INTEGRATED TUNING ELEMENTS FOR MILLIMETER AND SUB-MILLIMETER SIS MIXERS

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ABSTRACT

We present two different designs of integrated tuning elements for superconducting tunnel junction (SIS) mixers. The structures consist of a self-complementary log-periodic antenna, superconductor-insulator-superconductor (SIS) tunnel junction, and the broadband superconducting tuning circuit placed between the center of the antenna and the SIS junction. The designs allow good coupling to relatively large area (7 μm^2) junctions. The best design has 3-dB bandwidth of 102 GHz at the central frequency of 98 GHz, and 75 GHz at the central frequency of 492 GHz. Microwave scale model measurements show excellent agreement with the simulation results. Devices are being fabricated in niobium technology at Westinghouse Science & Technology Center.

INTRODUCTION

Superconducting tunnel junction (SIS) mixers are used in receivers for the detection of millimeter and sub-millimeter wavelength signals. They are low-noise detectors with sensitivity superior to all known alternatives in the frequency range from 100 GHz to 500 GHz [1].

At RF frequencies, the SIS junction can be represented as its parasitic capacitance (C_j) in parallel with its RF junction resistance (R_{rf}). Junction parameters C_j and R_{rf} are determined by the fabrication process parameters: the critical current density j_c [A/cm^2], the junction capacitance per unit area c_j [fF/μm^2] and the junction area A[μm^2]:

$$C_j\,[\mathrm{fF}] = c_j\,[\frac{\mathrm{fF}}{\mu\mathrm{m}^2}]\,A\,[\mu\mathrm{m}^2]$$

$$R_{rf}[\Omega] \approx R_N[\Omega] = \frac{\pi}{4}\,\frac{V_g\,[\mathrm{mV}]}{j_c\,[\frac{\mathrm{A}}{\mathrm{cm}^2}]\,A\,[\mu\mathrm{m}^2]}\,\mathrm{X}\,10^5$$

where V_g is the gap voltage, about 3 mV for niobium.

The typical junction sizes are 1-10 μm^2. Due to the very high junction capacitance one of the main problems at higher frequencies is poor coupling between the SIS junction and the source impedance. The mismatch between the junction and the source can be reduced if high current density and small area junctions are used, but this places special demands on the circuit fabrication processes. An alternative approach is to use integrated tuning elements to tune out the large capacitance and to transform the low junction impedance in large area junctions.

Two major goals in the design of tuning structures are small power loss at the central frequency and wide 3-dB bandwidth. A number of different designs of tuning elements were proposed for the frequency of 100 GHz and less [2-8]. However, most of them have either small bandwidth [2, 4, 7] or they are impractical at higher frequencies [3, 8]. We propose two different designs of tuning structures applicable both at lower and higher frequencies. Circuits presented here are designed for 98 GHz and 492 GHz.

BROAD-BAND TUNING STRUCTURES

In our quasi-optical receiver, the SIS junction is built integrally with a planar self-complementary log-periodic antenna on a dielectric substrate [9, 10, 15]. Antenna is placed on the back of the quartz hyperhemisphere. The hyperhemisphere and the teflon lens in front of it focuses the radiation into the antenna. The antenna impedance is frequency independent over several octaves and is equal to 119 Ω.

If no tuning structures are used the SIS junction is placed in the middle of the antenna. Coupling between the junction and the antenna is shown in Fig. 1 for the junction with different parameters. To achieve the best coupling, large current density and small area are needed. Even then, at high frequencies, most of the input power is not coupled into the SIS junction. The major reason for such a mismatch is the existence of the junction capacitance which shorts out higher frequencies.

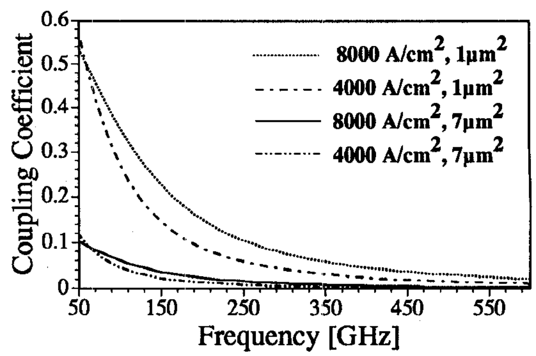


Fig. 1. Coupling coefficient between the SIS junction and the log-periodic antenna when no tuning structures are used.

By using integrated tuning elements, the mismatch problem can be resolved. Two different designs are shown in Fig. 2. In the first, a transmission line is placed between the junction and the antenna (Fig. 2.a). The Smith chart in the Fig. 3. shows how the transformation is performed. The SIS impedance is first rotated along the length of the transmission line to a small resistance R approximately equal to:

IF1

$$R \approx \frac{R_{rf}}{1 + (\omega C_j R_{rf})^2}$$

which is then transformed to a value equal to Z_{ant} at the end of the transmission line. The transmission line has a characteristic impedance between R and Z_{ant} and its length is greater than one quarter wavelength.

In the second design, junction capacitance C_j is first tuned out using the inductive element and then the junction RF resistance is transformed into the antenna impedance (Fig. 2.b). This design gives more degrees of freedom than the first one, since the transformation is done with more than one element. The designs will be discused in more detail below. These same tuning structures are applicable also to waveguide receivers.

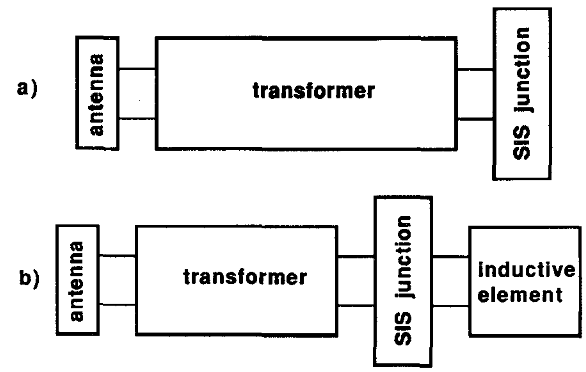


Fig. 2. The principal schematic of the two different tuning structures.

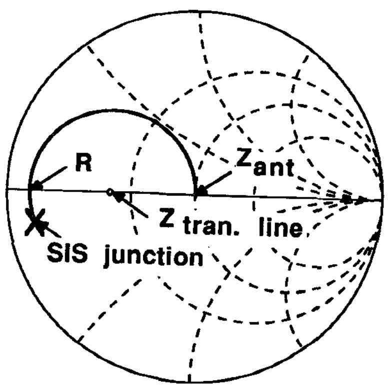


Fig. 3. The Smith chart impedance transformer description. Smith chart is normalized to the impedance of the antenna. The transmission line transformer impedance is equal to $\sqrt{Z_{ant} R}$.

All the designs presented here are done for the fabrication process with the following parameters:

$$j_C = 4000 \frac{A}{cm^2},\ c_j = 60 \frac{fF}{\mu m^2},\ A = 7\ \mu m^2.$$

These process parameters correspond to the junction parameters:

$$R_N = 8.6\ \Omega \text{ and } C_j = 420\ fF.$$

Without tuning structures, these junctions have less than one percent power coupling to the antenna at frequencies larger than 250 GHz (Fig. 1).

Tuning structures are built on the one arm of the antenna which is used as their ground plane (Fig. 4). Our scale model measurements show that the tuning structure placed on the antenna arm does not change the antenna impedance. Superconducting microstrip lines [11, 12], coplanar waveguides [12] and radial stubs [14] are used as the basic elements of the tuning structures. Their characteristic impedance and the effective wavelength are calculated taking into account the superconductor complex surface impedance [11, 13, 16, 19].

The tuning structure equivalent circuits are optimized in the "Touchstone" RF simulation program [17] to achieve the largest 3-dB bandwidth consistent with 75 % or more coupling at the central frequency.

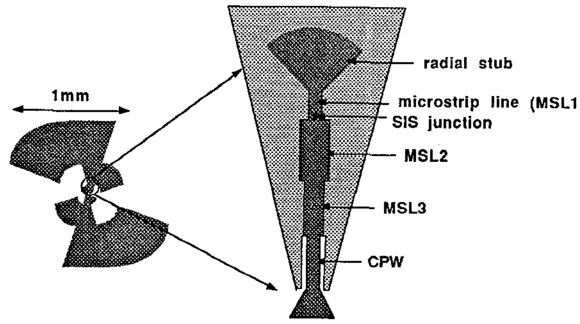


Fig. 4. Self-complementary log-periodic antenna with the superconducting tuning structure.

DESIGN #1

In the first design a superconducting microstrip line (MSL) is placed between the center of the antenna and the SIS junction (Fig. 5.a) [2]. The length and the characteristic impedance of

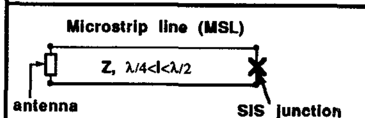
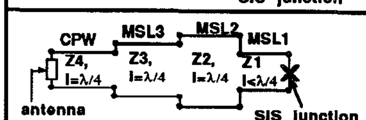
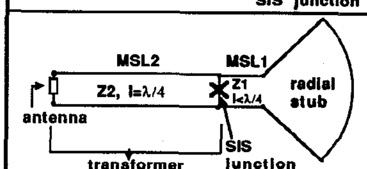
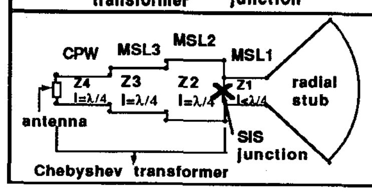
	electrical circuit	fractional bandwidth 98GHz	fractional bandwidth 492GHz
a)		0.26	0.07
b)		0.82	X
c)		0.34	0.13
d)		1.04	0.15

Fig. 5. Electrical schematic of tuning circuits with the expected fractional bandwidths at 98 GHz and 492 GHz. X means that the circuit is not realizable with our fabrication process.

the MSL are chosen to achieve high coupling and large bandwidth. The total length of the MSL is $\lambda/4<l<\lambda/2$ at the central frequency. This design is simple, but it has the smallest bandwidth of all proposed designs. Another disadvantage is that the only way to increase bandwidth is to allow some small mismatch or to use smaller area and higher current density junctions. The simulation data at 98 GHz and 492 GHz are shown in Fig. 6.a and Fig. 7.a respectively. At 98 GHz 3-dB bandwidth of 26 GHz with 90 % coupling at the central frequency is achieved. At 492 GHz we got the bandwidth of 36 GHz with 76 % coupling at the central frequency. This bandwidth is comparable or larger than previously designed tuning circuits [2, 3, 4, 8], but in comparison with the other designs discused here we consider it the small bandwidth. The reason for the small bandwidth is that the transformer is not a broadband structure due to the large transformation between the small resistance R and the large antenna impedance.

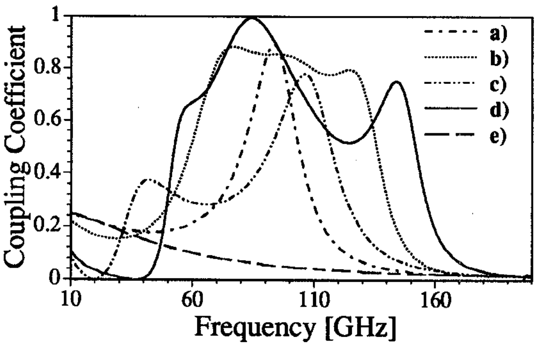


Fig. 6. Coupling coefficient between the SIS junction and the antenna for 98 GHz: curves a) to d) correspond to the designs from Fig. 5.a to 5.d; curve e) show coupling when no tuning elements are used.

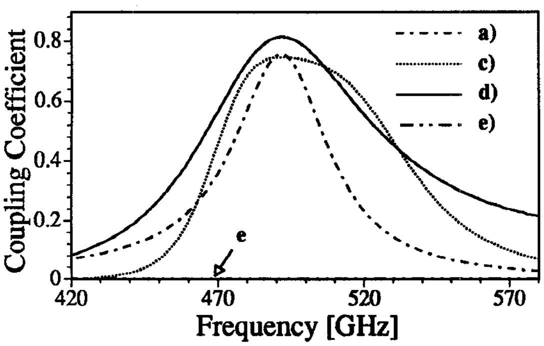


Fig. 7. Coupling coefficient between the SIS junction and the antenna for 492 GHz: curves a), c) and d) correspond to the designs in Fig.5.a, 5.c and 5.d, respectively and e), which is close to zero, is coupling when no tuning structures are used.

Significant improvement in the bandwidth can be achieved by replacing the part of the MSL (the quarter of the wavelength long MSL) by multiple section Chebyshev transformer (Fig. 5.b). In that case the bandwidth is larger if more sections are used. We use a three section Chebyshev transformer which consists of one coplanar waveguide (CPW) and two microstrip lines (MSL2 and MSL3), each one quarter wavelength long.

The reason for using coplanar waveguide is that the maximum characteristic impedance of microstrip line is two low. The range of the practically realizable characteristic impedances of any transmission line depends on: 1) fabrication process limitations such as the minimum line width and the minimum separation between two lines, 2) excitation of higher order modes which can be prevented by keeping the lateral dimensions of the transmission line below a quarter of the wavelength. In the case of the coplanar waveguide which always reaches the center of the antenna (Fig. 4) the limit values for the characteristic impedances are also defined by the dimension of the antenna at its center. The impedance limits calculated for this fabrication process are shown in Table 1.

Transmission line	$(Z_0)_{min}$ [Ω]	$(Z_0)_{max}$ [Ω]
Microstrip line	1.5 (m)	43 (d)
Coplanar waveguide	59 (d,a)	123 (d,a)

Table 1. Comparison of the characteristic impedance limits for the following fabrication process parameters (ε=4, h=0.3 μm, minimum line-width and spacing between two lines=2 μm, frequency=492 GHz). The letter 'm' indicates that the limit is caused by higher order modes, the letter 'd' that it is caused by minimum dimensions and the letter 'a' that the limit is due to the dimensions of the antenna.

The antenna excites the odd propagation mode in the coplanar waveguide that makes the connection between the CPW and MSL3 possible (Fig. 5.b)[12]. Although this design gives extremely large bandwidth at lower frequencies (80 GHz 3-dB bandwidth with the coupling of 85 % at 98 GHz Fig. 6.b) it is practical at higher frequencies only if high current density and small area junctions are used. If we were to use this design for 492 GHz the junction capacitance is so high that the resistance R is very small (Fig. 3.). As a result the third section of the Chebyshev transformer (MSL2 Fig. 5.b.) would need to have an impedance less than 1Ω which is out of the range shown in Table 1. However, the solution with Chebyshev transformer gives excellent results with small area and high current density junctions [18].

DESIGN #2

The best designs for high frequencies are shown if Fig. 5.c and 5.d. The tuning circuit is composed of two parts: 1) inductive element, and 2) transformer. The inductive element is a combination of an open-ended superconducting radial stub and short, high impedance superconducting microstrip line MSL1. The radial stub is designed to make a broadband short at RF frequency at the place where it is connected to MSL1. Instead of the radial stub an open-ended one quarter wavelength long MSL can be used. MSL1 provides a small inductance necessary to tune out junction capacitance at the central frequency. In Fig. 5.c one quarter wavelength long microstrip line (MSL2) is used as the transformer which transforms junction RF resistance to the impedance of the antenna.

The 3-dB bandwidth can be influenced independently by three elements: 1) radial stub, 2) short microstrip line (MSL1) and 3) transformer. To achieve bigger bandwidth larger angle radial stub and larger characteristic impedance microstrip line (MSL1) should be used. In Fig. 6.c it can be seen that the 3-dB bandwidth of 33 GHz with 79 % coupling is achieved at 98 GHz. At 492 GHz bandwidth is 66 GHz with 75 % coupling at the central frequency (Fig.7.c).

Similarly as with the design in Fig. 5.a, this design can be improved by using multiple section Chebyshev transformer. Although at smaller frequencies this improvement is really significant, at higher frequencies it does not make much change due to the large junction capacitance. As a result, extremely large bandwidth of 102 GHz with coupling of 79 % is obtained at 98 GHz, and bandwidth of 75 GHz and coupling of 81 % at

492 GHz (Fig.6. d. and Fig. 7. d. respectively).

SCALE MODEL MEASUREMENTS

We have made microwave scale model measurements of our tuning structures. The structures are modeled with copper tape and therefore the superconducting surface impedance is not included. We have built the tuning circuit on a 30 cm thick polyethylene block, with $\varepsilon_r = 2.2$. For the dielectric in microstrip lines we have used 1 mm thick mylar with $\varepsilon_r = 2.9$. The circuit parameters are first optimized by the "Touchstone" RF simulation program [17] to achieve large bandwidth. By using the available dielectric constants and optimized parameter values the physical dimensions of scale models are calculated. Measurements are done for the circuits designed with the central frequency of 295 MHz. Scale models are measured in the frequency range from 150 MHz to 550 MHz. To model the SIS junction, a parallel connection of resistor $R = 10\ \Omega$ and capacitor $C = 150$ pF is used. The structure is fed through 50 Ω coaxial cable and the reflection coefficient is measured with an HP8720B Vector Network Analyzer. We have made measurements using both a large ground plane and a log-periodic antenna shaped ground plane. Both measurements show excellent agreement with the simulation results. In Fig. 8 we present scale model results for the circuit shown in Fig. 5.c.

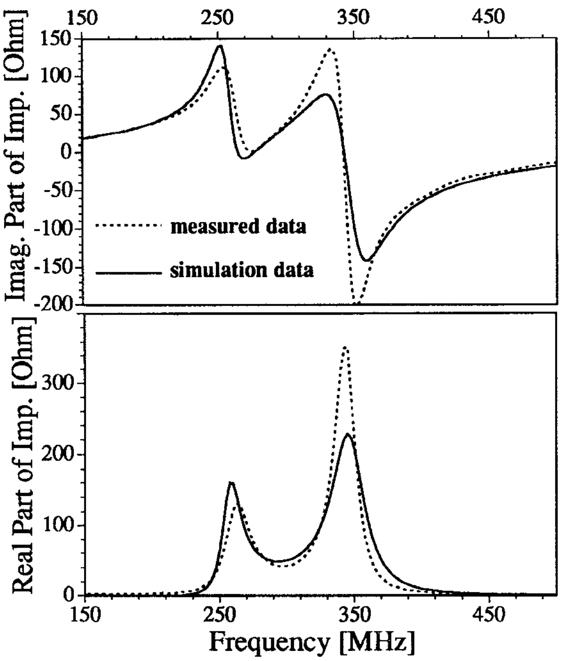


Fig. 8. Results of the scale model measurements compared to the simulation results for the structure with the equivalent circuit in Fig. 5.c.

CONCLUSION

Broadband tuning structures for 98 GHz and 492 GHz have been presented. The proposed configurations are suitable both for quasi-optical SIS mixers and SIS waveguide mixers. Proposed structures are realizable up to very high frequencies, and they give excellent coupling and large bandwidth even with large area junctions. By using larger current density and smaller area junctions, significantly larger bandwidth could be achieved. The scale model measurements done at the central frequency of 295 MHz have shown excellent agreement with the computer simulation data, which confirms our designs. The integrated mixers are being fabricated in cooperation with Westinghouse Science & Technology Center.

ACKNOWLEDGEMENTS

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19. Note that there is a mistake in reference [13]: equation (5) should be replaced with $Z = j2\pi f\mu_0 g_1 + 2Z_s(f)g_2$