beam pattern, which varies as the rf changes over a one octave range. It might also be due to some standing wave effects between the various optical elements between our mixer and the hot/cold load signals. A more involved discussion of what is needed to make measurements like this with high accuracy is given by [15]. No matter what the source of the extra loss, we can use the data shown in fig. 5 to make a conservative estimate of 18 GHz for the 3 dB bandwidth of our tuning circuit.

IV. CONCLUSION

Results for an SIS receiver with 38 K double sideband receiver noise temperature at 77 GHz, and less than 100 K noise over a 27 GHz bandwidth are presented. This is achieved using a quasioptical coupling scheme, a fairly noisy 15 K photodiode, and large area (12 μm²) SIS junctions.

Niobium integrated circuit technology has been developed commercially for digital circuit applications. We have shown that this technology can be used in SIS heterodyne mixers to achieve ultra-low noise performance. The digital technologies we used produced SIS’s with much larger area and much lower current density than most good SIS mixers use. However, accurate design of tuning structures taking advantage of the stability of commercial process can fully compensate for this limitation.

We conclude that commercial 1,000 A/cm² SIS technology is appropriate for ultra-low noise mixers at 100 GHz. With 5,000 A/cm² SIS junctions which are now commercially available from Hypres, we expect similar quality results up to 500 GHz.

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A Technique for Noise Measurements of SIS Receivers

Qing Ke and M. J. Feldman

Abstract—We present a simple new technique to determine the noise temperature of the rf input section of a superconducting quasiparticle heterodyne receiver. This quantity is difficult to measure by existing methods. The new technique uses standard hot/cold-load measurements, and the precision should be as good as the hot/cold-load determination of receiver noise temperature. For most receivers, correction terms will be much smaller than the quantum temperature $h
\omega/k$.

I. INTRODUCTION

Lossy components in the rf input section of a heterodyne receiver contribute to the receiver noise temperature. For cryogenic receivers this rf input section noise has proven extremely difficult to measure. This is because the noise is the aggregate of thermal radiation arising from quite small losses at various temperatures ranging from room temperature down to the cryogenic operating temperature. The magnitude of each loss, and the temperature at which it is incurred, cannot be accurately estimated. This is an important problem for practical superconductor-insulator-superconductor (SIS) quasiparticle receivers, because this rf input section noise can dominate the entire receiver noise temperature.

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The authors are with the Department of Electrical Engineering, University of Rochester, Rochester, NY 14627. IEEE Log Number 9216070.
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The receiving output power for an SIS receiver is plotted for hot and cold (LO) power levels. A straight line is drawn connecting these two points, for each of five different LO levels. This figure is copied from [2].

For example, in [1] the detailed sources of noise in a 114 GHz SIS receiver were quantitatively analyzed. The input section noise temperature was determined by two separate methods. In spite of careful and elaborate measurements, both methods yielded an uncertainty in the input section noise of about 16 K, which was by far the greatest source of uncertainty in inferring the mixer noise temperature. This problem can also be severe for submillimeter wavelength receivers, even though quasi-optical inputs are generally used. This is because measurement equipment and technique are more rudimentary, and material properties are less favorable and often less well determined at the higher frequencies.

We present a new technique to determine the input section noise temperature of an SIS receiver. Our method uses standard hot/cold-load measurements of a receiver driven by local oscillator power levels less than the receiver optimum. (The possibility of using such measurements as a diagnostic was suggested in [2].) The measurement is so simple that it may be performed during each experiment, to make sure that the input section noise temperature (e.g. beam-pattern matching, junction placement, etc.) has not changed. We present both analytic and computed results from the quantum theory of mixing to assess the accuracy and the precision of this new technique.

II. THE INTERSECTING LINES TECHNIQUE

The noise temperature of a microwave receiver is generally determined, quite accurately, by the so-called Y-factor method: Hot and cold matched loads, at temperatures $T_h$ and $T_c$, are alternately placed at the receiver input and the total receiver IF output powers, $P_h$ and $P_c$, respectively, are measured. The receiver noise temperature $T_R$ is then given by the equation

$$T_R = (T_h - Y T_c)/(Y - 1),$$

where $Y \equiv P_h/P_c$. This procedure can be performed graphically: The receiver output power is plotted against the load temperature, and the straight line connecting the points $(T_h, P_h)$ and $(T_c, P_c)$ is extrapolated back to intersect the load temperature axis. The temperature given by that intersection point is the negative of $T_R$.

Blundell, et al. [2] recently described a remarkable and curious property of SIS receivers. They showed that if the hot/cold-load graphical procedure is performed for a variety of local oscillator (LO) power levels less than the receiver optimum, the hot/cold-load straight lines for each LO level all intersect at a single point. Fig. 1, copied from [2], is an example. The measured receiver output power is plotted for $T_h = 295$ K and $T_c = 77$ K and a straight line is drawn connecting these two points, for each of five different LO levels. These five lines all quite precisely intersect.

Fig. 1. The output power from an SIS receiver is plotted for hot and cold loads and a straight line is drawn connecting these two points, for each of five different LO levels. This figure is copied from [2].

The intersecting lines technique relies on the fact that the SIS mixer output noise temperature [3] is largely independent of mixer gain for low local oscillator power. Fig. 2 shows a standard diagram of a heterodyne receiver. The receiver consists of three blocks: the rf input section, the mixer, and the IF amplifier; with respective gains $G_{RF}$, $G_M$, and $G_{IF}$; and with respective equivalent input noise temperatures $T_{RF}$, $T_M$, and $T_{IF}$. Cascading these blocks, the total output power $P_T$ from the receiver when a matched load with temperature $T$ placed at the receiver input is given by

$$P_T = [(T + T_{RF})G_{RF}G_M + T_{out} + T_{IF}]kG_{RF},$$

where we define the equivalent output noise temperature of the mixer

$$T_{out} \equiv T_M G_M.$$

In functional form, (2) simply states that $P_T(T)$ is a straight line. Equation 2 could also be written $P_T = (T + T_{RF})kG_{RF}G_MG_{IF}$, so that $P_T = 0$ for $T = -T_{RF}$, which is a restatement of the graphical Y-factor method of determining $T_R$.

Let us now hypothesize that $T_{out}$ is independent of the LO power, $P_{LO}$, for low LO power levels. (We will establish the accuracy and the range of validity of this hypothesis for SIS mixers in the next section.) Then in (2) only $G_M$ will depend upon $P_{LO}$, and so $P_T$ is independent of $P_{LO}$ for some (negative) input load temperature $T = -T_X = -T_{RF}$. Therefore our hypothesis implies that we can read $T_{RF} = T_X$ directly from the intersection point on a graph like Fig. 1. This is the "intersecting lines technique." Contributions to $P_T$ are sketched in Fig. 3.

Fig. 2. A standard block diagram of a heterodyne receiver.

What if our hypothesis is not completely true? Consider the possibility that $T_{out}$ also has a component which is proportional to $G_M$. Call this component $rG_M$. Then (2) shows that the $P_T(T)$ lines for various $P_{LO}$ still intersect at a point, but in this case $T_X$ is no longer equal to $T_{RF}$ but to $T_{RF} + r$ (we take $G_{RF} \approx 1$ for

Fig. 3. Contributions to the receiver output power for interpreting the intersecting lines technique are sketched.
convenience). Reference [2] asserts that $T_M$, not $T^{\text{out}}$, is independent of $P_L$, and if this were true then $\tau$ would be equal to $T_M$.

Consider further the possibility that $T^{\text{out}}$ has a component with some more complicated dependence upon $G_M$ as $P_L$ is changed. Then (2) shows that the $P_L(T)$ lines for various $P_L$ will not intersect at a point, so $T_X$ is not defined. This happens in fact for SIS mixers at higher $P_L$ levels. But when intersecting lines are observed for lower $P_L$, then $T^{\text{out}}$ consists at most of a constant plus a term proportional to $G_M$.

If $\tau$ were large the intersecting lines technique would not directly give $T_{RF}$. To see this, it is instructive to apply the technique to the ideal exponential Schottky diode mixer receiver [4]. In this case $T^{\text{out}} = (1 - 2GM)TD$, where the diode noise temperature $TD$ is constant [5]. Equation (2) then remarkably predicts that all lines will intersect at a point, for every value of $P_L$. The intersection point, however, gives

$$T_X = T_{RF} - 2TD$$ (Schottky receiver),

which can be considerably smaller than $T_{RF}$. Nevertheless, Schottky receiver engineers are likely to find the intersecting lines technique and (4) helpful in characterizing their receivers.

IV. ACCURACY OF THE TECHNIQUE

The LO power of an SIS mixer can be characterized by the parameter $\alpha = \varepsilon P_L / h c$, where $P_L$ is the LO power across the SIS junction and $h$ is the LO frequency. For any LO power it is appropriate to expand the equations from Tucker's quantum mixer theory [6] which describe the SIS mixer in a power series in $\alpha$. This is independent of $\varepsilon$. Second, there is the "correlated" LO-induced shot noise, conventionally represented by correlated noise sources placed at the mixer's signal, image, and IF terminations; this is proportional to $\alpha^2$ to lowest order. Third, $T^{\text{out}}$ can include quantum noise, but for now let us consider a double-sideband (DSB) mixer perfectly matched to the LO source, and does not appear as a component of $T^{\text{out}}$.

This discussion shows that $T^{\text{out}}$ contains no $\alpha^2$ term, and therefore it is independent of $P_L$ for low LO power levels for an LO-matched SIS mixer. We conclude that when the hot/cold-load straight lines intersect (for low LO power) then the equation $T_{RF} = T_X$ is exact: the intersecting lines technique works perfectly.

This is not precisely true. Several processes produce small correction terms. In this section we identify the components of $T^{\text{out}}$ which are proportional to $\alpha^2$ in the small-$\alpha$ limit expansion [7]. These terms, divided by $G_M$, describe the quantum mixer theory equations if the LO is perfectly matched. The LO match is characterized by the parameter

$$g = 2(h \omega G_s / e) (1 - L_{\text{IF}}),$$

where $G_s$ is the mixer's LO, signal, and image source conductance. The mixer is perfectly matched to the LO source to lowest order in $\alpha$ when $g = 1$.

The first correction arises because the quantum noise of a mixer increases if its LO is mismatched. This excess quantum noise appears as a component of $T^{\text{out}}$ in lowest order $\alpha^2$ and gives an effective input temperature

$$\tau_2 = (g - 1)^2 / 8g T_Q$$ (DSB).

Second, the leakage current shot noise which is coupled into the IF amplifier depends upon the output conductance of the mixer, and this is affected by the LO level. This produces an $\alpha^2$ term in $T^{\text{out}}$ with an effective input temperature

$$\tau_3 = \frac{kT_G}{eV'} G_s - G_0 - 1 - g^2 T_Q$$ (DSB),

where

$$V' = I_1 / G_1$$

is generally equal to the superconductor energy gap voltage.

We have only considered a DSB mixer, for which the LO, signal, and image imbedding impedances are equal. If the sideband ratio is not unity, then a portion of the quantum noise must be generated by internal mixer processes. This quantum noise appears at the input of the mixer independent of the gain, and thus constitutes an error in the determination of $T_{RF}$.

To summarize, the analytic results of this section show that the intersecting lines technique should give $T_{RF}$ very accurately. The intersection point gives $T_X = T_{RF} + \tau$ where $\tau$ is the error in degrees $K$, is the sum of the correction terms (5)-(8). Each of these correction terms is much smaller than the quantum temperature for most published SIS receivers. The exception is if the LO is severely mismatched, i.e., if $g$ is very large or very small. But even if $g$ is as large as ten, the correction term $\tau_2$ is just equal to $T_Q$, and the other correction terms are likely to be much smaller.

V. AN EXAMPLE

Fig. 4 shows a blow-up of the intersection region for a 225 GHz receiver using a six-junction array SIS mixer in a quasi-optical dewar. The data, the output power from the receiver using hot and cold matched loads for the LO power which optimizes the receiver (line 1) and for subsequently decreasing LO power (lines 2 to 6), was supplied by S.-K. Pan of the National Radio Astronomy Observatory. Line 1 gives $T_R = 73.1 \pm 0.6 K$, taking into account the uncertainty in each power measurement of $\pm 0.5 \mu$W. We see that lines 2 to 6
Fig. 4. A blow-up of the intersection region for a 235 GHz receiver, showing the experimental precision of the intersection. The data implies that the rf input section of this SIS receiver has noise temperature $= 45.3 \pm 1.7$ K.

intersect in a rather narrow region, and that line 1 gives somewhat larger output power, as expected.

The intersection of lines 2 to 6 in Fig. 4 gives $T_{RF} = 48.3 \pm 1.7$ K, again taking into account measurement uncertainty. In earlier work [8], the rf input section noise of this receiver was determined by detailed measurements of each component along the input signal path to be $41 \pm 7$ K.

More extensive data should narrow the uncertainty in the intersection of the lines in Fig. 4 from $\pm 1.7$ K to $\pm 0.6$ K. This is the same as the uncertainty of the Y-factor measurement of $T_{H}$. Systematic errors should also be the same as for the Y-factor measurement of $T_{H}$. Note, in Fig. 4 that intersecting lines result for a wide range of LO power, up to quite close to the optimum LO power. This is a general feature of all of our computer simulations as well.

VI. CONCLUSION

We have presented an extremely simple new technique to determine the noise temperature of the rf input section of an SIS receiver. Analytic and computed results based on Tucker's quantum theory of mixing predict correction terms which are much smaller than the quantum temperature $\frac{\hbar \omega}{k}$, except when the mixer is severely mismatched to the LO source. Otherwise, the accuracy and the precision of this technique should be as good as the hot/cold-load determination of receiver noise temperature.

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REFERENCES


Experimental Performance of a Back-to-Back Barrier-N-N⁺ Varactor Tripler at 200 GHz

Debabani Choudhury, Antti V. Räisänen, R. Peter Smith, Margaret A. Freking, Suzanne C. Martin, and John K. Liu

Abstract—This paper describes the performance of planar back-to-back Barrier-N-N⁺ (bbBN) devices for frequency multiplier applications. A tripling efficiency of 3.3% has been achieved using these devices in a 200 GHz crossed waveguide mount. This is the first experimental result with a bbBN waveguide frequency multiplier. A technique has been developed for characterizing planar bbBN devices with a network analyzer, which gives both the series resistance and voltage dependent capacitance of the device. The experimental results are compared with the theoretical multiplier performance, calculated using a large signal analysis approach.

I. INTRODUCTION

Planar devices are being developed to replace whisker contacted devices as frequency multipliers and mixers to improve the ruggedness of spaceborne submillimeter wave heterodyne receivers [1]. One candidate is the planar back-to-back Barrier-N-N⁺ (bbBN) varactor device. The BNN varactor provides an improvement for high frequency applications over the Barrier-Intrinsic-N⁺ (BIN) varactor [2] due to its lower series resistance. It exhibits a very sharp change in its capacitance versus voltage resulting in very efficient harmonic generation with small input power levels. The bbBN device has symmetric C-V and anti-symmetric I-V characteristics. Impedance nonlinearities symmetric to zero bias will generate only odd harmonics, thereby simplifying the frequency multiplier mount design. For instance, a tripler mount for a symmetric device will be similar in complexity to a doubler mount for a device without symmetry. It is predicted that bbBN devices can be made to operate efficiently at frequencies over one terahertz [3]-[5].

II. DEVICE DESCRIPTION

A conceptual diagram of the device is shown in Fig. 1. The semiconductor consists of several layers: the barrier, a sheet doping layer, a moderately doped layer and a highly doped region [6]. The layer thicknesses and compositions can be adjusted for optimum performance. The device structure from the top surface down is, 1) Schottky contacts, separated by a gap, 2) an AlGaAs layer that is sufficiently thick to preclude tunneling but sufficiently thin to allow large capacitance per unit area, 3) a highly doped (delta doped) region

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D. Choudhury, R. P. Smith, M. A. Freking, S. C. Martin, and J. K. Liu are with the Center for Space Microelectronics Technology, Jet Propulsion Laboratory, California Institute of Technology, Pasadena, CA 91109.

A. V. Räisänen is with Helsinki University of Technology, Radio Laboratory, SF-02150 Espoo, Finland.

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