An 800 GHz SIS mixer using Nb-Al_2O_3-Nb SIS junctions

C.E.Honingh, K.Jacobs, D. Hottgenroth, and S.Haas.

Kölner Observatorium für MM- und Submm- Astronomie

KOSMA
I. Physikalisches Institut der Universität zu Köln
Zülpicher Str. 77
D-50937 Köln
Germany

Introduction

The development of an 800 GHz mixer is part of a project to build a dual channel SIS receiver for submillimeter frequencies to be used at a ground based telescope. The two channels are integrated in one hybrid liquid helium dewar, and are each fed by one polarization of the incoming signal. Since most mixers are sensitive to one polarization only a factor of two in observing time is gained. Also the simultaneous observation at two frequencies improves the calibration accuracy of the two sets of data with respect to each other. This is especially important in observing spectral lines in molecular clouds of which the relative intensity at two frequencies is crucial to determination of their properties.The receiver will have one channel just for 450GHz-500 GHz, while the other channel can be used alternatively for 650-700 GHz or 800-820 GHz.
For the two lower frequencies superconductor-insulator-superconductor (SIS) quasi-particle mixers are the obvious choice. In the past few years there has been a rapid development of SIS mixers towards use at higher frequencies, and various groups have shown sensitivities as good as five to ten times the quantum limit in both bands.[1-5]. This development has been possible because high quality, high current density Nb-Al_2O_3-Nb SIS junctions became available, with areas small enough to be useful at higher frequencies.
A typical current density at 700 GHz is 10 kA/cm^2, with a leakage current below the gap voltage smaller than 10 μA. The area of the junction that is used depends on the method that is chosen to tune out the geometrical capacitance of the junction. A tuning that depends solely on adjustable tuners has been used up to 750 GHz [5]. The tuning of this type of mixer usually has to be adjusted at every frequency. Generally mixers that have a larger instantaneous bandwidth use a microstrip tuning structure integrated with the junction. The microstrip metallization is also made of niobium. This makes the tuning structures almost lossless (below the gap frequency of niobium), which is crucial to their excellent performance. This types of tuning allows somewhat larger junction areas, typically 1 μm^2 around 700 GHz, than the former type.
For the 800-820 GHz band the noise temperatures reported so far are about a factor of 10 higher than at 700 GHz. It has been shown [5], using a waveguide mixer with two mechanically

adjustable tuners and a junction with an area of 0.25 μm^2, that above the gap frequency of niobium (700 GHz) the Nb-Al_2O_3-Nb junctions still function very well as mixing elements. The tuning of the geometrical capacitance of the junction however is more complicated because integrated tuning structures will have losses and this rapidly deteriorates the sensitivity of the mixer [6].

1 Design considerations

SIS junctions and tuning

For the SIS junction we will take a simple equivalent circuit consisting of its normal state resistance (Rj) parallel to its geometrical capacitance (Cj). Apart from all other considerations the current density of an SIS junction to be used at 800 GHz should be equal or higher than 12 kA/cm^2 to have an ωRjCj product below 7. This is among the highest current densities shown so far, in junctions that still have a good SIS barrier. At even higher current densities the barrier becomes so thin that microshorts may occur, which potentially increase the shotnoise production of the junction.

The value of ωRjCj makes some kind of tuning necessary. Low loss waveguide tuners alone will not be sufficient. Although an E-beam lithography facility has been set up recently as part of this project, as a first approach we had to work with junctions that we could fabricate with standard UV lithography. This results in junction areas not much smaller than 1 μm^2 which require some kind of integrated tuning. We used traditional integrated tuning circuits, made of niobium, similar to those used below the gap frequency. In general the integrated tuning structures can be split in two types. Parallel resonant structures which connect an inductance parallel to junction capacitance, and series resonant structures, which put a small inductor in series with the junction. These latter structures are usually followed by a transformer to transform the very low impedance to a value that can be matched better to an antenna.

A quick estimate of the effect of a parallel resonant structure with a lossy inductor can be obtained from a lumped element equivalent circuit, given in Fig. 1A. Around the resonance frequency the loss in series(R_s) with the inductor(impedance X_s) can be transformed into a resistor (R_p) parallel to the junction resistance via $R_p=(Q^2+1)R_s$, where $Q=X_s/R_s$. The real and imaginary part of the inductive impedance are calculated from an ordinary transmission line model for lossy transmission lines.

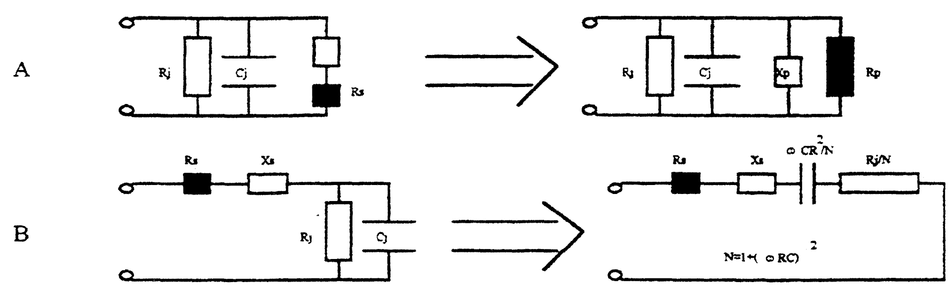


Fig. 1 Lumped element equivalent of a parallel and a series resonant circuit for an SIS junction Rj//Cj, with a lossy inductor (Rs + iXs).

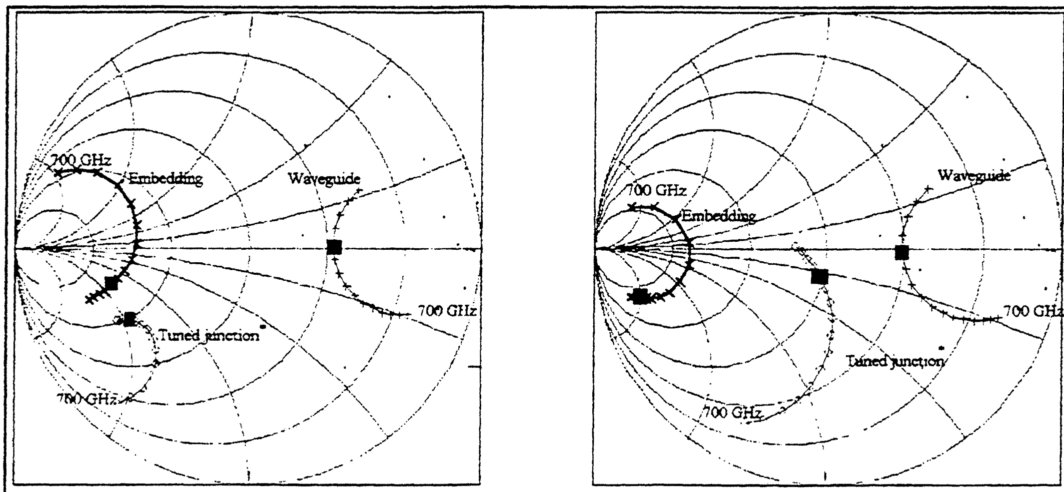
For a junction with area 0.95 μm^2, jc= 10 kA/cm^2 and Cj= 70 fF , Rp lies somewhere between 3.5 and 4.3 Ω, depending on the exact estimate of the loss in niobium. The transmission line is 4 μm wide, on a 200 nm thick SiO dielectric. The 4.3 Ω result from the Mattis Bardeen (MB) [7] theory in the extreme anomalous limit, which is certain to give a too low estimate of the loss [6, 8]. The MB calculation is made with a normal state conductivity of niobium of $1.25 \times 10^7 (\Omega\,m)^{-1}$, an energy gap 1.45 meV at zero Kelvin, and a critical temperature of 9.2K. The 3.5 Ω are found if the loss is taken 2.5 times as large. Comparing these resistances values with the junction normal state resistance of 23 Ω gives a maximum coupling to the junction of approximately 16%. Also the resonance frequency depends on the loss in the structure. Between the two extremes of loss that are used here the difference in resonance frequency is about 35 GHz. The situation can be improved by increasing the current density of the junction. To first order this reduces Rj without changing Cj, thus improving the ratio between Rp and Rj. Also the junction area can be made smaller, which reduces Cj, and leads to a higher value of Rp at resonance. Because Rj also increases the effect is not very large. For a junction with area 0.6 μm^2, jc= 12 kA/cm^2 and Cj= 45 fF, a maximum coupling between 20% and 26% is calculated this way.

For a series resonance, shown in Fig. 1B, a similar reasoning is possible. The SIS junction is now taken as a series circuit of a capacitance and a resistance. For a junction with area 0.95 μm^2, jc= 10 kA/cm^2 and Cj= 70 fF, this resistance is 0.33 Ω. A resonating series stub will have an impedance with a real part between 0.22 and 0.55 Ω for the two extremes of loss mentioned above. This leads to a coupling to the junction between 37% and 60%., much better than with the parallel stub. Unfortunately the resulting impedance is so low that a quarter wave transmission line transformer has to be used to transform the impedance to a reasonable level. This line attenuates the power by a factor exp $(2\alpha l)$, where α is the real part of the complex propagation constant along the line and l is the length of the transformer. The extra factor 0.5 to 0.8 leads to a total coupling between 20 % and 50 %.

As mentioned above, the transformed impedances have to be matched to a receiving antenna. Also the resulting embedding impedance of the junction, which determines its conversion and noise properties has to be evaluated. In Fig. 2 the embedding impedance and the (complex conjugate of) the transformed junction impedance are shown for a parallel(A) and series resonant(B) tuning circuit for a junction with area 0.95 μm^2, jc= 10 kA/cm^2 and Cj= 70 fF. As antenna impedance we chose the impedance of a standard waveguide mount. The loss in the tuning structure has the unavoidable effect that in the case of a series resonance structure the matching between the transformed junction impedance and the waveguide impedance is better than the parallel one, but the embedding impedance is worse. This leads to the question which effect has stronger influence on the overall receiver performance, the embedding impedance of the junction or the coupling between the waveguide and the junction.

Since the quantum theory of mixing [8] seems to predict mixer performance rather accurately even well above the gap frequency [6], we used the embedding impedance and the coupling efficiency shown above to calculate the overall receiver performance. The result is similar for the two structures and both temperatures lie between 1400K and 1800 K , assuming an IF frequency of 1.5 GHz, an IF noise temperature of 10 K and an IF input impedance of 50 Ω.

Accordingly we have implemented both types of tuning structures on the fabrication mask. An additional advantage of the series resonance structure might be that it is not as sensitive to alignment accuracy, with a length of around 30 μm and a width of 5 to 8 μm. The parallel structure needs a narrower line, almost as short as it is narrow, and an open circuited section to create an RF-short for which we used both radial and straight stubs. Because this last line section is much wider line its additional contribution to the loss is small. The resonance frequencies of both structures are dependent on the actual loss in the circuit. To reach receiver noise temperatures below 1000 K, assuming the same losses in the niobium tuning structure, we tentatively put some junctions with an area of approximately 0.6 μm^2 on the mask to be fabricated with a current density of 12 kA/cm^2.



A B

Fig. 2. Calculated embedding admittance (x), antenna admittance (+) and the complex conjugate of the tuned junction admittance, for a parallel (A) and a series (B) resonant tuning structure. The impedances are calculated from 700 GHz to 850 GHz, and the marked point is at 810 GHz. Both charts are normalized to 23 Ω.

Mixerblock

We use a traditional waveguide mixer block. A block with a backshort as well as an E-plane tuner would have been optimal, especially to reach the low impedance created by a parallel resonant tuning structure. To avoid the complication of fabricating a mixerblock with two perpendicular tuners, we built a block similar to the 500 GHz mixer [1] with a single backshort. The waveguide dimensions are 330 μm x 95 μm, and those of the substrate channel 100 μm x 95μm. The block is fabricated in split block technique to minimize the loss in the waveguide due to imperfect contact between the two halves. The waveguide and the substrate channel have been sawed into the brass mixer block, using a carbon hardened saw blade with a width of 95 μm.
The junction is fabricated on a fused quartz substrate with integrated RF blocking filters. The substrate is glued in the substrate channel with Crystal Bond in such a way that the junction is placed in the center of the waveguide, facing its narrow wall. Electrical contact with the end of the RF filters is made by wire bonding. A simple IF matching circuit is implemented in the mixer block. The reflection of the RF blocking filters is adapted to reach low waveguide impedances also

with one backshort. We use an integrated diagonal horn machined directly in line with the waveguide as described in Ref. [1].

2 Measurements

The DC IV-curve of the junction used in the measurements is shown in Fig. 3. The junction is fabricated at Cologne University, it has an area of 0.95μm^2 and a notably high current density of 15 kA/cm^2. The leakage current is about 20 μA. For mixer measurements a solid state local oscillator (Gunn + multiplier, 800-820 GHz) [9] is coupled to the mixer using a Martin-Puplett diplexer. The pumped curve is also shown in Fig. 3. It is taken at the maximum output power of the oscillator, which is 30 μW at 810 GHz according to the manufacturer. The junction used in the experiment had a simple parallel tuning structure, consisting of a short stub 4μm wide and 5μm long, followed by an open ended line of 18 μm wide and 20 μm long. This structure had a resonance in the DC IV curve at a bias voltage corresponding to approximately 780 GHz.

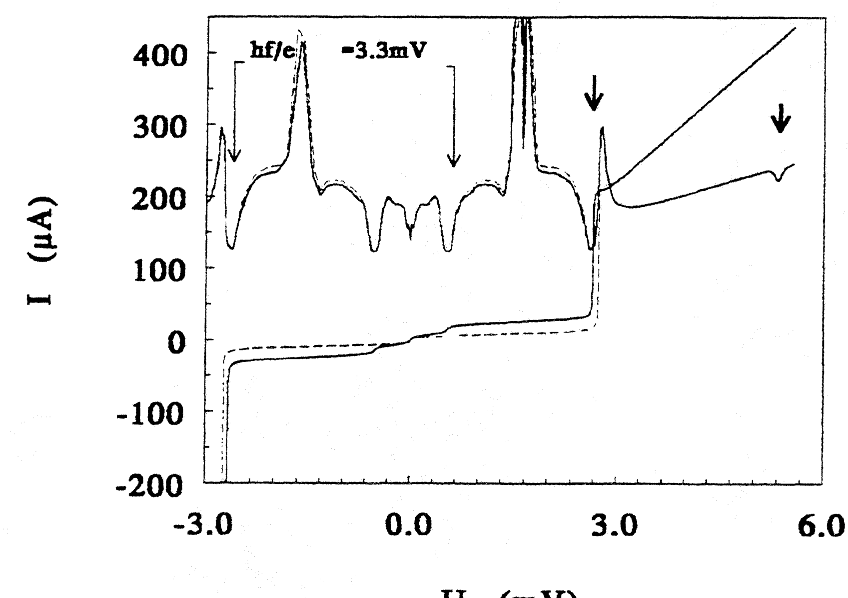


Fig. 3 DC IV-curve of the junction without and with 810 GHz radiation (quasi-particle pump step width is 3.3 mV).Notice the smaller step width above the gap voltage.The corresponding IF-ouput power at a 300K and a 77K input is given on the same voltage axis.

The mixer response to a hot and cold load is also shown in Fig. 3. The Josephson effect is suppressed as good as possible by a magnetic field. The receiver noise temperature calculated from this measurement is 4000 K DSB. Other junctions, also with parallel tuning structures using a radial stub, or slightly different line length and width have also been tested. The best noise temperatures all were between 3800K and 4200 K DSB at 810 GHz. At the other frequencies (800GHz to 820 GHz) the output power of the oscillator is lower. As can be seen in Fig. 3, the pump level at 810 GHz is already rather low for optimum mixer performance, so at the other frequencies the performance is probably worse due to LO-starvation. Junctions with series resonant tuning structures were unfortunately not available for measuring.

In all measurements there was a rather sharp dependence of the maximum input coupling on the backshort position. The slope of the pump step was not changed by the backshort setting. We

have however observed different pump step slopes for different tuning structures, indicating that the integrated tuning structures have some effect.
To get an impression of the quality of the tunnel barrier, the IF output power of a junction was measured as a function of bias voltage at 1.4 GHz in a 100 MHz bandwidth. The result is given Fig. 4 , together with the IF output calculated from the measured unpumped DC IV-curve. For the IF-amplifier a noise temperature of 10 K is determined. Apparently the junction does not show additional shotnoise below the gap voltage. At half the gap voltage a clear structure is observed, which is not affected by a magnetic field.
In a measurement to determine the cross polarization of the diagonal horn we found to our surprise that the mixer receives both polarizations equally well. For a standard diagonal horn a cross polarization detection level of 10-15 % is expected [10]. Scale model measurements at 100 GHz subsequently showed that a small misalignment , especially at the apex of the horn, destroys the symmetry of the horn in such a way that high cross polarization levels are easily achieved.
The misalignment in the scale model corresponds to 10 to 20 μm in the 800 GHz mixerblock, which was indeed observed. But even with a better alignment, tested in the 500 GHz mixerblock, we still observed a crosspolarization detection of 25-30% of the copolarization.

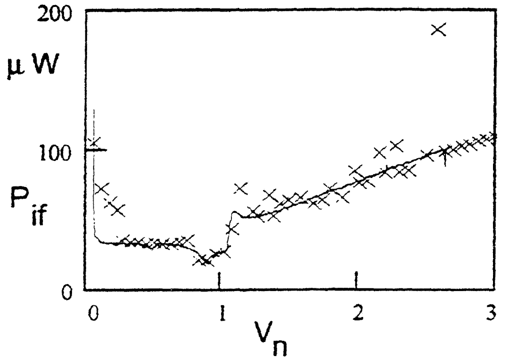


Fig.4 Measured (solid line) and calculated (crosses) IF output power of the unpumped SIS junction as a function of normalized bias voltage.

3 Discussion

The calculated noise temperature for the junction shown in Fig. 3 is approximately 1500 K.
This temperature is calculated using an estimation for the actual losses in the superconducting microstrip lines obtained from DC measurements. The resonance frequency of 40 different parallel resonant tuning circuits was determined from the bias voltage at which the current rise due to the AC-Josephson effect was observed in the DC IV curve at 4.2 K. Under the assumption that the current density is uniform over the whole wafer, the area of the junctions was calculated from the measured normal state resistance. For a specific junction capacitance of 75 fF/μm^2 the average loss in the stripline lies between 2 and 2.5 times the value calculated from MB, taking the material parameters mentioned in paragraph 1.
With this loss the embedding impedance of the junction is calculated from the geometry of the tuning structure, adapting the actual length of the tuning stub to the resonance frequency measured in the DC-IV curve. (the typical adaption of the length is smaller than 1μm, which is within the alignment accuracy)
Because the coupling of the LO and the signal are made with a polarizing diplexer, the high cross polarization level of the horn has an effect on the receiver noise temperature. If a loss of 40% is included in the signal path, the calculated noise temperature increases to 2000 K. Other influences, like the somewhat low pump level or the slight heating of the pumped curve, have only a small effect on the noise temperature.

We are still investigating various effects that might explain the measured noise temperature of 4000K. For example possible losses in the niobium RF-filters could influence the waveguide impedance. Equating the output power of the LO-source with the pump step height shows that only approximately 5 % of the available LO-power is coupled to the junction. We have not fitted an embedding impedance from the measured pumped IV-curve, but the flat slope of the pumped step indicates that the junction capacitance is at least partly tuned out. On the other hand the heating of the pumped IV-curve, as can be clearly seen from the reduction of the gap voltage, might be a sign of bad matching. Badly matched junctions at frequencies below the gap frequencies show a similar gap reduction

Even if we would have measured the theoretically predicted noise temperature however, the result is still only comparable with the best Schottky receivers at this frequency. A definite improvement can only be achieved if the loss in the tuning is reduced. It might be necessary to build a waveguide mixer with two tuners [5]. Aluminum electrodes might be better than niobium at frequencies above 800 GHz [6], and have the advantage that they require no drastic changes in the fabrication process. The propagation velocity along normal metal lines is different, which leads to a little bit longer structures. Really low loss tuning structures at these high frequencies can probably only be made of niobium nitride (NbN). In that respect the mixer results at 350 GHz [11], measured at IRAM, are very promising. A disadvantage of NbN is that the length of the tuning strucures is even shorter than that of the Nb stubs. This makes the alignment during fabrication very critical.

An alternative might be the use of a superconducting hot electron bolometer mixer, which is not limited by the gap frequency of the superconductor. It has a negligible capacitance, so no resonant tuning circuits are required. The noise temperature of this device is expected to be independent of frequency. Skalare et al.[12] reported a noise temperature of 650K DSB at 533 GHz at an IF frequency of 1.5 GHz.

We are currently working towards fabricating smaller junctions and superconducting hot electron bolometers to replace the present junctions. A new mixerblock is fabricated, which will have a corrugated horn or a Potter horn. Via a collaboration with IRAM [11] that has been set up two years ago for the development of NbN fabrication technology, we hope to get low loss tuning structures.

Acknowledgements

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