A BROAD BAND 230 GHZ ANTIPODAL FINLINE MIXER FOR ARRAY RECEIVERS

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ABSTRACT Finline transmission lines are easy to manufacture and to integrate with other planar components, but they have not, to our knowledge, been used for submillimetre SIS mixers. One of the main problems is the high characteristic impedance of these structures. In this paper we describe a design which overcomes this difficulty by including an antipodal finline section with overlapping fins on either side of a dielectric layer deposited on a thick quartz substrate. We have designed and tested an antipodal finline mixer with fully integrated tuning. Preliminary results confirm our predictions that the finline structure is suitable for making low noise and easy to manufacture SIS mixers, and hence is suitable for large-format imaging arrays.

INTRODUCTION

At the present time there is a considerable interest in developing mixers that are suitable for large-format imaging arrays. An important requirement is that these mixers should be compact and easy to manufacture. In this paper we describe a new mixer design which satisfies the above requirements. The mixer is fed by a waveguide diagonal horn which couples into an antipodal finline taper. This then couples into a microstrip line containing the junction. In this way the majority of the mixer is fabricated using planar circuit techniques; the only mechanical work required — manufacturing the horn — is straightforward. In particular, the design does not require either a back short or an E–plane tuner. Moreover, the IF assembly is in line with the axis of the horn so that the whole mixer is contained within the footprint defined by the aperture. Despite these simplifications, which become more rewarding at higher frequencies, we require that the noise performance of the mixer be comparable to that of more conventional probe-coupled waveguide mixers. This can be achieved by integrated tuning and careful design of the transmission lines so that mismatch losses, insertion and radiation losses are small. The mixer described in this paper was developed in the context of the MARS imaging array project (Padman and Blundell 1994), (Yassin, Padman and Withington 1994). The mixer uses the WR4 waveguide

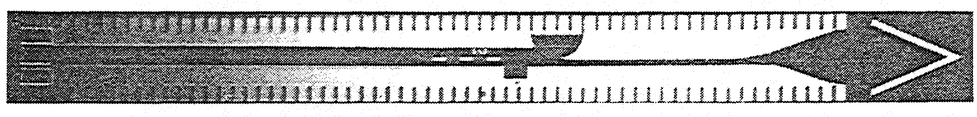

FIGURE I   A Photograph of the Mixer Chip

band (170–260 GHz), but it can easily be be modified to operate at much higher frequencies if desired.

A magnified view of the mixer chip is shown in Fig. 1. The RF transmission line comprises a transition from waveguide, to modified antipodal finline, to microstrip (Hinken, Niemeyer and Popel 1986). The fins (which constitute the base and the wiring layers) are made out of Nb and are separated by a 300 nm of SiO_2 and deposited on a 170 μm thick quartz substrate. It should be emphasized that the electrical properties of this section are mainly determined by the thin SiO_2 layer, rather than those of the quartz sheet. The latter is chosen relatively thick (20 % of the waveguide height) to allow a robust and an easy to handle structure. In the region before the fins overlap the thickness of the SiO_2 is much less than that of the quartz sheet, and the transmission line behaves like a unilateral finline on a quartz substrate. The impedance in this section is brought down from several hundred ohms to 80 Ω as the finline gap is reduced from 0.5 mm at the waveguide input to about one micron. As the fins overlap the structure it starts to behave like a parallel-plate waveguide with an effective width equal to that of the overlap region. When the width becomes large enough such that fringing effects can be ignored, a transition to microstrip is performed which in turn is tapered to the required width of 4 μm.

A feature of this design is that it yields low-impedance values as a result of the practical limitations on the SiO_2 thickness. This, on the one hand requires more effort in designing the transmission lines, but, in return, it allows for large area junctions which are easy to manufacture. Moreover the fringing becomes very low as a result of the large width to height micostrip ratio. The junction area required for a MARS device is approximately 1.5 μm^2. In designing the microstrip section, where the the SIS junction is located, we go for as high an impedance as practically possible. Taking a microstrip width of $w = 4\,\mu$m, dielectric thickness of $h = 300$ nm (dielectric constant of SiO_2=3.8), and metallization thickness $t = 300$ nm we obtain a source modal impedance of 12 Ω. This value is increased by a factor of 1.3 as a result of the field penetration when the line becomes superconducting (Yassin and Withington, 1995).

The RF choke is designed in the usual manner and the capacitance of the junction is tuned out by an inductive strip which is terminated by a half moon radial stub. The IF and bias board follow the chip and are separated from the bonding pads by a 200 μm gap. The resulting inductance of the bonding wires is then less than 0.5 nH which has negligible effect at 1.5 GHz (the IF frequency).

DESIGN OF THE TRANSMISSION LINE

The conventional way to taper a quasi-TEM transmission line is to taper the geometry according to the impedance profile. This method has two disadvantages. The first is that the characteristic impedance of a non–pure TEM mode is not uniquely defined, and the second is that calculation of the characteristic impedance of finlines with thick metallisation and for all gap dimensions is not easy. We designed the MARS transmission lines employing an "Optimum Taper Method" which only requires the knowledge of the propagation constant (and the cutoff frequency) as a function of the finline lateral dimension (Schieblich Piotrowski and Hinken 1984). This method is based on minimizing the coupling coefficient between the incident and reflected waves along a quasi–TEM transmission line (Sporleder and Unger 1979). The end product is a minimum-length high-pass section which gives a return loss lower than a specified design value $R_{\rm max}$ at frequencies above a design frequency f_0.

In the present work we employed the above approach to design the finline section although the overlapping-fin section was considered as a parallel plate waveguide in order to make a smooth transition to the microstrip section where the metallisation thickness is rigorously considered. This approximation was found to be satisfactory within the practical tolerances. The design of the microstrip taper was also based on the above method, but in in this case an analytical expression could be used to compute the impedance as a function of the longitudinal coordinate (McGinnis and Beyer 1988). This approach is convenient since the characteristic impedance of a microstrip is unambiguously defined and can be calculated accurately using the conformal mapping method.

Our computations showed that a total taper length of two wavelengths is sufficient to match the $300\,\Omega$ impedance of the loaded waveguide to the $15.5\,\Omega$ seen by the junction. In fact a much shorter taper should be sufficient provided the electrical properties of the section where the fins overlap are predicted accurately. Our design method was supported experimentally by model measurements at 5 GHz where a 15 dB return loss value was measured for a back to back assembly (Yassin, Padman and Withington 1994). The model measurements also highlighted the following important factors:

1. It is essential to employ a serrated choke in order to suppress the excitation of higher order modes in the slot which supports the quartz substrate. We found that it was necessary to leave a small gap between the metallization and the adjacent face of the slot. In this device the thickness of the substrate ($\approx 200\,\mu$m) is slightly less then the depth of the slot in the lower half of the split block, in which the chip is located.

2. The semicircular cavity which is formed when the two Nb layers are tapered to a microstrip (Ponchac and Doney 1988) did not present measurable resonances within the waveguide band. This may be the result of the right choice of the cavity diameter in addition to the fact that the microstrip fringing at the location of the cavity was negligible.

THE SIS JUNCTION

The SIS junction was fabricated at the same time as the transmission line, using a four-cell mask, and was located in the microstrip section (Blamire, *et al.* 1988) . At first an Nb/Al-oxide/Nb trilayer is sputtered on a 170 μm thick wafer. The trilayer is then coated with positive photoresist and one cell on the mask is used to define the base finline layer and the ground plane for the microstrip. After etching, the base layer is again coated with photoresist and a second cell on the mask is used to define the junction and the SiO_2 layer. The SiO_2 layer is then sputtered while the junction is still protected with photoresist, isolating the junction and at the same time forming the dielectric medium for the microstrip. The photoresist and the SiO_2 layer covering the junction area are then lifted off and a third cell is used to pattern the wiring layer which includes the microstrip, the RF choke and the top fin. Finally, the fourth cell is used to define the palladium bonding pads.

The junction employed in the present experiment had a normal resistance of 34 Ω, a current density of 2500 A/cm^2 and an area of 2.4 μm^2. Assuming a junction capacitance of 55 fF/μm^2 we obtain $\omega R_n C = 6.5$. Unfortunately however, the stub length was initially designed for a junction area of 1 μm^2, and hence its resonant frequency came out to be about 130 GHz instead of the required 230 GHz. The performance was therefore limited by this discrepancy rather than by the intrinsic properties of the mixer.

The RF choke was made of alternating high and low impedance $\lambda_g/4$ sections to provide the high rejection of RF signals over the operating band. In computing the required embedding impedance and the choke and tuning stub parameters, we included the effect of superconductivity on the modal values via the Mattis–Bardeen theory. Because of the large w/h value it was not necessary to consider the effect of the current non–uniformity over the microstrip width, but this could easily be incorporated in our formulation.

THE MIXER BLOCK AND IF ARRANGEMENT

The split mixer block comprises a diagonal horn (Withington and Murphy 1992) and a WR4 waveguide (waveguide dimensions 0.0215 in $\times$ 0 043 in and frequency band 170–260 GHz) as shown in Fig. 2. The width of the groove in the side of the waveguide was made 0.2 mm to accommodate the substrate and its depth (0.17 mm) was chosen to accommodate the serrated finline choke. The length of serrations was chosen to be slightly larger than $\frac{\lambda}{4}/\sqrt{\epsilon}$ where ϵ is the dielectric constant of quartz (Tomiyasu and Bolus 1956). The magnetic field required to suppress the Josephson effect was supplied by a coil wound on a flanged hollow tube, which plugs around the waveguide so that the junction is located inside the coil on its axis. We used a coil of 2500 turns, a wire diameter of 70 μm and a coil former of 6.5 mm diameter. With this arrangement a current sweep from 0–300 mA was sufficient to sweep through several Josephson nulls separated by approximately 60 mA.

The IF circuit consisted of a ground-backed coplanar waveguide (CPW). The CPW was preferred to the more commonly used microstrip mainly because the ground planes and the central conductor are in the same plane with the

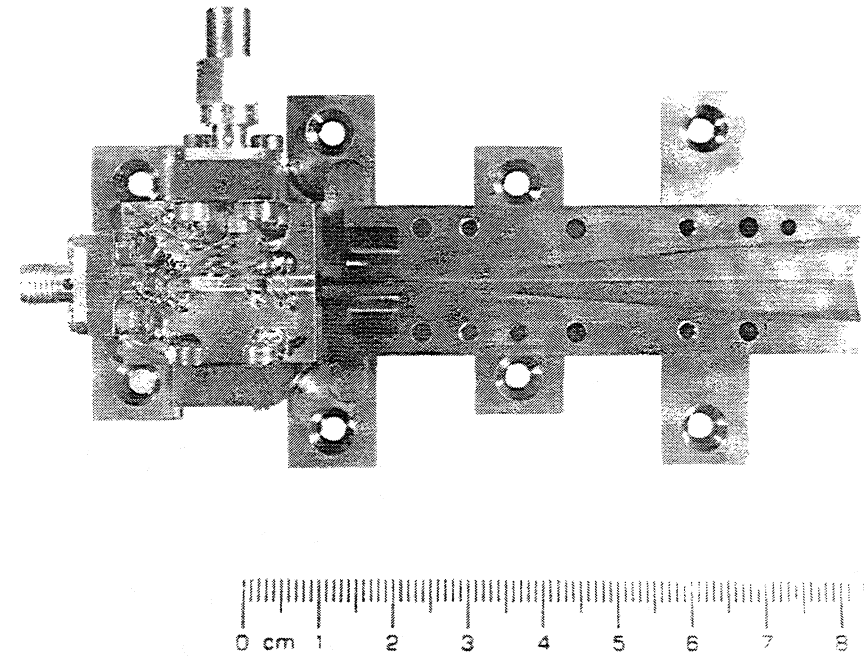

FIGURE II A Photograph of the Mixer Block Lower Half and the IF-bias board

bonding pads on the chip, which made connecting the pads to the IF circuit easier. The CPW circuit was patterned on a Duroid sheet of 1.27 mm thickness, and to ensure that the ground planes were equipotential, the CPW ground planes were mutually connected by copper bridges in several places along the strips. DC bias to the junction was supplied via an inductor which was connected to the central strip. Both the IF board and the mixer were bolted to a cradle fixed to the cold plate of the dewar. No effort was made to match the input impedance of the CPW ($50\,\Omega$) to the output impedance of the mixer or to the normal resistance of the device.

It should be emphasized that the above design was only intended to test the operation of the finline mixer. Several improvements can be made to this scheme, and some of them have already been implemented in a new design which is more suitable for the array receiver.

EXPERIMENTAL RESULTS

LO power was supplied by a 76 GHz Gunn Oscillator with a frequency tripler providing an output power of 0.5 mW. Power from the radiating horn was then focused onto a 75 mil. mylar splitter which was mounted at 45° relative to the normal to the dewar window. The LO signal is relayed to the mixer by a PTFE lens supported by a conical section bolted in front of the mixer horn. The IF signal was taken from the chip to the CPW via short wires (as explained earlier) and is then fed to an isolator, followed by the IF amplifier.

With this arrangement, we were able to obtain sufficient power to pump the junction. This is illustrated in Fig. 3 where we show the generation of photon steps, and the complete suppression of Josephson effect (at a coil current of $\sim$ 120 mA corresponding to the second null), at frequencies of 213, 218 and 240 GHz. The pumped I–V curves clearly show the effect of the tuning stub.

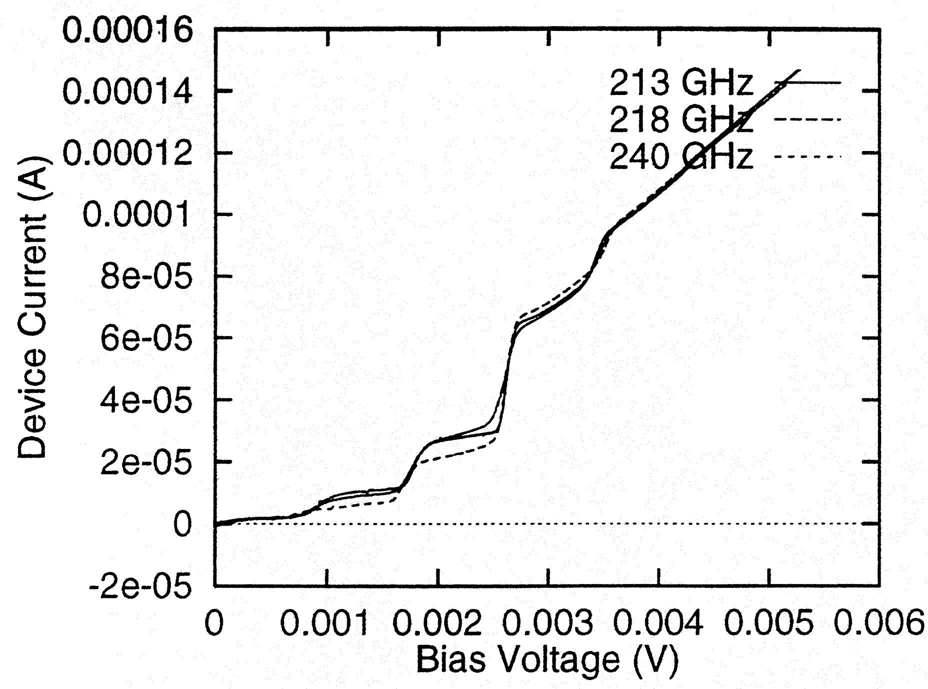

FIGURE III Pumped I-V Curves at 213, 218, 240 GHz with Suppressed Josephson Effect

From the slope of the photon steps we notice that the embedding impedance becomes less capacitive as the frequency decreases. This is consistent with the the stub being tuned to a much lower frequency. The fact that the stub had a significant effect on the embedding impedance despite the very large detuning demonstrates the wide–band nature of the stub.

The heterodyne performance of the mixer was next studied by feeding the output of the IF amplifier to a 20 dB gain 0.5–2 GHz (warm) amplifier and then to the detector. In this way we could display the IF output power as a function of bias voltage for a given LO frequency when the mixer looked into either a hot load (290 K) or a cold load (80 K). The noise temperature of the receiver could then be found by measuring the Y–factor. An example of these measurements at 218 GHz is shown in Fig. 4, where we compare the curves corresponding to hot and cold loads together with the unpumped output. Here also we notice the efficient suppression of the Josephson resonances, this time at a coil current of only 58 mA (corresponding to the first node).

The I–V and the IF curves can be used to evaluate the gain and the noise temperature of the mixer. This is done by calibrating the IF noise in terms of temperature using the linear portions of the curves where the junction is considered as a calibrated shot noise source (Woody, Miller and Wengler 1985). From hot and cold load measurements at 213.6 GHz we find that the receiver noise temperature is 270 K (DSB) which includes the loss of the mylar splitter and the lens (at a temperature of 27 K). Using the IF calibration we found that the conversion loss of the mixer is -5.3 dB and that the IF contribution to the receiver noise is 171 K. From these figures we are in principle able to evaluate the noise temperature of the mixer. Since the contribution of the IF to the system noise is so high, it is very difficult to come up with an exact figure, but clearly the noise temperature of the mixer is significantly below 100 K.

The high contribution of the IF system to the receiver noise is the cause

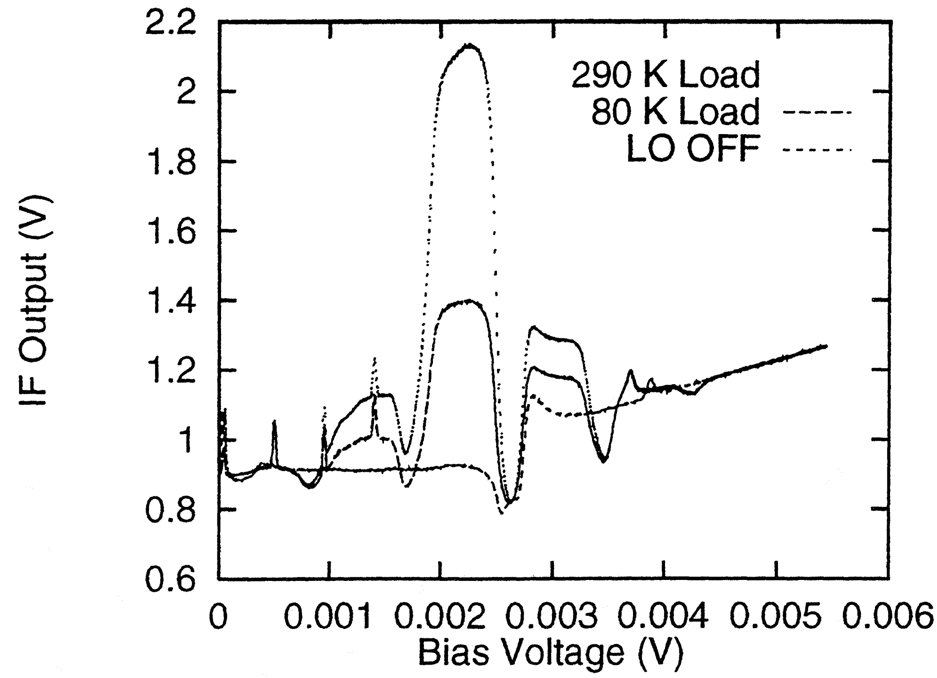

FIGURE IV Responses of the IF output of the Mixer to Hot (290 K) and Cold (80 K) Loads at 213 GHz

of the large DC offset in Fig.4. We attribute this to a large mismatch between the junction and the amplifier. The main contribution to this mismatch comes from the long and low impedance microstrip and the very high capacitances of the bonding pads. We tested this assumption by dicing out the long microstrip and bonding pads and connecting the chip to the IF board via two test pads on the top and bottom fins. We then compared the IF noise levels before and after the dicing and noticed that the calibrated DC offset dropped by a factor of two, confirming our initial predictions.

CONCLUSION

We have presented a new type of SIS mixer which combines the high beam directivity of metallic horns with the construction simplicity of RF planar circuits. The mixer is wide-band and tunerless and can present a wide range of source impedances to the junction. Although the noise performance of the receiver was degraded by a high IF contribution, our preliminary tests of the mixer support most of our design predictions. We have already made several significant changes to the mixer block and mask designs in order to improve the IF coupling circuit and to make the mixer suitable for the array receiver. Tests of the new mixer will be carried out shortly.

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