Seventh International Symposium on Space Terahertz Technology, Charlottesville, March 1996

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SUB-MILLIMETER DISTRIBUTED QUASIPARTICLE RECEIVER EMPLOYING A NON-LINEAR TRANSMISSION LINE

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Abstract

The non-linear quasiparticle current flowing in a superconductor- insulator-superconductor (SIS) transmission line has successfully been used in low-noise heterodyne receivers at sub-millimeter wavelengths. These distributed mixer receivers have been implemented in both the 400 and the 600 GHz frequency bands and double-side-band receiver noise temperatures of 66 K at 455 GHz and 210 K at 630 GHz have been measured. Such performance is comparable to, or slightly better than the corresponding lumped element SIS mixerreceivers. The transmission lines used are fabricated from Nb/Al/AlO$_x$/Nb trilayer material, and are typrically 200 nm wide and 1–2 λ_g long. The distributed mixer chips have been designed with a simple linear model. We present design rules derived from this model and discuss some difficulties in mixer implementation.

I. Introduction

The success in recent years of Nb/Al/AlOx/Nb Superconductor-Insulator-Superconductor (SIS) tunnel junction in low-noise sub-millimeter receiver technology is largely related to the development of integrated superconducting microstrip tuners [1-3] that tune out the geometrical capacitance of the SIS junction. Nearly all SIS receivers currently operated are equipped with mixers that have some form of on-chip tuner. Indeed, well designed integrated tuners have eliminated the need for mechanical waveguide tuners, and numerous broadband, fixed-tuned SIS receivers have been operated at frequencies below about 500 GHz [4-7].

At higher frequencies, however, the development of passive matching circuits to tune out the geometrical capacitance of the SIS junction becomes more difficult. This is because the ωCR product of the SIS junction increases with frequency for a given critical current density, J_c. In order to achieve a reasonable RF bandwidth, higher values of J_c and smaller device areas are generally required. Unfortunately, such devices are more difficult to fabricate and they tend to have increased leakage current, resulting in poorer mixer performance.

The trade-off between RF bandwidth and noise temperature in the design of an SIS receiver parallels the classical problem of designing broadband microwave amplifiers in which the capacitance of the transistor generally limits its bandwidth [8]. It has been known for decades that distributed amplification can be used to realize multi-octave bandwidths in microwave amplifiers [9]. In its most common form, a distributed amplifier contains an input transmission line and an output transmission line with Field Effect Transistors (FET) providing the active coupling between the lines. In 1965, McIver [10] proposed the traveling-wave transistor in which amplification was provided by a distributed trans-conductance action between the input and output transmission lines on a planar structure that resembled an elongated FET. In a similar manner, we have proposed and tested distributed mixing in a superconducting non-linear transmission line with the aim of increasing the RF bandwidth of the mixer [11].

In the proposed traveling-wave mixer, the specific capacitance of the long SIS junction is incorporated into the characteristic impedance of the non-linear transmission line. The need to tune out the junction capacitance is therefore replaced by a simple real impedance transformation. Consequently, large bandwidths can be obtained even with high values of ωCR, or equivalently with low critical current density junctions. At frequencies below the band gap frequency of the SIS junction, ohmic losses are generally negligible. It follows that this new class of distributed mixer may offer both high conversion efficiency and wide bandwidth.

II. Linear Transmission Model

Rigorous analysis of the distributed quasiparticle mixing process in a non-linear transmission line requires the simultaneous solution of both Werthamer's equation [12,13] and the transmission line equation. The computing resources required is quite formidable. Nevertheless, a simple linear model turns out to be a sufficient design tool.

Lee *et al* [14] studied the propagation characteristics of a lossless linear transmission line containing a tunnel barrier. A cross-sectional view of the transmission line is given in Fig. 1, showing a junction of width w_1 with symmetrical counter-electrode overhang of width w_2 each on either side. The thicknesses of the base and top electrodes are t_1 and t_2 respectively. Outside the tunnel barrier, the 2 electrodes are separated by a dielectric thin film of thickness d_2 and relative permittivity ε_r. Lee *et al* found that in the low frequency limit, the propagation mode is indistinguishable from a TEM mode. Starting from their model, we are going to derive both the phase velocity and the characteristic impedance of the line from its inductance per unit length, L_s, and its capacitance per unit length, C_p [15]:

$$\frac{1}{L_s} = \frac{1}{\mu_0}\left[\frac{w_1}{h_1} + \frac{2w_2}{h_2}\right], \tag{1}$$

$$C_p = C_j w_1 + \frac{2\varepsilon_0\varepsilon_r w_2}{d_2}, \tag{2}$$

where

$$h_1 = \lambda_L \coth\left(\frac{t_1}{\lambda_L}\right) + \lambda_L \coth\left(\frac{t_2}{\lambda_L}\right), \tag{3}$$

$$h_2 = d_2 + h_1, \tag{4}$$

and C_j is the specific capacitance per unit area of the SIS junction and λ_L is the penetration depth. The characteristic impedance of this lossless linear transmission line, Z_0, and the slow wave factor, n, the ratio of the velocity of propagation in the line to that in free space, c, are given by:

$$Z_0 = \sqrt{\frac{L_s}{C_p}}, \tag{5}$$

$$n = c\sqrt{L_s C_p}. \tag{6}$$

In Fig. 2, Z_0 and n are plotted against w_1 for the cases of $w_2 = 0$ and 1.5 μm. Here, we have assumed $C_j = 60$ fF/μm^2 for Nb/Al/AlO$_x$/Nb tunnel junctions of low critical current density and that $t_1, t_2 >> \lambda_L = 75$ nm. The insulator between the 2 electrodes is a layer of 200 nm thick SiO ($\varepsilon_r = 5.7$).

The non-linear quasiparticle conductance of the tunnel barrier introduces losses to the electromagnetic wave propagating down the line. Quantum susceptance from the non-linearity can also modify the values of Z_0 and n, but we have ignored this effect in this simple linear model. We also assume that the conductance is distributed uniformly along the entire length of the line. This can be considered as a zero-order approximation because the large signal RF voltage impressed by the Local Oscillator (LO) drive varies with position along the line. A more sophisticated model can be built by assuming, for example, a sinusoidal spatial variation in the large signal voltage.

Let G_p be the conductance per unit length distributed along the line. The propagation constant, γ, and the characteristic impedance, Z_c, of the lossy line are given by

$$\gamma = \sqrt{j\omega L_s(G_p + j\omega C_p)} = j\omega\sqrt{L_s C_p}\left[1 + \frac{G_p}{j\omega C_p}\right]^{1/2}, \tag{7}$$

$$Z_c = \sqrt{\frac{j\omega L_s}{G_p + j\omega C_p}} = \sqrt{\frac{L_s}{C_p}} \left[1 + \frac{G_p}{j\omega C_p}\right]^{-1/2} . \tag{8}$$

At sufficiently high frequencies, $\omega C_p >> G_p$. Inspection of (7) and (8) shows that the phase velocity and characteristic impedance should be approximately the same as for the lossless line. The attenuation coefficient is

$$\alpha = \frac{1}{2} G_p Z_0 . \tag{9}$$

Let l_j be the length of the junction and G_m be the total distributed conductance, that is

$$G_m = G_p l_j . \tag{10}$$

The one-way attenuation of the long junction is therefore

$$A = \exp(-\frac{1}{2} G_p Z_0 l_j) = \exp(-G_m Z_0 / 2) \tag{11}$$

It should be noted that this is independent of the length of the device, and only depends on G_m and Z_0.

III. Design Considerations

An infinitely long lossy transmission line provides a perfect match over any frequency bandwidth to a source impedance equal to the characteristic impedance of the line. However, in the case of a distributed mixer, two problems arise if the line is too long. Firstly, the IF output capacitance will be unneccessarily high. Secondly, towards the far end of the line, the LO drive voltage is small so that while that part of the device contributes little to down conversion, it adds output shot noise at the IF. It seems that a line length of between 1 and 2 guided wavelengths offers a good compromise between match and mixer efficiency.

We have chosen a simple open-ended line because no simple termination exists. At the open end of the line, the incident wave is reflected back towards the source. The reflection coefficient at the source is, therefore, A^2. In order to provide an input return loss smaller than -12 dB, or equivalently an input reflection coefficient of less than 0.25, we find that $A > 0.5$. Hence, using equation (11), we have

$$G_m Z_0 > 1.386 \tag{12}$$

This design rule implies that a high characteristic impedance, Z_0, is desirable for a given value of G_m. Referring to Fig. 2, we conclude that it is better to employ a narrow line without a counter-electrode overhang ($w_2 = 0$).

When the device is opimally pumped by the LO drive, the conversion efficiency should be close to that of a lumped element SIS mixer. It follows that the total distributed non-linear quasiparticle conductance, G_m, should be the same as the optimal source conductance of a lumped element SIS mixer, $G_{s,opt}$ [16]. Since $G_{s,opt}$ is proportional to G_N; for a given line length and width, the design rule establishes the minimum critical current density.

Let us consider the design process for a 460 GHz non-linear transmission line mixer. Suppose that a 0.2 μm wide line is to be used. From Fig. 2, Z_0 =8.9 Ω and $n = 31.9$. Using a line $2\lambda_g$ long, we need a line length of about 40 μm. We assume that $G_m = G_{s,opt} = G_N$ at the target frequency. It follows from the design rule in (11) that $G_N > 0.156$. Since the area of the device is $8\mu\text{m}^2$, we derive that $J_c > 3700$ A/cm^2. Note that this critical current density is substantially below that of a typical lumped element SIS mixer operating at the same frequency.

IV. Topology and Fabrication

Owing to the extremely narrow width of the lines (typically 200 nm), the junctions have to be fabricated using electron beam lithography. The fabrication process is essentially identical to that of lumped element SIS junctions [17,18] except for the interface between the wiring layer and the counter-electrode of the Nb/Al/AlOx/Nb trilayer. In a lumped element SIS mixer chip, the junction size is small and the wiring layer always covers the entire junction. However, in a distributed mixer chip, the wiring layer only contacts the end of the top electrode of the transmission line where electromagnetic energy is fed into the line.

A photograph of the non-linear transmission line and a schematic illustration of the layout are given in Fig. 3. A microstrip transformer section, of length L_t, matches the characteristic impedance of the long junction to the embedding impedance at the waveguide feed point. The length of the main non-linear transmission line is L_j. It does not have a counter-electrode overhang. However, the interface contact section between the transformer and the junction is actually a section of line of length L_c with counter-electrode overhang ($w_2 \neq 0$). For alignment purposes, the minimum value of L_c is 3 μm.

The base electrode, or the ground plane, of the non-linear transmission line is 150 nm thick Nb. The top electrode, or the strip portion of the line, is 120 nm thick Nb overlaid by 30 nm thick gold. Since the thickness of the two electrodes are only comparable to $2\lambda_L$, the values of Z_0 and n are slightly higher than the values given in Fig. 2.

V. Results in the 400 GHz Band

The first distributed mixer chip was designed to fit the 400–500 GHz fixed-tuned mixer block developed for the Sub-Millimeter Arrray (SMA). The design and characterization of a lumped element SIS mixer in this mixer block has been reported previously [5]. The non-linear transmission line section of the chip measures about 150 nm wide, and is 40 μm long. The contact section is 3 μm long with w_2 = 1.5 μm. A quarter wave transformer section matches the line impedance to an embedding source impedance of 35 Ω.

The critical current density of the junction is about 2500 A/cm^2 and the normal state resistance is about 13 Ω. The $G_N Z_0$ product is about unity. The leakage characteristics of the junctions are very good, due to the low J_c. The value of R_{sg}/R_N is about 20. The IF output capacitance of the mixer is about 0.5 pF, corresponding to a reactance of about 200 Ω at 1.5 GHz, or 60 Ω at 5 GHz. This output capacitance is large when compared to the 0.3 pF output capacitance for the lumped element SIS mixer developed for the SMA.

The noise temperature of the distributed mixer was measured in a setup identical to that of the lumped element mixer. The IF of the receiver was 5 GHz. Fig. 4 shows the DC I–V characteristics of the long junction without and in the presence of LO power at 455 GHz; the receiver IF output power in response to hot and cold loads is also shown in the figure as a function of bias voltage. From the figure, a $Y-$ factor of 2.52 was obtained, corresponding to a Double-Side-Band (DSB) noise temperature of about 66 K, or about 3 $h\nu/k$. The distributed mixer was operated with a magnetic field applied perpendicular to the transmission line to suppress the interfering pair current. The applied field to the long junction was only slightly smaller than in the case of the lumped element device, showing that tens of fluxons of magnetic flux was coupled to the distributed mixer element.

In Fig. 5, the DSB receiver noise temperature of both the distributed and the lumped element receivers are plotted as a function of LO frequency. While both receivers provide about 20% bandwidth, the distributed receiver is 15 – 20% lower in noise temperature than the non-distributed receiver at the band center. Note that the lumped element device has an $\omega C R_N$ product of slightly under 5 and a J_c of about 8000 A/cm^2, compared to 2500 A/cm^2 for the long junction.

VI. Results in the 600 GHz Band

The 460 GHz distributed mixer design was scaled up in frequency to 660 GHz. At this higher frequency, the fabrication constraint of $L_c = 3$ μm begins to affect the design. The contact section introduces an electrical length of about 24° at 660 GHz, and its characteristic impedance is considerably lower than that of the main non-linear transmission line. Therefore, two mixer chips were designed. On chip A, the layout is identical to that on the 460 GHz chip. The non-linear transmission line is fed at one end with a 3 μm contact section with $w_2 = 1.5$ μm. A transformer section provides a match to the source. On chip B, L_c is increased to 23 μm, corresponding to an electrical length of 180°. Hence, at the design center frequency, the effect of the contact section is removed , but the bandwidth of operation is expected to be reduced. Since L_c is quite long, L_j has to be shortened in order to maintain a reasonable value of R_N.

The distributed mixer chips were tested in the SMA 600–720 GHz fixed tuned mixer block [19]. Measurements were carried out in a setup that had been used to characterize the corresponding lumped element SIS mixer chip. The critical current density of the distributed SIS junctions is 4300 A/cm^2. A number of different chips have been tested. The best noise temperatures of both design A and design B are plotted as a function of LO frequency in Fig. 6, along with the data of the lumped element mixer receiver.

For design A, the optimal configuration is $w_1 = 320$ nm and $L_j = 23.5$ μm, corresponding to about $1.5\lambda_g$ in the 600 GHz frequency band. The receiver noise temperature is around 125 K, or about $5h\nu/k$, between 500 and 545 GHz. Above the water vapor absorption band, around 560 GHz, the noise temperature rises only gradually. The poorer performance around 610 GHz is due to the 0.435 mm thick Teflon vacuum window which is $5\lambda_d/4$ thick at that frequency. The noise temperature at 630 GHz is 210 K and is comparable to the non-distributed receiver up to 650 GHz. The frequency response shows that the design is tuned towards a lower frequency. It seems very likely that the contact section effectively increases the equivalent electrical length of the matching transformer section. A second iteration with shorter L_t is currently under development.

In the case of design B, the best noise is obtained from a chip with w_1 = 220 nm and L_j =17 μm, slightly longer than λ_g. Compared with chip A, this design offers a much reduced bandwidth, about 10% about a center frequency of 630 GHz with a noise temperature of about 230 K. The overall receiver performance is generally inferior to that of the lumped element receiver. The measured data suggest that the contact section does not contribute much to the mixing process and constitutes an obstacle to proper impedance matching of the main non-linear transmission line. This agrees well with our theoretical model that suggests that Z_0 should be as large as possible.

X. Conclusion

The use of a non-linear quasiparticle transmission line as distributed mixer elements in receivers at sub-millimeter wavelengths has been investigated. Using a simple linear model, we have established design rules to determine the optimal width and length of the transmission line and the required critical current density. Several designs have been tested. In the 400 GHz band, the measured noise temperature is around $3h\nu/k$, which is 10 – 15 % better than the corresponding lumped element mixer. At higher frequencies, the performance of the distributed mixer is affected by the section used to contact the main non-linear transmission line. With a short contact section, we achieve a receiver noise temperature of about $5h\nu/k$ around 520 GHz, and comparable noise temperature to the corresponding lumped element receiver up to 650 GHz. We expect that further refinement of the design will yield improved performance.

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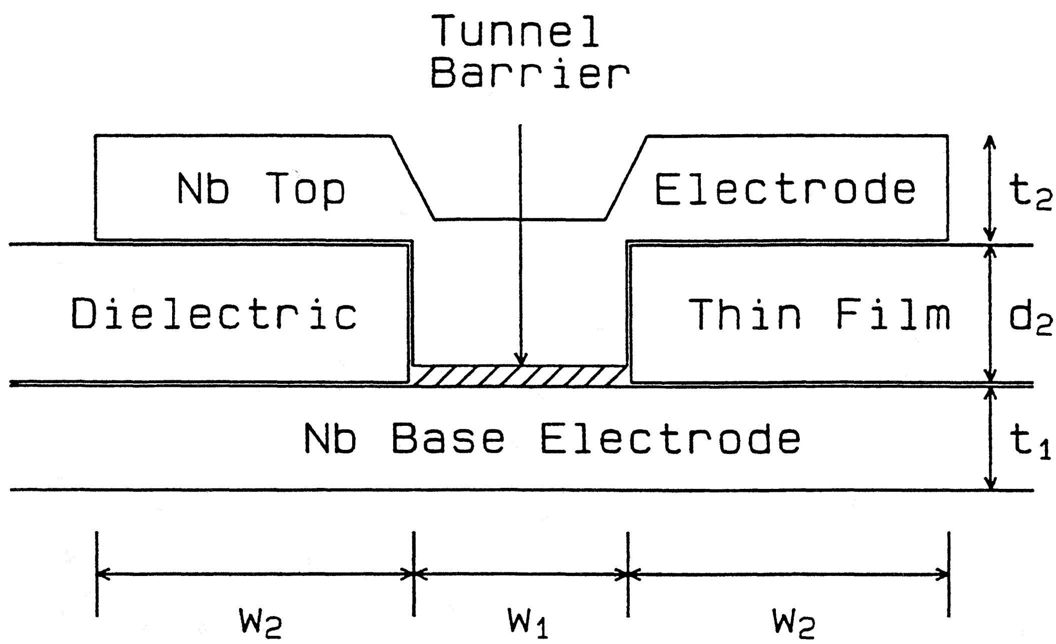

Fig. 1 Cross-sectional view of a niobium Superconductor-Insulator-Superconductor Transmission Line. The tunnel barrier has symmetrical counter-electrode overhang on either side.

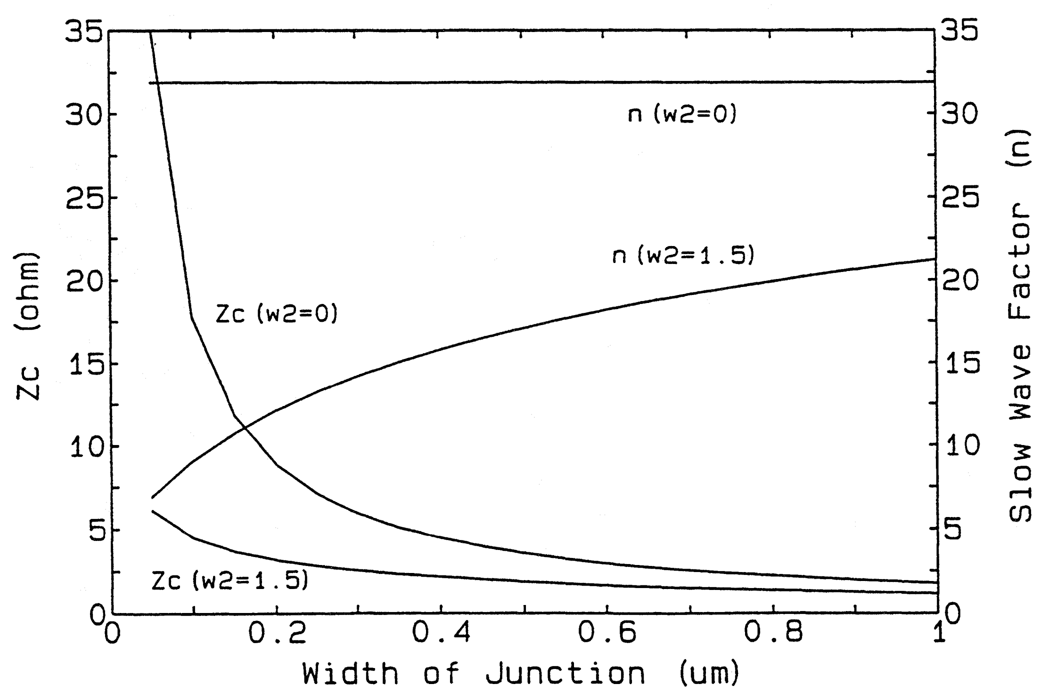

Fig. 2 Characteristic Impedance Zc and slow wave factor n of the lossless linear transmission line shown in Fig. 1 for the cases w2 = 0 and w2 = 1.5 μm.

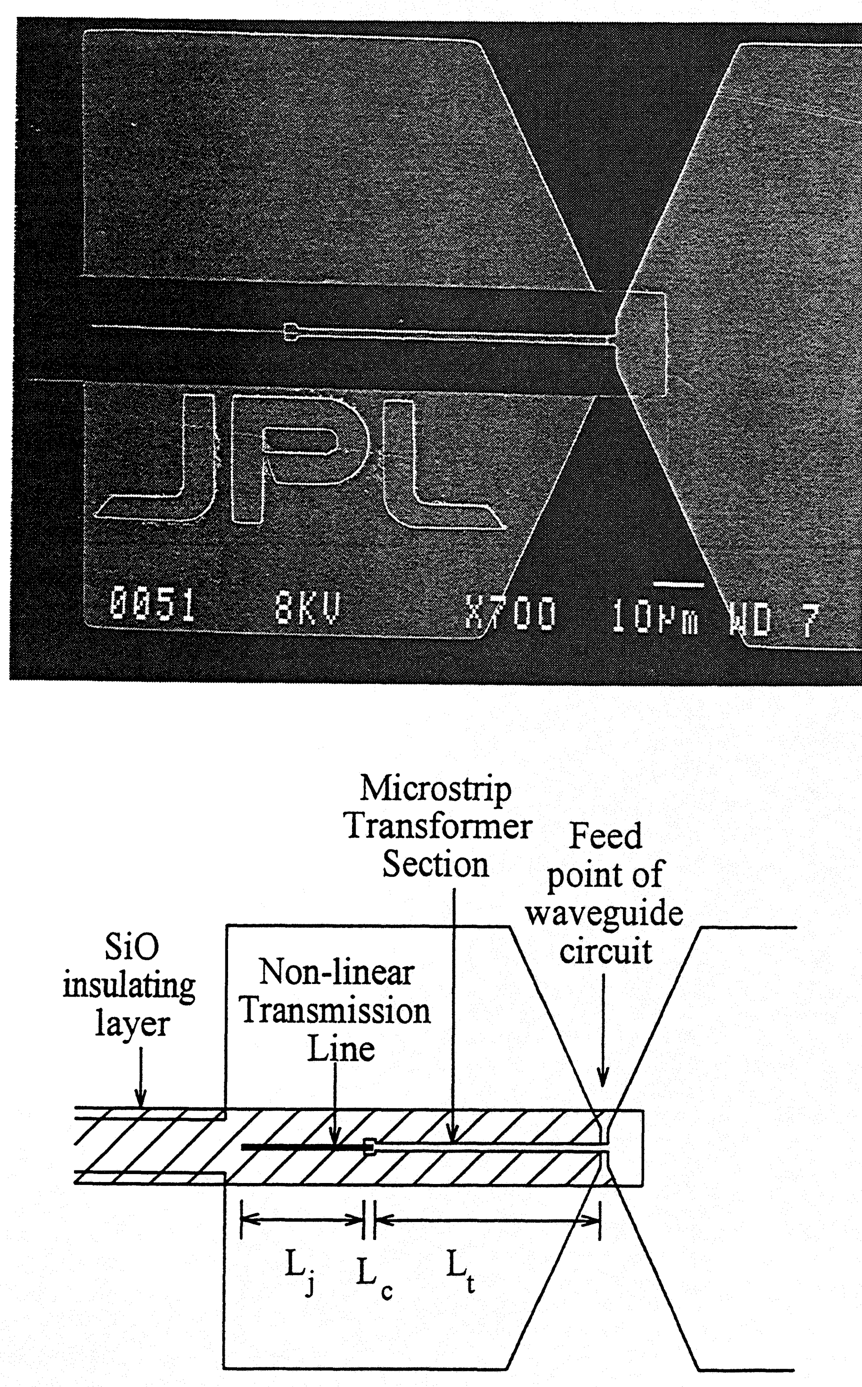

Fig. 3 SEM photograph (top) and line illustration (bottom) of the center part of the mixer chip carrying the non-linear superconducting transmission line.

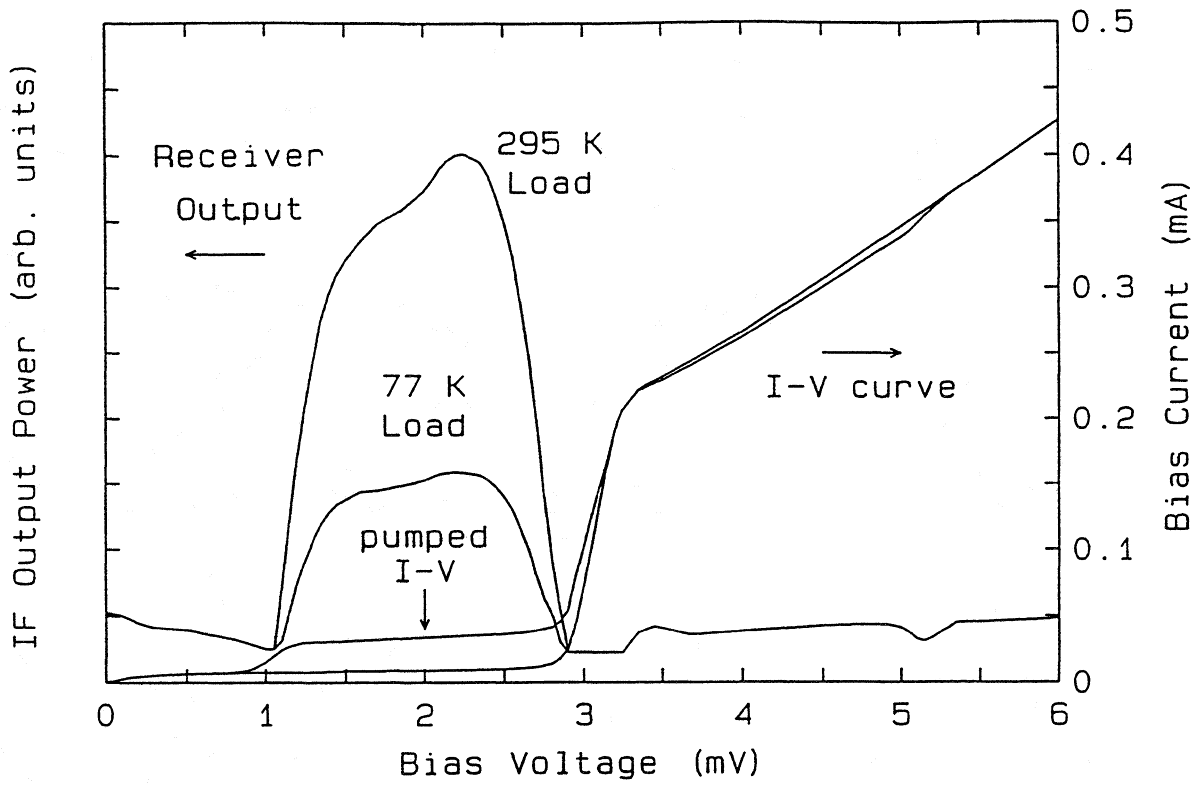

Fig. 4 DC current-voltage characteristics of the non-linear transmission line without and with incident LO power at 455 GHz. Also shown is the receiver IF output power against bias voltage for hot and cold input loads. A *Y*-factor of 2.52 is noted at a bias voltage of 2.4 mV.

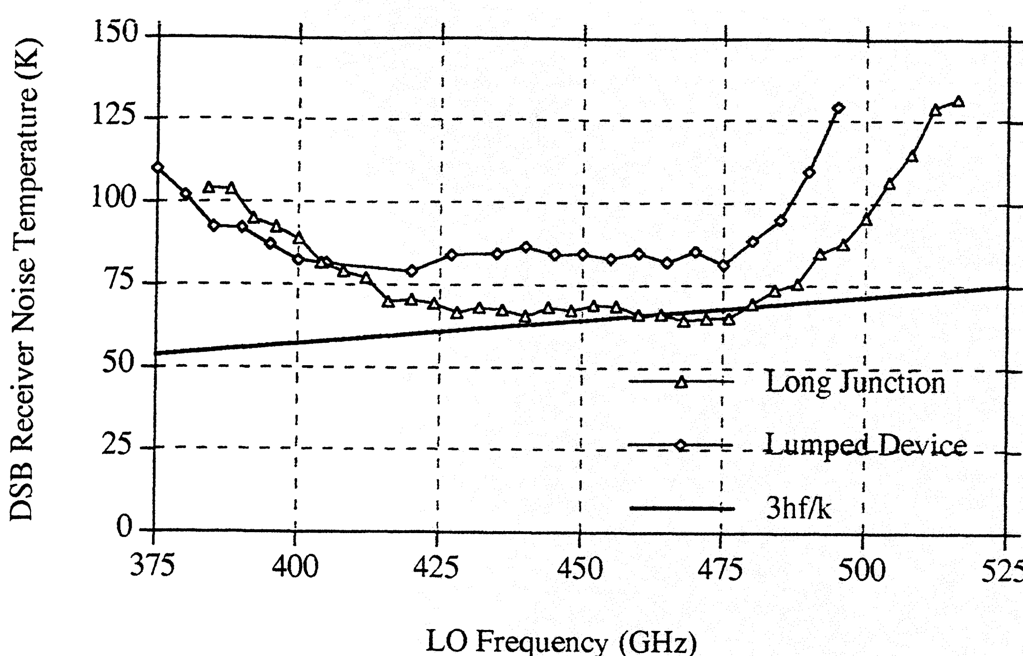

Fig. 5 Double-side-band (DSB) receiver noise temperature of the SMA 400 -- 500 GHz fixed-tuned SIS receiver, using (a) the 460 GHz distributed mixer chip with a long junction, and (b) a lumped element SIS junction [5]. Note that the noise temperature of the distributed receiver is close to 3hf/k at the center of the band.

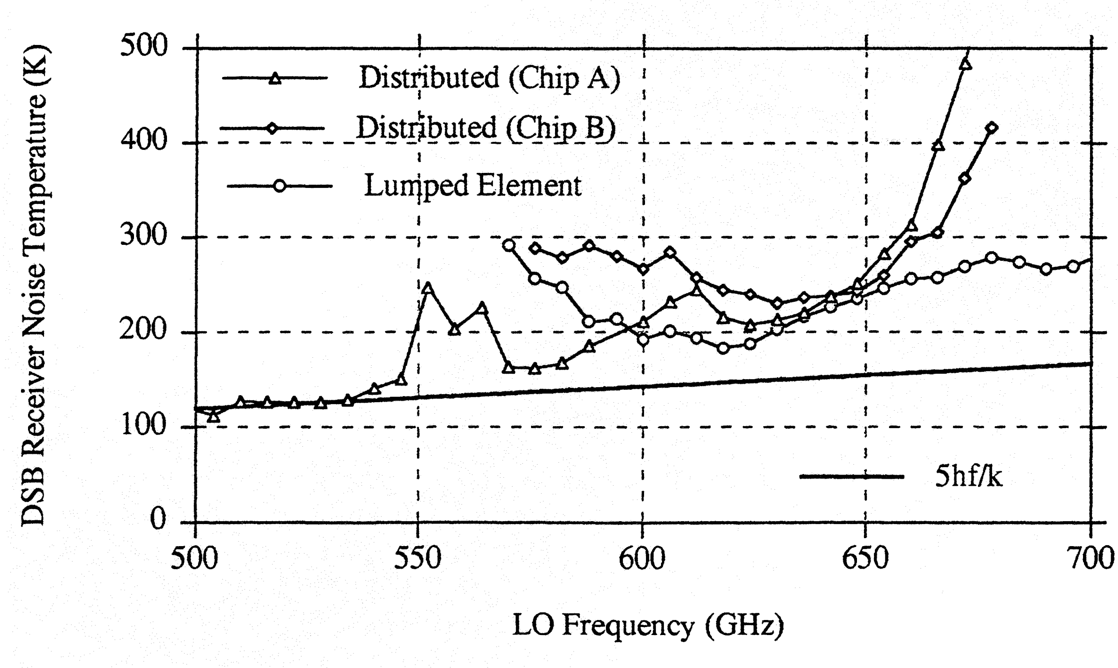

Fig. 6 Double-side-band (DSB) receiver noise temperature of the SMA 600 -- 700 GHz fixed-tuned SIS receiver, using (a) the distributed mixer chip A with Lc = 3 μm; (b) the distributed mixer chip B with Lc = 23.5 μm; and (c) a lumped element SIS junction [19].