BROADBAND QUASIOPTICAL SIS MIXERS WITH LARGE AREA JUNCTIONS

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Abstract

We have designed and tested a broadband quasi-optical superconducting tunnel junction (SIS) mixer with integrated tuning elements. We achieve state of the art results using low critical current density, large area niobium SIS's which are commercially available from Hypres, Inc. We have performed noise measurements in the frequency range from 70 GHz to 105 GHz. The best uncorrected double sideband receiver noise is 38 K at 77 GHz, with receiver noise temperatures less than 100 K from 75 to 102 GHz.

I. Introduction

Superconducting tunnel junctions (SIS's) are used in the most sensitive detectors at frequencies from 100 GHz to 750 GHz [1]. Much of the recent progress in improving sensitivity has relied on using very small area (submicron) junctions with very high critical current densities exceeding 10 kA/cm^2 [2-5]. The large current density allows higher frequency operation of junctions, while the small junction area is necessary to match large current density junctions to moderately high impedance radiation structures. Unfortunately, submicron lithography is very expensive and not widely available, so it is time consuming and expensive to develop submicron circuits. Similarly, 1 kA/cm^2 SIS technology is well developed for digital circuit research, but higher current densities require device development separate from the larger digitally oriented efforts.

We show in this paper that large area, low current density SIS's can be used to make state of the art SIS mixers. We do this by designing our mixers for Hypres, Inc.'s standard niobium integrated circuit process. We use inductive tuning to cancel out the large capacitance of large area SIS's, and a transformer to match to high impedance antennas. We are thus able to take advantage of all of the technology development effort that has gone into digital circuits.

II. Circuit Design and Simulations

In our quasi-optical SIS receiver, the SIS junction is built integrally with a planar self-complementary log-periodic antenna on the silicon substrate [6-8]. The antenna is placed on the back of a quartz hyperhemisphere [9]. The hyperhemisphere and a teflon lens in front of it focus the radiation onto the antenna. The antenna impedance is frequency independent over several octaves and is around 76 Ω.

Tuning structures are built on one arm of the antenna which is used as their ground plane (Fig. 1). The major goal in the design of integrated tuners is a large bandwidth of good coupling. A large bandwidth design has two advantages. First, it provides a sensitive receiver over a large frequency range. Second, it makes it likely that a particular frequency of interest will fall in the sensitive range of the receiver, even if some fabrication process parameters are different from those used in the design.

For SIS mixers without integrated tuning, it is necessary to use small area (submicron) SIS's with high current density in order to achieve good coupling over a reasonable bandwidth. Our work shows that with tuning structures, we do not require either submicron lithography or high current density to succeed. The SIS mixers we use were fabricated at Hypres, Inc. in their all-refractory niobium process. The Hypres $Nb/AlO_x/Nb$ SIS's have current density around 980 A/cm^2 and junction area of 12 μm^2. As our measurements show, we can achieve excellent results over broad bandwidths with these junctions.

The tuning circuit for the mixer we report here consists of two parts: an inductive part which tunes out junction capacitance, and a transformer which matches the junction resistance to the antenna impedance (Fig. 2). The inductive section is a combination of an open-ended radial stub and a short length high impedance microstrip line [10]. The large angle radial stub is used in order to get a broadband short on its other end. The high impedance microstrip line presents the inductance necessary to tune out the junction capacitance. As a transformer we use a quarter wavelength microstrip line which transforms the junction impedance into antenna impedance. The design presented here is applicable for different fabrication processes, and for both large and small area junctions. Table 1 shows the expected performance of the mixer with tuning circuit designed for different fabrication process parameters and different central frequencies. The 3 dB bandwidth shown is the frequency range over which the particular circuit has coupling to the antenna exceeding 0.5.

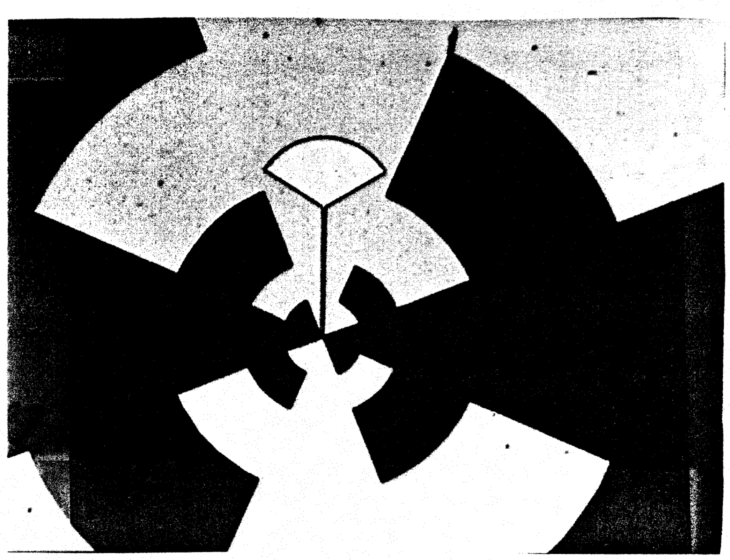

Fig. 1. Self-complementary log-periodic antenna with integrated superconducting tuning structure.

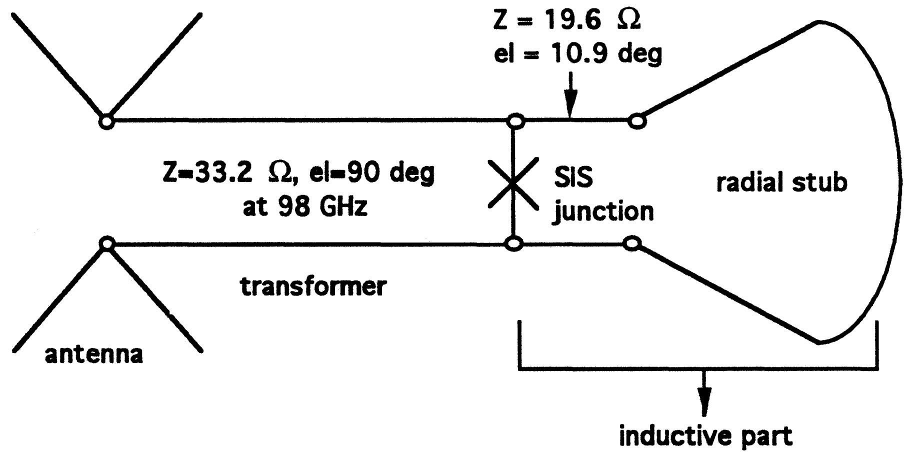

Fig. 2. The electrical circuit of the design, where Z is the characteristic impedance of the microstrip line and el is the electrical length at 98 GHz.

Central Frequency [GHz]	3 dB bandwidth [GHz]		
Jc A	1000 A/cm^2 12 μm^2	4000 A/cm^2 12 μm^2	4000 A/cm^2 1 μm^2
98	38	48	82
230	29	55	144
492	18	39	167

Table 1. Expected 3 dB bandwidth of the circuit shown in Fig. 2, designed for different fabrication process parameters and at different frequencies. Coupling at the central frequency is kept the same and it is equal to 0.9.

We used the RF simulation program *Libra* [11] to optimize circuit elements for the best coupling and the largest bandwidth. The junction is modeled as an impedance $1 / (1/R_{rf} + j\,\omega\,C_{rf})$. For typical biasing conditions R_{rf} has a value close to the normal state resistance (R_N), and the major part of the parasitic reactance is due to the geometrical junction capacitance (C_j), so that we used those values in the simulation. Junction parameters used in the design are $R_{rf} = R_N = 26.7\ \Omega$ and $C_{rf} = C_j = 432$ fF. The circuit was designed to have the best coupling of 91 % at 98 GHz and a 3 dB bandwidth of 34 GHz (Fig.3). A small mismatch at the central frequency is included to provide a larger bandwidth of the tuning circuit. After the dc IV curve of this circuit is measured, the actual junction parameters are used to run simulation program. The junction RF impedance calculated from the dc IV curve is $R_{rf} = 6\ \Omega$ and $C_{rf} = C_s + C_j = 544$ fF, where C_s is the capacitance calculated from the quantum susceptance, with the bias point in the middle of the first photon step and optimum local oscillator (LO) power [12]. Since the values for R_{rf} and C_s did not vary much at different frequencies, we assumed for simplicity R_{rf} and C_s to be constant. For the junction geometrical capacitance we have used 38 fF/μm^2, which is the typical value for the capacitance in Hypres fabrication process. The fabricated junction size is 12.2 μm^2 and the junction normal state resistance is 18.2 Ω. The coupling curve with the actual junction parameters is shown in Fig. 3.

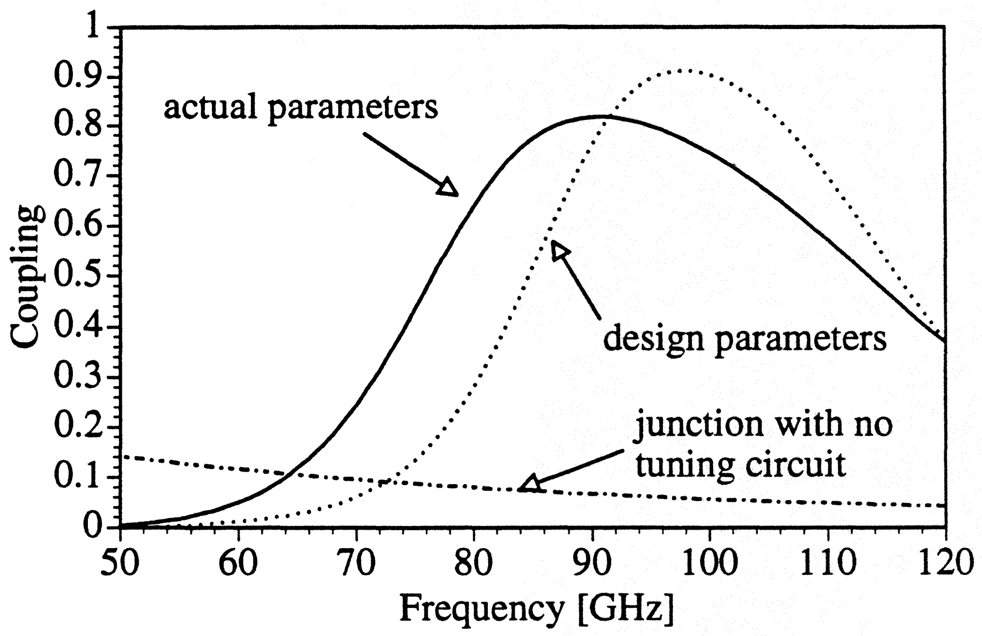

Fig. 3. Coupling coefficient between the SIS junction and the self-complementary log-periodic antenna

III. Results

The dc IV Curve Measurements

The simplest way to check the central resonance frequency of the tuning circuit is to measure its dc IV curve. We show the IV curve for dc voltage up to 0.97 mV in Fig. 4. The resonant step occurs because the tuning structure is impedance matching Josephson junction oscillations to the antenna. We translate the voltage range of this resonance step to a frequency range using the Josephson frequency relation $f_J = 2eV_0/h$. The dc IV curve in Fig. 4 shows a resonant step in the frequency range from 56 GHz - 98 GHz. The *Libra* simulation of the actual circuit shows 3 dB bandwidth from 76 GHz to 114 GHz (Fig. 3). There are a number of fabrication parameters in our real circuit that could account for the difference between theory and measurement. For example, a larger value of SIS junction capacitance than the one assumed (38 fF/μm^2) would account for the lower frequency resonance which we see.

Heterodyne Receiver Measurements

In the heterodyne receiver measurements we used a tunable Gunn oscillator as a LO, which has operation range from 70 GHz to 105 GHz. In our receiver we couple the input radiation through a set of lenses (Fig. 5). The beam from the LO is first focused by the teflon lens and then it reflected from the mylar beamsplitter toward the mylar window in the bottom of the receiver. Further on, the beam passes through the quartz window built on the 77 K shield, and through the teflon lens, 1 mil thick black polyethylene, and 40 mil thick quartz. Finally, the beam is focused by the quartz hyperhemisphere lens onto the center of the log-periodic antenna. For the hot / cold load measurements we used absorber at 295 K and 77 K. The receiver noise temperatures we report are those measured at the place of hot / cold signal source. We make no corrections for any beam losses between the hot and cold loads, and the mixer itself.

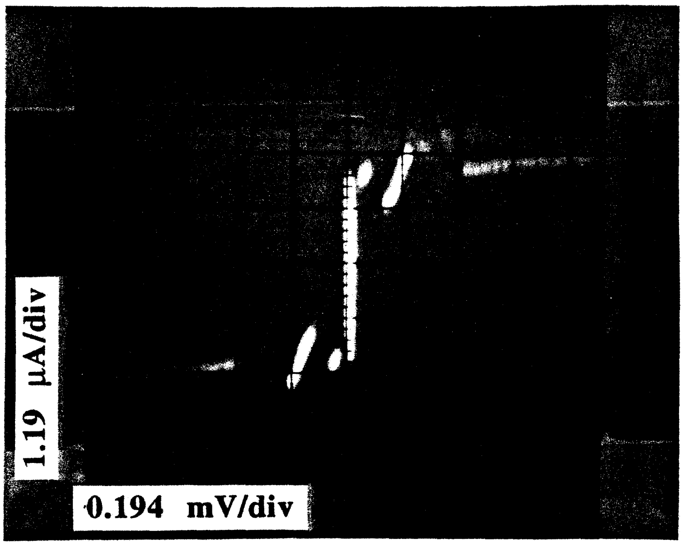

Fig. 4. The dc IV curve of the mixer with tuning structures. The self induced resonant step is in the frequency range from 56 GHz to 98 GHz.

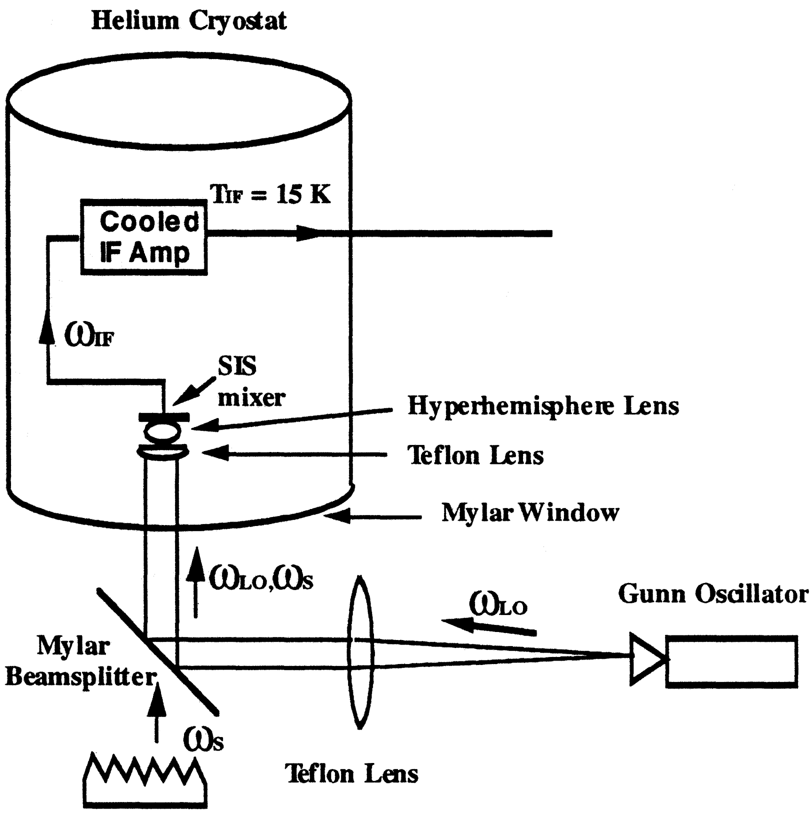

Fig. 5. The experimental setup for the hot / cold load measurements.

The dc IV curve of the mixer with the tuning structures shows flat photon steps at frequencies larger than ~ 95 GHz and steps with negative slope at some frequencies lower then 95 GHz (Fig. 6). On the first photon step we were not able to find the optimum noise temperature at all frequencies, due to the unstable bias at some regions of the LO power. Fig. 7 shows measured receiver noise temperature on the first photon step. All values that are shown in Fig. 7 are achieved with stable bias, but those marked with crosses are the frequencies where we had regions of unstable bias, so we could not obtain the noise temperature for all LO power values. The IF amplifier we use in the setup has noise temperature of T_{IF} = 15 K. The best double sideband receiver noise temperature we obtained is 38 K at the LO frequency of 77 GHz. The calibrated mixer noise at that point is 14 K with the mixer gain of -1.3 dB.

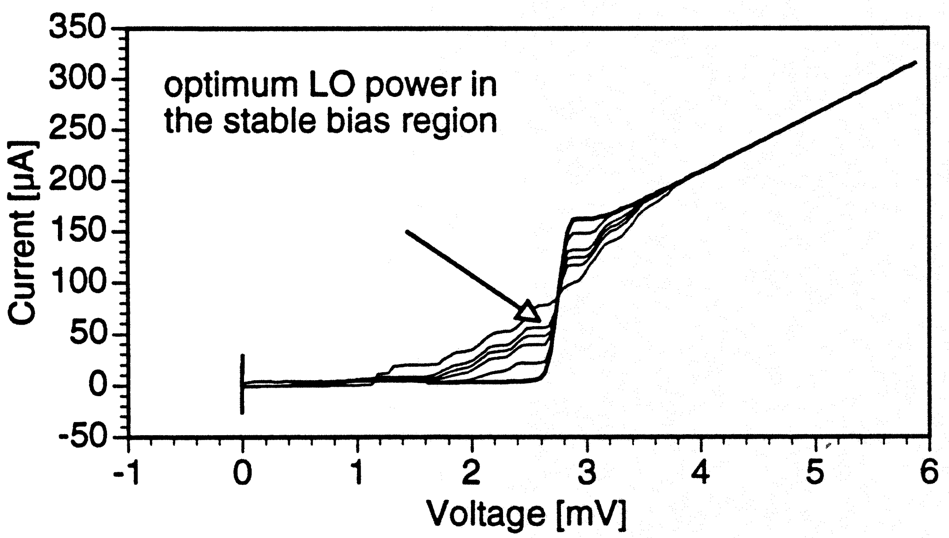

Fig. 6. The dc IV curve of the SIS mixer with integrated tuning structures. Bold line represents the dc IV curve when no LO is applied. The curve with the optimum LO power is marked on the chart. The LO frequency is 77 GHz. All measurements were made at 4.2 K.

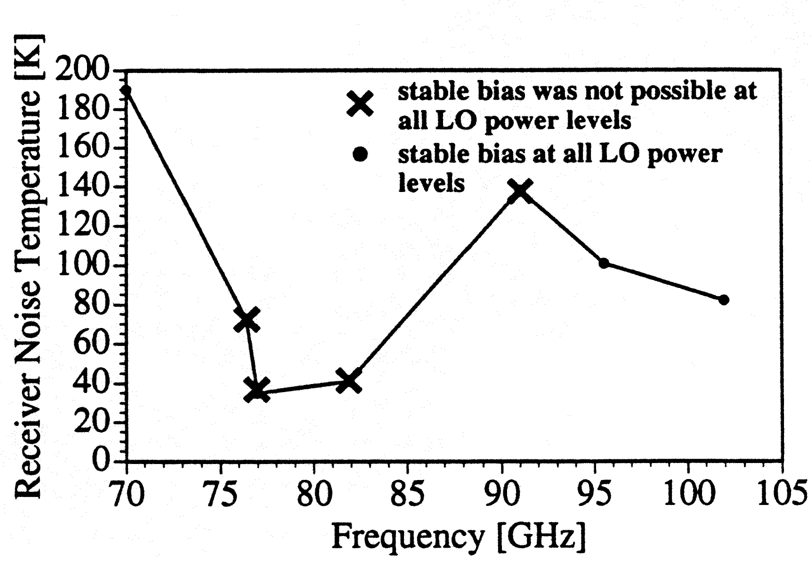

Fig. 7. Double sideband receiver noise temperature on the first photon step.

We measured the noise temperature on the second photon step where the bias was stable in the whole LO frequency range, except for one point which we skipped in the measurement. The best receiver noise temperature we obtained on the second photon step is $(T_R)_{DSB}$ = 180 K at the frequency of 89.7 GHz. The calibrated mixer noise is 141 K and the mixer gain is - 4.1 dB. As a reference, we measured a mixer with no tuning structures. In this case, the 12.2 μm^2 junction is fabricated directly at the center of the log-periodic antenna. Due to the large parasitic capacitance, photon induced steps are almost invisible in the dc IV curve. The minimum mixer noise temperature obtained with this circuit is 1478 K with the mixer gain of - 8.2 dB, when biased on the second photon step and at a frequency of 89 GHz.

The simple photodiode theory [1] predicts that the mixer noise temperature on the first photon step is inversely proportional to coupling between the SIS junction and the source. In Fig. 8 we compared the coupling obtained from the *Libra* simulation to C/T_{MIX} where T_{MIX} is the calibrated mixer noise temperature on the second photon step and C is the normalization constant. An increase in the mixer noise at the frequencies close to 70 GHz is probably caused by the performance of the antenna, which is designed to operate at a minimum frequency of ~ 72 GHz. The central frequency of the real coupling curve is, we believe, shifted toward smaller frequencies, as evidenced from the dc IV curve. These two curves do not fit very well, but they definitely show strong dependence between the noise temperature and the coupling. Using this measurement we estimated the 3 dB bandwidth of our tuning circuit to be ~ 18 GHz.

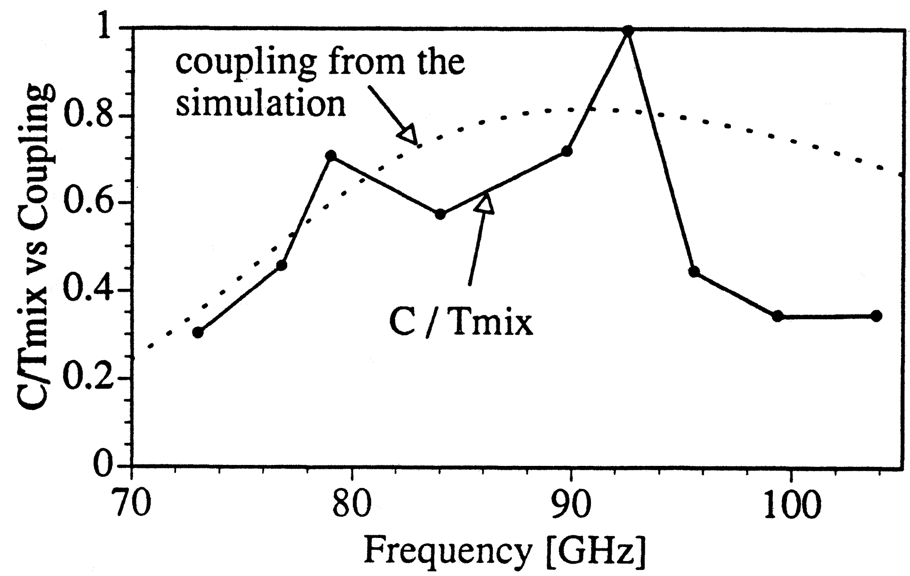

Fig. 7. Comparison between the coupling curve calculated from the *Libra* simulation program and C/T_{MIX} where T_{MIX} is the calibrated mixer noise temperature on the second photon step and C is the normalization constant.

Conclusion

We have presented results for an SIS receiver which achieves less than 100 K receiver noise over a 27 GHz bandwidth with the minimum values around 77 GHz. This is achieved using a quasioptical coupling scheme, and a fairly noisy 15 K IF amplifier. It is achieved using low critical current density (980 A/cm^2), large area (12 μm^2) SIS junctions.

We have demonstrated that extremely low noise, fairly broad band millimeter receivers can be achieved using large area, low current density SIS junctions. This should greatly reduce the cost of producing and maintaining SIS mixers. It should make them more attractive for space-based astronomical platforms.

We have achieved these results by careful design of appropriate superconducting tuning elements. In particular, we tune out the very large parasitic capacitance of the SIS, and we use a stripline transformer to match the low junction impedance to the higher antenna impedance. We have demonstrated that we can accurately design superconducting tuning structures for millimeter wave mixers using Hypres's commercial technology.

We believe that similarly excellent results will be possible using large area junctions up to frequencies of at least 500 GHz. As we get to higher frequencies, going to higher current density junctions will allow higher bandwidth designs, but we are still capable of 18 GHz bandwidth even with 1000 A/cm^2 junctions.

Acknowledgements

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