The Super-Schottky Diode

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***Abstract*—The super-Schottky-barrier diode, a superconductor-semiconductor tunneling junction, has been established as the most sensitive detector of microwaves. These record sensitivities were obtained in both the video and mixing modes of operation. Measurements at X-band have yielded a video NEP of 5×10^{-16} W/Hz$^{1/2}$ and a mixer input noise temperature of 6 K. The super-Schottky mixer provides a front-end component for ultralow-noise receivers that is superior in bandwidth to available parametric and maser amplifiers and yet has a comparable noise temperature. This article reports the design, fabrication, and measurement of Pb on p-GaAs super-Schottky diodes which perform as nearly ideal low-noise mixers at 9 GHz.**

I. Introduction

THE demands by the radio astronomy community and, more recently, by those in satellite communication have traditionally set the pace for improvements in microwave and millimeter-wave receivers in terms of high sensitivity and wide bandwidth. The front-end component for this type of receiver has most frequently been the Schottky-barrier diode because of its fundamentally wide bandwidth, sensitivity, and reliability. The performance of this device dominates the overall receiver quality.

The introduction of the super-Schottky-barrier diode [1]–[3], a superconductor–semiconductor tunneling junction, is seen to represent a major advance in low-noise diodes. It has achieved the lowest measured noise-equivalent power reported in the literature as a video detector and the lowest noise temperature operating as a mixer. The super-Schottky mixer possesses a bandwidth far in excess of the best available parametric and maser amplifiers and yet has a comparable noise temperature. However, since resistive mixers have a loss rather than a gain, the overall performance of a receiver incorporating a super-Schottky mixer will be critically dependent on the choice of an IF amplifier. Low-noise masers or paramps would be good candidates although the full bandwidths of the super Schottky might not be utilized. Another possibility is an all-superconducting receiver as proposed by Silver [2]. In this paper the results of extensive theoretical and experimental work are presented in which considerably improved diode performance is both theoretically predicted and experimentally achieved. Limitations to performance and potential further advances are also discussed.

The physical structure of the diode, similar to that of conventional Schottky diodes, consists of a metal contact on a semiconductor such as GaAs. However, in this case a superconducting metal and a heavily doped semiconductor are utilized. The doping is chosen sufficiently large that electron tunneling dominates the current–voltage (I–V) characteristics of the diode. For applied voltages less than the superconducting energy gap parameter Δ, this type of diode exhibits a high degree of nonlinearity in its I–V characteristic [4]. This resistive nonlinearity results from the effect of the superconducting energy gap on the tunneling of thermally excited carriers. It is this nonlinearity that is exploited by the super-Schottky diode. The high doping concentration also achieves the high current densities that are required for proper impedance matching.

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The performance of the super-Schottky mixer, like a conventional Schottky mixer diode, is directly related to the degree of nonlinearity of its I–V characteristic. The I–V behavior of either diode can be represented over the voltage range of interest approximately by [4]–[6]

$$I = i_0 \exp(SV) \tag{1}$$

where i_0 is determined by the area and material parameters of the diode. A more accurate expression for the I–V characteristic of super-Schottky diodes is shown in (11). The parameter S is a measure of the nonlinearity of the diode and, as such, is of central importance in the determination of the sensitivity of the diode as either a video detector or mixer. It has been determined empirically that for conventional Schottky diodes, S can be expressed approximately as [5]

$$S \simeq q/k(T + T_0) \tag{2}$$

where q is the electronic charge, k is the Boltzmann constant, T is the Kelvin temperature, and T_0 is an empirical constant. Numerically, (2) becomes

$$S \simeq 11\,600/(T + T_0)\ \text{V}^{-1}. \tag{3}$$

Both experimentally and theoretically for super-Schottky diodes reported here, $T_0 \ll 1$ K if $kT \ll q\Delta$; and therefore, $S \approx q/kT$ is a reasonable theoretical approximation. The value of T_0 for Schottky barriers on n-type GaAs is greater than 40 K [7], [8]. Comparing a conventional mixer [9], [10] with a super Schottky, both operating at 1 K, implies greater than a 40 to 1 improvement in noise temperature [8]. This reduction in noise would imply a corresponding improvement in sensitivity if the conversion loss of the super Schottky were equivalent to that of the conventional Schottky. As discussed in the following, conversion loss is a major issue for the super Schottky. Because of the relatively limited voltage range, the design criteria for obtaining low conversion losses are stringent.

II. Design Considerations

The noise temperature T_r of a heterodyne receiver, referred to the input of the mixer, is given by

$$T_r = L_c(T_D + T_{IF}) \tag{4}$$

where L_c is the single-sideband conversion loss of the mixer, T_D is the noise temperature of the mixer diode measured at the IF terminals of the mixer, and T_{IF} is the noise temperature of the IF amplifier, respectively. Hence, in fabricating a mixer, utmost attention must be paid to minimizing L_c.

Conversion loss is conveniently expressed as the product of two terms:

$$L_c = L_0 L_1 \tag{5}$$

where L_0 is the intrinsic loss associated with the frequency conversion process that occurs in the nonlinear resistance and L_1 is the loss arising from parasitic impedances associated with the junction [11]. The theory of L_0 and its functional dependence on local oscillator (LO) excitation, the type of imbedding circuit, and the impedances of that circuit have been examined by several authors [12], [13]. Assuming a Y-connected circuit [13] driven by an LO voltage of amplitude V_1, the broad-band (i.e., with image) single-sideband conversion loss L_0 is given by

$$L_0 = \frac{2}{\eta}[1 + \sqrt{1-\eta}]^2 \tag{6}$$

where

$$\eta = 2I_1^2/I_0(I_0 + I_2) \tag{7}$$

and $I_n \equiv I_n(SV_1)$ are modified Bessel functions of the first kind. Equation (6) predicts that an increase in S, brought about by a reduction in temperature, results in a reduced value of L_0 [12].

Equation (6) is valid if the signal (RF) and IF input conductances are matched to their circuit counterparts. This situation is satisfied for optimum signal and IF load conductances $g_{s,\mathrm{opt}}$ and $g_{\mathrm{IF,opt}}$, respectively, given by

$$g_{s,\mathrm{opt}} = g_a(I_0 + I_2)\sqrt{1-\eta} \tag{8}$$

$$g_{\mathrm{IF,opt}} = g_a I_0 \sqrt{1-\eta} \tag{9}$$

where

$$g_a = Si_0 \exp(SV_0) \tag{10}$$

represents the conductance of the diode at the dc bias V_0 with no LO voltage applied to the junction.[1]

Equations (6)–(10) assume the I–V characteristic has an exponential form over the entire swing in LO voltage. Since for the super-Schottky diode the exponential behavior ceases for voltages greater than Δ, the swing in LO voltage is limited to the range 0–Δ.[2] The limited range of voltage (Δ is in the order of millivolts) leads to some relatively rigid design constraints in the task of minimizing L_0 for the diode operating in a given circuit. Design equations will now be formulated in which the aforementioned mixer theory is combined with the specific I–V characteristic of this type of structure. An appropriate theoretical expression for the tunneling current is given by [4]

$$I = \sum_m K_m \sinh m\frac{qV}{kT}: |V| < \Delta. \tag{11}$$

For temperatures such that $\exp[q(|V| - \Delta)/kT] \ll 1$ and $qV \gg kT$ the conductance is given adequately by the expression

$$g = g_n C(\Delta,T)\exp[S(|V| - \Delta)] \tag{12}$$

where g_n is the zero bias conductance when the metal is normal, $S \approx q/kT$, and $C(\Delta,T) \approx (\pi\Delta/2kT)^{1/2}$. Because of the restriction on temperature and voltage required for the validity of (12), the highest voltage for which the exponential dependence holds is always less than Δ and decreases as T increases. (This temperature dependence is in addition to the temperature dependence of Δ [14].) Furthermore, at small voltages such that $qV \lesssim kT$, the I–V characteristic is linear and this linearity extends to higher and higher voltages as the temperature of the junction is increased. Hence the ratio of the maximum obtainable conductance $g_{\max}$ to the minimum obtainable conductance $g_{\min}$ is a strong function of temperature. The ratio $g_{\max}/g_{\min}$ can be numerically evaluated [15] as a function of $q\Delta/kT$ and then translated into a form compatible with the parameters in (6)–(12) in the following manner. The optimum LO drive $(SV_1)_{\mathrm{opt}}$ corresponds to a full sweep of the diode conductance characteristic and can be expressed as

$$(SV_1)_{\mathrm{opt}} = \tfrac{1}{2}\ln(g_{\max}/g_{\min}). \tag{13}$$

Equation (13) and the $g_{\max}/g_{\min}$ evaluation yield a one-to-one relationship between $(SV_1)_{\mathrm{opt}}$ and $q\Delta/kT$. A notable result of this analysis is the near equivalence of $(SV_1)_{\mathrm{opt}}$ and $q\Delta/2kT$ at low temperatures. This situation is equivalent to the biasing condition of $V_0 \approx V_1 \approx \Delta/2$ and a thermionic dependence of $S \approx q/kT$.

The results of the preceding analysis were used to plot Fig. 1, which shows L_0 as a function of $q\Delta/kT$. Therefore, Fig. 1 provides a quick evaluation of the theoretical capability of a super-Schottky mixer diode given the Δ of the superconductor and the temperature at which it is to be operated. For Pb contacts at 1 K, Fig. 1 predicts a theoretical lower limit of $L_0 = 4$ dB. Lower values of L_0 could be achieved by either decreasing temperature or increasing Δ. Increasing Δ, of course, requires the use of higher T_c materials.

The equivalence between $g_{\max}/g_{\min}$ and $(SV_1)_{\mathrm{opt}}$ as given by (13) is also particularly useful in evaluating the conversion loss of a diode by means of its measured I–V

[1] For a broad-band mixer $g_{s,\mathrm{opt}}$ is not the value which "matches" the mixer input impedances in the conventional sense. Saleh [13] has shown that for a broad-band mixer operating with optimum source impedance and a matched IF load, the input impedance to the mixer is not identical to the optimum source impedance. However, this mismatch is of little real significance in this case. For example, with the LO drive levels encountered in the experiments reported, the "reflected" power from an optimally terminated mixer would amount to less than 0.15 percent of the incident signal.

[2] The superconducting energy gap parameter Δ as used throughout this paper is defined to be a voltage corresponding to one-half of the voltage corresponding to the superconducting energy gap.

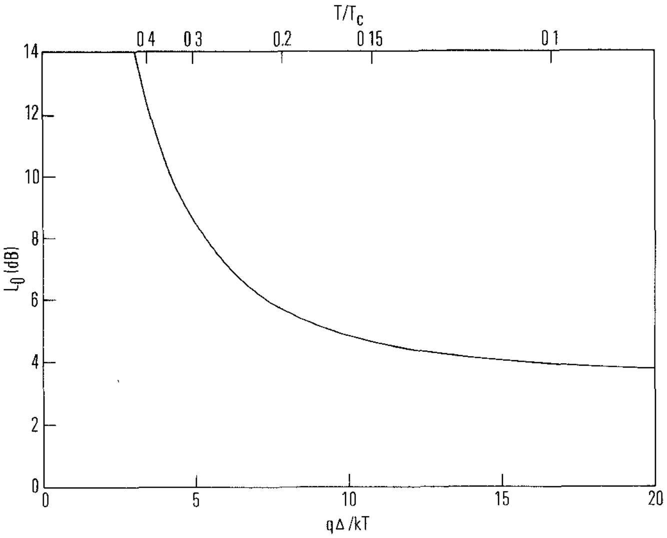

Fig. 1. Conversion loss L_0 as a function of the energy gap parameter Δ normalized to kT. The abscissa axis T/T_c is derived on the basis of the superconductor energy gap $2q\Delta$ being given by $3.5kT_c\sqrt{1 - T/T_c}$ where T_c is the superconducting transition temperature.

characteristic. The $g_{\max}/g_{\min}$ ratio can be determined from this measured conductance curve and converted to $(SV_1)_{\text{opt}}$ by means of (13). This allows a direct evaluation of L_0 from (6) and (7). For this reason, this conductance ratio is an important experimental parameter.

Having established the conversion losses which can be anticipated, the required circuit conductances will be determined. The geometric average conductance g_a can be written approximately as

$$g_a = \sqrt{g_{\max} g_{\min}} \approx g_n \exp\left[-(SV_1)_{\text{opt}}\right] \tag{14}$$

since $g_n \approx g_{\max}$ (e.g., for Pb at 1 K, $g_{\max} \approx 2.4g_n$). Combining (8), (9), and (14) yields the dependence of the optimum circuit conductances in terms of the optimum LO drive. This functional dependence is plotted in Fig. 2. The numerical evaluations of $g_{\max}/g_{\min}$ and (13) have been used to establish the second abscissa axis shown in Fig. 2. It is clear that cooling a super-Schottky mixer diode lowers the conductances of the diode and consequently lowers the circuit conductance required to achieve impedance matching.

It is apparent that matching the conductance of this type of mixer diode to its circuit involves many parameters, some of which are the dc and LO voltages, the temperature of the diode, and the area of the contact. However, the primary design factor, the one that sets the general impedance level of the super Schottky, is the normal conductance g_n. This quantity is strongly dependent on the impurity concentration N of the semiconductor [3], [16]. The tunneling distance for carriers is equal to the depletion width of the barrier which, in turn, is proportional to $N^{-1/2}$. Hence the tunneling probability, and $\log g_n$, is approximately linear with $N^{-1/2}$ as shown in Fig. 3 for p-type GaAs. This figure shows the results of several measurements of g_n as well as a theoretical calculation

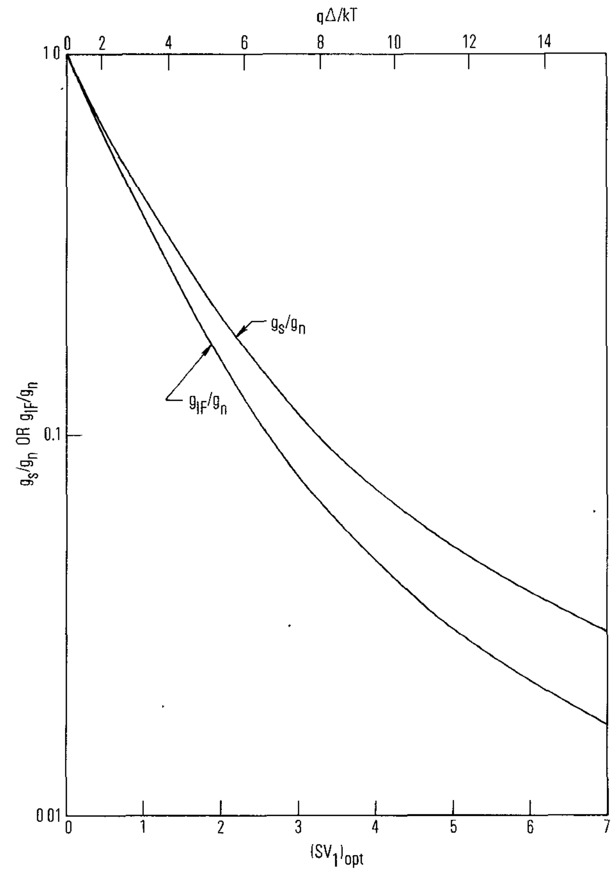

Fig. 2. Optimum source and load conductances g_s and g_L, respectively, normalized to the normal conductance g_n of the super-Schottky mixer diode as a function of optimum load oscillator drive and $q\Delta/kT$.

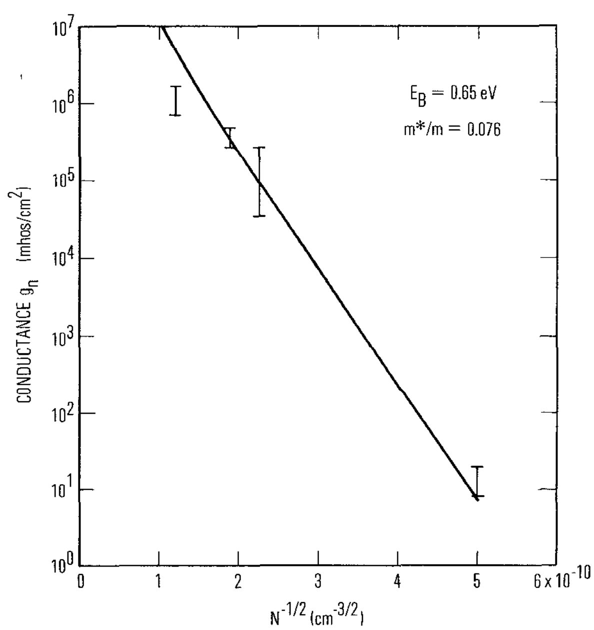

Fig. 3. Normal conductance g_n per unit area of Pb contacts to p-GaAs as a function of hole concentration N. The solid line is the theoretical result obtained [16] using a barrier height of 0.65 eV and an effective mass ratio of 0.076 for the tunneling carriers.

identical to that carried out in [16]. By the proper choice of doping and area, the impedance of the diode can be optimized.[3]

Leakage current near zero bias voltage is a major obstacle in achieving low values of L_0 since it can seriously reduce the range of nonlinearity. The diode characteristic shown in [1, fig. 1] is an example in which excess leakage current exists. In this figure, an S value of 5000 V^{-1} and a maximum-to-minimum conductance ratio g_{max}/g_{min} of 100 were obtained which would yield an intrinsic conversion loss of 7.1 dB. Larger values of this ratio are required to decrease this value of L_0. For the use intended at that time, i.e., a video detector, leakage was not a serious problem since leakage currents were negligible at the point on the I–V curve which had the greatest nonlinearity (and thus responsivity) and simultaneously the best RF impedance match. Leakage can be the result of various extraneous conduction mechanisms, but the major offender for these diodes results from the pressure of the whisker contact on the active area of the Pb electrode. This pressure produces a region of normal metal in contact to the semiconductor.

Avoiding the actively conducting area of the junction by contacting the superconductor at an inactive region reduces the leakage greatly. Experimentally, this is accomplished by a relatively small refinement in the fabrication of the diodes and the method of contacting them. The diodes are basically fabricated in standard fashion by electroplating Pb on the p-GaAs through an array of windows in an insulating layer of SiO_2. However, the plating is continued until the metallization pads almost come into contact with one another as shown in the inset of Fig. 4. A blunted whisker can then contact the edges of two or more neighboring diodes. Fig. 4 shows the I–V behavior that results from such a procedure on 2×10^{19} cm^{-3} p-type GaAs. Hence using the improved contacting procedure on diodes constructed from the same material increased the g_{max}/g_{min} ratios from 100 to over 4000.

The loss term L_1 is the consequence of signal power lost in the series spreading resistance R_s shown in Fig. 5. It is easily shown that L_1, the ratio of power absorbed by R and R_s to the power absorbed by R alone, is given by

$$L_1 = 1 + R_s/R + \omega^2 C^2 R R_s \tag{15}$$

where ω is the signal angular frequency, C is the junction capacitance, and R is the signal input impedance [12] of the LO-pumped nonlinear junction resistance. Several methods are available for minimizing this loss by reducing R_s. Two such methods involve the reduction of the resistivity ρ of the semiconductor. Since $(L_1 - 1)$ is proportional to R_s, which for a small area circular contact is given by $R_s = \rho/2D$ (D is the contact diameter) [12], L_1 can be reduced by choosing semiconductors that have large mobilities and, to a certain degree, are heavily doped. Heavy doping is entirely compatible with the super-Schottky diode as discussed previously. A negative contribution of the heavy

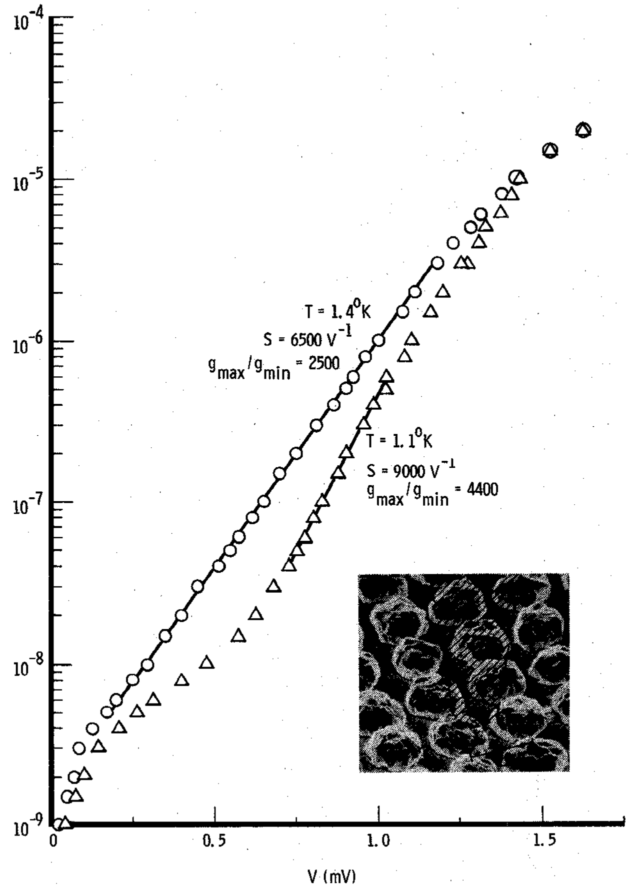

Fig. 4. Electrical behavior of a super-Schottky diode of Pb on 2×10^{19} cm^{-3} p-GaAs. The inset is an SEM view of several diodes illustrating the excess overplating. The photograph was taken at an angle to the surface; the diodes are not in contact with one another. The diameter of the metal overplate is approximately three times the junction diameter. The shaded area schematically illustrates a metal whisker contacting four neighboring junctions.

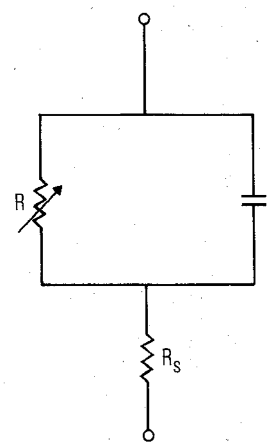

Fig. 5. Equivalent circuit of the super-Schottky diode.

[3] A frequently asked question is why not use n-type GaAs, a very high mobility material and one that is used for conventional high-frequency Schottky mixing rather than p-type GaAs whose mobility is much lower. The obstacle in using n-GaAs lies in the difficulty of obtaining a sufficiently high carrier concentration to obtain a usable impedance level for super-Schottky construction. That is, because of its larger barrier height, n-GaAs would require a larger doping than p-GaAs to obtain the same impedance. As it stands to date, the solubility of known donor impurities is smaller than that of acceptor impurities which are presently being used in GaAs.

doping is the increase in junction capacitance. High mobility materials, e.g., n-type InSb [17], would be the best candidates for millimeter wavelength applications. A third method to reduce L_1 involves the use of multiple contacts [18]. For the same total area of contact and hence impedance, $(L_1 - 1)$ is inversely proportional to the square root of the number of identical independent contacts connected in parallel.

A mixer based on the I–V characteristic of Fig. 4 would be an excellent converter only at frequencies well below 10 GHz. For example, using the 1.1 K data, calculations yield an L_0 of 4.8 dB and a signal input impedance of 1000 Ω. However, at a frequency of 10 GHz, (15) yields an L_1 of 5.1 dB which results in an L_c of about 10 dB. In an effort to reduce L_1 by reducing the signal impedance of the diode, dopings heavier than 2×10^{19} cm^{-3} were investigated.

With higher dopings, a new leakage mechanism dominates the I–V behavior, the origin of which is under investigation. However, 3×10^{19} cm^{-3} p-GaAs does provide a considerable improvement in L_1 for 10-GHz usage. Although g_{max}/g_{min} ratios of only 300 are typically obtained at 1 K, calculated loss factors are $L_0 = 6$ dB and $L_1 = 1.5$ dB (the reduction being due to a reduction in R); hence $L_c = 7.5$ dB, which represents an improvement of 2.5 dB. This latter conversion loss is in excellent agreement with the values reported in the following.

The fabrication of the diodes is similar to that described earlier [1]. An array pattern with 5-μm diameters and 20-μm center spacings was produced. Junctions consisted of Pb electroplated through windows in an SiO_2 coating on p-type GaAs. A pyrolysis of silane for 5–10 min at 360°C was substituted for the ethyl silicate process to allow the use of a lower temperature.

III. Experimental Apparatus

Since super-Schottky diodes are extremely low-noise devices, measurements of their properties require the use of careful radiometric techniques. Accordingly, the measurement apparatus includes 1) means for connecting two levels of thermal radiation (room temperature and liquid helium temperature) to the input of the mixer under test for determining conversion loss, 2) input X-band signal generators and a reflectometer setup for measuring the input impedance and the relative conversion loss of the two sidebands, 3) X-band noise diode for optimizing performance and for making broad-band reflection measurements over the signal passband, and 4) a sensitive IF radiometer for measuring conversion loss, mixer noise, and diode IF reflection coefficient.

The diode is mounted in 90-Ω ridge waveguide with a movable back short for tuning and a coaxial cable connector for applying dc bias and for conducting the IF output signal to the radiometer. Since the diode must operate in liquid helium, the mixer is mounted near the bottom of a liquid helium dewar with waveguide and coaxial cable connections to the room temperature environment. The back short position is adjustable using a drive shaft connected to the top of the dewar. The LO is connected to the RF signal line through a 30-dB directional coupler in order to reduce the thermal energy incident on the diode from the LO. The directional coupler is immersed in the liquid helium so that a helium-cooled attenuator can feed into the RF line in front of the LO connection. The dewar arrangement, together with the RF and IF measuring circuits, is shown in Fig. 6. The liquid helium-cooled attenuator is adjustable from the top of the dewar through a drive shaft to be either at a maximum position (~ 50 dB) or a minimum position (~ 0.3 dB). When in the maximum attenuation position the attenuator acts as a matched termination at liquid helium temperature on the input to the mixer. Since the thermal radiation coupled to the mixer is proportional to the temperature of the attenuator, a thermocouple and a carbon resistor have been included as a thermometer. This is particularly useful when the liquid helium level drops below the attenuator. When the cooled attenuator is switched to the minimum attenuation position, a termination at ambient temperature (~ 295 K) or a reflectometer circuit can be connected to the mixer input through the waveguide switch. The reflectometer is used to determine the reflection coefficient of the mixer at the LO frequency. In addition, an X-band noise diode signal can be injected into the line for use in optimizing signal-to-noise ratio and for obtaining an average reflection coefficient over the mixer sidebands. When the cooled attenuator is adjusted to its maximum value, the input to the mixer is changed from a thermal source of 254 K to

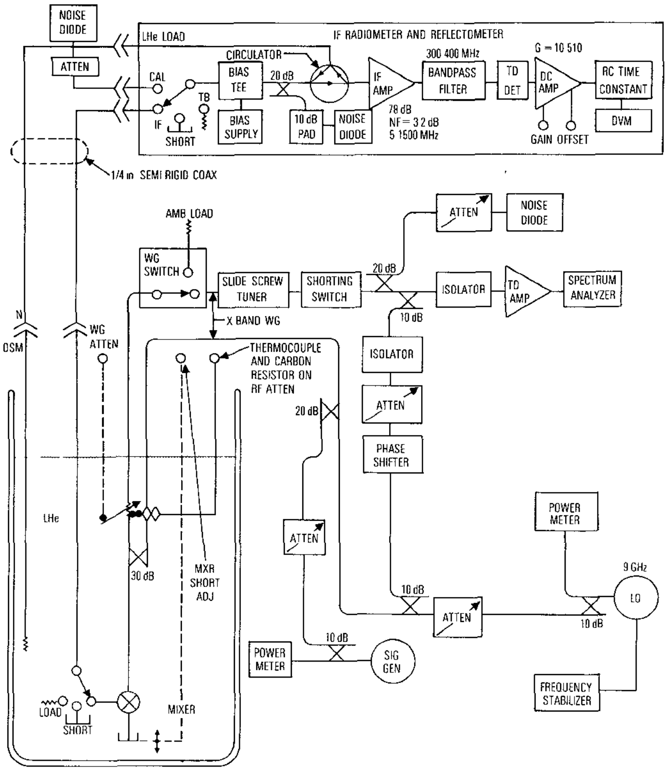

Fig. 6. Apparatus used to measure the properties of super-Schottky mixers. The principal components are a 350-MHz total power receiver, a liquid helium bath for cooling critical circuit components including the mixer mount, and a waveguide bridge circuit for measuring impedance and for introducing the LO and signal power.

a source of ~ 1 K (depending on liquid helium level). This change in input signal and the measured change in output IF power is a measure of the mixer conversion loss.

As shown in Fig. 6, the IF cable connects to the input of a total power receiver of the type used by Kerr [19]. Initially, a short section of coaxial line terminated in a matched load is connected to the IF port. Using this termination at ambient temperature and at liquid N_2 temperature permits setting gain and offsets of the receiver so that each millivolt of output signal corresponds to 1 K temperature on the input. A noise diode and ambient termination at the radiometer input furnish secondary standards which are used throughout the measurement period. A second noise diode inside the radiometer is connected to the switched input through a 20-dB directional coupler. It is used to measure the IF reflection coefficient of the various inputs. The short position is used to obtain a reference for these reflection measurements.

Another coaxial switch is located in the liquid helium near the mixer. It is used to provide a matched load and a short termination to the IF coaxial line. These terminations provide a reference for the measurement of the mixer IF reflection coefficient and permit separation of the effects of thermal noise originating in the coaxial line and the radiometer.

IV. Mixer Measurements

In analyzing the measurement results we have attempted to determine the intrinsic mixer performance which would be obtained if the diode were located in an optimum imbedding circuit for the diode. The results are referred to the model in which the mixer is considered to be a 3-port device with terminals at the signal, image, and IF frequencies. L_s and L_i represent the total conversion losses at signal and image frequencies and are the experimentally determined values for L_c defined in (5). The corresponding input terminations at these frequencies are at temperatures T_s and T_i. The noise contribution of the mixer can be expressed in terms of a diode noise temperature T_D when referred to the output or a mixer noise temperature T_M when referred to the input. For our purposes all of the mixer noise will be considered to be at the signal frequency. The thermal output of the mixer (i.e., noise from the mixer into the IF amplifier) can be expressed by

$$T_{\mathrm{IF}} = T_D + \frac{T_s}{L_s} + \frac{T_i}{L_i} \tag{16}$$

where

$$T_D \equiv \frac{T_M}{L_s}. \tag{17}$$

In the usual case $T_s = T_i$ so that

$$L_s T_{\mathrm{IF}} = T_M + T_s(1 + L_s/L_i). \tag{18}$$

The loss ratio L_s/L_i can be determined by relative loss measurements at the signal and image frequencies. Equation (18) is used to determine L_s by measuring ΔT_{IF}, the difference in T_{IF} which results when two values of T_s which differ by ΔT_s are used. These ΔT must be corrected for the diode impedance mismatch. Since T_M does not change, (18) becomes

$$L_s = (\Delta T_s/\Delta T_{\mathrm{IF}})(1 + L_s/L_i). \tag{19}$$

Measurement of T_M requires an absolute determination of T_{IF} for a known value of T_s. In addition, accurate determination of T_{IF} requires a determination of thermal energy from connecting cables which radiate into the radiometer directly or indirectly through reflection from the diode impedance mismatch.

V. Mixer Results

A considerable number of super-Schottky diodes have been tested as mixers with similar results in each case. The results presented as follows are mainly for one of these diodes for which we have the most complete data. The diode I–V curves are shown for several values of LO power in Fig. 7. Curves A and B indicate the mixer output with the microwave source termination on the mixer at 1.06 and 253.91 K, respectively. The difference in these curves is a direct measure of the total system conversion loss. At a bias voltage level of 0.6 mV, ΔT_{IF} has a maximum value of 54.85 K, compared with an input $\Delta T_s = 252.85$ K.

From measurements of the relative transmission through the mixer at the upper and lower sideband frequencies, L_s/L_i was estimated to be 0.74. Thus the raw single-sideband conversion loss (neglecting coaxial line attenuation) is from (19) $L_s \approx (252.85/54.85)(1 + 0.74) = 8.02$ or 9.04 dB. This quantity corrected for line losses is plotted in Fig. 8. Curve C of Fig. 7 is a measure of the IF power reflected from the mixer diode using a noise diode source. These data are used to calculate the IF reflection coefficient $|\Gamma_{\mathrm{IF}}|^2$ which, in turn, is used to obtain the IF mismatch loss $(1 - |\Gamma_{\mathrm{IF}}{}^2|)$ plotted in Fig. 8. This quantity can also be estimated from the slope of the I–V curve with the LO bias applied shown in Fig. 7. The RF mismatch loss $(1 - |\Gamma_{\mathrm{RF}}{}^2|)$ was obtained by measuring the reflection of X-band noise power from the mixer diode at signal and image frequencies. The single-sideband conversion loss corrected for RF and IF reflection losses is shown as the lower curve in Fig. 8. The diode noise temperature T_D was calculated and plotted in Fig. 8 for several values of bias voltage. It can be seen that these points fall very close to the line representing the bath temperature. The lowest value of T_M occurs at a bias of 0.6 mV with $T_D = 1.2$ K and $L_s \approx 7$ dB resulting in $T_M = 6$ K.

As previously stated, the variation in resistance of a super-Schottky diode with a change in applied voltage gets smaller as temperature is increased. (This is apparent from Fig. 4 which shows the I–V curves for a diode at 1.1 and 1.4 K.) It is to be expected that the performance of the diode as a mixer correspondingly decreases. Fig. 9 shows mixer characteristics as a function of temperature for one particular diode. It is to be noted that the single-sideband conversion loss and diode temperature both increase as the temperature is increased. Lower temperature operation for a super-Schottky mixer is clearly advantageous.

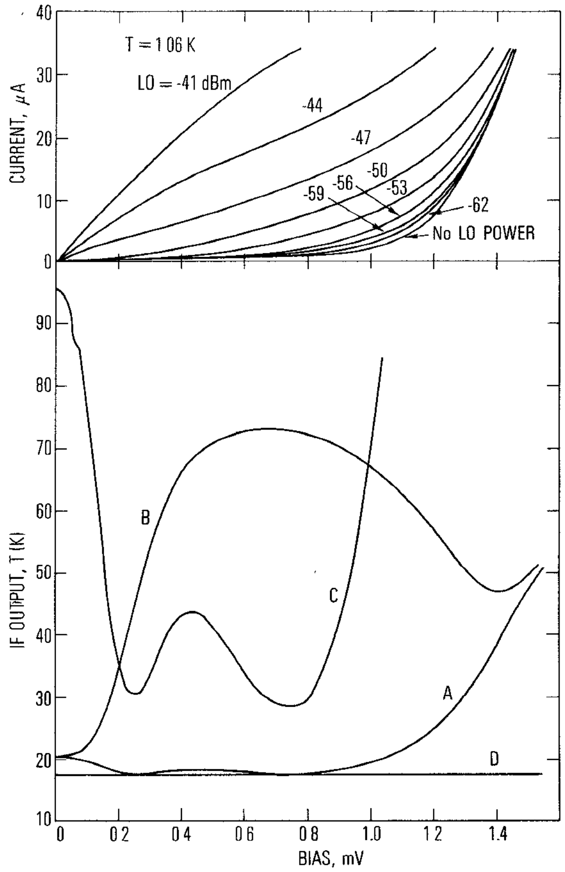

Fig. 7. Pb/p-GaAs super-Schottky diode characteristics measured as functions of bias voltage at 1.06 K. The upper graph shows the dc I–V curve variation with 9-GHz LO power. Optimum mixing is obtained with the LO level maintained near −47 dBm where the slope, which closely approximates the diode IF conductance, is about 50 Ω over a wide range of bias voltage. The graph also indicates the sensitivity to low-level signals in the video mode. The lower graph is the recorded IF radiometer power output calibrated directly in degrees Kelvin when −47 dBm of LO power at 9 GHz is applied. Curves A and B are with 1.06 and 253.91 K microwave source terminations, respectively. The difference between curves A and B is a measure of the conversion loss. Curve C is the power reflected from the mixer diode using an IF noise source at approximately 2000 K. The difference between curves C and A is thus a measure of the diode IF mismatch. Curve D is a measure of the radiation from the IF coax when it is terminated in a matched load. The difference between curves A and D is caused mainly by room-temperature radiation reflected from the diode.

VI. Video Detector

The extremely sharp curvature of the I–V curves for super-Schottky diodes can also be utilized for video detection. Since the curvature gets sharper as the temperature is lowered, the responsivity and NEP also improve. The NEP is given by the expression

$$\mathrm{NEP} = \langle i_n^2 \rangle^{1/2} / \mathscr{R} \tag{20}$$

where

- $\langle i_n^2 \rangle$ $= 4kT_D B/R_D$ = mean square noise current, ampere^2;
- k Boltzmann constant, 1.38×10^{-23} J/K;
- T_D diode noise temperature, degrees Kelvin;
- B noise bandwidth, hertz;
- $\mathscr{R}$ responsivity, amperes/watt;
- R_D diode incremental resistance, ohms.

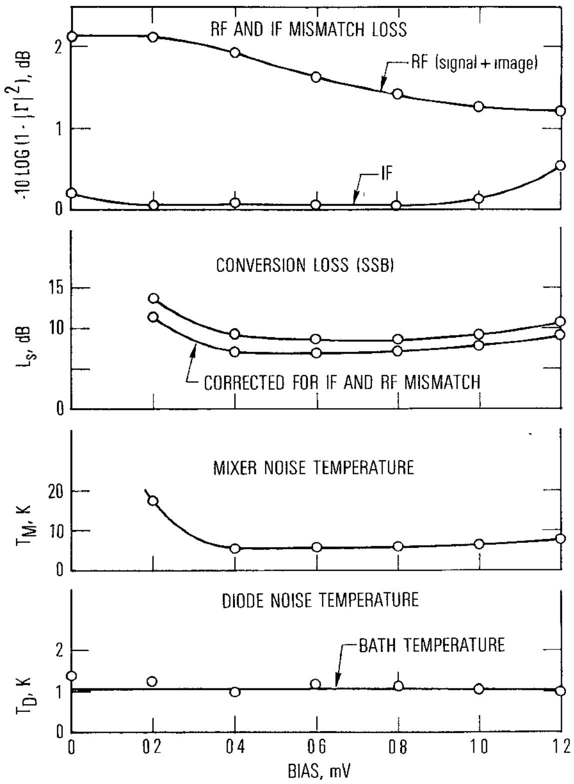

Fig. 8. Parameters describing mixer performance as a function of bias voltage at $T = 1.06$ K. Notice that the diode noise temperature is essentially the bath temperature and the IF mismatch loss is very low. When the single-sideband conversion loss is corrected for mismatch losses, the measured results agree very well with predictions based on dc characteristics.

Fig. 10 shows responsivity versus bias voltage for various temperatures derived from data similar to those shown in the upper part of Fig. 7. It can be seen that $\mathscr{R}$ is a sensitive function of the bias voltage. This arises primarily because of the large variation in impedance over the I–V curve at low RF power levels. T_D was measured for the same diode as a function of bias voltage and temperature using the 350-MHz radiometer. When the effect of impedance mismatch is taken into account, the NEP in femtowatts per root hertz is shown in Fig. 10. As expected, the NEP improves as the temperature is decreased. The best value obtained was 5.4×10^{-16} W/Hz$^{1/2}$ at a temperature of 1.06 K. This is the highest sensitivity reported to date for a video detector. Theoretically, for this type of diode where $I(V)$ is represented by a simple exponential, the responsivity is given by $\mathscr{R} = S/2$ where $S \equiv q/kT$ [1]. When parasitic loss is present in the form of a series resistance, the responsivity is modified to be

$$\mathscr{R} = S/(2L_p) \tag{21}$$

where L_p is defined in the same manner as L_1 in (15) with the exception that R is replaced by R_D, the incremental diode resistance. At 1.06 K and a bias voltage of 1.0 mV, where the minimum NEP was found, Fig. 10 indicates $\mathscr{R}_{\mathrm{exp}} = 2200$ A/W. Using (21) with the theoretical value of $S = 10950$ V^{-1} yields the estimate $\mathscr{R}_{\mathrm{theor}} \approx 2400$ A/W which agrees well with the experimental value. However, in the absence of parasitic loss $\mathscr{R}_{\mathrm{theor}} \approx 5500$ A/W. Thus,

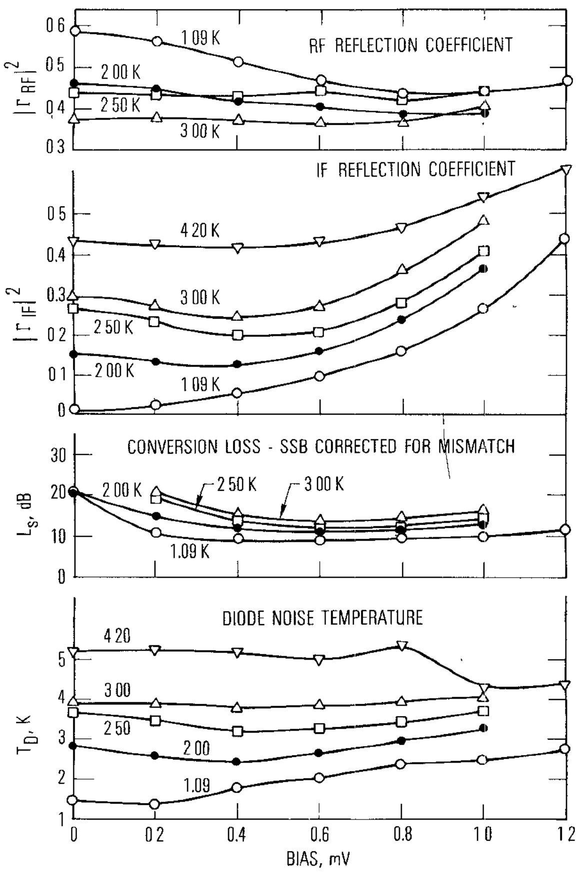

Fig. 9. Variation of diode noise temperature, conversion loss, and reflection coefficient with changing bath temperature. In general, performance is best at the lowest temperature.

if parasitic losses were eliminated, the NEP would be decreased by a factor of (5500/2200) yielding an estimated NEP $\approx 2.2 \times 10^{-16}$ W/Hz$^{1/2}$.

VII. Conclusion

The super Schottky is the best resistive diode operated either as a mixer or a video detector. In both cases the intrinsic noise is, within experimental error, the noise appropriate to the bath temperature. Further improvements in device operation must be effected by reducing the parasitic loss L_1 and in the case of mixer operation, the intrinsic conversion loss, L_0. Materials engineering approaches to accomplish this are feasible. Further improvement in conversion loss would be useful at X band and will be required at higher frequencies. The primary method of reducing L_0 is increasing SV_1. The range of useful LO voltage is restricted to the highly nonlinear region controlled by Δ on the high-voltage end and shunt leakage on the low-voltage end. Thus superconductors with high critical temperatures and hence larger gap voltages can be used; some of the high T_c alloys can probably be sputter deposited. L_1 is controlled by the semiconductor and the geometry. High mobility semiconductors with low effective masses are ideal if suitable Schottky barriers exist. Some low band gap materials, notably n-InSb [17] and n-InAs [20], [21], are characterized by near-zero barrier heights with most metal contacts. The ternary material $In_{1-x}Ga_xSb$ combines the advantage of a high mobility [22] with a strong possibility

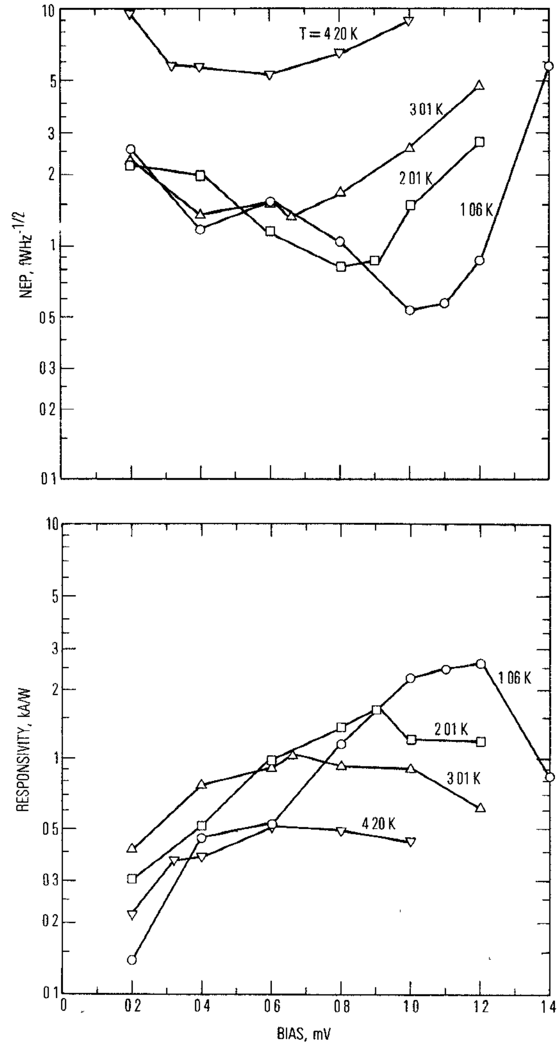

Fig. 10. Measured values of noise equivalent power and responsivity for a super-Schottky diode operated in the video mode at 9.0 GHz. The values plotted are intrinsic to the diode since corrections for mismatch have been made. However, the corrections are fairly small near optimum bias. For example, at 1.06 K and in the bias voltage range of 1.0–1.2 mV the uncorrected values of NEP and responsivity differ by less than 50 percent from the corrected values. Also to be noted is that while NEP and responsivity improve with decreasing temperature, the values are still quite good at 4.2 K (5×10^{-15} W/Hz$^{1/2}$ and 500 A/W).

for controlling the barrier height by the crystal mixture [23]. Thus these crystals would be capable of providing a suitable barrier height for a variety of superconducting metals and, simultaneously, permit the achievement of lower series resistances. A multiple contact geometry can also be used to advantage. These approaches are under investigation in this laboratory.

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The Measurement of Noise in Microwave Transmitters

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Invited Paper

***Abstract*—A tutorial review of the basis for transmitter noise measurements shows that noise is best described and measured as AM and FM noise. The determination of RF spectrum is done by calculation after the AM and FM noise are known. The contribution of AM noise to RF spectrum shape is determined by the power spectral density shape of the AM noise. The contribution of FM noise to RF spectrum is to make the shape that of an *RLC* circuit resonant response rather than a δ function with a sideband structure. The measurement of AM noise is done with a direct detector diode. The measurement of FM noise for frequencies above 5 GHz is done with a discriminator based on a one-port cavity resonator. The measurement of FM noise below 5 GHz is done with an improved transmission line discriminator which is described in detail. Measurement of low-power low-noise signal sources is made possible with an injection-locked oscillator for a preamplifier to the discriminator. The most widely used baseband analyzer is the constant bandwidth superheterodyne wave or spectrum analyzer. Most differences in measurement results are resolved by understanding the baseband analyzers. At least the baseband spectrum analysis of transmitter noise measurements can be automated with worthwhile savings in time and improvement of documentation.**

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I. Background Concepts

NOISE in a microwave transmitter is an unwanted kind of signal which degrades the ability of the system to transmit communications or other information. We must include both stochastic signals and coherent spurious signals—usually related to bias supply ripple or mechanical vibration—in our thinking about transmitter noise. In the first stages of receiving systems, we are usually concerned with a ratio of desired signal to noise in the order of unity. In a transmitter, the ratio of carrier to noise is on the order of 10^3 or more. Most of the amplifying or oscillating devices in a transmitter are operating in a saturated mode; thus some of the linear system theory ideas that are so useful in the study of receiving systems must be used with care in studying a transmitter system.

The user of a microwave transmitter also will be using a detection system which may respond to either amplitude modulation or angle modulation (an FM discriminator), but not often both. His essential question is, "How much of the noise in a transmitter will come out of my detection system?" The kind of a detector being used will determine if the user is concerned with what has come to be called "FM noise"