A 547 GHz SIS RECEIVER EMPLOYING A SUBMICRON Nb JUNCTION WITH AN INTEGRATED MATCHING CIRCUIT

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ABSTRACT

A heterodyne receiver using an SIS waveguide mixer with two mechanical tuners has been built and characterized over a frequency range of 460 GHz to 630 GHz. The mixer uses high current density, 0.25 μm^2, Nb/AlO$_x$/Nb SIS tunnel junctions with integrated superconductive RF circuits to tune the junction capacitance. A DSB receiver noise temperature as low as 200 ± 17 K has been obtained at 540 GHz. In addition, negative differential resistance has been observed in the DC I-V curve at 487-491 GHz.

INTRODUCTION

The most sensitive heterodyne receivers used for millimeter wave and submillimeter wave radioastronomy employ superconductor-insulator-superconductor (SIS) tunnel junctions as the nonlinear mixing element. Good performance has recently been reported for SIS junctions used in planar mixer circuits [1, 2] and waveguide mixers [3-8] from about 300 GHz to 500 GHz. In general, however very few SIS mixers have been demonstrated at these high frequencies. We have developed a submillimeter wave SIS heterodyne receiver for observing important rotational transitions of molecules in the interstellar medium, and in particular the ground state transition of $H_2{}^{18}O$ at 547 GHz. This receiver is based on a waveguide mixer with an adjustable backshort and E-plane tuner [9]. The mixer uses a high current density, submicron area Nb-AlO$_x$-Nb tunnel junction. The large capacitive susceptance of the junction at high frequencies will shunt the signal away from the nonlinear conductance and hence must be properly tuned for optimum performance. This is accomplished here through the use of carefully designed superconductive microstrip circuits to match the complex impedance of the junction to the available tuning range of the waveguide mount. The receiver performance has been measured over the frequency range 460 GHz to 630 GHz. A DSB receiver noise temperature as low as 200 ± 17 K has been achieved at 540 GHz, and represents one of the best values reported to date at this frequency. In addition, negative differential resistance has been observed in the DC I-V curve at frequencies around 491 GHz. These results indicate that the superconductive Nb microstrip transmission lines used in the tuning circuits are low-loss and perform well up to at least 80% of the energy gap frequency.

SIS JUNCTIONS WITH INTEGRATED TUNING CIRCUITS

High quality Nb-AlO$_x$-Nb tunnel junctions have been fabricated using a trilayer deposition and self-aligned insulator lift-off process [10]. The junction area of 0.25 μm^2 is defined by electron beam lithography, while the junction leads and RF filter structure are produced by conventional photolithography. The Nb films are DC sputter deposited. The base electrode is 1600 Å thick and the top electrode is 2300 Å. The junctions have an estimated specific capacitance of 85-100 fF/μm^2, current densities in the range 7,000-13,000 A/cm^2, and normal state resistances R_N in the range 60 Ω - 120 Ω. The integrated RF tuning circuits are defined by the Nb wiring electrode on a 2000 Å thick SiO insulating layer. Figure 5 shows a typical I-V curve.

Two different integrated RF tuning circuits have been designed and tested. One is a 3-section series microstrip transformer [11] as shown in Figure 1. The other is a parallel microstrip line terminated in a radial stub as shown in Figure 2. For the series transformer, the first section of superconducting microstrip line (starting at the SIS junction) transforms the SIS junction complex impedance to a low real impedance of ~ 2 Ω. The characteristic impedance Z_o of this line was chosen close to $(2\pi f_s C_J)^{-1}$ to optimize the bandwidth (f_s is the signal frequency and C_J is the junction capacitance). The low real impedance is then transformed to 50 Ω with a 2-step Chebyscheff transformer. Each step is one quarter of a guide wavelength, λg, long. The adjustable waveguide circuit (discussed below), can readily provide 50 Ω to couple to the microstrip circuit.

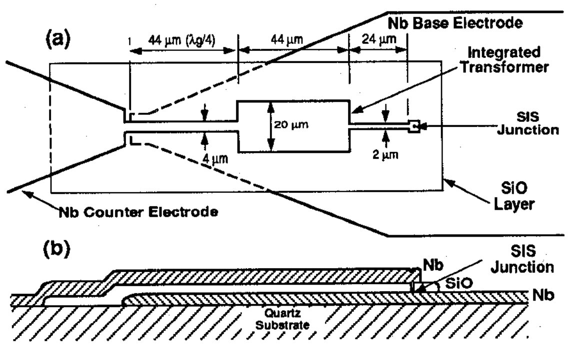

Figure 1: SIS tunnel junction with an integrated series microstrip transformer circuit.
(a) Top view showing transmission line dimensions.
(b) Cross section view showing film topology.

For the parallel circuit, shown in Figure 2, a radial stub is used to provide an RF short over a broad bandwidth. The stub dimensions were designed using an effective dielectric constant, which accounts for the penetration of the magnetic field into the superconductor. The short is then transformed to an inductance by an appropriate section of microstrip line to compensate the capacitance of the junction.

The superconducting microstrip line is a slow-wave line due to the penetration of the magnetic field into the Nb films over a distance comparable to the SiO dielectric thickness. In order to properly design our microstrip circuits, the characteristic impedance Z_o and phase velocity v of the line must be known.

Z

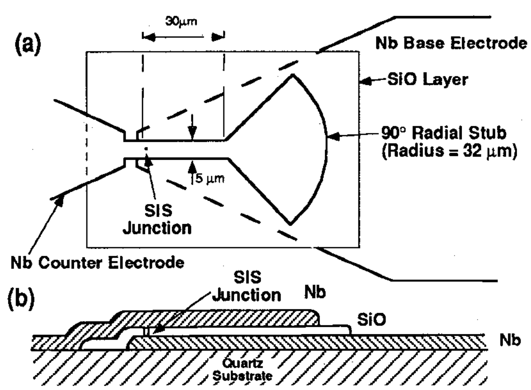

Figure 2: SIS tunnel junction with an integrated parallel microstrip tuning circuit.
(a) Top view showing transmission line dimensions.
(b) Cross section view showing film topology.

These are given by $Z_0 = \sqrt{L/C}$ and $v = 1/\sqrt{LC}$ where L is the inductance per unit length of line and C is the capacitance per unit length. Convenient expressions [12, 13] valid below the gap frequency, are given for these line properties by:

$$C = k\varepsilon_r\varepsilon_0 \, {}^{w}\!/_{t_d} \qquad (1)$$

and

$$L = \frac{\mu_0}{kw}\left[t_d + \lambda\left[\coth\left({}^{t_1}\!/_{\lambda}\right) + \coth\left({}^{t_2}\!/_{\lambda}\right)\right]\right] \qquad (2)$$

where w is the width of the microstrip, t_1 and t_2 are the thicknesses of the top and bottom Nb films respectively, t_d is the thickness of the SiO insulting layer, $\varepsilon_r = 5.5$ is the dielectric constant of SiO, and k is a fringing field factor. k varies between 1.05 and 1.37 as a function of the width of the line [14]. Equations (1) and (2) are valid for ${}^{w}\!/_{t_d} >> 1$. A penetration depth $\lambda = 750$ Å [15] was used as the nominal value.

As indicated by both theory and experiment [15], the dispersion and loss in these Nb transmission lines can be neglected in this application up to about 550 GHz. Thus we have not included these effects in our design. In addition, the step and end discontinuities on the microstrip line have also been neglected since the large width-to-height ratio makes these corrections very small. These circuits have been designed to resonate the junction capacitance based on a specific capacitance ranging from 60 to 100 fF/μm^2; a value which is dependent on the current density of the junction. A range of RF tuning circuits has been designed based on this uncertainty, and a systematic study of mixer performance as a function of junction capacitance will be published at a later date.

RECEIVER DESIGN AND MEASUREMENT TECHNIQUES

The SIS tunnel junction, integrated tuning circuit, and low-pass RF filter are fabricated on 50 μm thick quartz which is cut to a width of 150 μm. This substrate is installed into the waveguide mixer mount and wire bonded to the 50 Ω IF output connector. The waveguide mixer was designed using a low-frequency model to maximize the accessible region of impedances on the Smith chart over an equivalent frequency range of 500 GHz to 600 GHz [9]. This mixer has an adjustable backshort and E-plane tuner which provide a wide range of impedances to the SIS junction. Radiation is coupled into the waveguide mount by a dual mode conical horn [16].

Figure 3 shows a block diagram of the receiver. The local oscillator (LO) source consists of two whisker-contacted Schottky varactor frequency multipliers (x2x3) [17] driven by a Gunn oscillator at 91.5 GHz, the center frequency to detect $H_2{}^{18}O$. We varied this Gunn oscillator frequency and used two other Gunn sources to make measurements in the 460 GHz to 630 GHz frequency range. The signal and LO are combined in a folded Fabry-Perot diplexer and injected into the cryostat through a differentially-pumped window using two 125 μm thick mylar vacuum windows. Fluorogold far IR filters, one wavelength thick at 547 GHz, on the 77 K and 4 K stages, block room temperature radiation from saturating the mixer. An off-axis elliptical mirror reflects the combined radiation into the mixer which is installed on the 4 K stage of the cryostat. The 1.4 GHz IF output of the mixer is transformed to the required 50 Ω input impedance of the low noise HEMT amplifier by a 2 step Chebyscheff microstrip transformer made on 1.27 mm thick Duroid substrate with $\varepsilon_r = 10.5$. The IF is further amplified by two high gain room temperature amplifiers. The bandwidth for noise measurements is 300 MHz. A superconducting magnet is used to suppress unwanted Josephson interference and thus improve receiver performance. It has been designed to provide a magnetic field of about 800 gauss near the junction for a current through the coil of 1 ampere. This is sufficient since 1 flux quantum in our junction corresponds to about 340 gauss.

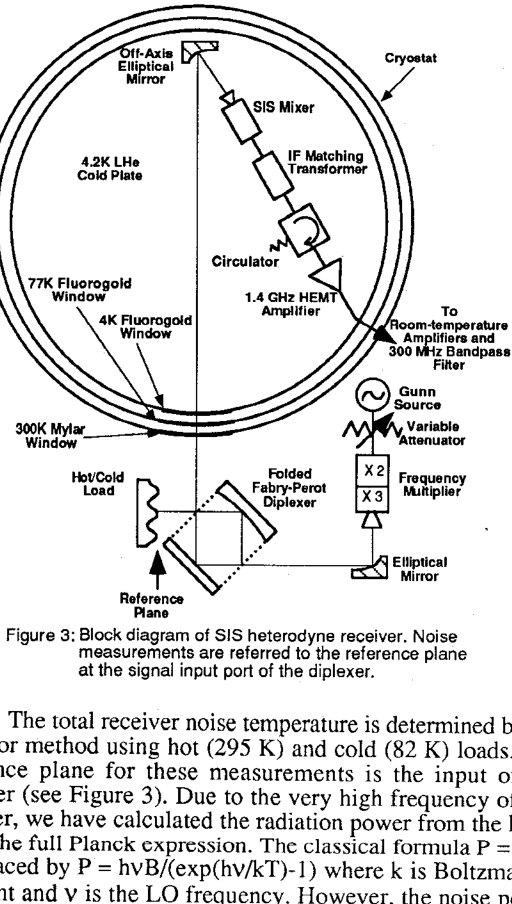

Figure 3: Block diagram of SIS heterodyne receiver. Noise measurements are referred to the reference plane at the signal input port of the diplexer.

The total receiver noise temperature is determined by the Y-factor method using hot (295 K) and cold (82 K) loads. The reference plane for these measurements is the input of the diplexer (see Figure 3). Due to the very high frequency of this receiver, we have calculated the radiation power from the loads using the full Planck expression. The classical formula P = kTB is replaced by P = hνB/(exp(hν/kT)-1) where k is Boltzmann's constant and ν is the LO frequency. However, the noise power per unit bandwidth of the receiver is expressed as a temperature using the classical expression.

RESULTS AND DISCUSSION

Receiver performance

The receiver performance has been measured over an LO frequency range from 460 GHz to 630 GHz. Three different SIS junctions with integrated tuning circuits have been measured. Figure 4 shows the DSB receiver noise temperature as a function of the LO frequency. Junction A used the 3-section series transformer (see Figure 1). While junction B and C used the parallel tuning circuit with the radial stub (see Figure 2). For each data point in Figure 4, the waveguide backshort and E-plane tuner, LO level, and DC bias voltage were optimized. The best performance was for junction C which gave T_R = 200 K (Y-factor = 1.78) at 540 GHz. The receiver performance was very good with this junction, yielding $T_R \leq 300$ K over the frequency range from 463 GHz to 549 GHz. The only exception was a few points near 490 GHz, where the noise increased due to the appearance of negative differential resistance (this is discussed in more detail below). The receiver also performed well up to 627 GHz with a noise temperature of 1100 K, which is the best result today at this frequency. This junction, which had the lowest normal state resistance, $R_N = 63\ \Omega$ gave better results than the two other junctions with $R_N = 110\ \Omega$. One reason is that the junction IF impedance, which is about three times R_N, is better coupled by the IF transformer to the 50 Ω IF system, than the high impedance junctions. This difference in coupling efficiency, however, does not fully explain the observed difference in noise performance. Further theoretical analysis of the mixer performance is required.

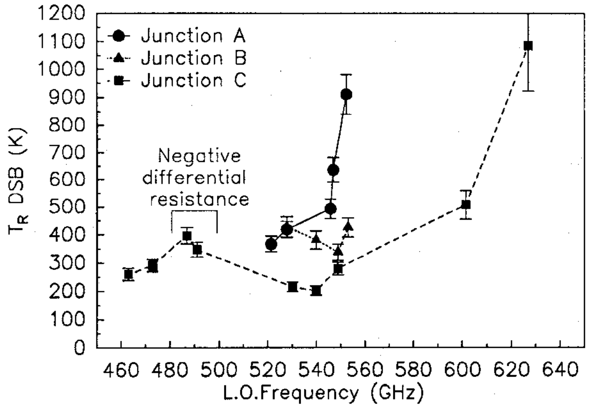

Figure 4: Receiver noise temperature as a function of the LO frequency for 3 different SIS junctions. Junction A (R_N = 108 Ω) uses a series transformer tuning circuit and junctions B (R_N = 114 Ω) and C (R_N = 63 Ω) use a parallel circuit with a radial stub (see text). Very good performance is obtained from 463 GHz to 627 GHz, with the best result T_R = 200 K at 540 GHz for junction C. Negative resistance is also observed near 490 GHz. The IF is 1.5 GHz and the bandwidth for noise measurements is 300 MHz.

Figure 5 shows the unpumped and LO pumped I-V curves, at 540 GHz of junction C, which gave the best performance. The unpumped curve shows very low subgap current and a sharp gap structure at 2.85 mV. The photon step is clearly seen on the pumped I-V curve. Also shown in this figure is the IF output power from the receiver for both hot and cold loads as broadband signal sources at the RF input. It can be seen that the IF power is low near zero voltage which indicates that the Josephson current has been almost completely suppressed with the magnetic field of the superconducting coil corresponding to about 2 flux quanta. However, there is some structure in the IF power curves near 1.1 mV corresponding to the first RF induced Shapiro step, but it does not create any noticeable additive noise or instabilities at the operating bias voltage at about 2 mV. It may, however, suggest that the SIS junction barrier is not uniform over its surface. This effect has not been observed with the two other junctions of lower current densities. These junctions may have a more uniform barrier and a Josephson current which is easier to suppress with an external magnetic field.

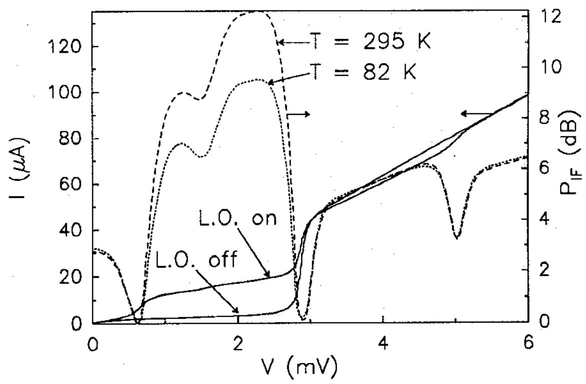

Figure 5: Current vs. voltage characteristic (solid lines) for junction C with and without LO power applied at 540 GHz. Receiver IF output power, P_{IF} (dashed lines) for broadband hot and cold load signals at the RF input.

Negative differential resistance

A DC negative differential resistance has been observed on the 1st photon step at LO frequencies of 487 GHz and 491 GHz. This is a quantum mechanical effect which is not predicted by classical theory. Also, negative resistance implies that the available mixer gain can be infinite in theory. When the mixer is biased in the negative resistance region, the IF output power is very high and unstable, probably due to low frequency oscillations. As a result, the receiver is unstable and noisier at these frequencies. However, since the negative resistance did not extend completely across the photon step, it was possible to bias the junction at these frequencies though not at an optimum point. As seen in Figure 4, the receiver noise is higher by 100 K - 200 K in this region.

The negative resistance at 491 GHz could not be eliminated by adjusting the backshort and E-plane tuner. This indicated that the microstrip tuning circuit had fully compensated the junction capacitance and was very strongly coupled to the junction. Also, the microstrip losses must be very low. A similar result has been previously reported at 230 GHz [18]. The LO source does not cover the range 500-520 GHz, so the full bandwidth over which this tuning circuit resonates well is not known. This circuit was nominally designed for 547 GHz. This result shows that it actually resonates at a lower frequency than expected (in agreement with Josephson resonances, as discussed below). These results also suggest that it is not desireable to operate at the optimum resonant frequency due to possible instabilities caused by negative resistance.

Mixer noise temperature and conversion efficiency

The shot-noise produced by the linear resistance of the SIS junction I-V curve for voltages higher than the gap voltage was used to calibrate the IF system. This is a simple technique to calibrate the IF system noise and gain, thus allowing the mixer noise and conversion efficiency to be estimated from the measured receiver performance. The SIS junction acts as a variable temperature load when the junction is biased on the linear part of its I-V curve. The IF system output power as a function of this temperature is a straight line: its slope is the gain of the IF system and the temperature at zero IF power is the IF system noise. From this calibration, the junction which exhibited 200 K noise temperature at 540 GHz, has a mixer DSB noise temperature of about 170 K with a coupled conversion efficiency (including the IF mismatch) of about -7 dB. The largest contribution to the receiver noise comes from the mixer noise. The mixer noise includes the RF mismatch and all the losses in the optics, the windows and the Fabry-Perot.

High frequency limits

As seen in Figure 4, the receiver noise begins to increase near 550 GHz for each junction tested. We conjecture this is due, in part, to the bandwidth of the tuning circuits. These circuits were designed to resonate the junction capacitance near 547 GHz, however, resonant peaks in the DC I-V curve suggest the circuits resonate at lower frequencies. That is, an SIS junction biased at a voltage V_0 produces a high frequency oscillation given by f = 484 GHz x V_0 (mV) due to the AC Josephson effect. Generally, at high frequencies, the capacitance of the junction shunts these oscillations. But at the frequencies where the tuning circuit resonates the capacitance i.e. at the DC voltages corresponding to these frequencies, these Josephson oscillations interact with the non-linear resistance of the SIS junction and give a DC current peak on the I-V curve [19]. Peaks we observed at about 0.95 mV, corresponding to 460 GHz, for junctions with the series transformer and at about 1.08 mV, corresponding to 520 GHz for the junctions B and C with the parallel tuning circuits. Thus for frequencies well above these (i.e. 550-600 GHz), the waveguide backshort and E-plane tuner are not able to compensate the junction capacitance and the receiver noise begins to increase due to a poor RF match.

Another source of noise increase near 545-550 GHz is due to the presence of a strong water line at 557 GHz. The line absorbs part of the signal coming from the cold load and emits at room-temperature. By putting the cold load at different positions: close to the Fabry-Perot diplexer and up to 9 cm away, a difference in the receiver temperature is clearly observed. From this increase of temperature, it is possible to determine the absorption due to the water line. An absorption of about 0.037 dB/cm at 549 GHz was deduced. This measured value is in reasonable agreement with the theoretically calculated absorption given the inaccuracy of the technique. Such an absorption can add at least 30 K at 549 GHz to the receiver temperature due to the signal path length up to the cryostat window. All junctions show some increase in the receiver temperature above about 545 GHz.

CONCLUSION

A receiver for radioastronomy applications in the range 460-630 GHz has been demonstrated. The best result is a DSB receiver temperature as low as 200 K at 540 GHz. The mixer noise is about 170 K and the conversion efficiency is around -7 dB. The optics have been designed to work in the 500-600 GHz RF frequency range. Nevertheless, this receiver was tested at lower and higher frequencies to determine the behavior of the integrated tuning circuits which resonate with the SIS junction capacitance. A differential negative resistance has been observed at 487 GHz and 491 GHz which indicates that the circuits work better at frequencies lower than designed. This is also confirmed by the resonances resulting from the AC Josephson current and observed on the unpumped I-V curves. At frequencies higher than 600 GHz, the receiver noise increases quickly with the frequency which is partly due to the roll off of the tuning circuit.

In addition, we have begun measuring the performance of this receiver using a series array of two junctions made by optical lithography techniques [20]. A noise temperature T_R (DSB) = 330 K has so far been achieved at 491 GHz. This work is in progress and will be reported later.

ACKNOWLEDGEMENTS

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