is beneficial to increase the diameter to reduce R_s.

It is also possible to reduce the R_sC_{jo} product, at the cost of an increase in V_o, by increasing the doping density and reducing the thickness of the epilayer. The minimum possible value of the R_sC_{jo} product will yield the best performance at high frequencies (above 600 GHz), while the minimum value of V_o is best below 300 GHz.

C. Comparison to Experimental Results

Our predictions are supported by experimental measurements of mixer performance at various frequencies by a variety of researchers. The most sensitive Schottky diode heterodyne receivers at frequencies from 100 GHz to 2.5 THz are listed in Table I together with the parameters of the diode used. All of these diodes have been developed and fabricated at the University of Virginia. The best results at 100 GHz were achieved with a diode that was optimized for the lowest possible value of V_o (8.5 mV at 20 K). This diode (2P9–300) has a relatively large R_sC_{jo} product (78 Ω fF at 20 K). The best results from 300 to 600 GHz were obtained with diodes having larger values of V_o, but significantly improved R_sC_{jo} products. At 2.5 THz, the highest frequency considered, the best mixer performance was achieved with a diode designed for the minimum possible R_sC_{jo} product (18 Ω fF) and the smallest zero-biased junction capacitance available at that time (approximately 1.5 fF). These results, which were obtained in different receivers with vastly different mixer mounts and a variety of diodes, clearly demonstrate the trends which are predicted by our analysis.

The relationship between the receiver noise temperature and the mixer noise temperature should be discussed briefly. The noise temperature of a heterodyne receiver can be expressed in terms of T_M as [35]

$$T_R = T_M + LT_{IF} \tag{30}$$

where L is the conversion loss of the mixer and T_{IF} is the noise temperature of the IF amplifier. Values of T_{IF} vary from about 70 K for a typical room-temperature amplifier, to less than 10 K for a cryogenically cooled amplifier [36]. It is important to consider whether the conversion loss will have a significant effect on the trends in T_R, and thus affect the diode optimization. The conversion loss at the dc bias point and LO voltage amplitude that yielded the minimum T_M is plotted in Fig. 7 versus the diode parameters for the three-port mixer at 100 and 1000 GHz. At 100

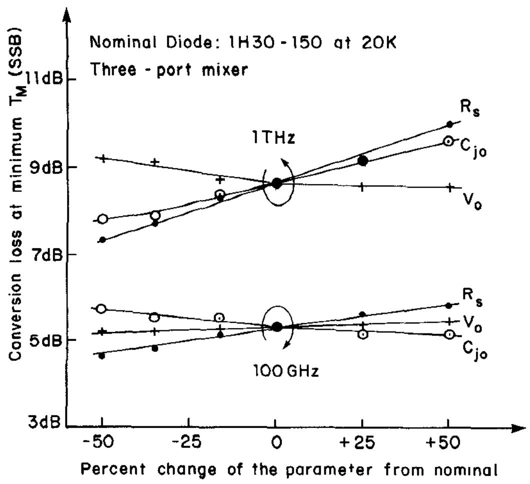

Fig. 7. The predicted conversion loss [14] of a 1H30–150-type diode at 20 K for the V_{dc}, V_{LO} values that yield the minimum T_M versus the three diode parameters considered in this study. Results at 100 GHz and 1 THz are shown, and the three-port mixer circuit defined in Section III-B is assumed.

GHz, the range of L is roughly 1.3 dB (ratio of 1.35). Therefore, if $T_{IF}=10$ K, the variation of T_R due to LT_{IF} is about 13 K. At 1 THz, the range of L is about 2.7 dB (1.9 ratio), and even if T_{IF} is as high as 70 K, the variation of LT_{IF} is only 130 K. In both of these cases, the quantity LT_{IF} varies much more slowly than T_M. Thus, our analysis of T_M can be used to accurately optimize the diode parameters for the best receiver performance. However, in special cases where the LT_{IF} is more significant, this term must be taken into account. This is a fairly trivial task provided the noise temperature of the IF amplifier is known.

IV. Conclusions

The GaAs Schottky barrier mixer diode is the key element of most receivers operating in the millimeter- and submillimeter-wavelength ranges. Improved sensitivity will be achievable if the diodes are properly designed for the frequency of operation. Of critical importance are the tradeoffs between the R_sC_{jo} product and the diode slope parameter $V_o=\eta k_B T/q$ and between R_s and C_{jo}. These tradeoffs have been investigated by a numerical analysis of the mixer circuit. The analysis used is similar to that of previous studies [8], [9], but it has been extended to include the correlation of the frequency components of the noise in the series resistance. The correlation occurs due to the time-varying hot-electron noise. Our treatment shows that the terms of the series-resistance noise correlation matrix depend on the Fourier coefficients of the diode current squared, and not the diode current to the first power, as has been previously assumed [10].

Through the use of the mixer analysis and an accurate model of the diode junction, we have shown that a significant decrease in the R_sC_{jo} product, at the expense of an increase in V_o, can yield a significant improvement in the performance of submillimeter-wavelength receivers. Furthermore, for a given R_sC_{jo} product, the values of R_s and C_{jo} are very important. At high frequencies (above about 1 THz), it is beneficial to minimize C_{jo} as much as technologically possible. At lower frequencies (less than about 300 GHz), the best results will be obtained if V_o is minimized at the expense of R_s and C_{jo}. A survey of the best Schottky receivers worldwide verifies the importance of these tradeoffs.

Acknowledgment

The authors wish to thank A. R. Kerr of the National Radio Astronomy Observatory in Charlottesville, VA, for his substantial contributions to this research.

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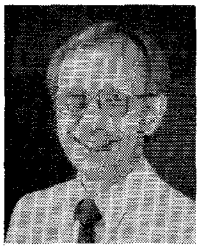

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