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Chapter 2.5

QUASIPARTICLE MIXERS AND DETECTORS

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2.5.1 Introduction

The beautiful phenomena of superconducting tunneling were discovered nearly 30 years ago. Shortly thereafter, interest developed in the use of these effects for the detection and mixing of infrared and millimeter wave radiation. The first attempts made use of Josephson pair tunneling. The experience with Josephson detectors and mixers will be mentioned briefly before turning to the more successful quasiparticle detectors and mixers.

Josephson currents respond to very high frequencies, are very nonlinear, and oscillate at a frequency $\omega_J = 2eV/\hbar$, which is proportional to the applied voltage V. They occur at LHe temperatures, so devices based on them can be expected to have low noise. The Josephson effect devices explored included a direct detector, both an internally pumped and externally pumped heterodyne mixer, several types of internally and externally pumped parametric amplifiers and an oscillator. Of these, only the externally pumped parametric amplifier and the millimeter wave oscillator are under active development today. Neither has yet seen extensive applications outside of the laboratory.

The Josephson effect detector and mixer require tunnel junctions whose capacitance is negligible at the energy gap frequency. In practice, this means point contact junctions, which are unstable and difficult to reproduce. Even with a reliable junction technology, however, these devices would not be very satisfactory. For both devices, the response to applied power scales as ω^{-2}, so decreases rapidly with increasing frequency. As a consequence, the direct detector cannot compete with the widely used composite bolometer at submillimeter wavelengths where sensitive direct detectors are needed. The noise in the Josephson heterodyne mixer is significantly larger than might be expected from the operating temperature because harmonic mixing at all frequencies up to the energy gap downconverts noise into the output. Also, because of the ac Josephson oscillations, both the Josephson detector and mixer are sensitive to the embedding admittances over a wide range of frequencies.

A different type of superconducting mixer was explored that is based on the nonlinearity of single particle or quasiparticle tunneling through a Schottky barrier between a superconductor and a semiconductor. This super Schottky diode gave very low noise at microwave frequencies. Because of the series resistance of the semiconductor, however, extension of this good performance to millimeter wavelengths proved difficult.

Faced with these problems, workers in the field turned to the nonlinearity of the quasiparticle tunneling in thin film superconductor-insulator-superconductor (SIS) tunnel junctions. The first SIS quasiparticle mixer made had lower noise than the best Josephson mixer ever made (Richards et al., 1979). The quantum theory of mixing (Tucker, 1979) developed to explain the properties of the super Schottky diode gave a complete theoretical treatment of the SIS quasiparticle mixer and direct detector. Within two years of its invention, the SIS mixer was in active use on radio telescopes (Phillips and Woody, 1982). This application has expanded and continued to the present.

The practical success of the SIS quasiparticle mixer rests on its low noise and high conversion efficiency, but also on its ability to use tunnel junctions which have significant capacitance at the signal frequency. Such junctions can be produced by optical lithography and have been developed for Josephson effect digital applications. Quasiparticle devices are biased close to the gap voltage so that the ac Josephson frequencies are comparable to the gap frequency. These high frequency Josephson currents are shorted by the junction capacitance, so do not play an essential role in the operation of the SIS quasiparticle mixer or direct detector. If mixer operation is required at a significant fraction of the gap frequency, however, Josephson effects can be troublesome. Some work has been done on mixing in superconductor-insulator-normal metal (SIN) junctions which have a weaker nonlinearity, but no pair tunneling. This SIN mixer is similar to the super Schottky diode, but with less series resistance than for the semiconductor.

This review will focus on the SIS mixer. The SIS direct detector will also be described. Frequent reference will be made to the extensive review by Tucker and Feldman (1985) when describing early work. Readers interested in more information about Josephson devices are referred to the early review by Richards (1977) and the update by Richards and Hu (1989).

Frequent reference will be made throughout this review to well known properties and figures of merit for infrared and millimeter wave devices. Direct detectors are characterized in terms of the dark current or the detector noise limit to the noise equivalent power (NEP), the responsivity S, and the dynamic range, or saturation power P_{sat}. Heterodyne receivers are described by the mixer noise temperature T_M, the mixer gain G, the IF amplifier noise temperature T_{IF}, the receiver noise temperature $T_R=T_M+T_{IF}/G$, and the saturation power P_{sat}. Receivers based on direct detectors are used for broad bandwidths and high frequencies. Receivers which use heterodyne mixers are used for narrow bandwidths and low frequencies. The crossover between these approaches is at ~300 GHz for a fractional bandwidth of 10^{-1} and ~2 THz for a fractional bandwidth of 10^{-5}. Coherent mixers, as well as direct detectors that are coupled to the electromagnetic field by an antenna, are sensitive to only one electromagnetic mode with throughput $A\Omega=\lambda^2$. Here A is the area of the focal spot and Ω the solid angle of the focus. Other detectors, such as the superconducting bolometer have a sensitive area of arbitrary size, so can accept any number of electromagnetic modes. Readers not familiar with these concepts may wish to refer to a general review of detection technology (Richards and Greenberg, 1982).

2.5.2 **Photon-assisted tunneling**

The first step in understanding the performance of quasiparticle detectors and mixers is to compute the response of the quasiparticle current to a coherent local oscillator (LO) source with constant amplitude voltage (V_{LO}).

$$V = V_o + V_{LO} \cos \omega t \; . \tag{2.1}$$

After complicated calculations (Tucker and Feldman, 1985) the tunneling current as a function of time can be written

$$I(t) = a_o + \sum_{m=1}^{\infty} (2a_m \cos m\omega t + 2b_m \sin m\omega t), \tag{2.2}$$

where

$$2a_m = \sum_{n=-\infty}^{\infty} J_n(\alpha)[J_{n+m}(\alpha) + J_{n-m}(\alpha)] \; I_{dc}(V_o + n\hbar\omega/e),$$

and (2.3)

$$2b_m = \sum_{n=-\infty}^{\infty} J_n(\alpha)[J_{n+m}(\alpha) - J_{n-m}(\alpha)] \; I_{KK}(V_o + n\hbar\omega/e) \; .$$

Here α is a dimensionless parameter proportional to the local oscillator voltage $\alpha = eV_{LO}/\hbar\omega$, and I_{KK} is the Kramers-Kronig transform of the dc I-V curve, which is defined as

$$I_{KK}(V) = P \int_{-\infty}^{\infty} \frac{dV'}{\pi} \frac{I_{dc}(V') - V'/R_n}{V' - V} \tag{2.4}$$

where $P \int_{-\infty}^{\infty}$ is the Cauchy principle value. A dc I-V curve for a Nb/Pb-alloy junction is shown in Fig. 1(a).

The time dependent current I(t) from Eqs. (2.2)-(2.4) depends only on the I-V curve of the unpumped junction $I_{dc}(V)$ and the amplitude V_{LO} of the pump. The static response of the pumped junction is

$$I = a_o = \sum_{n=-\infty}^{\infty} J_n^2(\alpha)\, I_{dc}(V_o + n\hbar\omega/e) \ . \tag{2.5}$$

This I-V curve has the form of a sum over the dc I-V curves $I_{dc}(V)$ observed without the LO pump, each displaced in voltage by an amount $n\hbar\omega_{LO}/e$, where $n=0,\pm1...$ The amplitude of each term in the sum is $J_n{}^2(\alpha)$ where $\alpha = eV_{LO}/\hbar\omega$. The sharp onset of quasiparticle tunneling thus causes a series of steps as is shown in Fig. 1(b). These steps are due to photon assisted tunneling. Whenever a multiple $n\hbar\omega$ of the photon energy makes up the difference $(eV-2\Delta)$ there will be an enhancement $(eV<2\Delta)$ or reduction $(eV>2\Delta)$ of the tunneling current. The quasiparticle is said to tunnel with the absorption or emission of one or more photons. Distinct photon assisted tunneling steps are seen only when the onset of quasiparticle tunneling is sharp on the voltage scale $\hbar\omega/e$. At low frequencies, or for rounded I-V curves, Eq. (2.5) approaches the classical limit of a dc I-V curve averaged over the LO voltage swing.

From Eqs. (2.2) and (2.3) we can see that the tunneling current has nondissipative out-of-phase components as well as dissipative in-phase components. These out-of-phase currents are a quantum-mechanical "sloshing" between coupled states whose energies cannot be matched through the absorption or emission of an integral number of photons. They produce a quantum susceptance which has no analogy in classical resistive mixers. This quantum susceptance is only important when the onset of $I_{dc}(V)$ is sharp in the voltage scale $\hbar\omega/e$. It can be either inductive or capacitive, depends on the junction properties and the bias conditions. The quantum susceptance complicates the design and optimization of SIS mixers.

When the admittance of the LO source is small, the voltage amplitude V_{LO} of the pump depends on the RF admittance of the junction which, in turn, is a function of the bias voltage for a constant pump power. Negative resistance can occur on photon assisted tunneling steps when the RF admittance of the pumped junction varies in such a way that V_{LO} decreases with increasing dc bias. The quantum theory of quasiparticle detection and mixing (Tucker and Feldman, 1985) includes the formalism required to calculate pumped I-V curves with arbitrary LO source admittance. The nonlinear quantum susceptance, which is the counterpart of the nonlinear quasiparticle conductance, plays an important role in this theory. The results of such a calculation are shown in Fig. 1(c) for a case in which negative resistance is seen (Mears et al., 1989).

2.5.3 **Quasiparticle Mixing**

A heterodyne mixer is generally used to down-convert signals from some high RF frequency to a lower frequency where amplification and further signal processing is convenient. The classical mixer makes use of the nonlinear resistance of a Schottky barrier diode that is strongly pumped by a local oscillator at ω_{LO}. The mixer produces a linear response at the intermediate frequency ω_{IF} when a small signal is supplied at the signal frequency $\omega_S=\omega_{LO}+\omega_{IF}$ or the image frequency $\omega_I=-\omega_{LO}+\omega_{IF}$. In general, currents flow in the mixer at all frequencies $\omega_m=m\omega_{LO}+\omega_{IF}$, $m=0,\pm1...$, where $\omega_S=\omega_1$, $\omega_I=\omega_{-1}$, and $\omega_{IF}=\omega_0$. The mixer performance depends on the admittances Y_m which terminate the pumped junction at all of these frequencies, or ports, as is shown in Fig. 2. In practical mixers the junction capacitance usually provides a high enough admittance that the RF voltage is zero for all ports $|m|>1$. In this case, a three-port theory, which includes only ω_S, ω_I and ω_{IF} is a very useful approximation.

The classical theory of microwave mixers assumes that the instantaneous current I(t) in a nonlinear resistance is determined by the instantaneous voltage V(t). The single sideband (SSB) conversion efficiency or gain of such a mixer is always less than 0.5. The quantum theory of mixing (Tucker and Feldman, 1985) includes classical mixer theory as a limiting case when the I-V curve is not sharp on the voltage scale $\hbar\omega/e$. When quantum effects are important, this theory predicts many unusual properties for quasiparticle mixers, including large gain and very low noise.

The quantum mixer theory is formulated in terms of an admittance matrix $Y_{mm'}$ which relates the current i_m in port m to the voltage $V_{m'}$ in port m'. A prescription is given for calculating $Y_{mm'}$ from V_{LO} and the dc I-V curve of the junction (Tucker and Feldman, 1985).

$$Y_{mm'} = G_{mm'} + iB_{mm'}, \tag{3.1}$$

where

$$G_{mm'} = \frac{e}{2\hbar\omega_{m'}} \sum_{n,n'=-\infty}^{\infty} J_n(\alpha)\, J_{n'}(\alpha)\, \delta_{m-m',n'-n}\{[I_{dc}(V_o+n'\hbar\omega/e+\hbar\omega_{m'}/e)$$

$$-I_{dc}(V_o+n'\hbar\omega/e)]+[I_{dc}(V_o+n\hbar\omega/e)-I_{dc}(V_o+n\hbar\omega/e-\hbar\omega_{m'}/e)]\}\ , \tag{3.2}$$

and

$$B_{mm'} = \frac{e}{2\hbar\omega_{m'}} \sum_{n,n'=-\infty}^{\infty} J_n(\alpha)\, J_{n'}(\alpha)\, \delta_{m-m',n'-n}\{[I_{KK}(V_o+n'\hbar\omega/e+\hbar\omega_{m'}/e)$$

$$-I_{KK}(V_o+n'\hbar\omega/e)] - [I_{KK}(V_o+n\hbar\omega/e)-I_{KK}(V_o+n\hbar\omega/e-\hbar\omega_{m'}/e)]\}\ . \tag{3.3}$$

The coupled mixer gain, $G_{\pm 1}$, is defined as the power coupled to the IF load divided by the power available from an RF source at the signal port (+1) or the image port (-1). It can be expressed in terms of the Z-matrix $[Z_{mm'}] = [Y_{mm'}+Y_m\, \delta_{mm'}]^{-1}$,

$$G_{\pm 1}(\mathrm{SSB}) = 4\mathrm{Re}(Y_{\pm 1})\mathrm{Re}(Y_0)|Z_{0\pm 1}|^2 \ . \tag{3.4}$$

This result holds for coupled mixer gain from mth port to the output if we replace the subscript "±1" with "m". It is instructive to express the coupled gain $G_{\pm 1}$ in Eq. (3.4) as a product of the available mixer gain $G_{\pm 1}{}^{o}$ and the IF coupling coefficient C_{IF},

$$G_{\pm 1} = G_{\pm 1}^{o}\, C_{IF}, \tag{3.5}$$

where

$$G_{\pm 1}^{o} = R_D \mathrm{Re}(Y_{\pm 1}) \left| \frac{Z'_{0\pm 1}}{Z'_{00}} \right|^2 \ . \tag{3.6}$$

Here the dynamic resistance R_D is dV/dI from the pumped I-V curve, and

$$C_{IF} = 1 - \left| \frac{1/Z'_{00} - Y_0^*}{1/Z'_{00} + Y_0} \right|^2 \ . \tag{3.7}$$

In Eqs. (3.6) and (3.7), Z' is the Z-matrix calculated with an open output $Y_0=0$. Eq. (3.6) suggests that the available mixer gain is approximately proportional to the dynamic resistance R_D of the pumped junction. The available gain can become infinite for infinite or negative R_D. The input admittance of an SIS mixer is of order R_N^{-1}. When the gain is large, the output admittance $(Z'_{00})^{-1} \approx R_D^{-1}$ is $\ll R_N^{-1}$. Large coupled gain is observed, therefore, only when a transformer is used to provide an IF load impedance significantly larger than 50 Ω (Räisänen et al., 1987).

Many of these predictions regarding mixer gain can be understood from a simple picture. The mixing between the RF signal and the LO can be viewed as a small modulation of the LO amplitude at ω_{IF}. This modulation produces a current I_{IF} in the junction and provides an available IF power of $I_{IF}^2 R_D/4$, resulting in a linear dependence of the mixer gain on R_D. Since R_D is a maximum at the center of each photon assisted tunneling step, the gain is expected to have peaks at these bias points. These features are clear in Fig. 3 which shows the IF output of an SIS mixer as a function of bias voltage.

2.5.4 **Mixer Noise**

The contribution of shot noise in the tunneling current of the pumped junction to mixer noise was first calculated using classical radiation fields (Tucker and Feldman, 1985). The components of the broad-band shot noise that appear at the frequencies ω_m, $|m|\geq 1$ are downconverted and superimposed on the shot noise in the IF band. Because these contributions to the output noise are correlated, destructive interference can occur. It can be shown (Feldman 1986, Wengler and Woody 1986, Devyatov et al., 1986) that for a properly tuned SIS mixer, the mixer noise temperature due to shot noise can be as low as $\hbar\omega/2k$ for a SSB mixer, and zero for a DSB mixer or a mixer pumped with two local oscillators (Wengler and Bocko, 1989).

This noise theory is incomplete, however, because it does not include the effects of photon shot noise in the input circuit. General arguments related to the Heisenberg uncertainty relation $\Delta n\, \Delta\phi > 1/2$ for photon number n and phase ϕ show that for signals in a coherent state, any phase preserving linear amplifier will add a noise power $\hbar\omega\, B/2$ for each input port (Caves, 1982).

Since frequency down-conversion with unity power gain corresponds to a large gain in photon number, these results are valid for heterodyne mixers. An exception to this general result can occur when most of the uncertainties are squeezed into one variable, leaving the fluctuations in the other variable relatively small (Yurke et al., 1988). Calculations have shown (Wengler and Bocko, 1989) that an SIS mixer pumped by two local oscillators can be operated in a phase-sensitive mode, and can thus be used to detect squeezed state signals.

When the noise power per unit bandwidth at the output of a mixer is calculated from a complete quantum mechanical treatment (Wengler and Woody, 1986) it has the form

$$\frac{P_N}{B} = \mathrm{Re}(Y_0) \sum_{mm'} Z_{0m} Z^*_{0m'} H_{mm'} + \sum_m G_m(\tfrac{1}{2}\hbar\omega_{m'}) , \tag{4.1}$$

where G_m is the conversion gain from the mth port to the output as defined in Eq. (3.4), and

$$H_{mm'} = e \sum_{n,n'=-\infty}^{\infty} J_n(\alpha)\, J_{n'}(\alpha)\, \delta_{m-m',n-n'} \{\coth[\beta(eV_0+n'\hbar\omega+\hbar\omega_{m'})/2] \times$$

$$I_{dc}(V_0+n'\hbar\omega/e+\hbar\omega_{m'}/e)+\coth[\beta(eV_0+n\hbar\omega-\hbar\omega_{m'})/2] \times$$

$$N_{dc}(V_0+n\hbar\omega/e-\hbar\omega_{m'}/e)\} . \tag{4.2}$$

The mixer noise temperature T_M referred to the input is defined as,

$$T_M(SSB) = \frac{P_N}{kG_1B} ,$$

and

$$T_M(DSB) = \frac{P_N}{k(G_1+G_{-1})B} . \qquad (4.3)$$

The first term in (4.1) is the contribution from shot noise in the junction current. The second term arises from quantization of the external circuit. It has a minimum noise power of $\hbar\omega/2$ per bandwidth at each of the mixer input ports. Assuming that the input ports of the mixer are terminated by a blackbody with physical temperature T, then in addition to the shot noise term, the minimum total power at each of the input ports is,

$$\frac{P_N}{B} = [\frac{1}{2} + \frac{1}{2} \coth(\frac{\hbar\omega}{2kT})]\hbar\omega . \qquad (4.4)$$

At T=0, Eq. (4.4) gives a minimum noise power $\hbar\omega B$ at each port of the mixer input. This minimum noise power equals the radiated power from a blackbody at temperature $T_Q=\hbar\omega/k\ln 3$. Some authors define a noise temperature T_M', which is linear in noise power, by setting $\hbar\omega B$ equal to $kT_M'B$. For this case $T_Q'=\hbar\omega/k$. Mixers have been built at 36 and 114 GHz that have measured noise temperatures within a factor 2 of T_Q (McGrath et al., 1988, Pan et al., 1988).

In addition to the intrinsic noise discussed above, SIS mixers can suffer from limitations associated with the Josephson effect. To avoid instabilities caused by the Josephson current at low bias voltages, an SIS mixer has to be biased above a threshold voltage,

$$V_T = V_{LO} + K \left(\frac{\hbar\omega 2\Delta}{e^2\omega R_N C}\right)^{1/2} , \qquad (4.5)$$

where K is a constant close to unity (Tucker and Feldman, 1985). Below this threshold the noise is high and the bias point is unstable as is seen in Fig. 3. Because an SIS junction is usually biased at $\sim\hbar\omega/2e$ below the gap, this threshold is a problem for SIS mixers at high frequencies. To avoid this instability a magnetic field is often used to quench the Josephson currents. Other possibilities include using SIS junctions with magnetic impurities in the tunnel barrier to suppress the pair tunneling, and using SIN junctions which do not have pair tunneling.

2.5.5 **Imbedding Admittance and Computer Modeling**

The importance of the imbedding admittance can be understood with the help of the equivalent circuit of an SIS mixer shown in Fig. 4. In this circuit the signal source is represented by an RF current source in parallel with a source admittance Y_S. The imbedding admittance is represented by a series admittance Y_L and a parallel admittance Y. The SIS junction is represented with its RF admittance Y_{RF} in parallel with its geometric capacitance C. The RF coupling coefficient C_{RF} characterizes the impedance matching between the source and the mixer can be calculated from the formula

$$C_{RF} = 1 - \left| \frac{Y_S - Y_J^*}{Y_S + Y_J} \right|^2 , \tag{5.1}$$

where Y_J is the total admittance on the right side of Fig. 4. The mixer gain referred to the signal source is approximately proportional to C_{RF}, and the mixer noise temperature is inversely proportional to C_{RF}. Thus C_{RF} is a key parameter for the design of SIS mixers. Since the invention of SIS mixers, much experimental work has been focused on optimizing the RF imbedding admittances.

For optimum SSB gain, the imbedding susceptance at ω_S should resonate the susceptance of the pumped junction, which arises from both the nonlinear quantum susceptance and the geometrical junction capacitance. Also, the imbedding conductance at ω_S should match the RF conductance. For DSB mixers, these conditions must be met at both ω_S and ω_I. Waveguide mixers generally have $\omega_S R_N C \gtrsim 2$ so only three mixer ports are important, and one or more mechanical tuning elements are used to obtain good RF coupling. Planar lithographed quasi-optical mixers are sometimes operated with $\omega_S R_N C \sim 1$ to provide good coupling to a resistive RF source. In this case harmonic response can be important. Alternatively they operate with $\omega_S R_N C \gtrsim 2$ and are provided with lithographed RF matching structures.

The quantum theory of mixing (Tucker and Feldman, 1985) can be used to compute the performance of an SIS mixer if the I-V curve, the dc and LO bias and the imbedding admittances are known. The unpumped I-V curve, dc bias, and available LO power can be measured directly. The theory includes a prescription for calculating V_{LO} from the available LO power. As with classical mixers, however, it is difficult to obtain adequate information about the IF and RF imbedding admittances.

Waveguide mixers for millimeter waves are often designed using lower frequency (3-12 GHz) measurements on large scale models. Scaled modeling has not yet been used for planar lithographed mixers which use RF matching structures that rely on the properties of superconductors. The geometries used in these structures, however, are often selected to facilitate direct calculations of the RF imbedding admittances.

The dependence of the shape of the pumped I-V curve on the LO source admittance shown in Fig. 1(b) and (c) can be used to deduce values for the RF admittance. The original technique (Shen, 1981) used the available pump power and the dc current to obtain allowed values of Y_{LO} in the form of circles in the admittance plane. If the input data are very precise, Y_{LO} can be obtained from the intersection of several circles. More recently, considerable success has been obtained with an automated computer search for the value of Y_{LO} which produces the best fit to an experimental pumped I-V curve, as is shown in Fig. 1c) (Mears et al., 1989).

A number of attempts have been made to compare calculations of mixer performance from quantum mixer theory with direct measurements of gain and noise. All of the qualitative effects predicted by theory have been observed. Predictions of gain have been quite successful (Feldman et al., 1983) when the onset of quasiparticle tunneling is not very sharp on the voltage scale $\hbar\omega/e$. Predictions of noise are less successful, but still frequently agree within a factor 2. Substantial disagreements between theory and experiment are often found when the I-V curve is very sharp, especially with regard to the conditions under which infinite gain is available (McGrath et al., 1988). It is possible that comparisons in the quantum limit are particularly sensitive to errors in the imbedding admittances.

2.5.6 LO Power and Saturation

An estimate of the V_{LO} required to pump a mixer biased on the nth photon step can be obtained from the value of $\alpha_n = eV_{LO}/\hbar\omega$ which corresponds to the first maximum of $J_n(\alpha)$. Since the input impedance of an SIS mixer is of the order of its normal resistance R_N,

$$P_{LO} \approx (N\hbar\omega\alpha_n/e)^2/2R_N \quad . \qquad (6.1)$$

This equation includes the case of a mixer which uses a series array of N junction with total normal resistance R_N. It is found to account for the observed P_{LO} to within 2dB for experimental values which range from 1nW for single junction mixers to 30µW for array mixers (Tucker and Feldman, 1985). These low values of P_{LO} are a great convenience for SIS mixers, especially at submillimeter wave frequencies. It appears possible that the Josephson local oscillator can be used to pump an SIS mixer (Lukens et al., 1988).

Because of the small values of P_{LO} required, quasiparticle mixers saturate at relatively low signal levels. For this reason, SIS mixers are limited to small signal applications. Saturation first occurs in the mixer output because of the rapid dependence of mixer gain on bias voltage shown in Fig. 3. The IF response of the mixer can be viewed as a modulation of the instantaneous bias point at frequency ω_{IF}. When the amplitude V_{IF} of this IF voltage swing reaches some fraction γ of the width $N\hbar\omega/e$ of the gain peak, the average gain is suppressed. If the mixer is matched at the IF output, the input RF power that will cause saturation can be written

$$P_{sat} = (\gamma N\hbar\omega/e)^2/2GR_D \ . \qquad (6.2)$$

For a single junction mixer with R_D=50Ω, G=3dB and γ=0.1 (which corresponds to 0.2dB gain compression) this expression gives P_{sat}=2pW, in agreement with a measured value of 1.5pW (Smith and Richards, 1982). For a quantum noise limited mixer with unity gain and a 500 MHz IF bandwidth, the dynamic range is 20dB at 36 GHz. This is large enough for most astronomical applications. However, it might be insufficient for radar and communication systems.

Although problems with saturation from 300K noise can occur in broad-band SIS mixers, such extreme problems can be avoided by several techniques. The use of a series array of N junctions will increase P_{sat} by N^2. Also, if the coupled RF bandwidth is larger than ω_{IF}, then the V_{IF} that comes from a broad-band signal can be reduced by the use of a low pass filter that shorts the mixer output for frequencies above ω_{IF} (Wengler et al., 1985).

2.5.7 **Series Arrays of Junctions**

Some freedom is introduced into the design of quasiparticle mixers by the possibility of using arrays of junctions in series. If the RF currents have the same phase in N identical junctions, then the equivalent circuit of the N-junction array can be reduced to that of a single effective junction with normal state resistance and series inductance increased by the factor N and capacitance decreased by the same factor. The voltage scale of the I-V curve is increased by the factor N. If junctions with the same tunnel barrier are used, the response time $R_N C$ is unaffected. In order to retain impedance matching at RF and IF, the junctions in the array should have areas N times larger than for the single junction mixer. Measurements of the performance of array mixers scaled in this way show that mixer gain and noise can be independent of N up to at least N=25 (Crété et al., 1987).

Advantages of array mixers include a saturation level and dynamic range which scale as N^2, and junction areas which scale as N. The relaxation of fabrication requirements for the larger junctions is partly offset, however, by the need for nearly identical junctions. One disadvantage of array mixers is that the series inductance L of the array scales as N. Limits to the operating frequency ω_S of SIS mixers set by $(LC)^{-1/2}$ or R_N/L can be troublesome when arrays are used at high frequencies.

2.5.8 **Types of Junctions**

The SIS quasiparticle mixer depends on the availability of tunnel junctions with a well defined onset of quasiparticle current at $V=2\Delta/e$, values of resistance $20 \leq R_N \leq 100\ \Omega$ that can be matched at RF and IF frequencies, and small enough capacitance that the relaxation time R_NC can meet the criterion $1 \leq \omega_S R_N C \leq 10$. When $\omega R_N C$ is held fixed, the Josephson critical current density, which is an exponential function of barrier thickness, scales directly with frequency. This easily measured parameter is ~500A/cm^2 at 100 GHz. Since the spread of useful barrier thicknesses is only 10%, the specific capacitance depends only on the type of barrier used. The value of 40 fF/μm^2 for the oxides of Pb-In alloys (Magerlein, 1981) and 45 fF/μm^2 for Al_2O_3 (Gurvitch et al., 1983) and the value of 140 fF/μm^2 for the higher dielectric constant oxides of Nb (Magerlein, 1981) and Ta (McGrath et al., 1988) can be used for design purposes. The junction areas required scale inversely with operating frequency and are typically 1-4 μm^2 at 100 GHz.

A variety of approaches have been used to fabricate the small junction areas required. These include photoresist lift-off to produce window junctions with areas of 1-4 μm^2, photoresist bridge masks to produce overlap junctions with areas of 0.5-2μm^2, and edge techniques for areas less than 0.5μm^2.

Most early mixer experiments made use of Pb-alloy junctions with In-oxide barriers. These junctions usually degrade gradually when stored at room temperature. Junctions made from Nb/Nb-oxide/Pb-alloy are more durable, but suffer from the higher dielectric constant of the Nb-oxide. All-Nb junctions with artificial barriers such as Al_2O_3, MgO and α-Si are becoming available that combine ruggedness with a low barrier dielectric constant. A few experiments have been done with Sn/Sn-oxide/Sn and Ta/Ta-oxide/Pb-alloy junctions which have very low leakage current and a very sharp onset of quasiparticle tunneling.

The requirement of small current flow at voltages below $2\Delta/e$ sets an upper limit of $\sim T_c/2$ on the operating temperatures that can be used for SIS mixers. Since there are significant conveniences to operation with unpumped LHe at 4.2K, or with mechanical refrigerators (which achieve temperatures below ~4.5K only with difficulty) there are benefits from the use of Nb junctions with T_c=9K compared with ~7K for the Pb alloys. In the future, the cryogenic problem will be eased by the availability of NbN junctions with T_c=15K.

It is interesting to speculate on the usefulness of SIS quasiparticle mixers made from the new superconductors with much higher values of T_c. In the radio astronomy applications, higher operating temperatures would be an advantage only if the noise temperature does not also increase. Since noise temperatures of receivers that use cooled Schottky diode mixers are less than 10 times those of the best SIS receivers at W-band, a millimeter wave high T_c mixer would have competition if T_M is degraded significantly. The energy gap limitation to the operating frequency of SIS mixers could be significantly relaxed by the use of high T_c superconductors. Operation at frequencies above one THz, however, will require extremely small junctions with area $<0.1 \mu m^2$ and very high current densities $\sim 10^5 A/cm^2$. At present there is no appropriate high T_c SIS junction technology.

2.5.9 **Quasiparticle Mixer Measurements**

Measurements of the performance of SIS receivers are generally made by coupling in signals from hot and cold loads at ~300K and 77K, and by measuring the output of the cold IF amplifier on a spectrum analyzer. Coherent sources are used to test the relative gains for the signal and image ports. A bi-directional coupler is frequently introduced at the output of the mixer to measure the impedance mismatch at the output, which is important for receiver optimization. A coherent IF signal from an external source can then be reflected from the mixer output to evaluate the IF coupling and signals can be introduced to measure the gain and bandwidth of the IF amplifier. The complications of cryogenic operation make it difficult to obtain the accurate measurements of the performance of the isolated mixer that are required to test quantum mixer theory. Special techniques such as cryogenic hot-cold loads at both the RF and IF ports have been developed for this purpose (McGrath et al., 1986).

Performance of Waveguide SIS Mixers and Receivers

Soon after the first mixing experiments were reported in 1979 (Richards et al., 1979, Dolan et al., 1979), SIS quasiparticle mixers began to replace Schottky diode mixers in coherent receivers for molecular line radio astronomy. These receivers are now in daily use on millimeter wave telescopes and interferometers in at least six observatories. Portable line receivers for submillimeter wavelengths are being developed for use on mountain top and airborne telescopes. Other applications include atmospheric line measurements and radiometers for measurements of the anisotropy of the cosmic microwave background (Timbie and Wilkinson, 1988). The SIS quasiparticle heterodyne mixer is now the technology of choice for sensitive coherent receivers from the ~40 GHz upper limit of high electron mobility transistor (HEMT) amplifiers (Pospieszalski et al., 1988) to more than 800 GHz.

More than one hundred papers have been published describing the performance of SIS quasiparticle mixers and receivers. A few highlights of these developments will be described here, starting with the waveguide mixers which are typically used at frequencies below ~400 GHz.

Good coupling to both the real and imaginary parts of the RF mixer admittance is most easily obtained by the use of a waveguide mount with two mechanical adjustments. Early evaluations of the potential of SIS mixers done in this way gave significantly better performance than was obtained from early mixer blocks with one mechanical adjustment.

Much attention has been given to careful optimization of W-band mixers (75-110 GHz) to achieve broad instantaneous bandwidth, broad tuning range and optimum termination of the image and harmonic ports (Woody et al., 1985, Räisänen et al., 1986, Pan et al., 1988). Current practice often makes use of integrated tuning elements lithographed on the junction substrate, which can take the form of lumped or distributed circuit elements (D'Addario, 1984, Räisänen et al., 1986, Kerr et al., 1988). An example of such a mixer with two mechanical tuning elements (Pan et al., 1988) is shown in Fig. 5. A broad tuning range with good instantaneous bandwidth has also been obtained with only one mechanical adjustment by the use of suitable lithographed tuning elements (Räisänen et al., 1986).

The lowest noise and highest gain thus far obtained from SIS mixers have come from waveguide devides. One experiment in K_A band at 36 GHz with very high quality Ta junctions and two experiments at 100 GHz with Pb-alloy junctions gave noise temperatures of T_m(SSB)=3.6K (McGrath et al., 1988), T_m(DSB)=5.6K (Pan et al., 1988), and 6.6K (Räisänen et al., 1986), respectively, which are within a factor two of the quantum limits for these frequencies. Measurements of a W-band mixer with a small IF load admittance gave values of coupled gain as large as 12.5dB (Räisänen et al., 1987). The observation of such large coupled gain is an interesting confirmation of quantum mixer theory. Because of the low noise available from HEMT IF amplifiers, however, gains of order unity are more appropriate for practical receivers.

Despite the progress that has been made in optimizing W-band SIS quasiparticle receivers, the noise temperatures of the receivers on telescopes are a factor 10 or more above the quantum limit. There is room for improvement before the sky temperature limit is reached. A summary (Büttgenbach et al., 1988) of some of the best reported results is shown in Fig. 6.

Waveguide SIS mixers have been constructed at frequencies up to 345 GHz by several groups (Ibruegger et al., 1987, Blundell et al., 1988, Sutton, 1988). Because of increased waveguide loss and increased difficulty of fabricating precise structures for these frequencies, simpler mixer blocks are often used with a single mechanical adjustment and a circular waveguide as is shown in Fig. 7. Many of the best results have been obtained by operating at $\omega R_N C \sim 1$ with submicron junctions and no lithographed tuning elements. Although the noise temperatures of these receivers shown in Fig. 6 do not approach the quantum limit as closely as do the W-band receivers, their performance is good enough to produce very valuable astronomical data.

Interference from Josephson tunneling phenomena becomes increasingly troublesome as the operating frequency is increased. Even in a magnetic field, the value of P_{LO} must sometimes be limited so that the instability described in Eq.(4.5) does not interfere with operation on the first photon step below the gap. An encouraging noise temperature of $T_M(DSB) \approx 200K$ has been obtained at 230 GHz with an SIN mixer (Blundell and Gundlach, 1987) which avoids this problem.

Quasioptical SIS Receivers

Thin film SIS tunnel junctions are compatible with other lithographed superconducting receiver components such as planar antennas, transmission lines, and filters. It is therefore attractive to use optical lithography to make integrated planar quasioptical receivers at high frequencies so as to avoid the fabrication problems associated with waveguide structures. Since a planar antenna located on a dielectric surface radiates primarily into the dielectric, the RF signals are introduced through the back surface of the dielectric, which is curved to form a lens as is shown in Fig. 8. Early work on planar integrated SIS mixers began with bow-tie antennas, but attention has shifted to log-periodic and spiral antennas which have more symmetrical central lobes and so can couple more efficiently to telescopes. All three are self-complementary and so have real impedances of 120 Ω when deposited on quartz (Rutledge et al., 1983).

As is shown in Fig. 6, very good performance has been obtained over the extremely broad bandwidth from 100 to 760 GHz with a planar quasioptical SIS receiver (Buttgenbach et al., 1988). This mixer used a single Pb-alloy junction with $\omega_S R_N C=1$ at 300 GHz and a spiral antenna. Saturation on 300K noise was avoided by shorting the mixer output for frequencies above ω_{IF}.

The future appears very bright for planar quasioptical SIS mixers for frequencies up to and beyond 1 THz, especially if junctions can be made from superconductors such as NbN with small enough areas to match the antenna resistance. These severe requirements on junction fabrication can be eased by the use of lithographed matching structures such as those shown in Fig. 9. Such structures have been used on planar quasioptical mixers with bow-tie and log-periodic antennas at frequencies from 90 to 270 GHz (Li et al., 1988, Hu et al., 1989). They are used to resonate the junction capacitance over RF bandwidths of 5-25 percent and thus to permit the efficient use of junctions with $\omega_S R_N C \gtrsim 10$. Since conventional microwave test apparatus is not available at such high frequencies, special techniques are used to evaluate the RF coupling provided by such structures. These have included using a Fourier transform far-infrared spectrometer as a sweeper and the mixer junction a direct detector (Hu et al., 1988).

2.5.10 **Quasiparticle Direct Detector**

Theory

The quasiparticle direct detector, also called a video or square-law detector, uses the nonlinearity of the quasiparticle I-V curve of an SIS junction to rectify the coupled RF signal. The current responsivity $S_I = \Delta I_{dc}/P_S$ of such a detector is defined as the induced change in the static current, ΔI_{dc} divided by the available signal power P_S. In the quantum theory (Tucker and Feldman, 1985) the current responsivity is obtained from the small signal limit of the theory of the pumped I-V curve discussed in Section 2.5.2,

$$S_I = \frac{e}{\hbar\omega} \left| \frac{I_{dc}(V_o+\hbar\omega/e) - 2I_{dc}(V_o)+I_{dc}(V_o-\hbar\omega/e)}{I_{dc}(V_o+\hbar\omega/e) - I_{dc}(V_o-\hbar\omega/e)} \right| . \qquad (10.1)$$

If the RF source is not matched to the detector, the current responsivity will be reduced from (10.1) by the RF coupling coefficient C_{RF} defined in Eq.(5.1). In the small signal limit, the RF conductance and susceptance of an SIS junction can be calculated analytically,

$$G_{RF} = \frac{e}{2\hbar\omega} [I_{dc}(V_o + \hbar\omega/e) - I_{dc}(V_o - \hbar\omega/e)] , \quad (10.2)$$

$$B_{RF} = \frac{e}{2\hbar\omega} [I_{KK}(V_o + \hbar\omega/e) - 2I_{KK}(V_o) + I_{KK}(V_o - \hbar\omega/e)] , \quad (10.3)$$

where I_{KK} is the Kramers-Kronig transform of the dc current defined in Eq.(2.4).

The quantity in the square brackets in Eq.(10.1) is the second difference of the unpumped I-V curve computed for the three points $V=V_o$ and $V_o \pm \hbar\omega/e$, divided by the first difference computed between $V=V_o \pm \hbar\omega/e$. In the classical limit, where the current changes slowly on the voltage scale $\hbar\omega/e$, the differential approximation gives the usual result for a diode detector $S_I=(d^2I/dV^2)/2\ (dI/dV)$. If the I-V curve is sharp enough that the current rise at 2Δ occurs within the voltage scale $\hbar\omega/e$ and if the bias voltage V_o is just below $2\Delta/e$, then Eq.(10.1) becomes $S_I=e/\hbar\omega$. This quantum limit to the responsivity corresponds to one extra tunneling electron for each coupled photon. The SIS direct detector makes a continuous transition between the classical energy detector and the quantum photon detector.

Since direct detectors do not preserve phase, there is no quantum limit to the detector noise analogous to that for the mixer. The intrinsic noise limit of the quasiparticle direct detector is the shot noise $<I_N^2>=2eI_{dc}(V_o)B$ in the dark current I at the bias point. Here B is the post-detection bandwidth. The noise equivalent power (NEP) in $WHz^{-1/2}$ of an RF-matched SIS direct detector is then

$$NEP = <I_N^2>^{1/2}/S_I B^{1/2} = [2eI_{dc}(V_o)]^{1/2}/S_I \ . \qquad (10.4)$$

In the quantum limit, the NEP increases linearly with signal frequency ω_S.

When the signal power is increased, the responsivity of an SIS detector falls (Feldman and D'Addario, 1987) as $1-P_S/P_{sat}$, where

$$P_{sat} \approx 16[(\hbar\omega/e)^2/2\ R_{RF}] \ . \qquad (10.5)$$

This expression for saturation power is $16G/\gamma^2N^2$ times the corresponding Eq.(6.2) for an SIS mixer. For a single junction mixer this factor is ~10^3. It can be significantly smaller for mixers made with many junctions in series. Series arrays of junctions are not useful for SIS detectors, because the responsivity is reduced and the NEP increased by the factor N.

Detector Performance

The first experimental test of an SIS direct detector (Richards et al., 1980) showed excellent agreement with the quantum theory. The current responsivity of 3.6×10^{3}A/W was within a factor 2 of the quantum-limited value $e/\hbar\omega$ at 36 GHz. Figure 10 shows the measured and calculated current responsivity S_I as a function of bias voltage. Similar results are reported at W-band (Feldman and D'Addario, 1987). In these experiments, the shot noise was measured with amplifiers at 50 MHz and 1.4 GHz which were designed for use as IF amplifiers for heterodyne mixers and found to agree with theory. The NEP was deduced to be 2.6×10^{-16}WHz$^{-1/2}$ at 36 GHz (Richards et al., 1980), which is essentially equal to the performance of a millimeter wave astronomical radiometer based on the ^{3}He-cooled composite bolometer. In the usual radiometric applications the signal is modulated at some low frequency $1<f<100$ Hz. A receiver for such signals is sensitive to 1/f noise at the frequency f, which is commonly observed in tunnel junctions. Also, an amplifier at frequency f must be used that does not contribute significant excess noise for a source resistance of a few hundred ohms. Even if these problems are solved, an astronomical radiometer based on the SIS direct detector will not be significantly more sensitive at millimeter wavelengths than the SIS heterodyne radiometer which uses a Schottky diode detector at the output of the IF amplifier (Weinreb, 1986). It will be less sensitive at submillimeter wavelengths than the ^{3}He-cooled bolometric radiometer.

The potential remains for applications of the SIS direct detector at near-millimeter and submillimeter wavelengths which benefit from its higher operating temperature and faster speed when compared with the composite bolometer. It is easier to make in planar arrays than either the composite bolometer or the SIS mixer.

A novel radiometer configuration has been suggested (Richards, 1989) which uses one SIS junction pumped with an LO as a heterodyne downconverter followed by a second SIS junction used as a photon detector. From one viewpoint, the SIS mixer is a preamplifier for the photon detector. It amplifies the photon rate by the product of its power gain and the frequency downconversion ratio. The SIS photon detector is used at the relatively low IF frequency where its NEP is better than at the RF. From another viewpoint, this system is a simplified version of a conventional heterodyne radiometer. Because of the conversion gain of the SIS mixer and the excellent performance of the SIS direct detector, there is no need for an IF amplifier. This configuration appears to have higher sensitivity than the SIS direct detector while retaining its high speed and operating temperature. It appears to be simple enough that planar integrated arrays of detectors are possible.

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FIGURE CAPTIONS

Fig. 1. Line (a) is the experimental I-V curve for a Nb/Pb-alloy SIS junction at $T \ll T_c$ traced in the direction of decreasing current. The Josephson current at zero voltage seen when the I-V curve is traced in the direction of increasing current is not shown.

Dotted line (b) is the same I-V curve pumped at 90 GHz with an LO source whose admittance is high compared with R_N^{-1} (constant voltage source).

Solid line (b) is the photon assisted tunneling calculated from the dc I-V curve and Eq.(2.2)

Dotted line (c) is the same junction pumped from a 90 GHz LO source whose admittance is small compared with R_N^{-1}.

Solid line (c) is the photon assisted tunneling calculated from the quantum theory of mixing with the admittance at the LO port adjusted for the best fit (Mears et al., 1989).

Fig. 2. Functional circuit for a mixer made using a nonlinear diode pumped at ω_{LO}. The imbedding admittances of the mixer at frequencies $\omega_m = m\omega_{LO} + \omega_0$, where $\omega_0 = \omega_{IF}$, $\omega_1 = \omega_S$ and $\omega_{-1} = \omega_I$, are given by Y_m. There is a small signal applied at $\omega_1 = \omega_S$. Noise can drive the mixer at other frequencies.

Fig. 3. Curves describing the performance of an SIS quasiparticle mixer (Ibruegger et al., 1984). The line labeled 1 is the unpumped I-V curve. Line 2 is the I-V curve when pumped with a LO at 150 GHz. Line a is the IF output power with a 300 K load at the input and line b is the output power for a 77 K

load. The mixer is unstable for bias V<2.2 mV because of Josephson effects. Measured values of T_R(DSB) were 102 K on the first photon peak and 81 K on the second photon peak below $2\Delta/e$.

Fig. 4. Equivalent circuit of an SIS mixer. The signal source is represented by a current source in parallel with its admittance Y_S. The SIS junction is represented by its quasiparticle RF admittance Y_{RF} in parallel with its geometric capacitance C. The parasitic inductance Y_L arises mainly from the inductance of the leads to the junction, and Y is the total admittance provided by the tuning structure.

Fig. 5. Cross section of a W-band SIS quasiparticle mixer with two mechanical adjustments for RF matching (Pan et al., 1988). The SIS junction with a lithographed RF matching structure is deposited on a Si chip that is bonded to the suspended stripline used to provide dc bias and IF output.

Fig. 6. A summary of some of the best results for the single sideband noise temperature of SIS quasiparticle and Schottky diode heterodyne receivers (Büttgenbach et al., 1988).

Fig. 7. Scalar feed horn and mixer block used for an SIS quasiparticle mixer from 85 to 115 GHz (Woody et al., 1985). This mixer uses circular waveguide and a single mechanical tuning element for RF matching.

Fig. 8. Cross section of the optical system used with a planar lithographed SIS quasiparticle mixer (Wengler et al., 1985). The junction and antenna are located on a quartz substrate attached to the back surface of a quartz lens.

Fig. 9. Layout of window junctions and lithographed RF matching structures used in quasioptical SIS quasiparticle mixers from 90 to 270 GHz (Hu et al., 1989).

Diagram a) shows a series array of five junctions with a parallel wire inductor terminated in an open-ended $\lambda/4$ microstrip stub.

b) Shows a single junction with an inductive open-ended $3\lambda/8$ stub.

c) Shows a single junction with an inductive $\lambda/8$ stub whose end is RF shorted by an open-ended $\lambda/4$ stub.

Fig. 10. (a) Measured dc I-V curve of an ~4 μm^2 Pb-In-Au alloy SIS tunnel junction at 1.4 K.

(b) Measured and calculated responsivities of an SIS quasiparticle direct detector made from the above junction as a function of bias voltage.

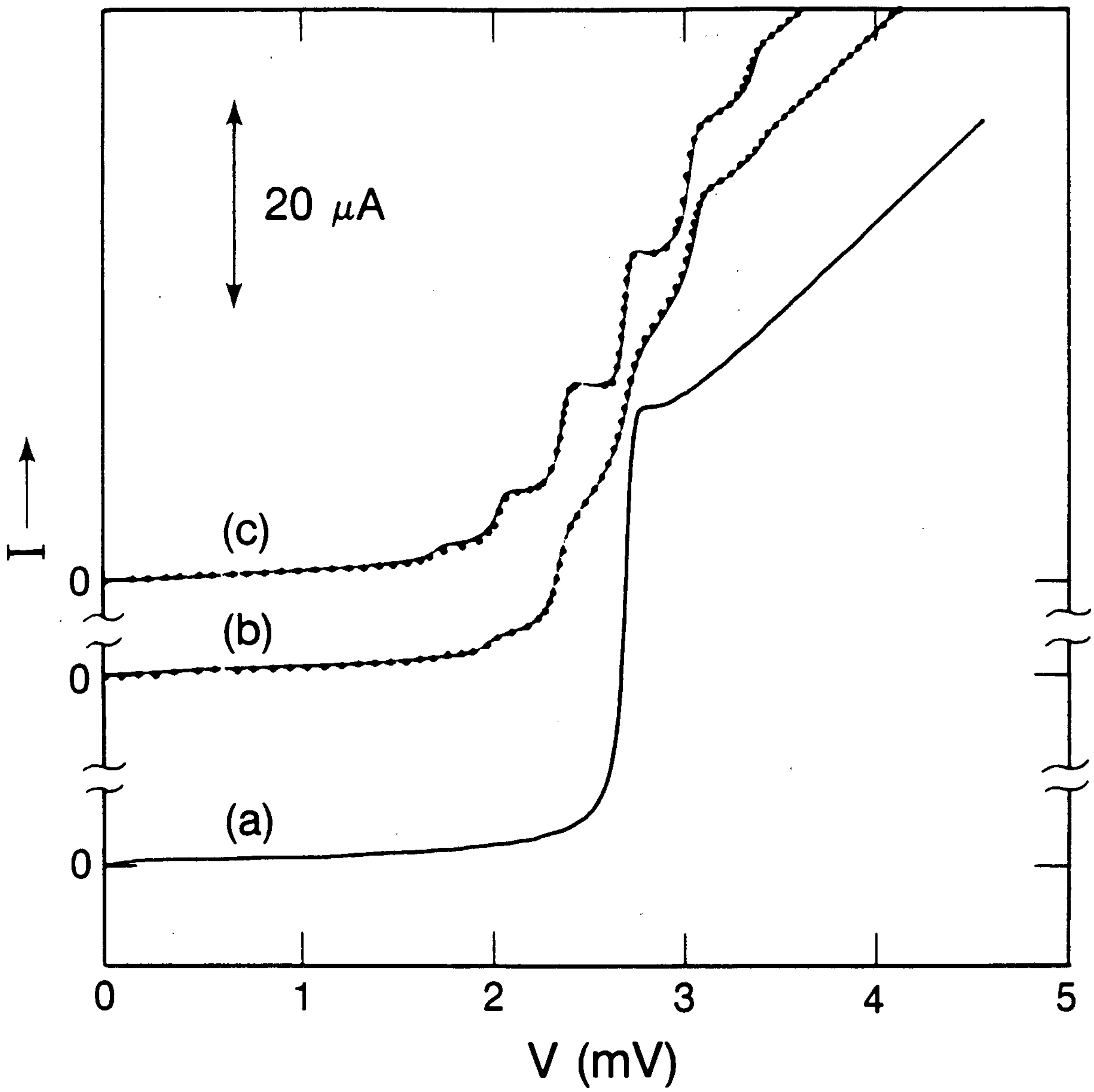

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FIGURE 1

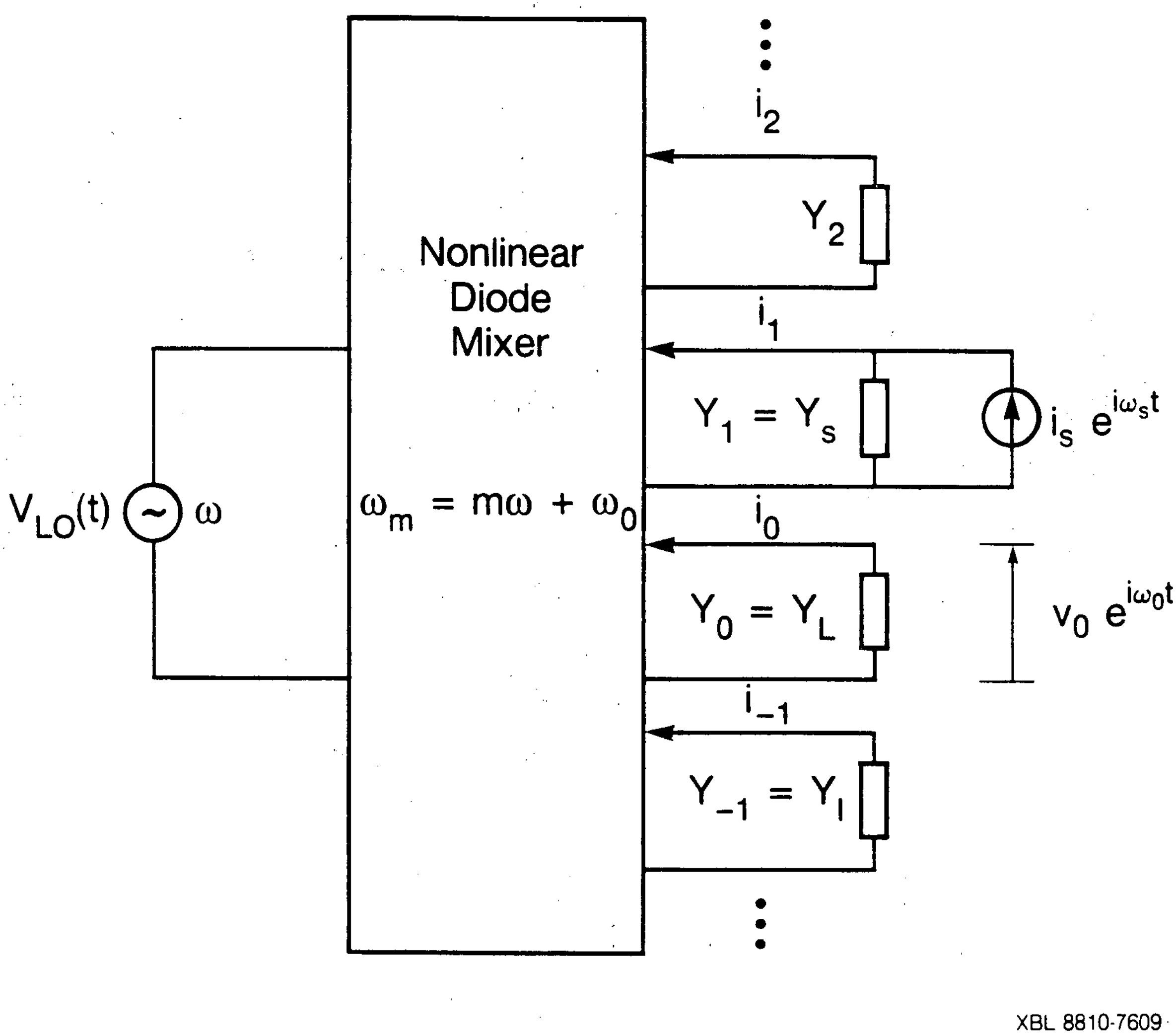

Nonlinear
Diode
Mixer
ωm = mω + ω0
VLO(t)
ω
i2
Y2
i1
Y1 = Ys
is e^iωst
i0
Y0 = YL
v0 e^iω0t
i−1
Y−1 = YI
XBL 8810-7609

FIGURE 2

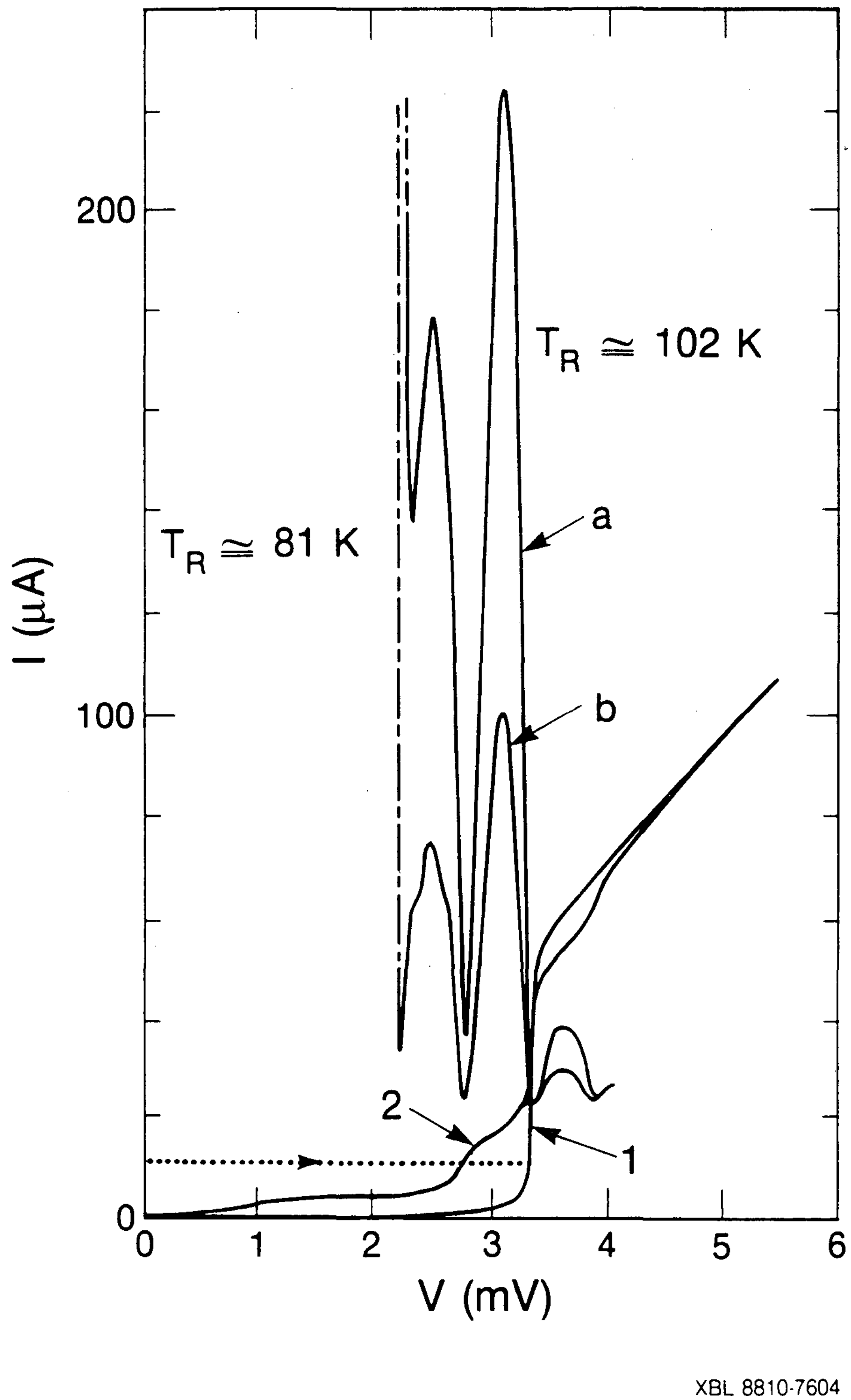
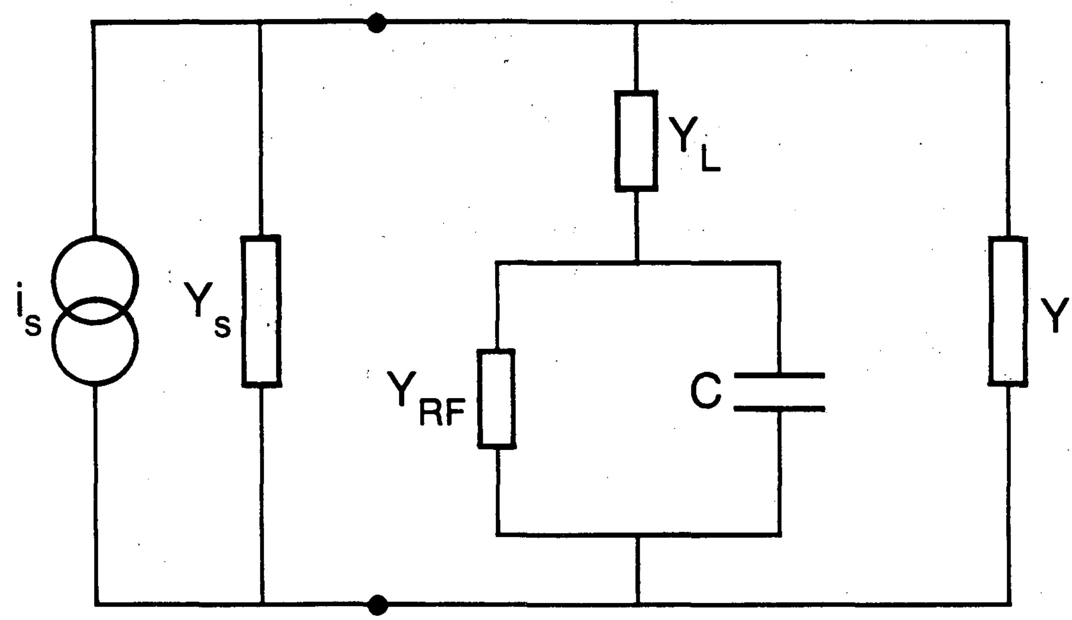
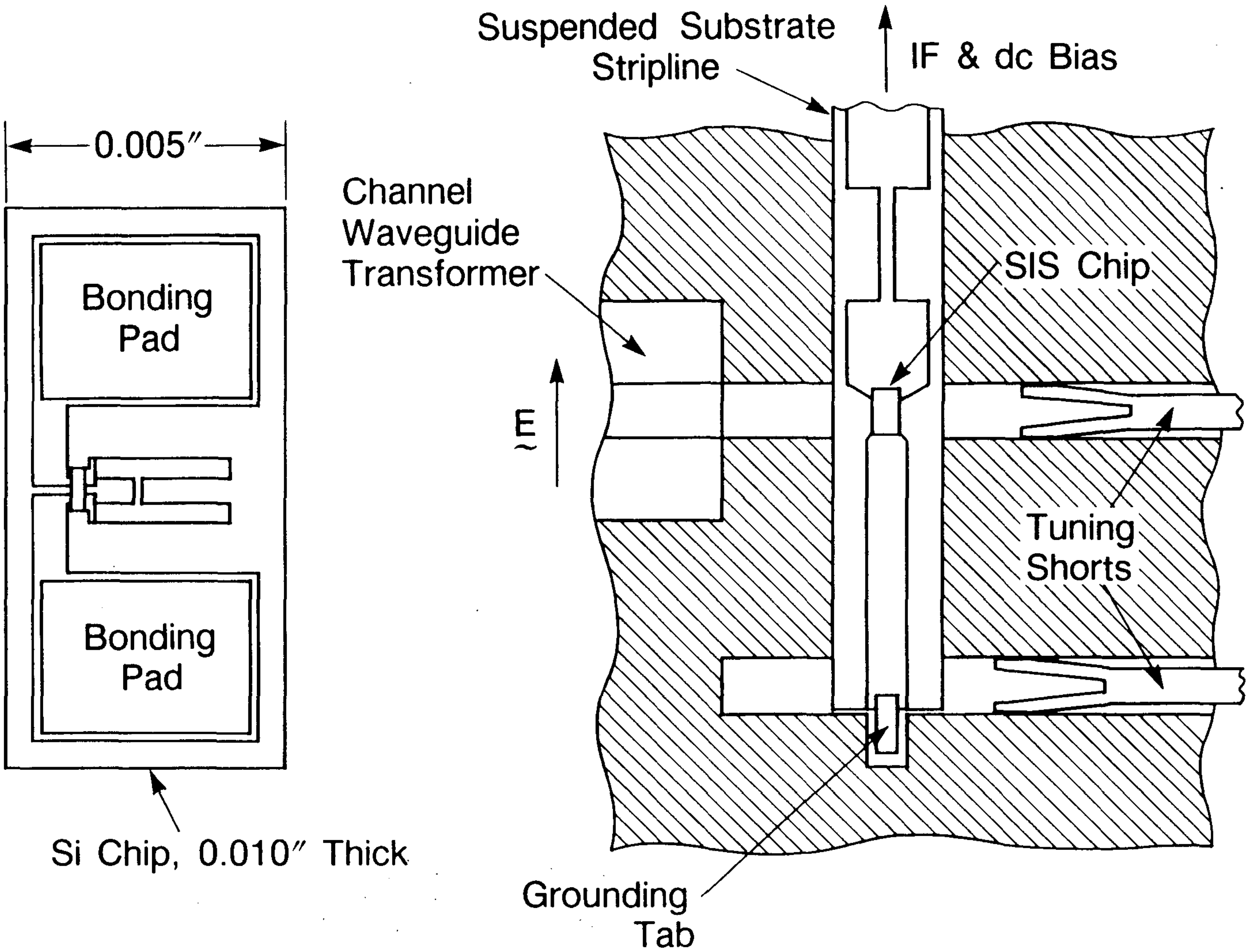


FIGURE 3



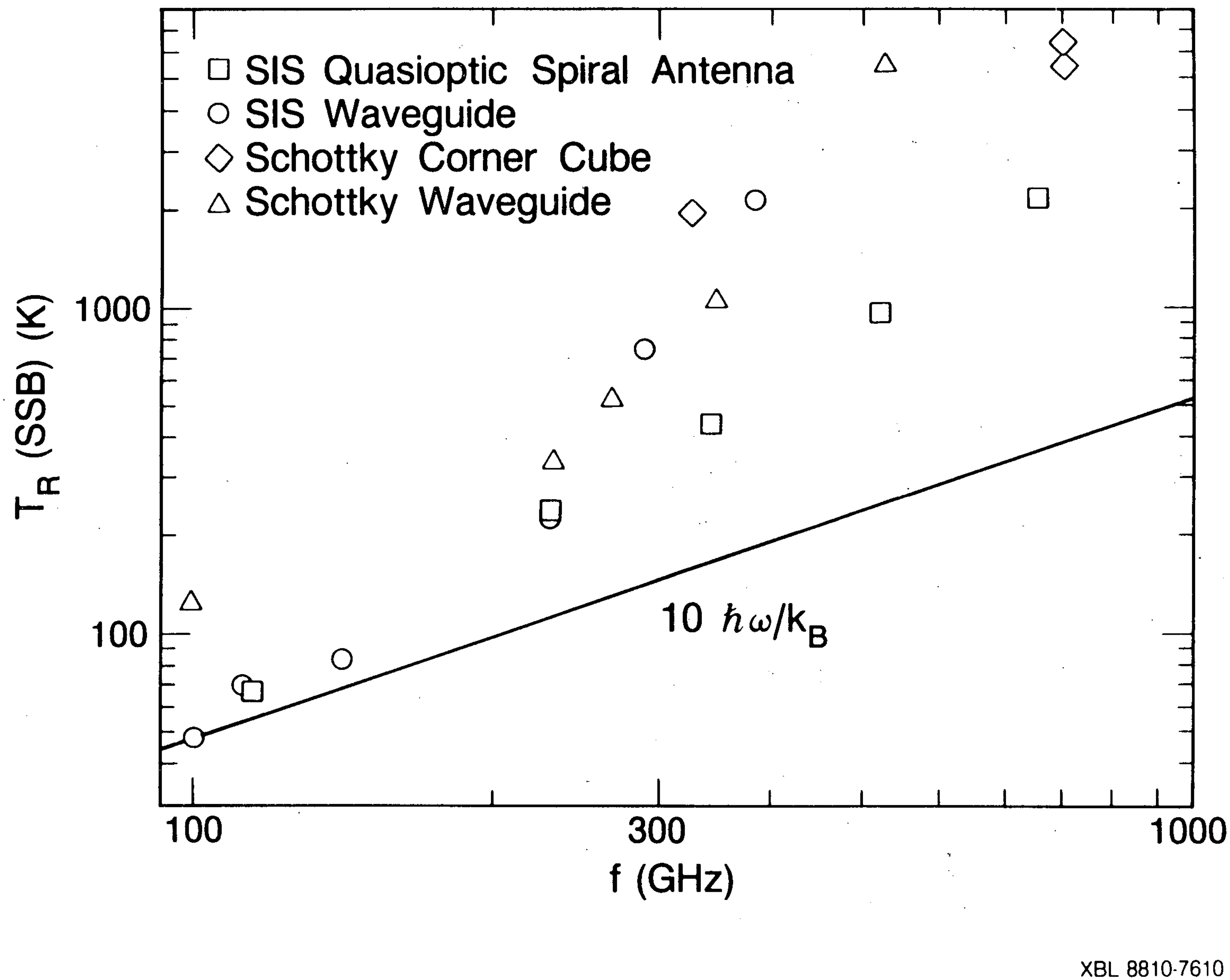
XBL 8810-7606

FIGURE 4



XBL 8810-7608

FIGURE 5



SIS Quasioptic Spiral Antenna
SIS Waveguide
Schottky Corner Cube
Schottky Waveguide
1000
100
T_R (SSB) (K)
10 $\hbar\omega/k_B$
100
300
1000
f (GHz)
XBL 8810-7610

FIGURE 6

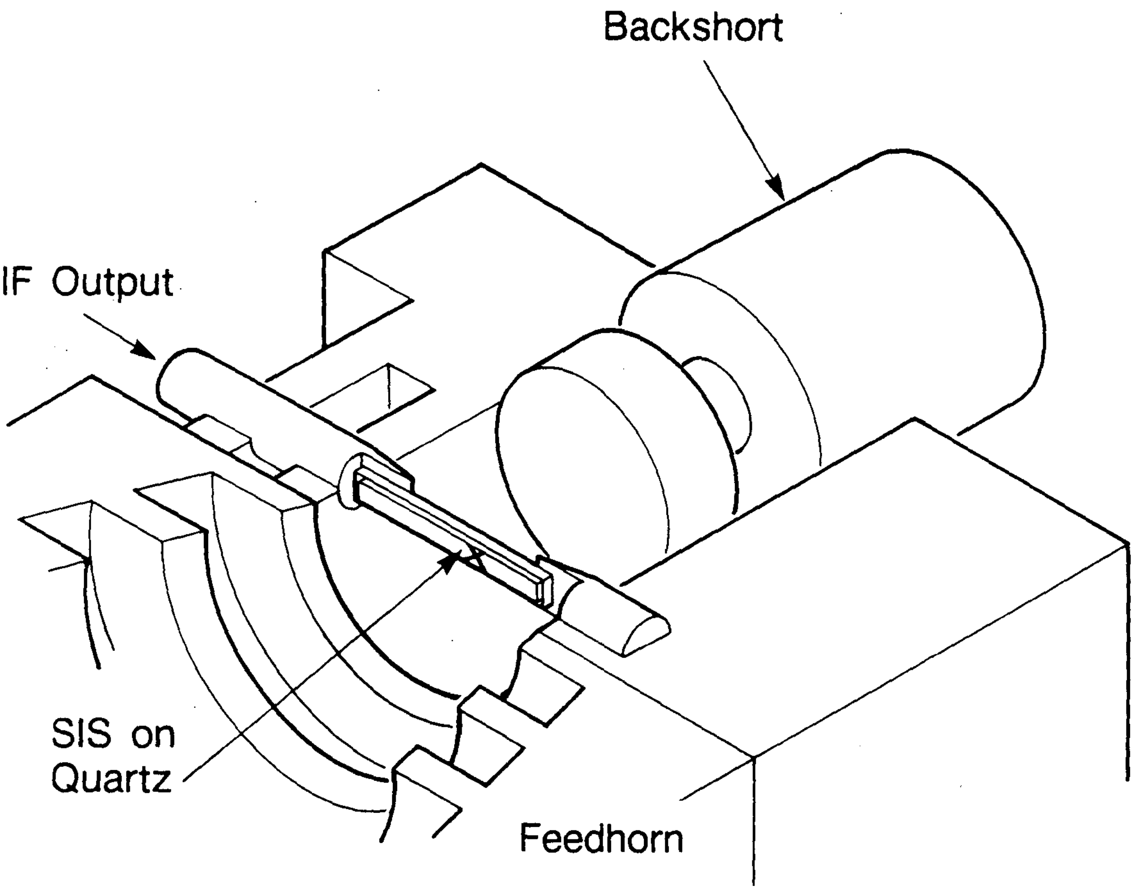


FIGURE 7

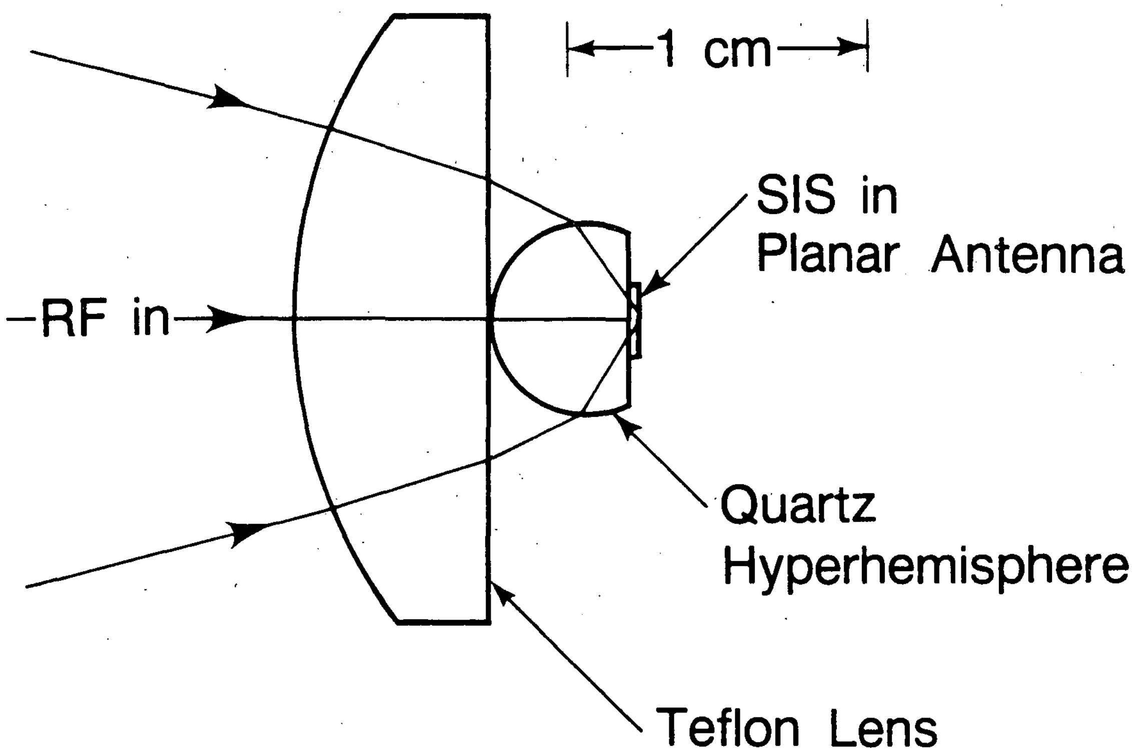
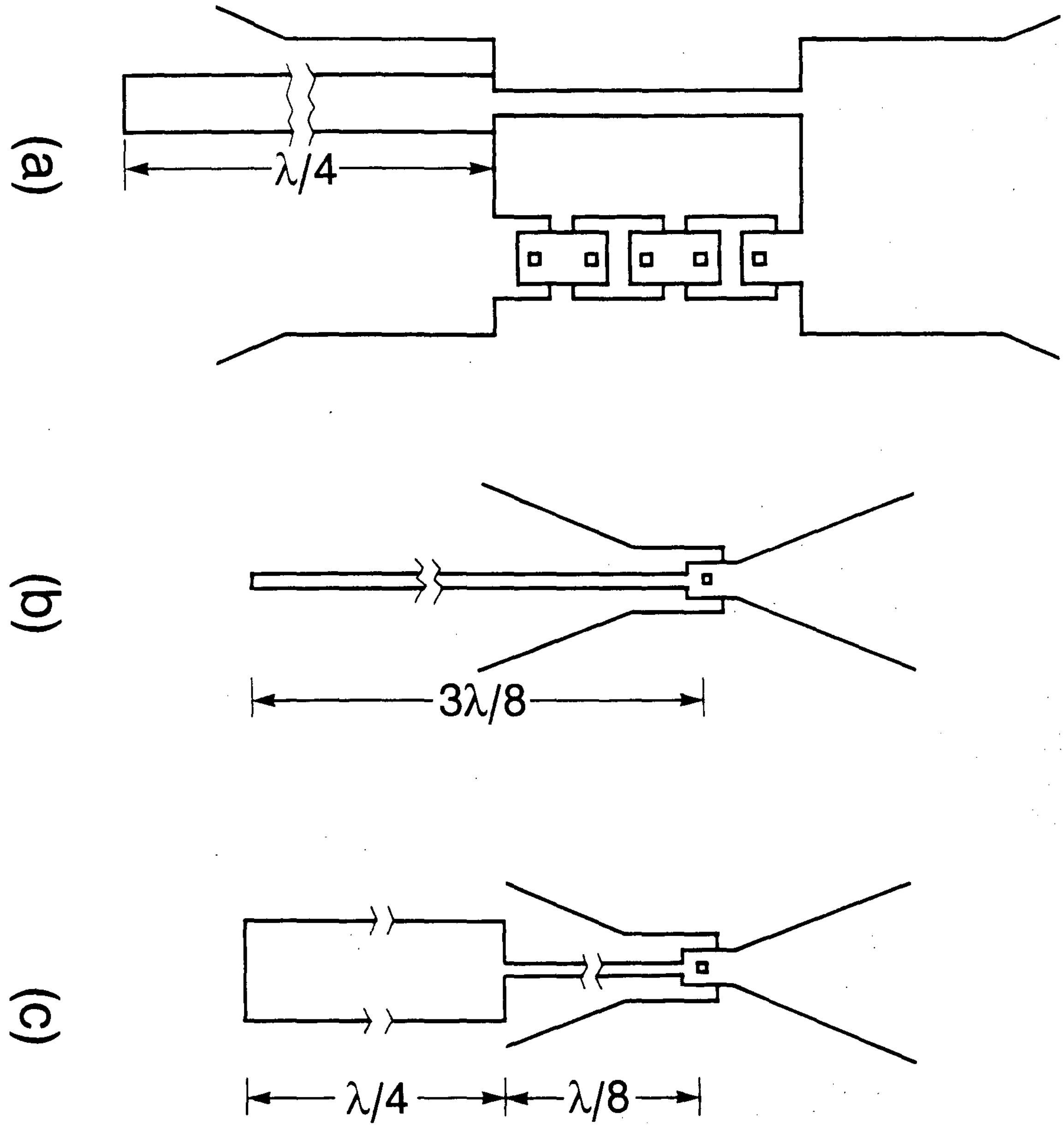
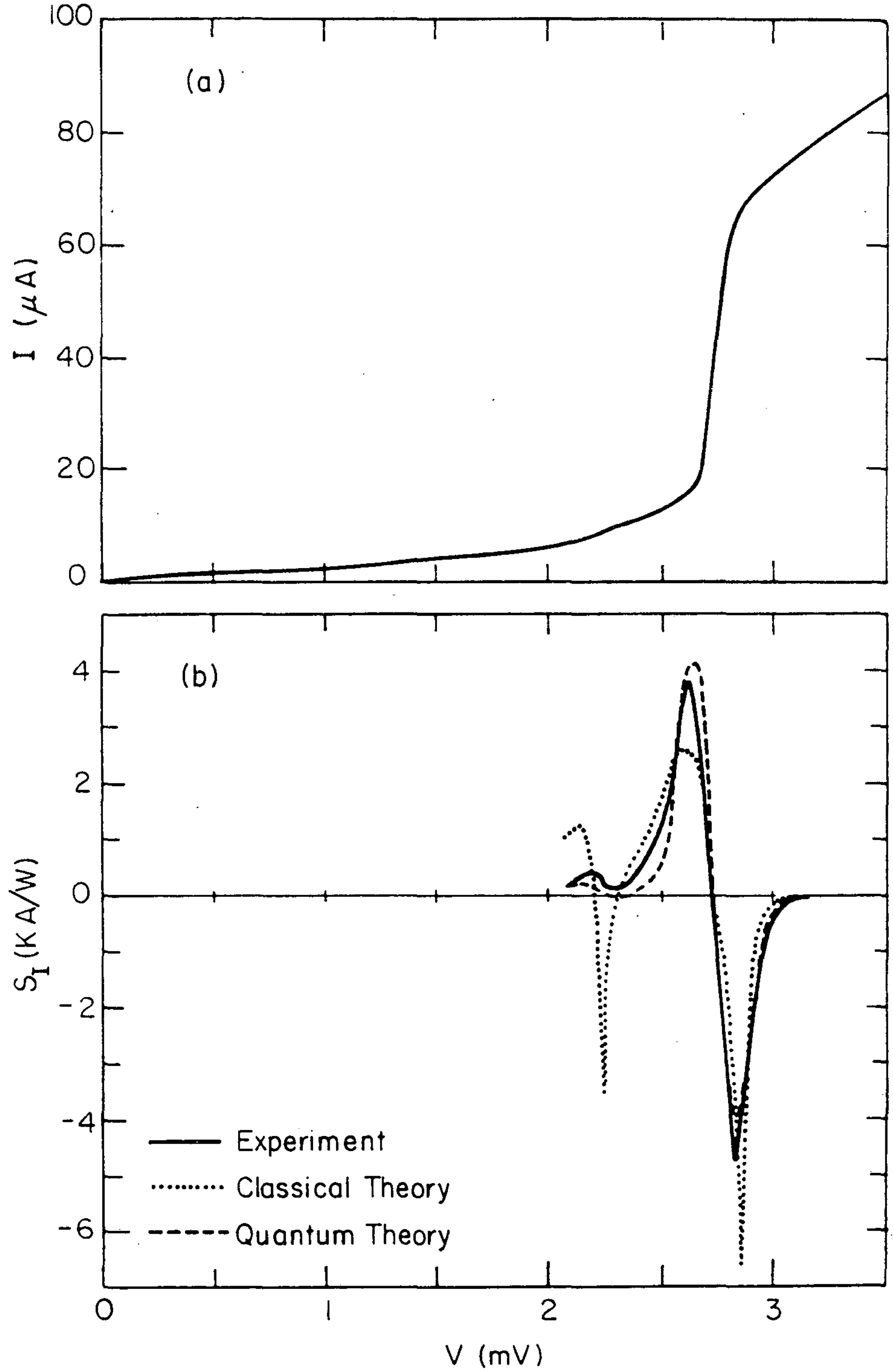

1 cm
SIS in
Planar Antenna
-RF in
Quartz
Hyperhemisphere
Teflon Lens
XBL 8810-7611

FIGURE 8



XBL 888-7506

FIGURE 9



XBL 798-6875A

FIGURE 10

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