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Development of Advanced Superconductor-Insulator-Superconductor Mixers for Terahertz Radio Astronomy

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Development of Advanced Superconductor-Insulator- Superconductor Mixers for Terahertz Radio Astronomy

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CHAPTER 1

I The thesis introduction

I.1 INTRODUCTION

Since ancient times, mankind has been trying to understand the nature of the universe by improving its skills and tools of observation. Due to the characteristics of human sensory organs, the range of visible light was the most natural for observation, but allowed detection of only bright sources of radiation, such as stars and planets. In the coldest areas of the interstellar medium, there are dense clusters of interstellar dust and molecular gas. The largest of these, known as Giant Molecular Clouds (GMCs), can extend up to ~ 310 light-years in diameter and have masses ~ 100 times the mass of the Sun. GMC clouds are so cold that they do not emit light in the visible spectrum, and stellar dust blocks the background radiation of stars. The birth of stars, such as the Sun, likely took place in the depths of GMCs. Cold molecular gas was locally compressed and heated before the start of a thermonuclear reaction. Then, the gravity and ionizing radiation of the young star destroyed the formation of molecular gas.

For a correct study of the physical processes occurring in the GMC, a molecular gas must be observed directly as it is illuminated, heated, and ionized by a new star. For this, measurement of the spectral emission lines is carried out. For a cold molecular gas, the maximum wavelength in the emission spectrum is in the millimeter and submillimeter range, which corresponds to GHz and THz frequencies. This is due to the excitation of rotational and vibrational degrees of freedom of complex chemical compounds. For example, molecular hydrogen (H_2), has a very high excitation temperature ($T \sim 300$ K), which makes it of little use as a test gas in GMC ($T \sim 10$ K). Carbon monoxide CO, on the contrary, has a spectral emission line in the submillimeter range and is quite common in gas interstellar clouds. By measuring the spectral profile of CO, H_2O , HCl, NH, CH, CO, OH, and other compounds, it is possible to make a conclusion not only about the gas composition of cold molecular clouds, but also about the physical processes that occur in GMCs and other cold space objects.

Astronomical observations of submillimeter waves are difficult because atmospheric water vapor has many spectral absorption lines (Figure I-1) and the studied signals are usually very weak. This demands development of low

noise detectors. When deeply cooled to the temperature of liquid helium, the intrinsic thermal fluctuations (Johnson-Nyquist noise) in the receiving system can become lower than quantum fluctuations. This results in receiving systems operating near the fundamental noise limit. Superconducting receiving systems with very high non-linearity are the best suited to be low-noise THz radiation detectors. According to the theory of noise, the time required to register super-weak signals is proportional to the noise temperature squared. Therefore, such receivers can significantly reduce the time of observation, which is an extremely important condition for modern astronomical observations.

Figure I-1 The transparency windows in the Earth's atmosphere at 5040 m (online: <https://almascience.eso.org/about-alma/atmosphere-model/>)

Molecular spectral lines are very narrow and the measured frequency spectrum is often crowded. Due to the Doppler effect, the spectral lines can be shifted. Therefore, to conduct high-quality astronomical observations, which makes it possible to determine the structure of a distant object, a receiver with a high spectral resolution, of the order of $R \equiv \nu / \Delta\nu \sim 10^4\text{--}10^8$, is required. For example, it is possible to determine the approximate radial distribution of a matter of a rotating object (galaxy, double galaxy) by analyzing the spectrum profile.

The placement of THz radiation receivers on high mountains (Figure I-2) or in outer space increases the signal-to-noise ratio due to decreased

atmospheric influence. The layout of a heterodyne laboratory test receiver system is shown in Figure I-3.

Figure I-2 The ALMA telescope array (online: <https://www.space.com/25534-alma.html>). The instrument is placed in one of the driest places on Earth. Each single dish telescope is being moved by a transport machine to a prepared basement, which has the necessary connections to the central correlator. These instruments are installed at different positions to form an array, which works as an interferometer with different baseline configurations.

Figure I-3 A heterodyne receiver test system scheme.

I.2 LOCAL OSCILLATORS IN SUBMM REGION

A local oscillator (LO) is a critically important part of a heterodyne receiver. Tuneability, frequency band coverage, LO stability, and LO noise are very important properties of an LO source. The following sub-chapters include a brief description of different LO types.

I.2.1A BACKWARD WAVE OSCILLATOR (BWO)

The backward wave oscillator (BWO) is one of the most successful microwave sources [1]. In the past, the frequency coverage of BWOs ranged from microwave to sub-microwave. The construction of the generator is similar to a travelling wave amplifier; however, in a BWO, a special decelerating structure is used to maintain a feedback loop. A BWO is a complex electric vacuum device in which emitted electrons from the cathode are accelerated toward the anode through a periodic deceleration structure similar to a ridge, a comb, or a spiral. An external magnetic uniform field is commonly used to form a collimated electron beam. There are two types of BWOs, O and M. In type O, the kinetic energy of the electrons is transferred to the radiofrequency (RF) wave energy. In type M, the potential energy of the electrons is transferred in the RF energy and the average kinetic energy remains constant. An important advantage of BWOs is their tunability [2]. A BWO type M can reach 40–60% efficiency, while a BWO type O has an efficiency of only a few percent. Significant disadvantages of BWOs are that they are bulky and heavy high-voltage power supplies (some kVs) require a strong magnetic field bias that leads to relatively high LO system weight and supply power.

I.2.2SCHOTTKY DIODE MULTIPLIERS

For heterodyne detection, THz sources that are low-noise and easily tunable in a wide frequency band are required. The 300 GHz–10 THz segment lies between the infrared (IR) and RF ranges, where the efficiency of generating devices is significantly reduced. It is sometimes referred to as a “terahertz gap”. Historically, THz generators have been based on BWOs that require a high-voltage power supply and a strong (~ 1 Tesla or higher), heavy magnet. For laboratory studies, gas lasers and other devices based on quantum effects that are large in size, extremely sensitive to environmental conditions, difficult to manufacture, and expensive, have also been used. More recently, multistage multipliers based on Schottky diodes [3]–[7] have undergone significant development. For their operation in a wide frequency range, broadband low frequency amplifiers with power on the order of tens to hundreds of milliwatts (mW) are required. These amplifiers are commercially available. The operational principle of the Schottky diode

multiplier LO is based on generation of higher harmonics by a strongly nonlinear element. Schottky diodes are highly suitable in this role as they have a high nonlinearity due to a semiconductor-metal barrier. This source type [3]–[6] can produce tens of hundreds of milliwatts of power at room temperature and has a compact size, which is convenient for application in a normal lab environment, at a telescope, or in space instruments.

To work efficiently, filtering and integrated tuning systems are required for the multipliers. The following components are used in the typical LO chain: an integrated tuning structure for a Schottky diode capacitance; a low pass filter to prevent the leakage of high frequency signal into an amplifier feedback loop chain; a band pass filter to prevent low frequency signal from leaking into the output, and filters to prevent signal leakage into bias chains. Commonly, these components can be assembled on a waveguide-based structure with mechanical tuners and impedance transformers that allow broadband operation. The highest harmonics generated on a Schottky diode are less powered than the input base signal, leading to possible difficulties in generating sufficient power for ~ 1 THz signals. The generation of a stable low frequency signal is also necessary. There are candidate technologies for this purpose, including Gunn diodes, which generate hundreds of milliwatts of power. This type of source can be tuned electronically within a narrow frequency band and can be tuned in a relatively wide frequency range using mechanical tuners.

High electron mobility transistor (HEMT) technology is used in production of high frequency ~ 100 – 150 GHz amplifiers. A Gunn's oscillator, an amplifier based on HEMTs and Schottky diode multipliers, is used to create an LO with stable output continuous wave signal. This LO source has enough power for a superconductor-insulator-superconductor (SIS) mixer of up to ~ 1.5 – 2 THz.

I.2.3 PHOTONIC MIXER OSCILLATOR

There are advances in photomixer manufacturing technology (e.g., LT-GaAs) and laser diodes, as well as in state of the art optical fiber technology (~ 0.2 dB losses per km) that have led to development of a room temperature LO in the GHz range that has acceptable noise and power levels and a broad frequency tuneability [8]. This LO type was demonstrated as a source for an SIS mixer. The main operating principle of this LO type is the mixing of signals from two lasers tuned to slightly different frequencies. This source type has fundamental limitations. The LO noise level can be limited by its intrinsic noises. The photomixer process includes signal and wide band zero-point fluctuation noise, as well as intrinsic photomixer noise. All of these processes lead to a wide band signal in the LO output.

In conclusion, the current state of the art and future photomixers and laser sources will likely be good alternatives to conventional, Schottky-diode based multipliers for room temperature LOs.

I.2.4 FLUX FLOW OSCILLATOR

The Flux-flow Josephson oscillator (FFO) is a tunable oscillator (see Chapter 5 for details). The oscillator is based on a long ($\sim$ hundreds of microns) Josephson SIS junction. The electrical bias current, which runs through the long SIS structure, leads to creation of a vortex chain inside the FFO. The "control current" is a lateral current that runs parallel and under the long SIS, causing the vortex chain to move toward the FFO long junction short edge. The emitted RF radiation frequency is defined by the second Josephson law through the applied bias voltage. An emitter microchip is glued to the collecting lens and has a characteristic size on the order of a square centimeter. The LO module dimensions, together with the control board and cylindrical magnetic screens made of permalloy and lead, are 4 cm in diameter and 10 cm in length. At the same time, the characteristic power consumption of the device does not exceed 0.1 mW, since the operating currents are less than 100 mA, with bias voltages on the cryogenic elements not exceeding ~ 1 mV. This is negligible compared to the power consumption of traditional generators (multiplier chain LOs, Gunn diodes). Due to its construction and compatibility of manufacturing processes, an FFO oscillator and a SIS mixer can be integrated on one chip, indicating great potential for further reduction of the weight and power consumption of SIS receivers.

Thus, the development of THz generators with a wide operating range, ease of frequency tuning, low power consumption, a narrow spectral line of emission, and highly sensitive THz range receivers is an important task.

I.3 SIS/FFO MANUFACTURING TECHNOLOGY

The manufacturing process of superconducting integrated structures based on Nb-AlN-NbN Josephson junctions is briefly described as follows. The SIS junctions were fabricated using the Selective Niobium Etching and Anodization Process (SNEAP). First, the Al_2O_3 buffer was sputtered on a silicon (Si) substrate. The purpose of the layer was to protect the Si substrate during the etching process, during which the bottom Nb wiring was formed. Over the 80 nm Al_2O_3 buffer layer, the 3-layer (Nb-AlN-NbN) structure was formed in a single vacuum run, as described in Chapter 2. The Nb, Al, and NbN layers were sputtered using direct current (DC) magnetrons from the Nb and Al targets. To form the SIS and FFO tunnel junctions, the reactive ion etching process was used. During the process, the structure may warm up significantly. To avoid unwanted heating, the Si substrate was placed on

a cooling metal plate-holder. The prepared structure was then anodized in liquid to protect the SIS junction from a possible short during the next sputtering operation. Next, the dielectric SiO_2 layer (400 nm) or layers (200 + 200 nm) and the top metal (Nb, 450 nm) wiring were formed by rf-magnetron, reactive DC magnetron sputtering, and lift-off and/or etching procedures. The surface of Nb reacts with O_2 in the Earth's atmosphere. Therefore, to avoid issues with electrical connection to the manufactured samples, the gold contact pads were formed at the corresponding places in the Nb wiring. Figure I-4 below shows a schematic representation of the SIS junction.

Figure I-4. A cross-section in the vicinity of a SIS junction. The junction itself is a very thin barrier illustrated at the figure center ($x=1$).

I.4 REGULAR HIGH FREQUENCY TRANSMISSION LINES THEORY

Consider a long two-wire line or a symmetrical microstrip. An equivalent circuit of such a line is presented in Figure I-5.

Figure I-5 An equivalent circuit of a long two-wire line section.

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Here, we introduce the specific (per unit length) capacitance dC (F/m), conductivity dG (1 / Ohm * m), inductance dL (H/m), and linear resistance dR (Ohm / m). The linear resistance depends on the quality of the thin boundary layer of the electrode adjacent to the dielectric; the conductivity depends on the quality and thickness of the dielectric layer; and the capacitance and inductance depend on the relative positions and sizes of the electrodes and the dielectric. In this case, dz is the length of the elementary section of the line. According to the definition, the elementary section of a two-wire line in electrodynamics is considered the one in which the approximation of instantaneous currents and voltages is physically correct. In other words, the $dz \ll \lambda$, which is the wavelength of the microwave signal transmitted by the line.

$$dR = Rdz, \quad dG = Gdz$$

$$dC = Cdz, \quad dL = Ldz$$

Increments of current and voltage in this section of the line are:

$$dU = I(dR + i\omega dL)$$

$$dI = U(dG + i\omega dC)$$

By performing substitutions and differentiating with conditions of constant linear complex resistances and conductivity, we obtain homogeneous wave equations of the long line:

$$\frac{d^2 U}{dz^2} = \gamma^2 U$$

$$\frac{d^2 I}{dz^2} = \gamma^2 I,$$

where γ is the wave propagation coefficient in the line:

$$\gamma = \sqrt{ZY} = \sqrt{(R + i\omega L)(G + i\omega C)} = \alpha + i\beta$$

The solution of these wave equations can be expressed as:

$$U(z) = A \exp(\gamma z) + B \exp(-\gamma z),$$

$$I(z) = \frac{(A \exp(\gamma z) - B \exp(-\gamma z))}{\sqrt{\frac{Z}{Y}}}$$

The phase parameter in these equations is responsible for the spatial distribution of the energy of the electromagnetic (EM) wave. In addition, the impedance of a line segment is a complex quantity that strongly depends on the length of the line segment and the frequency of the microwave signal transmitted along the line. Therefore, by changing the line length, it is possible to achieve inductive or capacitive impedance. This important physical consequence is the basis of microwave matching of circuit elements and underlies the transforming properties of the line. Below, we note some important special cases of lossless lines.

1) Open line (zero boundary conditions for current at the end of the line):

$$U = U_0 \cos\left(\frac{2\pi z}{\lambda}\right),$$

$$Z_{in} = \frac{U}{I} = iX_{in} = -i \cot g\left(\frac{2\pi z}{\lambda}\right)$$

2) Loaded line (zero boundary conditions for voltage at the end of the line):

$$I = iI_0 \sin\left(\frac{2\pi z}{\lambda}\right), \quad Z_{in} = \frac{U}{I} = iX_{in} = i \cot g\left(\frac{2\pi z}{\lambda}\right)$$

3) Line loaded by resistance R_n :

$$Z_{in} = R_n \frac{\cos\left(\frac{z2\pi}{\lambda}\right) + i \frac{Z_0}{R_n} \sin\left(\frac{z2\pi}{\lambda}\right)}{\cos\left(\frac{z2\pi}{\lambda}\right) + i \frac{R_n}{Z_0} \sin\left(\frac{z2\pi}{\lambda}\right)}$$

The Z_{in} is a source impedance, the Z_0 is an intrinsic line impedance $\left(\sqrt{\frac{Z}{Y}}\right)$.

In the sections of the line where the voltage reaches maximum or minimum, the input impedance is purely active. Moreover, a careful analysis of the above relationships shows that, under conditions of purely active load, the traveling wave mode and the highest efficiency of signal transmission to the load are realized, and the following formula is correct:

$$Z_{in}(x) * Z_{in}\left(x + \frac{\lambda}{4}\right) = Z_0^2$$

Consider a circuit consisting of a generator with an output impedance Z_2 and a load Z_1 . The ratio of the power released in the load to the power released in the generator is called the power coupling coefficient; it is sometimes called "coupling" or "matching", and is described by the formula:

$$cpl = \frac{4Re(Z_1) * Re(Z_2)}{(Im(Z_1) + Im(Z_2))^2 + (Re(Z_1) + Re(Z_2))^2}$$

This expression characterizes the matching microwave structures"by power"..

I.4.1 NUMERICAL-ANALYTICAL MODEL WITHOUT LOSSES

The model is based on analytical and empirical formulas for thin microstrip lines without taking dielectric and conductive losses into account [9]. The formulas were derived from analytically calculated linear inductances and capacitances, and the numerical coefficients in front of significant parameters were obtained through generalization of experimental data. However, the upper microstrip wiring is quite thick, because each successive layer should be thicker than the previous one. Since the capacitance of the ends of the film becomes comparable with the capacitance between electrodes of the film, it is necessary to consider correction for edge effects. The effective width of the microstrip depends on the line width (W), the thickness of the dielectric (Hi), the strip thickness ($d2$), and the correction for the thickness of the anodic oxide (ha):

$$\begin{aligned} We(W, Hi) &= (Hi + ha) \times \\ &\times \left(\left(\frac{W}{Hi + ha} \right) + \left[\frac{d2}{\pi * (Hi + ha)} \right] \times \right. \\ &\quad \left. \times \left(2 + \ln \left(2 * \frac{Hi + ha}{d2} \right) \right) \right) \end{aligned}$$

The effective dielectric constant depends on the material dielectric constant - ϵ_i , adjusted for the anodic oxide - ha :

$$Eea(\epsilon i, W, Hi) = \frac{\epsilon_{12}(\epsilon i, Hi, \epsilon a, ha) + 1}{2} + \frac{\epsilon_{12}(\epsilon i, Hi, \epsilon a, ha) - 1}{2} \times$$

$$\frac{1}{\sqrt{\left[1 + \frac{12(Hi + ha)}{we(W, Hi + ha)}\right]}}$$

The above formulas can be derived through a complex (conformal) mapping of a "thick electrode" into a "thin one". The new charge distribution can then be determined and the resulting linear capacitance recalculated as the capacitance of a thin microstrip line, but with an effective width and an effective dielectric parameter.

It is important to note that the effective permittivity of a two-layer dielectric with layer thicknesses of $h1$ and $h2$ can be calculated using the formula:

$$\epsilon_{12}(\epsilon_1, h1, \epsilon_2, h2) = \epsilon_1 \epsilon_2 \frac{(h1 + h2)}{(\epsilon_1 h2 + \epsilon_2 h1)}$$

Ultimately, the wave impedance, considering the above amendments, is found via:

$$Zo(\epsilon, W, Hi) = \frac{120\pi}{\sqrt{Eea(\epsilon, W, Hi)}} \times$$

$$\times \frac{1}{\left(\frac{we(W, Hi)}{Hi + ha} + 1.393 + 0.667 \times \ln\left(\frac{we(W, Hi)}{Hi + ha} + 1.4444\right)\right)}$$

Formulas for microstrip lines do not account for the superconducting properties of the material, so "correction" is needed. For example, in ordinary conductors, a high-frequency current flows in a layer of thickness =

$\sqrt{\frac{2}{2\pi f \mu_0 \sigma}}$, but the magnetic field is pushed out of the superconductor

(Meissner effect); the penetration depth of the magnetic field (t_m) into the superconducting electrode is given by the formula:

$$t_m = \lambda \coth\left(\frac{d}{\lambda}\right)$$

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Here, d is the thickness of a superconducting electrode made of a superconductor with a London penetration depth λ . Thus, the nature of the superconductivity led to a change in the linear inductance. Indeed, the limited depth of field penetration into superconducting electrodes sets the value of the effective thickness of the electrode wiring. The EM fields and currents are distributed in this layer. Therefore, the following formula is used to calculate the linear inductance for a "thin" ($d \ll h$) microstrip:

$$L = \mu_0 \frac{1}{w} (\lambda_1 \coth\left(\frac{d_1}{\lambda_1}\right) + \lambda_2 \coth\left(\frac{d_2}{\lambda_2}\right) + h),$$

where λ_i are the London penetration depths of the magnetic field in 1 and 2 wiring electrodes, d_i are the thicknesses of the 1st and 2nd electrodes, and h is the thickness of the dielectric.

For a "thick" strip, accounting for the Meissner effect led to correction factor A in the wave resistance and complex part γ .

$$A(h) = \sqrt{\frac{\lambda_1 \coth\left(\frac{d_1}{\lambda_1}\right) + \lambda_2 \coth\left(\frac{d_2}{\lambda_2}\right) + h}{h}}$$

$$\beta = 2\pi\nu A(h), \quad K = \sqrt{E\epsilon a\epsilon_0\mu_0},$$

where K is a phase coefficient.

I.4.2 NUMERICAL-ANALYTICAL MODEL WITH FREQUENCY DEPENDENT LOSSES

Before proceeding with the description of the model, we consider the necessary theoretical concept of surface impedance.

Let there be given a Cartesian coordinate system with x, y, z axes and an infinite conductive plate of thickness d placed in such a way that the x and y axes lie on the infinite surface of the plate, and the z axis is directed normal to the surface of the plate inside. When a microwave sinusoidal wave passes orthogonally to the z axis (the Poynting vector is parallel to the z axis), a current flows in the plate. The surface impedance is then given by the formula:

$$Z_s = \frac{E_x(0, \omega)}{\int_0^d dz J_x(z, \omega)},$$

which is equivalent to $\vec{E}_t(\vec{r}, \omega) = Z_s[\vec{H}_t(\vec{r}, \omega) \times \vec{n}]$. Figure I-6 illustrates the definition and equivalent one-dimensional diagram showing the equivalence of the above definitions.

Figure I-6 Graphic illustration of surface impedance determination (CST help). The left part shows the electromagnetic (EM) model and the right part shows the schematic illustration.

As mentioned previously in this chapter, one of the ways to describe the propagation of EM waves in a microstrip line is based on the concepts of line impedance Z_0 and the wave propagation coefficient in the line $\gamma = \alpha + j\beta$. A schematic representation of an asymmetric microstrip line is shown in Figure I-7.

An EM wavelength is described by the formula $\lambda = \frac{2\pi}{\text{Im}(\gamma)}$. The real part of the propagation coefficient has the meaning of the attenuation coefficient and is measured in Napier per unit length $\alpha = \text{Re}(\gamma)$. Shunting admittance per unit length in the System International (SI system) is given by the formula $Y = \frac{j\omega w \epsilon \epsilon_0}{d}$, where d is the thickness of the dielectric layer and ϵ the dielectric constant of the microstrip line. The serial impedance per unit length in the SI system for a microstrip line of width W and dielectric layer thickness d is described by the formula $Z = \frac{j\omega \mu_0 d}{w} + \frac{Z_{s1} + Z_{s2}}{w}$, where Z_{s1} and Z_{s2} are the surface impedances of the lower and upper electrodes. The first term in the above sum describes the geometric inductance in the microstrip line, and the second term describes the material parameters of the wiring electrodes.

Figure I-7 Schematic representation of an asymmetric microstrip line; 3D illustration.

The precise analytical expression for the surface impedance depends on how strongly the charge carriers have “been influenced” by the EM field. For normal metals, the mean free path of an electron is defined as the average distance between the scattering centers in the metal lattice. In a superconductor, at a temperature different from 0, there are always both quasiparticles and Cooper pairs. For a mixed type of conductivity (quasiparticles and Cooper pairs), according to Pippard [10], the coherence length can be introduced as $\xi = \frac{1}{(\frac{1}{l_0} + \frac{1}{\xi_0})}$. In metals with a normal skin effect,

the mean free path is comparable to, or substantially less than, the skin layer. Two limiting cases for surface impedance are distinguished, depending on the purity of the material and the frequency of EM radiation: the local limit and the anomalous limit. In the local limit ($\xi \ll \lambda$, $l_0 \ll \delta$), at least several collisions with scattering centers occur during one period of the external EM field. In this case, Ohm's local law is valid ($J = \sigma E$) and the surface impedance is described by the formula:

$$Z_s(f) = \sqrt{\frac{j2\pi f \mu_0}{\sigma}} \coth(\sqrt{j2\pi f \mu_0 \sigma d}),$$

where f is the radiation frequency, d is the wiring thickness, and j is the complex meaning.

In the anomalous limit ($\xi \gg \lambda, l_0 \gg \delta$), many periods of external radiation occur during one collision event. In this case, the surface impedance is given by the formula:

$$Z_s(f) = \frac{1 + j\sqrt{3}}{\sqrt{3}\sqrt{\pi}} \sqrt[3]{\frac{9(2\pi f)^2 \mu_0^2 l_0}{16\sigma}}$$

When $\xi \sim \lambda, l_0 \sim \delta$ there is an anomalous regime; the surface impedance formulas for it are somewhat more complicated and are given in [11], [12]. Thus, due to the high-frequency current flowing in the surface layer, the impedance of a conductor can significantly differ from the impedance of bulk wiring by direct current. In addition, the presence of irregularities in the near-surface layer can significantly increase losses at a high frequency. For materials with frequency-dependent losses, it is possible to introduce the complex conductivity, $\sigma_s = \sigma_1 - j\sigma_2$, which should be substituted in the above written surface impedance formulas instead of direct current conductivity. This technique allows us to describe inductive additives to the surface impedance of lines made of superconductors. The values σ_1 and σ_2 are obtained via the Mattis-Bardoin theory [13], [14]:

$$\begin{aligned} \frac{\sigma_1}{\sigma_n} &= \frac{2}{\hbar\omega} \int_{\Delta}^{\infty} d\epsilon [f(\epsilon) - f(\epsilon + \hbar\omega)] \times \frac{\epsilon^2 + \Delta^2 + \hbar\omega\epsilon}{\sqrt{\epsilon^2 - \Delta^2} \sqrt{(\epsilon + \hbar\omega)^2 - \Delta^2}} + \\ &+ \frac{1}{\hbar\omega} \int_{\Delta}^{\hbar\omega - \Delta} d\epsilon [1 - 2f(\hbar\omega - \epsilon)] \times \frac{\hbar\omega\epsilon - \Delta^2 - \epsilon^2}{\sqrt{\epsilon^2 - \Delta^2} \sqrt{(\hbar\omega - \epsilon)^2 - \Delta^2}} \\ \frac{\sigma_2}{\sigma_n} &= \frac{2}{\hbar\omega} \int_{\text{Max}[-\Delta; \Delta - \hbar\omega]}^{\Delta} d\epsilon [1 - f(\epsilon + \hbar\omega)] \times \frac{\epsilon^2 + \Delta^2 + \hbar\omega\epsilon}{\sqrt{\epsilon^2 - \Delta^2} \sqrt{(\epsilon + \hbar\omega)^2 - \Delta^2}}; \end{aligned}$$

where $f(E)$ is the equilibrium Fermi-Dirac distribution at temperature T , and σ_n is the conductivity of the superconductor in the normal state, at a temperature slightly above the transition to the superconducting state. The first integral in the sum for calculation of σ_1 describes quasiparticle temperature excitations, and the second integral describes quasiparticle excitations caused by the destruction of Cooper pairs; therefore, it is equal to zero at $\hbar\omega \leq 2\Delta$.

The formulas for calculating σ_2 describe the kinetic inductance of a superconductor due to the presence of Cooper pairs. It is important to note that in the above formulas it is necessary to use the gap width at a given temperature. In general, to determine it, the integral equation must be solved [14]. Knowing all the above parameters, it is possible to calculate matching of any microstrip structures using the sequential load method described at the beginning of this chapter. Taking losses into account, the impedance transformation formula of a microstrip line of length L takes the form:

$$Z = Z_0 \frac{Z_l \cosh(\gamma L) + Z_o \sinh(\gamma L)}{Z_o \cosh(\gamma L) + Z_l \sinh(\gamma L)}$$

However, the above approach, due to the complex-valued gamma having a complicated form, and due to the peculiarities of numerical calculations, it is a complex numerical problem; therefore, it is reasonable to use the ABCD matrix method described below.

I.4.3 ABCD MATRIX APPROACH

Let us consider a quadrupole, as shown in Figure I-8; the currents and voltages at its inputs can be described by a matrix:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}, \text{ or } \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} = \begin{pmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{pmatrix} \begin{bmatrix} V_1 \\ -I_1 \end{bmatrix}$$

The matrix consisting of "b" elements represents the right ABCD matrix, and the matrix consisting of "a" elements represents the left matrix. The calculation of the series connection of several four-terminal networks is reduced to the direct multiplication of their ABCD matrices. The coordination coefficient for the power of the load and the source connected by a certain four pole is given by the formula:

$$KK = \frac{4r_1}{r_2} \left(\frac{r_2 a_{22} + a_{12} + r_1 r_2 a_{21} + a_{22} r_1}{r_2} \right)^{-2}$$

Connecting the left-to-left impedance Z of a four-terminal network with the left matrix transforms the impedance according to the following rule:

$$Z_{out} = \frac{Z a_{11} + a_{12}}{Z a_{21} + a_{22}}$$

Figure I-8 Four terminal element described by the ABCD matrix approach. An element is presented as a block with input and output V and I parameters without focusing on the element's nature.

Using the above formulas, sequential calculation of the matrix elements of various microstrip components was performed, allowing complete calculation (taking frequency-dependent losses into account) of a complex structure based on microstrip lines.

I.4.4A 3D CALCULATIONS TAKING INTO ACCOUNT FREQUENCY-DEPENDENT LOSSES

The above calculation methods allow a relatively quick calculation; however, due to the geometric features of the structures being calculated, not all areas of EM wave propagation are considered in numerical-analytical methods of calculation, leading to decreased calculation accuracy. Therefore, the use of fully numerical calculation programs, such as Microwave Studio Computer Simulation Technology (MWS CST), can improve accuracy by solving the electrodynamic problem. A CST output is a set of S parameters.

The core methods used by CST solvers include the finite element method, the finite integration technique, and the transmission line matrix. The program splits the entire volume into tetrahedrons. Their size should be so small that the EM fields on the size of the tetrahedron have a simple shape and can be described by a linear set of functions with coefficients sought from the boundary conditions of the Maxwell equations. After that, the calculation module excites the simulated system in special places called ports. As a result, the electrodynamic system reduces to solving an algebraic linear system of equations for unknown coefficients in the expansion of the EM field on the elements of the grid-partition. The CST program was not developed for the design of superconducting electronics, but its latest version has the function of specifying material through the table of surface impedance. This has allowed the calculation of superconducting elements. As shown, the surface impedance connects the tangential component of the electric field with the current or with the magnetic tangential field. Since the

computational module operates with the expansion of the amplitudes of the EM fields on the constructed numerical grid, there are no differences between the physical nature of the materials for the program.

I.5 MODELS OF JOSEPHSON JUNCTIONS: THE IMPACT OF MICROWAVE RADIATION ON THE SIS JUNCTION

A small direct current can flow through the Josephson junction without voltage drop. This phenomenon is called the stationary Josephson effect. In a simple model, it is considered:

$$I = I_c \sin(\varphi),$$

where I is the direct current crossing the SIS junction isolator in the absence of any external EM field, I_c is its critical current, and φ is the phase difference of the wave functions on the edges of the junction. When the current exceeds the critical value I_c , the second (non-stationary) Josephson law is “triggered” and an alternating voltage arises at the junction, the frequency of which is determined by the constant voltage at the Josephson junction. This process is called the Josephson generation:

$$2eV = \frac{h}{2\pi} \frac{\partial \varphi}{\partial t},$$

where e is the electron charge, V is the voltage at the Josephson junction, and φ is the phase difference of the wave functions.

I.6 WEAK LINK JUNCTION OVERVIEW

It is convenient to begin consideration of the SIS junction with a resistive model (Figure I-9). In this model, the Josephson junction is represented by a special superconducting element (the current passing through is described first by the DC Josephson equation), the resistor (the current passing through it which is described by Ohm's normal law with $R \sim R(V)$), and the junction capacitance.

Figure I-9 Schematic of a simple SIS.

Let this current be constant and equal to I_{dc} ; $R(V)=R$; the balance equation (Kirchhoff) can then be written as follows:

$$I_{dc} = C \frac{dV(t)}{dt} + GV(t) + I_c \sin \phi$$

$$G = \frac{1}{R}$$

The first term describes the current that “flows” through the capacitance (C). The second term shows the current flowing through the resistance. In a resistive model, the resistance is assumed to be constant. The last term describes the current owing to Cooper pairs tunneling; it flows without dissipation. Using the first Josephson relation, and moving on to dimensionless parameters, we derive the expressions:

$$\tau = \omega_j t$$

$$\beta_j = \frac{1}{\omega_j} \frac{1}{RC}$$

$$\omega_j = \sqrt{\frac{2e2\pi I_c}{hC}}$$

$$\gamma_0 = \frac{I_{dc}}{I_c}$$

ω_j is the plasma frequency and γ_0 is the normalized current.

$$\gamma_0 = \sin \phi + \frac{d\phi}{d\tau} * \beta_j + \frac{d^2 \phi}{d\tau^2}$$

Consider the case of small capacitance; using such an approximation, the problem was solved by Pippard:

$$\varphi(\tau) = 2c \tan\left(\sqrt{\frac{\alpha^2 - 1}{\alpha^2}} \tan\left(\frac{\tau\sqrt{\alpha^2 - 1}}{2\beta_j}\right) - \alpha\right).$$

Before proceeding to the consideration of the current-voltage (IV) characteristics of the SIS junction with a finite capacitance, we consider the limiting case of a very large capacitance. Let the SIS junction be connected to some current bias system. As soon as the system current exceeds the critical current value for the SIS junction, a voltage will appear on the resistor. This voltage will cause an instantaneous phase change, which will change the value of the tunneling current (the DC Josephson effect). The capacitance impedance depends on the frequency; therefore, a very large capacitance shunts the alternating current component. Consider a Josephson contact with a finite capacitance; hysteresis appears on its IV. This occurs after the current exceeds a critical value, and a "normal" component of the current appears; it flows through the resistor R on an equivalent circuit. The capacitor impedance is inversely proportional to the frequency, so an alternating signal will be more efficiently shunted by the SIS junction's capacitance the higher its frequency. The plasma frequency precisely determines the Josephson voltage at the tunnel junction, operating in the current bias mode, at which a reset occurs from the reverse hysteresis branch to the critical current region. Hereafter, we will refer to the reverse branch of the IV characteristic as a characteristic measured when the current or voltage changes from a maximum value to zero in absolute value. Indeed, on the reverse branch of the IV characteristic in the mode of setting the current, there is such a current of the master system at which the amplitude of the instantaneous value of the Josephson current is equal to the instantaneous value of the component of the direct current flowing through the resistive part of the junction impedance. As soon as this happens, the phase of the superconducting current will stop changing and the SIS junction will go to the direct branch of the IV characteristic. Thus, the final capacitance of the Josephson junction adds a hysteresis branch to the IV characteristic. Figure I-10 illustrates a typical IV characteristic of the SIS junction in the current bias mode.

Figure I-10 SIS IV curves in current bias mode. The hysteresis is visible on the IV due to the chosen bias mode.

I.7 THE IMPACT OF MICROWAVE ON THE SIS JUNCTION

To detect microwave signal, a nonlinear element is used. The quasiparticle nonlinearity is shown on the IV characteristic of the SIS mixer. This mechanism was considered in detail in the works of Tucker and Feldman [15].

A response to DC can be written as follows:

$$I(V) = \sum_n J_n^2(\alpha) I_{dc}(V + n\hbar\omega),$$

where $\alpha = \frac{eV\omega}{\hbar\omega}$, $j(V)$ is the function of the current response. This approach was used to estimate the THz emission power. In this case, the signal power detected by the SIS junction was estimated. Figure I-11 shows the theoretically calculated IV characteristics for different powers of incident radiation on the SIS junction. Note that the quasiparticle "steps" on the theoretical curves are proportional to the squares of the corresponding Bessel functions of the dimensionless parameter α .

Figure I-11 The calculated IV curves for different LO power levels. Alpha (α) indicates the SIS "pumping" level. The SIS capacitance was tuned.

I.8 MODELING OF SIS JUNCTION IN THE 3D SIMULATOR

Due to quantum nature, the correct representation of the SIS junction for numeric design calculation is a complex modeling problem. In the simplest numerical calculation model, the SIS junction can be represented as a parallel-connected capacitance and a lumped port with an internal resistance. The junction capacitance is determined by its geometrical parameters, such as the thickness of the tunnel layer and its linear size. The impedance of the SIS junction is highly frequency dependent [16]. Analysis of various experimental data showed that, for calculation in the frequency range of 500–1200 GHz, the input resistance (RF resistance, RRF) of the SIS junction is approximately R_n , which can be measured in DC measurements.

As an example, Figure I-12 shows the calculation RRF for an Nb-AlN-NbN SIS junction under EM radiation at 240 GHz frequency.

Figure I-12 RRF/R_n dependences as a function of different parameters, including radiation power (α), SIS bias point, and radiation frequency. As a basis for the calculations, a measured IV curve was used. The default bias point and frequency were taken as 3.1 mV, and 240 GHz, respectively.

CHAPTER 2

II A heterodyne waveguide receiver in the 700–950 GHz range

We have designed, fabricated, and tested a waveguide superconductor-insulator-superconductor (SIS) mixer based on Nb/AlN/NbN twin tunnel junctions operating in a frequency range of 700–950 GHz. Mixer performance EM simulations and measurement results are presented. SIS junctions have a high gap voltage of 3.15 mV and a high current density of approximately 30 kA/cm², providing a wide receiver instantaneous bandwidth as confirmed by Fourier transform spectrometer (FTS) and noise temperature measurements. The corrected receiver noise temperature varies from 240 K at ~810 GHz to 550 K at the high end of the frequency band. We observed junction heating caused by Andreev reflection, and its influence on the SIS junction characteristics was studied. We found that the heating effect does not have a critical impact on mixer performance.

II.1 INTRODUCTION

In 2006, the CHAMP+ heterodyne array [17] was installed on the APEX (Atacama Pathfinder EXperiment) telescope [18]. The array is composed of 14 pixels, divided into two sub-arrays of 7-pixels each, arranged in a hexagonal configuration. The RF tuning range is 620–720 GHz for the 450 μ m sub-array and 790–950 GHz for the 350 μ m sub-array.

The low frequency band array is based on the then state of the art SIS mixers developed for the ALMA band 9 receivers—the result of a few years of research and development. For the high band mixers, a one-off design was made without further modification or iteration. As a result, the performance of the high band mixers was not as well-optimized as was that of the low band mixers, and it cannot be considered as state of the art compared to similar devices [19], [20].

Here, we present our work on a new design of dual sideband high gap SIS mixers for 790–950 GHz, which are intended as an upgrade of the existing CHAMP+ high band array. An additional interest is to gain experience with high gap mixer technology for later use at frequencies above 1 THz; [21].

II.2 THE FIRST ITERATION, DESIGN AND FABRICATION

To make a wide band receiver, we have used twin SIS junctions [22], [23] (each with an area of $0.5 \mu\text{m}^2$), which are coupled by a waveguide probe to the E-field of a rectangular waveguide of $300 \times 75 \mu\text{m}$ (Figure II-1). We have developed a SIS mixer based on high critical current density Nb/AlN/NbN tunnel junctions incorporated in a microstrip line consisting of a 300 nm thick bottom electrode (ground plane) made of NbTiN and a 500 nm thick top electrode made of Al [24]. The microstrip electrodes are separated by a 250 nm SiO_2 isolator. The Nb layer of the SIS junction is deposited on the NbTiN film, while the NbN layer is in contact with the Al top electrode. The detailed procedure used in circuit fabrication is described below.

First, a NbTiN film was deposited on a fused quartz substrate at room temperature by DC sputtering with a NbTi target in a nitrogen atmosphere. For the NbTiN film, the critical temperature T_c was measured at 14.1 K, and the room temperature resistivity estimated to be $85 \mu\Omega\cdot\text{cm}$. The geometry of the ground electrode was defined by a reactive ion etching (RIE) process. Tunnel junctions were fabricated from a Nb/AlN/NbN tri-layer [24] with a normal state resistance-area product $R_n A = 7 \Omega\cdot\mu\text{m}^2$, which corresponds to a current density of $J_c = 30 \text{ kA}/\text{cm}^2$; the Nb and NbN layers have thicknesses of 100 nm. An AlN tunnel barrier was grown immediately after deposition of a 7 nm Al layer using an RF magnetron discharge. Samples were attached to the grounded substrate table, maintained at 20°C , and positioned directly above a 5-inch Al magnetron RF source with a holder to source distance of 14 cm. For fabrication of the high current density tri-layer, we initiated a plasma discharge using a power density of $0.7 \text{ W}/\text{cm}^2$ at a nitrogen pressure of 4.5 Pa; nitridation time was approximately 40 seconds. Afterwards, the NbN layer was deposited by DC reactive magnetron sputtering at ambient temperature with $1.8 \text{ W}/\text{cm}^2$ power density using an Ar + 9% N_2 gas mixture. Circular-shape junctions with an area of about $0.5 \mu\text{m}^2$ were defined by deep ultraviolet photolithography. SIS junctions were patterned from the Nb/AlN/NbN tri-layer by successive RIE of the NbN layer using CF_4 , followed by RF sputtering of the AlN/Al film in Ar plasma, and subsequent RIE of the Nb layer using CF_4 . A dielectric layer for junction insulation consisted of 250 nm SiO_2 , defined in a self-aligned lift-off procedure. At the final step, a 500 nm thick top microstrip electrode made of Al was deposited by DC magnetron sputtering. The thickness of the quartz substrate was then reduced to 40 μm by mechanical polishing.

Nb/AlN/NbN junctions of $0.5 \mu\text{m}^2$ in area, fabricated on a Si substrate with a 200 nm Nb layer and 100 nm NbN film demonstrated a gap voltage (V_g) = 3.5 mV and a quality factor (ratio of the sub-gap to normal state resistance) $R_j/R_n > 30$ for the $R_n A = 28 \Omega \cdot \mu\text{m}^2$, while showing $V_g = 3.4$ mV and $R_j/R_n = 23$ for higher current density ($R_n A = 7 \Omega \cdot \mu\text{m}^2$) [25], [26]. Current SIS junctions that were made with considerably thinner Nb electrodes (100 nm) and fabricated on NbTiN film instead of Nb, have $V_g = 3.15$ mV and quality R_j/R_n between 13 and 16 for $R_n A = 7 \Omega \cdot \mu\text{m}^2$. We cannot yet explain the cause of such a decrease in the junction gap voltage. The V_g level is uniform over the wafer and the yield of the working SIS junctions is high (> 90%), although there was a variation of about 20% in junction area across the wafer. The frequency of 950 GHz corresponds to 3.9 mV photon step, which exceeds the V_g of both fabricated SIS junctions and classical junctions with Nb electrodes ($V_g = 2.8$ mV). Consequently, the voltage range available for SIS mixer biasing is wider by about 0.7 mV for the mixers described here compared with Nb mixers. The wider voltage range affords a big advantage in mixer operation based on the presence of a noise feature due to a Shapiro step right in the middle of the photon step (see Figure II-4).

Figure II-1 Photo of the SIS mixer (including the antenna and the filter structure) installed in a waveguide. The central part containing the SIS twin junction is magnified.

Due to the high current density of the produced AlN barrier, the lower R_n gives a higher $1/R_n C$ ratio for the junctions (C = junction capacitance) and provides a wider receiver band. The tuning structure and waveguide mount design is conceptually similar to the original design of the CHAMP+ high band devices. SIS junctions are embedded in a $4.5 \mu\text{m}$ wide microstrip line with a

6.5 μm inter-junction distance and coupled to the antenna by a $7 \times 27 \mu\text{m}$ impedance transformer (Figure II-1) tuned specifically for the high current density junctions.

II.3 MEASUREMENTS

This section includes experimental results and analyses. The main experiments are noise temperature measurements. Typically, this experiment involves a lot of preparation and setup assembly time; therefore, only one mixer was measured per day. To optimize the measurement process, samples from other batches were tested with more efficient methods, such as the Michelson Fourier Transform Spectrometer (FTS) technique, which led to 3–5 possible “pre-tests” per day. These “pre-tests” were used to select samples for noise temperature characterisation.

II.3.1 FTS RESULTS

To evaluate the instantaneous bandwidth from the coupling of radiation to a SIS junction, a Michelson FTS technique was used. A wide band glow bar source was coupled to the Michelson interferometer, where a SIS mixer was used as a detector. The direct response of the SIS mixer current was measured vs. moveable mirror position. A Fourier Transform of the measured interferogram gave the frequency response shown in Figure II-2.

In addition, EM simulations of the full mixer structure were performed using CST Microwave Studio. The calculated response of the mixer is presented in Figure II-2. In the case of junctions with AlN barrier there are still uncertainties concerning the specific capacitance C_s of junctions as a function of current density. In other studies, it has been assumed to be $C_s = 60 \text{ fF}/\mu\text{m}^2$ for a critical current density of $56 \text{ kA}/\text{cm}^2$ [27]. On the other hand, it was measured as $C_s = 85 \text{ fF}/\text{cm}^2$ for $J_c = 50 \text{ kA}/\text{cm}^2$ in [28] and, estimated as $C_s > 100 \text{ fF}/\text{cm}^2$ for $J_c > 20 \text{ kA}/\text{cm}^2$ in [29] and [30]. To deal with this uncertainty, as well as with variations in microstrip dimensions and junction parameters due to manufacturing tolerances, a set of design parameter modifications was implemented on the wafer.

Figure II-2 Measured FTS response of the twin junction mixer – dotted red line; simulations of the response – solid blue line; simulated response of each single SIS mixer of the twin structure – dashed lines.

The measured FTS response of the receiver is presented in Figure II-2, together with results of 3D EM simulations on a comprehensive model of the mixer. The mixer model includes the superconducting tuning structure, the substrate, and the waveguide backshort. According to the best fit between the data and the simulations, the specific capacitance was found to be 80 fF/cm^2 . The receiver response is wider than 200 GHz; however, the design should be corrected to shift the response to higher frequencies for better coverage of the desired 790–950 GHz range. The twin junction design helps to widen the bandwidth, as demonstrated by the distribution of the energy between the junctions shown as dashed curves in Figure II-2.

II.3.2 NOISE TEMPERATURE MEASUREMENT

Noise temperature was measured using the Y-factor method, a room temperature (300 K) absorber as a hot load, and an absorber immersed in liquid nitrogen (78 K) as a cold load.

The results of the noise temperature (T_n) measurements are depicted in Figure II-3 [31]. The data are corrected for the 88% transparent $12 \mu\text{m}$ mylar beam splitter used for LO injection. Each point on the graph represents the dual sideband (DSB) noise temperature averaged over a 4–12 GHz intermediate frequency band. The T_n level is about 230 K at low frequencies and rises to 550 K at high frequencies in the range of 790–950 GHz (ALMA band 10). The presented frequency range was limited by the LO used. The

noise temperature at 720 GHz were measured using a different LO, yielding values of 350 K, and confirming an extremely wide mixer bandwidth (exceeding 200 GHz). The measured T_n is in good agreement with the FTS data in Figure II-2.

Figure II-3 DSB noise temperature, corrected for the loss in the LO insertion beam splitter, measured for the first (squares) and second (circles) minima of critical current suppression.

The critical current suppression of twin SIS mixers of sub-micron junction area can be a problem due to shape and area differences between the junctions in a twin structure. For better suppression, it is preferable to use the second or even the third minimum of the critical current vs. magnetic field, since these minima are typically shallower and wider than the first one. However, a magnetic field that is too high reduces the gap voltage (V_g) and can even decrease the response of the mixer at high frequencies [32]; both effects having a negative influence on mixer performance. Our mixers are of similar size and require about the same magnetic field level for critical current suppression compared to prior results [32]; in contrast to prior studies, however, for our mixers the second minimum not only provides a better critical current suppression compared to the first one (smaller Shapiro features for the Y-factor measurements, Figure II-4 [31]), but also gives a 6% lower T_n (Figure II-3). This is due to the relatively high T_c and higher gap frequency of the NbTiN film, and the change of film surface impedance under action of the H-field on the takes place at frequencies above the mixer operating frequency range. Our results are confirmed by the observation that the FTS response does not degrade at higher H-fields, unlike the results from a prior study [32]. An additional T_n improvement can be related to the reduction of the sub-gap current for higher H-fields (about 5% at 2 mV as

Figure II-4 IV-curves of one of the SIS junctions: "unpumped" – no LO power is applied (solid lines), "pumped" – LO power providing an optimal T_n is applied (dashed lines). The uncorrected Y-factor measured using 77 K/300 K loads are shown by dotted curves. Index "1" corresponds to the 1st minimum of the critical current suppression and related curves are grey; index "2" - corresponds to the 2nd minimum and related curves are black. The inset at the top shows the details of gap reduction due to junction heating by LO power and an increased magnetic field. The inset at the bottom right demonstrates sub-gap current variation around 2 mV for different minima of the critical current suppression. The LO frequency is 810 GHz.

shown in Figure II-4). The sub-gap current decrease is the result of better suppression of the Josephson current.

The noise temperature for the third minimum was not measured because the critical current suppression was not stable there. The critical current vs. H-field dependence $I_c(H)$ was not smooth above the second minimum and showed sharp current jumps, as if the junction switched back and forth (sometimes with hysteresis) between different states described by different $I_c(H)$ curves. This behavior, which we have observed before for some high current density AlN-barrier mixers with both Nb electrodes produced for ALMA band 9 receivers, was determined to be junction related. It may be related to flux trapping or inhomogeneous distribution of the supercurrent in

the junction, leading to a local field-dependent suppression of the superconductivity.

II.3.3 HEATING OF THE JUNCTION

With the choice of the Nb/AlN/NbN technology for the SIS junction fabrication with the NbTiN/SiO₂/Al three-layer for the embedding microstrip circuit, we expected an extra heating of the SIS junction, similar to one described in [33], [24] which is caused by trapping of thermal electrons in the Nb layer by two higher-gap superconductors: NbTiN located below and NbN placed on the top after a thin AlN barrier). In this case the charge would be transported from the Nb layer junction into the NbTiN and NbN, while the thermal energy transport would be blocked (a process known as Andreev reflection [34]) causing a local heating of the electron gas. This is confirmed by observation of back bending effect of unpumped IV curves for a large area (5 μm^2) test SIS junctions. This effect is not present in the case for mixers corresponding to Figure II-4 due to their much smaller area. The back bending and RF-power-dependent gap suppression (see Figure II-4) could be explained by the Andreev reflection [34]. On another hand, in our case ([31] © 2011 IEEE), the electron assemblies of the Nb and NbN layers are separated by the barrier and can be biased relative to each other by the bias voltage. It will change the transport condition for heat electrons in Nb layer and, basically, should cancel the Andreev reflection in dependence on bias polarity. That should provide asymmetry in the IV curves, which is certainly not the case, because they were measured to be perfectly symmetrical. As result, heating seems to be related not to Andreev reflection but to resistive properties of the mixer layers. The gap voltage is reduced by $dV_g \approx 0.1$ mV for the LO power level giving the best noise temperature (see curves "pumped" compared to "unpumped" on Figure II-4, also in the insert). A comparison of this result to similar mixer technologies, both with AlN and Al₂O₃ SIS insulator barriers, is represented in the Table II-1 (here our high gap mixers are called "CHAMP+ Type1"). For the small dV_g levels that we observe, we can assume that dV_g scales both linearly with the junction area and with the current density J_c (i.e., it scales with the total current). With this correction, shown in Table II-1 by dV_g^* , the difference in heating between high gap technology and the others becomes smaller. Despite the expectation of heating due to trapped hot electrons, one can see that the dV_g level of our device is not dramatically different from the ones without trapping mechanism.

To have an additional estimation of the heating effect in the high gap mixers we have fabricated SIS mixers by the same technology but with the NbN layers replaced by a Nb layer. These devices are presented in Table II-1 as "CHAMP+ Type2" [31]. The difference in dV_g , after normalisation on the

current density J_c , is only 25%, showing that contribution of NbN layer in junction heating is more than for Nb, but it is not a real bottleneck. This difference seems to be related to difference in the thermal conductivity and the RF electrical resistivity of the Nb and NbN films.

Name, (range, GHz)	SIS microstrip	Area, μm^2 <i>mixer type</i>	J_c , kA/cm ²	dV_{g1} , mV	dV_{g^*} , mV
ALMA B9 Type1 (600-720)	Nb/AlN/Nb Nb/SiO ₂ /Al	0.25 <i>single</i>	56	0.03	0.06
ALMA B9 Type 2 (600-720)	Nb/ Al ₂ O ₃ /Nb Nb/SiO ₂ /Al	0.25 <i>single</i>	30	0.01	0.04
ALMA B10 (787-950)	Nb/ Al ₂ O ₃ /Nb NbTiN/SiO ₂ /Al	0.86 <i>twin</i>	10	0.06	0.21
CHAMP+ Type1 (790-950)	Nb/AlN/NbN NbTiN/SiO ₂ /Al	0.5 <i>twin</i>	30	0.1	0.2
CHAMP+ Type2 (790-950)	Nb/AlN/Nb NbTiN/SiO ₂ /Al	0.5 <i>twin</i>	21	0.05	0.14

Table II-1 Comparison of junction heating by LO signal for different mixer technologies

ALMA B9 Type1 corresponds to ALMA Band 9 SIS mixers with Al₂O₃ barrier; ALMA B9 Type2 - Band 9 mixers with AlN barrier (data is taken from the lab measurements of cartridges delivered to ALMA); ALMA B10 - ALMA Band 10 SIS mixer [18]; CHAMP+ Type1 - high gap SIS mixer described in this paper, CHAMP+ Type2 - copy of CHAMP+ Type1, but with an Nb layer instead of NbN to avoid the heating effect. Here dV_{g^*} is a dV_g normalized to an area of $1 \mu\text{m}^2$ and to the current density (J_c) of 30 kA/cm^2 . The area is given per individual SIS junction, as there are two SIS junctions in a twin structure.

It is interesting to mention that the optimal pumping level (ratio of LO induced current to gap current [I_g]) is only 16–18%, which is lower than the typical 20–25% for regular single junction SIS mixers. This can be explained by unequal LO pumping of the individual SIS junctions of the twin mixer. The

same effect is observed in other twin SIS junction structures, as reported previously [35].

Ultimately, due to the high V_g , the reduction of the photon-assisted tunneling step by $2dV_g$ is not critical for the operation of the mixer; even for the highest frequency of 950 GHz, the step is reduced from 2.4 mV to only 2.2 mV, keeping the optimum bias point far away from V_g .

To estimate the deterioration of the noise temperature (T_n) due to heating of the junction, we measured it at lower bath temperatures. By cooling down the mixer to a temperature of 2 K, V_g was increased by 0.1 mV, thereby compensating for the gap reduction caused by the LO power, again, at the optimal pumping level. In this case, the measured noise temperature is 10% less than that at 4.2 K (Figure II-5) [31]. However, this measurement gives only a rough estimate, because although the average pumped gap value at 2 K comes out the same as the unpumped gap value at 4.2 K, the temperature increase of the SIS electrodes may be different in both situations. For a 2 K bath temperature with the applied LO power, the NbN energy gap still corresponds to 2 K physical temperature since that part is well-cooled through the Al layer, i.e., the energy gap is higher than that at 4.2 K. The energy gap of the Nb layer is evidently reduced with respect to its value at 4.2 K (due to heating by the LO signal), as we see the same average V_g . It should be mentioned that the sub-gap current around 2 mV was reduced by 10–13% at the 2 K bath temperature, compared to 4.2 K, which is comparable to the T_n change.

Figure II-5, DSB noise temperature, corrected for beam splitter loss, measured at bath temperatures of 4.2 K (squares) and 2 K (circles).

II.3.4 A PRELIMINARY CONCLUSION

The measured noise temperature of the new high-gap junctions is already 10% better than that of the average CHAMP+ high band mixers currently installed on the APEX telescope [17]. As demonstrated, the heating of the junctions is relatively small and the Nb/AlN/NbN SIS junction embedded in a NbTiN/SiO₂/Al microstrip circuit is a promising technology. In view of the extremely wide FTS response shown by the twin junction design, it is attractive to consider switching to a much easier to operate single-junction design, providing sufficient bandwidth. The final goal is to reach the same, or better, sensitivity as the state-of-the-art mixers for ALMA band 10 [19], which still demonstrate about 1.5 times better noise temperature than the junctions presented here. The difference in noise temperature may be caused by losses at high frequencies in the NbTiN and Al microstrip electrodes. Further research is needed to assess the quality of these films. The design of the mixer should also be modified slightly to shift the response to higher frequencies.

In conclusion, the presented mixer technology, compared to Nb/AlO_x/Nb mixers, appears promising as: 1) higher V_g provides a 0.7 mV wider photon step for high frequencies (allowing a nearly 1.5 mV wide operating bias range for a frequency of 1.15 THz); 2) high current density allows design of a very wideband mixer; 3) the high T_c of the NbTiN film makes it beneficial to use

the second minimum of the critical current suppression; and 4) junction heating is not critical for mixer operation.

II.4 TECHNOLOGY PARAMETERS ESTIMATIONS

The recent optimization of used manufacturing equipment has led to a change in the manufacturing process. This can lead to modifications in key parameters of superconducting film that may alter its behavior at high frequencies and affect mixer performance. Changes in production can be characterised by evaluating the EM parameters of the manufactured thin layers. To optimize design, it is necessary to estimate the relevant parameters (λ , σ , ϵ).

Effective parameters of fabricated films that are needed for the calculation of microwave circuits [22], [23] are: the permittivity of the insulating layer in the microstrip line; London penetration depth; conductivity of the superconductor measured with DC current in the normal state in the vicinity of the transition temperature; and superconducting transition temperature (T_c). The parameters T_c and DC conductivity are also needed for the models that take into account frequency-dependent losses [23], [24], whereas the λ -parameter is needed for the simplified models that disregard such losses. Note that all these parameters are interrelated.

The above parameters substantially depend on the production technology and often differ from the values in the literature. For example, an insulating layer may consist of a mixture of silicon oxide and dioxide at unknown relative contents rather than contain pure silicon dioxide. In addition, the presence of a boundary layer between the metal electrode and the insulator results in variation in the effective permittivity.

Figure II-6 A test planar capacitor layout. The center of the chip is a capacitor between the top and bottom electrodes. The blue rectangles show the top wiring and red indicates the bottom wiring.

II.4.1 A DIELECTRIC LAYER

EFFECTIVE PERMITTIVITY ESTIMATION

To determine the permittivity, we fabricated several test samples (Figure II-6) that represent plane capacitors based on the Nb-SiO₂-Nb structure. To correct for systematic measurement errors, we fabricated planar capacitors with different areas and identical (for each substrate) thicknesses. A Tencor Alpha Step 200 Profilometer was used to accurately measure the thickness of the dielectric layer. Precision measurements of the area were not necessary, since the relative error of linear sizes of fabricated items is smaller than 1%. The sample was fixed in the measurement unit at the end of a long metal tube and immersed in liquid helium. The measurements were performed using a two wires method, thus the resistance and the spurious capacitance of wires in the metal tube, which was dipped in a liquid helium, was taken into account. A multimeter was used to generate a low-frequency signal with fixed amplitude, which was needed for measurement of the external circuit resistance. The estimated capacitance of the capacitors was several hundreds of pF, and the effect of the resistance of connecting wires and contacts at a frequency of 100 Hz was insignificant. To account for spurious capacitances, we performed the measurements on the setup with liquid helium in the absence of the sample under study. Then, we measured a reference capacitance of 330 pF at the liquid helium temperature (spurious capacitance of connecting wires, metal tube, and measurement unit was 100

	T, K	DC conductivity, S/m	T _c , K
Nb	10	$2.21 \cdot 10^7$	9
NbTiN	15	$9 \cdot 10^5$	14
Al	4.2	$1.89 \cdot 10^8$	

Table II-2 Film properties.

pF). Based on the experimental results, we determined the permittivity of the dielectric layer to be 4.1. This result is physically consistent, since the permittivities of pure silicon oxide and dioxide are 5.6 and 3.8, respectively.

II.4.2 AL, NBTiN, NB FILMS DC-CONDUCTIVITY

We also determined the DC-conductivity of normal metal (Al) and the conductivity in the normal state for superconductors (Nb or NbTiN), which was measured at a temperature slightly above the critical temperature (T_c) of Nb and NbTiN. To determine all these parameters, we fabricated several Nb and NbTiN test film strips with different thicknesses deposited on the Al_2O_3 or SiO_2 buffer sublayer on the silicon substrate. The thickness and width of the samples were measured using a profilometer; the resistance was measured using a four-point scheme. Table II-2 presents the resulting conductivities. A semiconductor thermometer in the measurement unit is used to determine the critical temperature (T_c ; Table II-2).

II.5 DESIGN OPTIMIZATION, BASED ON SINGLE SIS JUNCTION

The knowledge of the layer parameters determined with the aid of the above methods makes it possible to decrease the number of unknown variables in the numerical model from six to one. As a result, only variation in the specific capacitance in the calculation model can be used to reach the best agreement between the theoretical model and experimental response curves. Figure II-2 shows the experimental and theoretical responses of the receiving element based on a structure with two SIS junctions, and the relative contribution of each tunneling junction to the total response. The analysis of the results shows that the receiving element based on a single SIS junction has a lower noise temperature due to better matching and decreased shot-noise level, but it has a narrower frequency band compared with the receiving unit based on twin junctions. However, the width of the frequency response of the receiving unit based on a single SIS junction is still sufficient for operation in the required frequency interval of 790–950 GHz.

A receiving element based on a single SIS junction is designed and fabricated using the parameters listed above. The first experiments showed that such a receiving element is extremely sensitive to deviations in the technological process, namely the mutual alignment of the junction and wiring layer, as it is difficult to realize the desired response band of 790–950 GHz. Figure II-7 shows the first measurements of the noise temperature of receiving elements fabricated on a single substrate and based on one or two tunneling junctions. The receiving element based on two tunneling junctions is less sensitive to technological deviations in the elements area. The optimization of technological procedures for fabrication of the receiving element based on two SIS junctions makes it possible to reach an even lower receiver noise temperature (~ 120 K at a frequency of 725 GHz); one that exceeds a quantum limit of $hf/(2k_B)$ by only a factor of 6.

Figure II-7 Mixer noise temperature based on 1 SIS junction (1 - dashed line) and 2 SIS junctions (2 - solid line). Both mixer types were taken from the same batch.

II.6 T_c MEASUREMENTS OF THIN FILMS USED IN MIXERS

For application of the Mattis-Bardeen theory, it is necessary to know material parameters, such as conductivity (σ), superconductor energy gap (Δ) and a temperature (T). The Bardeen–Cooper–Schrieffer (BCS) theory shows that the Δ is a function of a Δ_0 and a temperature, where the Δ_0 is an energy gap at 0 K. According to a prior report [36], there is a consistent estimation $2 * \Delta_0 = 3.52 * K_b * T_c$, where T_c is the critical temperature of the

superconductor. The film T_c measurements were performed using a DC bias current. The aim of the experiment was to measure film resistance as a function of temperature. The measurement setup included an Ohm meter, a cryo-thermometer, and a heater. The measured film was placed on a copper-aluminum holder block. Very good thermal contact between the block and a cryostat cold plate that was bolted through stainless-steel washers enabled regulation of block temperature with relatively low use of electrical power. Film samples were placed on a copper holder and pressed down by a spring to provide sufficient thermal contact. A thermometer was mounted on the copper holder. The assembly was covered by a thermal isolation shield. Figure II-8, Figure II-9, and Figure II-10 show the construction of the experimental block.

Figure II-8 Assembly for T_c measurements. There are 2 thermal shields marked as A and B. The A shield covers the sample holder and the thermometer, while the B shield covers two heaters (~ 100 Ohm resistors) to decrease the radiation from them that could be critical due to low thermal capacity of metals at liquid He temperatures. C and D show normal cryostat wires for the film resistivity measurements and heater supply.

Figure II-9 Assembly for T_c measurements (alternate view). A indicates washers that were placed under the block assembly. B indicates low thermal conductivity wires that were used for the thermal sensor and the film resistance measurement. D indicates the heater. C indicates the thermometer, which is attached to the block by a screw.

Figure II-10 Assembly for T_c measurements (alternate view). A indicates regular (not low thermal conductive) wires for the 4-point connection. B indicates the film on a Si substrate.

Figure II-11 Measured resistances as a function of temperature for 200 nm thick NbTiN from different batches, including solid and gridded films, deposited on Silicon or Quartz substrates. There are fringes below the NbTiN T_c that are related to a bit noise, or that could be due to a surface multilayer with a lower T_c than that of the NbTiN film. There was no significant difference observed between samples based on Silicon versus Quartz substrates. The T_c for NbTiN solid film fabricated on a fused Quartz substrate is $14.74 \text{ K} \pm 0.05 \text{ K}$.

Metal thermal capacity is a linear function of temperature for the low temperature limit; the thermal conductivity for wires is proportional to T^3 . Hence a low thermal conducting wire should be chosen to decrease parasitic heating or cooling of the film. Another important aspect is control of film heating by a measurement current. Additionally, there is a thermal electric effect for cryostat wire connection and contact resistances. To exclude these factors, a 4-point electric connection was used and the bias current was kept low, at $\sim 100 \mu A$.

During the measurement cycle, each film sample was cooled down, warmed up, and then cooled down again. This approach allowed estimate of the $\sigma(T)$ hysteresis. Figure II-11 illustrates the experimental results, which confirm low T_c and indicate that NbTiN film fabrication technology should be improved.

II.7 RF PROPERTIES OF THIN FILMS

EM fields are concentrated mainly between the top and bottom electrodes in a microstrip line. In a film made of a superconductor, the magnetic field penetrates the material for a London's λ distance in accordance with London's model. If the material electrode is made of a normal metal, the field penetrates it for a distance called the skin depth. This implies that the material quality and electrical conductivity of the top part of the bottom electrode and the bottom part of the top electrode of the microstrip line are very important. To estimate NbTiN conductivity properly, measurements should be performed at RF frequencies using films that are fabricated with the same technology used to fabricate the mixer chip. For the $\sigma(\sigma_0, f)$ estimation at RF frequencies, a dispersive FTS (Fourier Michelson Interferometer) was used. A sample (NbTiN grid or solid film) was installed in a cryostat at the position of a fixed mirror of a Mickelson Interferometer set-up. Brief description of the method is as follows:

In the first step, we calculate the contrast value, which is given by the relation $\frac{P_c - P_h}{P_h}$, where P_c and P_h are power spectra measured for the film temperatures below and above T_{c1} , respectively. The film temperature was controlled during the measurements by a script-based feedback loop system, which adjusted the heater current until the holder block temperature was stabilized.

The second step was a surface resistance film calculation using the Z_s formula from chapter 1.4.2. The complex impedance can be rewritten as

$R+iX$, where R is the resistance (a real part) and X is the reactance (a complex part of the Z_s).

In the third step, the solid film reflectivity is estimated as $\frac{Z-Z_0}{Z+Z_0}$, where Z is the surface impedance and Z_0 is the free vacuum space impedance. In our case, the MWS CST was used for reflectivity coefficient simulation. Due to the large size of the grid ($\sim 2 \times 2 \text{ cm}^2$ in total; square grid was formed by $3.3 \mu\text{m}$ wire with $18 \mu\text{m}$ spacing), the periodical Floquet mode solution was used. Consequently, a frequency-dependent reflection coefficient for every R and X couple from step 2 (described above) could be calculated.

In the fourth step, for every set of variables in the Mattis-Bardeen (MB) theory (T , σ_0 , Δ and f), the reflection coefficient was taken from the CST data. Next, the theoretical contrast was calculated in cycle until the contrast derived from the experimental data was fitted to the numerical calculation, thereby generating an estimate of the experimental dependence $\sigma(f, T)$ (Figure II-12).

The Nb and NbTiN films were tested using the described method. We were able to measure the gap frequencies of actual conductive layers in the structures, which were found to be much lower than corresponding values for bulk superconductors. This allowed us to optimize the fabrication process by choosing an optimal NbTiN sputtering target. We also observed that further film processing (etching or lift off) could lead to degradation of the resulting film RF losses and could explain significant batch to batch differences of experimental response (FTS) of devices with the same design layout. After optimization of manufacturing technology, the estimated conductivity σ was determined, which was used for later numerical calculations.

Figure II-12 NbTiN film (350 nm thickness) surface impedance calculated using MB theory and its experimentally estimated parameters. This data set was used for a SIS mixer (780–950 GHz prototype) CST calculation.

II.8 CST NUMERIC CALCULATIONS FOR CHIP PARAMETER DEVIATIONS

During the project, the manufacturing technology was developed, but only a subset of all new samples generated resulted in improvement. In fact, many of them led to new issues in measurement. This sub-chapter includes some of the intermediate results of the technology optimization process. Due to the long evaluation time for 3D EM simulation for each parameter set, and a large number of relevant parameters, the brute force approach of sweeping all possible parameter values would result in a prohibitively long calculation time. Thus, the analysis was done manually. There are many model parameters, e.g., film thickness and conductivity, the rule for calculating the impedance of the conductivity in the film for sublayers, and possible sample shifts in a backpiece (BP). While model sensitivity to all parameters was estimated, here we report only the most significant parameters. For comparison of the FTS measurements with the CST EM simulations, Fresnel formulas and frequency multiplication were used.

Figure II-13 An eBeam photo of the 780–950 GHz frequency range test mixer with TWIN SIS junctions. These junctions are discriminated as two points on a top wiring. Material granules on the photo are discriminated.

To clarify the as-made details of the sample geometry, several samples (see Figure II-14) were scanned using an electron microscope (Figure II-13). The effects of height shift of the sample in the waveguide channel and a lateral shift during installation were analyzed. The influence of the deviation of the substrate thickness from the nominal value and the influence of changes in the dielectric constant from the nominal value were also calculated.

Figure II-14 A microphotograph of the design of the test mixer with TWIN SIS for the 780–950 GHz frequency range.

Figure II-15 The 3D CST model layout for the test mixer in the 780–950 GHz frequency range.

The mixer CST model (Figure II-15) included material parameters and the SIS junctions were simulated as lumped ports with connected parallel capacitors, while the superconductor and normal metal films were simulated as materials with a tabulated surface impedance. To model the superconductor wiring, Mattis-Bardin's theory was used with the parameters determined in the reflection experiment described above. Figure II-16 shows

matching curves calculated for various NbTiN layer thicknesses (from 1700 to 3300 Å) and a corrected FTS response curve.

Figure II-16 The numerical response of the model with different NbTiN layer thicknesses and the experimental FTS response.

With careful installation, the measured FTS for similar samples were comparable. Figure II-17 shows the measured FTS for the same design, but with different samples.

Figure II-17 The measured response (FTS) for similar design and SIS using two different test samples.

Noise temperature measurements take more time than an FTS experiment, and we have been comparing the FTS response and the CST calculations. Therefore, we tried to find parameter values to make a match between the CST's numerical calculation and the noise temperatures. To do this, we remeasured a known band 9 mixer on the same setup as a band 10 mixer

(Figure II-18) and made full 3D CST models of each mixer. By comparing the measured noise temperature and the corresponding calculated curves for the band 9 receiver, we obtained a coefficient for “recalculating” the noise temperature of the receiving element from 1 SIS to the matching level in CST model (Figure II-19).

Figure II-18 Reference interpolated noise temperature for band 9 and band 10 mixers.

Figure II-19 Measured and numerically estimated noise temperature vs. LO frequency. Numerical estimates were based on CST results (see text for more detail).

To obtain a mismatch of 0.6 at a frequency of 700 GHz, it was necessary to increase the conductivity of the metal film in the numerical model. The conductivity of the superconductor is not calculated in the CST model; therefore, the losses were artificially added into the aluminum layer. (Sample a06, 1st minimum of Josephson current magnetic field, batch is 16-01;

Aluminum conductivity is in S/m, see Figure II-20). Results from application of this algorithm to another batch are shown in Figure II-21.

Figure II-20 Numerical power coupling evaluated by CST with different Al conductivity for the top wiring layer. The figure also includes the measured FTS response and curves from Figure II-19. Data are from batch 16-01.

Figure II-21 Power coupling calculated by CST with different Al conductivity for the top wiring layer. The figure also includes the FTS response and curves from Figure II-19. Data are from batch 16-02.

If we assume that NbTiN is mixed with another low T_c superconductor, e.g., Nb, in the production process, then we can model the structure as two layers (NbTiN and Nb) with a thickness ratio of 0.1 to 0.9. With this assumption, we obtained typical matching curves as shown in Figure II-22.

Figure II-22 Numerical power matching by CST with different lateral SIS structure shifts (μm) on the chip, e.g., a photolithography mask was shifted during manufacturing. The sample shift in height was 0 in the back piece. The mixed conductivity proportion used was 0.1 Nb/0.9 NbTiN.

There were some batches with nonoptimal SIS parameters. Consequently, the noise temperature was not optimal, yet the FTS response shape could be reasonably well-explained. For some samples, the FTS was wider and dependent on the used mixer block due to differences in mixer block waveguide geometry. Indeed, variations in distance between an EM tuner and a probe alters the coupling profile vs. frequency. The second reason for the discrepancy was an uncontrollable gluing of the sample. To control these effects, we characterized/measured mixer block geometry and controlled the mounting process. Figure II-23 and Figure II-24 show the calculated response differences. The final simulation is shown in Figure II-25. A striking similarity between model calculation and measurement data was observed.

Figure II-23 Numerical responses for a 780–950 GHz frequency range mixer with different dielectric layer parameters. The experimental power response is also presented.

Figure II-24 Numerical responses for a 780–950 GHz frequency range mixer with different dielectric layer parameters and TWIN SIS junctions shifted into the probe side by 1 μm . The experimental power response is also presented.

Figure II-25 Numerical power matching by CST and an experimental power response (FTS) for a 700–950 GHz frequency range mixer after whole mixer block and layer characterisation.

II.9 A NEW 700–950 GHz DESIGN STRUCTURE

Development of the 700–950 GHz mixer was occurring simultaneously with that of the 211–275 GHz frequency range mixer described in the next chapter. Reanalyzing the matching principles and considering the latest technology parameter estimation, we produced a new test design. The probe was modified to a low impedance of $\sim 40\text{--}50$ Ohm and the mixer was tuned with estimated technology parameters. A comparison between the old and new designs is shown in Figure II-26. The new chips are fully compatible with the hardware developed for the previous design.

II.10 CONCLUSION

We have developed a DSB mixer for the 700–950 GHz frequency range. The SIS mixer was designed, manufactured and measured. Different chip mixer structures have been developed and used for characterizing and optimizing the manufacturing technology process.

The first significant result is production of a mixer set for updating the frequency range (ALMA band 10) receiver on the APEX telescope. The noise temperatures before and after the replacement mixers were installed are presented in Figure II-28 [37]. The optimization led to significantly decreased noise temperature of the instrument.

Figure II-26 Numerical power matching by CST for the old and new chip designs. The peaks observed at frequencies higher than 1000 GHz are artifacts from calculation model simplification.

The second significant result is an accumulated experience with SIS HF structure development. During the R&D process, we have improved our calculation approaches and manufacturing technology, which has been challenging and time-intensive due to many material technology obstacles, including the production of high-quality films and high-gap SIS junctions (0.5 μm^2 area) with $\sim 30 \text{ kA/cm}^2$ current density without e-beam lithography facilities. The demonstrated calculation approach numerical simulations performed for many design parameters have helped reduce the manufacturing and experimental time cycle and led to improvement in NbTiN film manufacturing technology.

Mixers produced during this project have been used later for receivers with Side Band Separation [38]. The experience and knowledge acquired throughout the project have helped in design of a DSB mixer for a lower frequency band through single EM design iteration, as described in the next chapter.

Figure II-27 DSB mixer noise temperature for different samples based on TWIN (2 SIS junctions) structure. The DSB noise temperature is averaged over the 4–12 GHz band.

Figure II-28 DSB noise temperature comparison of the new mixers, prepared for delivery to APEX, and old mixers delivered from the telescope. The noise temperature is corrected for approximately 10% insertion loss of a beam splitter and for the hot LO (300 K) contribution.

CHAPTER 3

III A heterodyne waveguide receiver in the 211–275 GHz range

We have designed, simulated, manufactured and evaluated a Double Side Band (DSB) receiver, which is operational in the frequency range of 211 to 275 GHz; the Intermediate Frequency (IF) band is 4 to 8.4 GHz. We employed high quality Nb/AlO_x/Nb superconductor-insulator-superconductor (SIS) tunnel junctions to achieve sensitivity close to limit. The DSB mixer noise temperature, as low as 16 K ($\sim 1.3 hf/2k_B$), was measured at 255 GHz using an averaged IF band of 5 to 8 GHz with no loss correction of the optical elements, which are present in the mixer input. The measured noise temperature at 220 GHz was 32 K, which decreased until 255 GHz where we obtained 16 K, and rose after the optimum point, achieving 22 K at 275 GHz. The testing campaign included the evaluation of the Single-Side Band (SSB) mixer design along with the DSB architecture.

III.1 INTRODUCTION

The development of ultra-sensitive heterodyne Terahertz (THz) receivers is currently one of the most successfully explored areas in astronomical instruments developing. SIS mixers have been used successfully for space missions, such as that involving the Hershel HIFI instrument, as well as for several ground-based telescopes, including the largest ALMA multi-element interferometer [39], the European NOEMA interferometer [40], and the APEX telescope [41]. Mixers based on SIS tunnel junction technology are highly sensitive devices at frequencies from 0.1 to approximately 1.2 THz. The noise temperature for DSB mixers is theoretically limited only by the quantum value $hf/(2k_B)$ [42], where f is the frequency, h is the Planck constant, and k_B is the Boltzmann constant.

Some DSB mixers allow exceptional sensitivity, achieving a noise temperature between 10 to 15 K. There is limited application in large array observatories due to the mechanical matching tuner [43], as it requires fine mechanical RF tuning in each receiver to achieve high performance for each frequency point. The ALMA observatory operates at band 6 (211–275 GHz),

using a receiver deployed by NRAO. The receiver has a SSB architecture and a noise temperature of around 40–60 K [44]. Another telescope array – NOEMA – has receivers operating at 206–267 GHz with ~ 38 –50 K of noise temperature and an SSB configuration [45]. Noise temperature at ~ 40 K was also achieved by a mixer developed at the University of Oxford [46].

The mission of the Russian space observatory – *Millimetron* [47] – is to observe various sources in the universe at millimeter and infrared frequency ranges, which will provide data for scientific studies. It is equipped with a 10 meter space telescope that has two operational modes: single-dish observation and Space-Earth interferometry. The second mode is aimed to observe extremely compact objects, e.g., the immediate vicinity of black holes, which requires ultra-high angular resolution (up to tens of billionths of an arc-second). The high angular resolution is directly dependent on the interferometer base line, which is defined by the *Millimetron* orbital configuration located near the Lagrange point L2 at 1.5 million kilometers from Earth. The sensitivity of the instruments and low noise temperature are also important variables in a high-resolution system and, for that reason, the band 6 receiver must perform a noise temperature below 45 K. The Sideband-Separating (2SB) mixer block and its SIS junctions are currently under development.

This chapter presents the results of the development and characterization of a DSB SIS heterodyne waveguide receiver test prototype that is operational at band 6. The developed receivers could be used for many future radio astronomy projects, such as LLAMA [48] and SUFFA [49].

Figure III-1 A 3D design of a double-side mixer layout, positioned inside a waveguide channel. The backshort is presented in a back view. The input horn is placed in the front of the waveguide mixer block (Waveguide in). The design includes a single SIS junction incorporated into Nb wiring.

III.2 MIXER DESIGN

We placed the mixer chip into a waveguide orthogonally to the propagation direction (Figure III-1), following the successful modular construction methodology of the ALMA Band 9 receiver [50, p. 9]. The Nb/AIO_x/Nb SIS junction was placed into a planar Nb/SiO₂/Nb tuning structure. The tuning structure was deposited on a 125 μm thick quartz substrate. The receiving chip (width 150 μm) is located in a rectangular waveguide (1000 x 500 μm, A x B, respectively; (Figure III-1) at a distance of 230 μm from the backshort in the waveguide. The mixer block, shown in Figure III-2, has several elements: a central part that includes the waveguide, a magnet block unit with two magnet pins to suppress the Josephson critical current, a back piece (BP) unit where the mixer chip is installed, and an input horn.

Figure III-2 An assembled mixed block, which includes a horn, a magnet block with two magnet pins, a central body block, and a backpiece.

The quartz chip, placed in the waveguide channel, is itself a dielectrically filled waveguide with a cut-off frequency of approximately 320 GHz, which corresponds to the first mode (Figure III-3). Low-pass RF filters (Figure III-1) were used to prevent leakage of the RF signal through this dielectrically filled waveguide. The receiving device in its working range operates in a single-mode regime. The main TE_{10} mode has a cut-off of frequency of approximately 150 GHz and the second mode has a cut-off frequency of approximately 305 GHz. The tunnel junction should be matched properly with the input waveguide impedance, which is dependent on RF frequencies, and with a 50 Ohm amplifier impedance in the entire IF frequency band toward achieving good SIS mixer performance.

Figure III-3 The 3D model of the backpiece. A test PCB plate with a 50 Ohm MSL is incorporated. PCB, printed circuit board.

To match the waveguide impedance (approximately 400 Ohm) and the RF SIS impedance, a waveguide probe and a tuning structure were used. The probe has a quasi-triangular shape because it allows the real impedance to be matched in a wide range of frequencies [51]. The calculated impedance of the current probe design is shown in Figure III-4. The calculations were performed with consideration of the finite thickness of the superconducting and dielectric layers in the strip lines.

Figure III-4 Calculated impedance of the waveguide probe: Re (solid line) and Im (dashed line).

A combination of the coplanar waveguide (CPW) and the microstrip lines (MSLs) was used to tune out the intrinsic SIS capacitance and allow matching of the resulting SIS impedance with the waveguide at the RF. This approach allowed us to make the tuning structures more compact.

The wavelengths in the tuning lines at the IF range are much larger than the dimensions of the structure, therefore the tuning of the microstrip line adds some capacitance to the sample impedance at IF. For this reason, the output receiver impedance must be adjusted by a printed circuit board (PCB), with a dielectric constant of 9.8, at the mixer output. In this work, we used a 50 Ohm line printed on the PCB without additional tuning elements (Figure III-3). The embedded impedance of the SIS junction [52], [53], shown in Figure III-5, was calculated via CST Microwave Studio (CST).

Figure III-5 The calculated embedded impedance of the SIS junction in a Smith Chart. The S parameter was normalized to 20 Ohms.

The London penetration depth for Nb films, approximately 85 nm, was used to account for the superconducting properties of microstrips and the antenna in the calculations. The design was developed and simulated using CST; the numerical matching is shown in Figure III-6.

Figure III-6 Experimental FTS results (solid line) and calculated matching at the main waveguide mode for the double-side design (dashed line). The features of the experimental and calculated curves at frequencies around 320 GHz are related to a resonance in the sample substrate.

Based on the DC IV tests, the SIS parameters of the tunnel junction were determined (S , R_n). The geometric dimensions of the sample were also measured. The measured parameters were used to correct the CST model of the sample, with the calculation accounting for the phenomenon of superconductivity.

The real and imaginary parts of the embedding impedance satisfy the stability criteria [52], where the real part of the complex admittance in the mixer output is always positive at the operation bias point, and at the optimal LO power for every RF point. A moderate mixer conversion gain of 1–3 dB is expected for mixer operating frequencies.

III.3 TECHNOLOGY AND MEASUREMENT RESULTS

This section includes the main experimental results and is divided into logical milestones.

III.3.1 MANUFACTURING OF THE SIS MIXER CIRCUITS AND ITS TEST AT DC

SIS tunnel junctions with extremely small leakage current under the gap voltage and minimal energy gap spreading (δV_g) are required to realize a quantum-limited performance. This is especially important for relatively low-frequency devices ($f \sim 200\text{--}300$ GHz), since the δV_g has to be much smaller than the size of the quasiparticle step hf/e , while the leakage current at a bias voltage of approximately $V_g - hf/2e$ determines the noise of the mixer.

The fabrication technology of a Nb- AlO_x -Nb tunnel junction is based on the fact that a very thin Al layer can completely cover the Nb base-layer [54]–[56], effectively creating a planar Al layer on top of the column-like structure of the Nb film. The Al layer is subsequently oxidized and another Nb layer is deposited on the oxidized layer to form a tri-layer structure. The SIS tunnel junctions can be fabricated from the tri-layer structure by using the selective niobium anodization process (SNAP) [57] or the selective niobium etching and anodization process (SNEAP) [55], [58]. The stress, surface morphology, superconducting characteristics, and crystal structure of sputtered Nb films, as well as the surface coverage of the thin Al layers deposited on Nb films with different sputtering parameters, were evaluated for fabrication of high-quality Nb- AlO_x -Nb junctions [59]–[62].

SNEAP technology was used in this work for fabrication of the SIS receiving structures based on Nb/Al- AlO_x /Nb circuits [25], [63], [64]. The substrate, where the tri-layer structure is deposited, consists of a quartz plate with a thickness of 125 μm . A “monitor” layer of Nb, approximately 100 nm thick, is deposited all over the substrate by DC magnetron sputtering to prevent etching of the quartz plate. The next step is the deposition of a tri-layer structure; it is carried out using DC magnetron sputtering in a single vacuum run. First, a layer of Nb, called electrode, with a thickness of 200 nm is deposited, followed by the deposition of a barrier Al layer with a thickness of 7 nm. Next, the aluminum surface is oxidized in an atmosphere of pure oxygen and the second electrode layer of Nb, with a thickness of 100 nm, is deposited. The geometry of the base electrode is formed by a lift-off process.

The SIS junctions are formed by plasma-chemical etching in CF_4 when the top Nb layer of the tri-layer structure is removed, according to the designed geometry, using a photoresist mask. The AlO_x barrier layer acts as a stop layer, which prevents further etching of the structure. The structure is anodized, up to 10 V; then an insulating SiO_2 layer, typically with a thickness

of 250 nm, is deposited by RF magnetron sputtering. The opening of contacts to the junctions is carried out by a lift-off process. The top electrode is also formed by lift-off from the sputtered 400–450 nm thick Nb layer. The gold contact pads of the base and top electrodes are formed in a similar way.

The current-voltage characteristic (IVC) of a Nb/Al-AIO_x/Nb SIS mixing element with an area of approximately 1 μm^2 is shown in Figure III-7.

Figure III-7 The pumped (dashed green line) and unpumped (solid black line) IV curves generated using data from the 241.2 GHz LO injection. The IF powers for the mixer under the hot (dashed red line) and cold (solid blue line) loads are also shown. The LO power was tuned to an optimal level. The IV curve displays a "knee" feature at voltages just higher than the gap voltage (2.75 mV). Furthermore, there is a small leakage below the gap voltage. Due to the high mixer gain, these features may manifest in the power vs. bias curve. It is important to emphasize that we used the central part of the photon assistant step, which is not influenced by those effects.

The IVC was measured in voltage-bias mode and the critical current of the SIS junction was suppressed by a magnetic field. The normal resistance of the SIS junction is $R_n \sim 34 \Omega$. The quality parameter was characterized by the ratio of the resistances under and above the gap, resulting in $R_{sg}/R_n \sim 36$, with the gap voltage (V_g) ~ 2.75 mV, and energy gap spreading $\delta V \sim 0.1$ mV. The pronounced "knee-like" feature on the IVC at voltages slightly higher than the V_g was due to the presence of a normal aluminum layer near the tunnel barrier. Its presence substantially modified the density

of electron states in the superconducting electrode. A theoretical model of such a structure [65] is built on the quasi-classical Usadel equations on the dirty limit. In that limit, the electron mean free path in a normal metal and in a superconductor is much less than their coherence lengths. It was experimentally demonstrated that the “knee-structure” can be suppressed considerably by introducing an additional Al layer in the base electrode [63]; we used this technique in subsequent experiments.

III.4 FTS AND T_{NOISE} MEASUREMENTS

The Michelson Fourier Transform Spectrometer (FTS) technique was used to evaluate wideband radiation matching of the SIS mixer at RF. A wideband GHz-THz source, a glow bar, was matched with the FTS, which was loaded to the SIS mixer as a detector. The mixer was voltage biased at 2.3 mV, then a direct current response was measured versus a mirror position; we applied the Fourier transform on those data to obtain the mixer response on the frequency domain. The experimental data are represented by the solid line in Figure III-6; the results of numerical simulations represented by the dashed line demonstrate a good fit with the experiment.

The noise temperature was measured via the Y-factor method using an absorber placed in liquid nitrogen (78 ± 0.2 K) as a cold load, and a room temperature (296.4 ± 1 K) absorber as a hot load. Figure III-8 shows the measured DSB uncorrected noise temperature over the RF band in comparison to a quantum limit level (Callen & Welton) [42].

Figure III-8 The noise temperature is presented without any correction for losses in the input optical elements (T_n curves) in comparison with a quantum limit level (Callen & Welton; dashed lines). The measurement uncertainty (~ 2 K) is indicated by error bars. The best T_n values measured in a narrow IF band (40 MHz) are indicated by the solid blue line. The T_n values averaged over the entire IF band (4–8.32 GHz), for each measured LO injection frequency, are indicated by the dashed red line. The range, 227–275 GHz, was limited by the parameters of the LO source used.

The primary experimental scheme is presented in Figure III-9. The signal from hot/cold loads passes through the 6 μm thick mylar beam splitter and the cryostat window, similar to the one used for ALMA band 6 [66]. Grid 1 is used to regulate LO power coupled to the mixer and grid 2 is inserted only during the receiver linearity measurement. An electronically tunable bandpass YIG filter was used to evaluate noise temperature distribution in the IF band. The experimental results are presented in Figure III-10. Each point on the graph represents a DSB noise temperature at specific IF averaged over a 60 MHz bandwidth of the YIG filter. The noise temperature measured for the double-side design achieved values below the hf/k limit at IF of around 8 GHz for a LO frequency of 241.2 GHz. Figure III-11 shows the IF receiver response in a narrow IF band with higher frequency resolution than the one presented in Figure III-10.

Figure III-9 The figure shows a RF and an IF principal experiment scheme. Grid 2 was installed for the linearity test measurement. Ordinary the grid 2 was removed.

Figure III-10 The uncorrected noise temperature from 4–12 YIG GHz (solid red line) compared to a quantum limit level (Callen & Welton; dashed lines). The LO frequency is 241.2 GHz. Originally, an inductor-capacitor (LC)-tuning structure on a PCB was planned. The results presented in this figure correspond to a 50 Ohm line (see Figure III-3) instead of an LC-tuning structure, which led to standing waves in the IF circuit. Therefore, the SIS junction impedance in is also changing and, as a result, the power matching is decreasing.

Figure III-11 Uncorrected noise temperature across an IF band compared to a quantum limit level (Callen & Welton). Measurement uncertainty is ~ 2 K, as shown by error bars. The data demonstrate that the IF circuit is not tuned and there are standing waves. A tuning IF structure on a PCB plate in the BP was replaced by a 50 Ohm line on the PCB. The LO frequency is 241.2 GHz.

The essential condition for the Y factor method is receiver linearity. Therefore, the experimental RF scheme was changed to check that receiver property [67]. The additional motorized grid 2 was inserted in the receiver beam at the position of the hot/cold load, as shown in Figure III-9. The IF receiver output power was measured as a function of the grid angle. In this experiment, the mixer received not only cold/hot load radiation from the load side, but also room temperature radiation reflection from the grid, which was placed in between the optical path of the cold/hot loads and the mixer. To check the measured linearity response of the receiver vs. the grid angle, the grid angle was fitted to the theoretical grid attenuation level as shown in Figure III-12.

Figure III-12 A fitting of the cold power level with respect to grid angle dependence. Unfortunately, the correct fitting at 0–180 degrees or 180–360 degrees is available due to unflattening of the grid. The curves shown represent 180–360 degree experimental data.

The grid inhomogeneity led to a slight difference in transmittance between 0–180 and 180–360 degree grid angles. Therefore, the experimental curves can be fitted well at the 180–360 or 0–180 degree ranges. The comparison shows that the mixer has slight compression, on the order of 1%, at a 300 K load, which is a typical number for SIS mixers in this frequency range. The correction gives 32.8 K instead of 31 K noise temperature at 241.2 GHz LO injection frequency at 6.66 GHz IF; likewise, the correction gives 25.5 K instead 23.9 K at 8.36 GHz IF for the same LO injection frequency.

The receiver stability at 241.2 GHz of LO was evaluated by measuring receiver output vs. time under the same operating conditions of the noise temperature measurements. The hot load was presented as from the receiver. A 60 MHz band pass filter was used to limit the signal bandwidth, and an Allan time [68] of ~ 10 s was registered (Figure III-13).

Figure III-13 The graph shows Allan Variance curves measured at 241 GHz (dotted blue line) and 261 GHz (solid red line).

III.5 NOISE BREAKDOWN

As we are close to the fundamental sensitivity limit, we performed noise temperature analysis of the receiver system and assessed the contribution of different components. We analyzed the receiver state as presented in Figure III-9 (without grid 2), based on the measured data and conditions of Figure III-7. As presented in Table III-1, the system consists of a 6 micron mylar beam splitter, a receiver window, an SIS mixer, and an IF circuit, which is a combination of a Pamtech cryogenic 4–12 GHz isolator and an YEBES cryogenic amplifier [69]. Each receiver element is characterized by its T_{noise} , which is an expression of noise temperature in reference to its input and gain. The parameters of the beam splitter were evaluated using Fresnel formulas. The cryostat window contribution was evaluated by measuring its transmission in an FTS set-up. The IF system contribution was evaluated using SIS junction as source of shot noise, which is proportional to its bias current. When the normal resistance is taken into account, we use the coefficient 5.8 K/mV and two bias points above the gap voltage to calibrate out the IF system gain and evaluate the IF system noise. Finally, the contribution of each element to the system input (T_{input}) was calculated using a cascaded element connection formula. The sum over T_{input} column in Table III-1 should correspond to the measured receiver noise temperature of 8.9 ± 2 K. Using this condition, the SIS mixer contribution (T_{input}) and its

corresponding T_{noise} was evaluated. The break down appears to be self-consistent within ± 2 K overall accuracy. The noise contribution of the SIS mixer is very close to the fundamental quantum limit of $hf/(2k_b)$, which is demonstrated experimentally.

NOISE BREAKDOWN			
Element	Tnoise, K	Gain	Tinput, K
Beam Splitter	0.5926	0.9980	0.5926
Window	0.5	1	0.501
SIS	5.94	3.08	5.9544
IF	6	1	1.952

Table III-1 The noise breakdown. The $hf/(2k_b)$ quantum limit is 5.78785 K at 241.2 GHz.

It must be emphasized that the measured SIS mixer gain is 3, which is a very high value. This is due to the choice of embedding impedance of the SIS mixer and the influence of a "knee" structure above the gap. Despite high gain, we have experimentally demonstrated that the receiver is linear and stable at an integral input signal mode. Due to high mixer gain and low noise contribution of the input elements, the overall receiver uncorrected noise temperature is also very close to the quantum limit. This is an important experimental confirmation that will inform improvement of SIS receivers for radio astronomical applications.

III.6 NO "KNEE" SIS MIXER ITERATION

With new lab facilities, we have now been able to carry out a clear "testone" experiment. The method consisted of measuring a weak signal (a testone) by the mixer. The LO and the weak signal source were tuned to detect the testone signal image in the IF mixer output. If the testone power is weak enough and the mixer works in a linear operation mode, the response value (a peak amplitude relative to the IF background), measured on IF, does not depend on a hot/load background.

The manufactured mixer, unfortunately, works without stability in a heterodyne mode with a testone, despite an almost linear mode with a significant Allan variance time. For this reason, the device cannot be used in

an astronomical instrument like a telescope or a satellite as a heterodyne receiver. We suggest that the "knee structure" is the reason for that behavior.

In order to use the mixers for astronomical instrumentation, a new batch of mixers based on SIS junctions with an additional 50 nm Al layer placed in the base electrode [63] was manufactured and tested. This layer not only improved the morphology of the niobium film near the barrier, but also significantly changed the electron distribution function in the electrode near the barrier, which led to the complete suppression of the knee-like feature on the IVCs of the Nb/Al/Nb-AlO_x-Nb tunnel junctions. The new manufacturing technology made it possible to realize an extremely low leakage current at voltages below the gap (the ratio of resistances under and above the gap R_j/R_n is greater than 35). This is close to the maximum possible value for junctions based on niobium, making it possible to achieve noise temperatures limited only by the quantum values.

To perform these measurements, a python based open-source script written to automate the test set-up was used. It allowed us to optimize the measurement time. Figure III-14 shows the optimal bias points that were found during the measurements using the full IF bandwidth. The charts show a smooth tuning of the mixer. After the tuning, the mixer response was measured for every LO frequency in the IF spectrum, for a cold (liquid nitrogen) and hot (room temperature) load.

To decrease noise temperature uncertainty due to spatial variations in room temperature, a thermometer was placed very close to the load. Figure III-15 shows the uncorrected noise temperature at the IF range for different LO frequencies. There are some gaps around 4.3 GHz for some LO frequencies due to a slight compression at those points. This feature will be studied further in future analysis.

It is important that the receiver does not have a proper LC-IF tuner after the mixer, but rather a 50 Ohm line. This could explain the standing waves in the IF system. Furthermore, the system is matched with the output backpiece port and, according to the numerical simulations, the output SIS impedance declines at frequencies higher than ~8-9 GHz if a proper tuner is not present in the backpiece.

Figure III-16 and Figure III-17 show measured noise temperatures with the Y-factor for the target frequency range. Considering the measurement experience with the similar mixer previously described (with the "knee"), the testone measurement was performed. There were some limitations with the

slight frequency tuning of the frequency sources used; therefore, the testone was not performed over the full LO range. Figure III-18 shows some of the measured results.

The SIS mixer was tuned at 247 GHz or 255 GHz and the second source was tuned for close frequencies for the purpose of measuring a testone spectrum line in the IF range of the mixer response. The mixer was operated in a linear mode, and the compression level was not achieved 1%. After the linearity test described above was performed, the Allan Variance was calculated for different M-sample variances. Figure III-19 illustrates the result. The Allan Time was defined as ~ 10 sec. Tables III-2,3 show the noise breakdown for this receiver configuration.

Figure III-14 Uncorrected mixer noise temperature as a function of SIS bias for different LO frequencies. The noise temperature value was measured by Y factor. The IF power for the hot/cold measurement included a full 4–12 GHz frequency range.

Figure III-15 Uncorrected noise temperature from 4–12 YIG GHz compared to a quantum limit level (Callen & Welton). LO frequency is 247 GHz for the top graph and 255 GHz for the bottom graph. There is only a 50 Ohm line (see Figure III-3) instead of an LC-tuning structure on a PCB, which was originally planned but led to standing waves in the IF circuit and a mismatch at the 9–12 GHz frequency range. The SIS junction impedance was changing, which resulted in decreased power matching.

Figure III-16 Uncorrected noise temperature (solid lines) compared with quantum limit levels (Callen & Welton; dashed lines). Measurement uncertainty is ~ 1.5 K, as indicated by error bars. The T_n values averaged over the IF 5–8 GHz and 5–12 GHz bands for each measured LO injection frequency are shown by solid blue and red lines, respectively.

Figure III-17 Uncorrected noise temperature (solid lines) compared with quantum limit levels (Callen & Welton; dashed lines). Measurement uncertainty is ~ 2 K; as indicated by error bars. The T_n values averaged over the IF 5–8 GHz and 4–8 GHz bands for each measured LO injection frequency are shown by solid red and blue lines, respectively.

Figure III-18 SIS mixer responses for a testone signal (240.8 GHz or 249.5 GHz) on 247 GHz (top) or 255 GHz (bottom). There was a slight compression of less than 1% (0.88% and 0.72%) for the 247 GHz and 255 GHz LO frequencies, respectively.

NOISE BREAKDOWN			
Element	Tnoise, K	Gain	Tinput, K
Beam Splitter	0.5926	0.9980	0.5926
Window	0.5	1	0.501
SIS	16.454	2.19256	16.487
IF	7	1	3.19901

Table III-2 The noise breakdown. LO frequency is 247 GHz. IF frequency bandwidth is 4–8 GHz.

Figure III-19 An Allan Variance curve measured at 241 GHz.

NOISE BREAKDOWN			
Element	Tnoise, K	Gain	Tinput, K
Beam Splitter	0.5926	0.9980	0.5926
Window	0.5	1	0.501
SIS	11.4293	3.10567	11.4522
IF	7	1	2.25846

Table III-3 The noise breakdown. LO frequency is 255 GHz. IF frequency bandwidth is 4–8 GHz.

III.7 CONCLUSION

We have designed a DSB SIS mixer for the frequency range of 211–275 GHz. The SIS mixer was manufactured and tested. The uncorrected DSB receiver noise temperature was measured to be 16 ± 1.5 K, on average, over the IF 5–8 GHz range, which is close to the fundamental quantum limit of $hf/(2k_b)$ at 255 GHz of LO frequency.

The receiver noise breakdown analysis showed that SIS mixer contribution was ~ 2.77 and ~ 1.87 $hf/2k$, corresponding to 247 GHz and 255 GHz of LO range, while the receiver was confirmed to be linear and stable. The measured DSB noise temperature, averaged over the 4–8 GHz IF frequency range, was lower than 30 K over 227–270 GHz. Further development is needed to extend the mixer IF range to 4–12 GHz.

CHAPTER 4

IV A 600–720 GHz frequency range receiver design

The information presented in this chapter describes the current ALMA band 9 mixer chip design and typical measured mixer response curves. A full 3D EM wave calculation of Nb-based superconducting structures was performed and compared with measurement results, which validate the chosen calculation method. Subsequently, this method was used to optimize superconducting structure parameters, taking into account requirements of the available fabrication process at Chalmers University (GUARD). It was found that a change only in the superconducting chip layout is sufficient.

IV.1 ALMA BAND 9 MIXER DESIGN

ALMA band 9 covers an RF frequency range of 602–720 GHz and utilizes superconducting SIS junctions based on DSB type SIS mixers. The cartridge receiver design and performance has been described in detail [50]. The ALMA band 9 mixer block is a waveguide type DSB mixer that contains a corrugated feed horn and a waveguide back-short cavity integrated into a backpiece. The superconducting SIS junction of the mixer is integrated on a quartz substrate with dimensions of 2 x 0.040 x 0.080 mm and has Nb superconducting tuning and filter elements based on Nb/SiO₂/Nb superconductor/dielectric/superconductor three-layer structure with Nb/AlO_x/Nb or Nb/AlN/Nb SIS junctions integrated into a SiO₂ layer between the top and bottom Nb electrodes. SIS junctions with AlO_x tunnel barrier and an R_nA product of 25 Ohm/μm² have been used for the first 8 preproduction series of ALMA band 9 receivers. However, the low current density of the AlO_x barrier junction limits the RF and IF bandwidth of the SIS mixer, and also leads to increased matching losses in Nb tuning structures because the ALMA band 9 top frequency is close to the gap frequency of superconducting material [27]. For this reason, higher current density in AlN tunnel barrier SIS junctions were successfully developed at TU Delft and implemented into the design. AlN SIS barrier design had a superior RF and IF bandwidth and better noise performance. Thus, it was implemented for all subsequent

receivers developed for ALMA band 9. All receivers fulfilled ALMA noise, IF, and RF bandwidth and stability specifications.

IV.2 TYPICAL ALMA BAND 9 FTS MEASUREMENTS

To proceed with the re-optimization of ALMA band 9 design, we needed to verify the calculation method by comparing a typical ALMA band 9 direct detection response, measured by means of Fourier Transform Spectrometer, with calculated results. Figure IV-1 shows the typical FTS response measured for several SIS mixers used in ALMA band 9 cartridge receivers.

In fact, several superconducting chip layouts were used throughout ALMA band 9 production. The most commonly used and best performing design type is presented in Figure IV-1. We subsequently used the solid curve in Figure IV-1 as the reference performance comparator. We observed that production parameter deviation and mounting uncertainties caused slight differences in FTS response between different implementations of SIS mixers (Figure IV-1).

Figure IV-1 Summary of FTS results of representative series of ALMA band 9 cartridges (dashed lines) and average FTS to be used for comparison (solid line).

Due to presence of superconducting gap frequency in ALMA band 9, additional parameters must be considered in comparison with lower frequency Nb-based SIS mixers. Therefore, we show magnetic field response dependence of an FTS spectrum of a typical ALMA band 9 mixer in Figure IV-2.

Figure IV-2 FTS response of one of the ALMA band 9 mixers as a function of magnetic field current (see figure legend). Clear magnetic field dependence of frequency response is visible.

This effect has been discovered during ALMA band 9 development and has been reported and discussed in detail [32]. The main reason for magnetic field dependence is the change in the effective superconducting gap value in the bottom superconducting layer of the microstrip-based tuning structure; this is due to proximity effect. Proximity effect seems to be caused by normal metal/low T_c layers in the vicinity of an Nb layer, due to anodization and Nb/AlN/Nb tri-layer etch. This effect will not be considered for the optimization and modeling below, and low magnetic field conditions will always be assumed.

IV.3 ALMA BAND 9 MIXER 3D LAYOUT AND MODELING APPROACH

The ALMA band 9 mixer 3D layout is shown in Figure IV-3. All geometrical dimensions were taken from design drawings and correspond to the mixer design performance presented in Figure IV-1. The entire structure, including

the superconducting Nb layer, was modeled using the CST microwave studio simulator. The waveguide was modeled as a perfect electrical conductor and the chip layer parameters are summarized in Table IV-1. The silicon dioxide of the fused quartz substrate was modeled as a lossless material with $\epsilon_r=3.8$.

Figure IV-3 ALMA band 9 mixer design as modelled in CST Microwave studio. Left: 3D representation where signal comes to the waveguide from the direction of the viewer. The SiO_2 substrate with filter structure, waveguide probe, and integrated tuning structure is shown in green. The vacuum-filled space of the waveguide is shaded in grey. Right: Top view of the superconducting chip shown at scale.

name	Layer thickness, μm	comments
substrate	40	fused Quartz; $\epsilon_{ps}=3,8$; no Loss
bottom wiring	0,2	Nb by MB($V_{gap}=2.75\text{ mV}$; $\sigma_N=2.21 \cdot 10^8\text{ S}$)
dielectric	0,25	SiO_2 ; $\epsilon_{ps}=3,8$ $\tan\Delta=0,004$ at 500 GHz
top wiring	0,4	Nb; by MB($V_{gap}=2.75\text{ mV}$; $\sigma_N=2.21 \cdot 10^8\text{ S}$)

Table IV-1 Parameters of chip layers as used in MWS CST simulations.

The Nb superconducting layers were modeled by introducing surface impedance into CST MWS simulation.

The frequency-dependent surface conductivity of Nb was calculated using the Mattis Bardeen (MB) theory [13], referred in the text and in Table IV-2 as MB. Typical values for Nb film (see Table IV-1), which were verified by additional FTS surface impedance measurements using FTS, are used throughout the calculations. A typical surface impedance curve for Nb is shown in Figure IV-4. This allows consideration of both 3D distribution of electric and magnetic fields in the structure, and the frequency-dependent losses and kinetic inductance of superconducting materials. The SIS junction was modeled as a lumped element port with certain capacitance and inductance, as calculated from the SIS junction area and $R_n A$ value. The specific capacitance of a SIS junction was assumed to be $0.85 \text{ pF}/\mu\text{m}^2$, the $R_n A$ value is $10 \text{ Ohm}/\mu\text{m}^2$, and a typical junction area is $0.6 \mu\text{m}^2$. This calculation method represents the most accurate modeling of the ALMA band 9 structure in comparison to prior design work.

Figure IV-4 Complex surface impedance [10, 11] of Nb film calculated using the Mattis Bardeen theory for typical Nb film and used in CST MWS for modelling of the Nb component of the structure. $T_c \text{ Nb} \sim 9.1 \text{ K}$.

Critical dimension	Specification of the design rule
SIS junction pattern: Junction size Junction width	min $3\mu\text{m}^2$ (standard process) min $1.3\mu\text{m}$ (standard process)
Other patterns: Spacing Width Overlap/spacing of layers	min $4.0\mu\text{m}$ min $10\mu\text{m}$ (lines longer than $200\mu\text{m}$) min $20\mu\text{m}$ (lines longer than $1000\mu\text{m}$) min $4.0\mu\text{m}$
Layer alignment accuracy	min $1\mu\text{m}$

Table IV-2 Minimal dimensions and spacing specified by GUARD for SIS process rules.

IV.4 CALCULATION RESULT AND CONCLUSION

The FTS response of a typical ALMA band 9 SIS mixer was calculated and is shown in Figure IV-5, compared with a typical measured FTS curve. Several SIS junction RF impedance ($R+jX$) values were used to evaluate stability of the design with respect to wafer-to-wafer current density changes due to production tolerances.

Figure IV-5 Calculated and measured reference ALMA band 9 mixer FTS response. $A_{\text{sis}}=0.65\mu\text{m}^2$; the substrate thickness is $40\mu\text{m}$.

A very good description of measured response by the model is achieved when taking into account that: a) corrugated horn matching section is not included in the calculation and may change the in-band response and b) complex superconducting layer gap structure, as indicated indirectly in magnetic dependence of FTS, is not considered. The presence of b) may lead to additional RF losses at the high frequency limit of the ALMA band 9 receiver. The IF match between the SIS mixer in operating bias point and the 50 Ohm load, representing IF isolator/cryogenic amplifier input, is shown in Figure IV-6. The IF match is defined as power matching between two ports: the SIS output port and the IF circuit output port. In fact, the entire $|S_{31}|$ (see Figure IV-7) circuit parameter was calculated for an IF circuit starting from the SIS junction, passing through parasitic impedances of tuner and environment and toward the IF transmission line connected to a 50 Ohm port. An intrinsic IF bandwidth of more than 30 GHz can be achieved with an appropriate cryogenic IF amplifier, with total match close to 1.

Figure IV-6 IF match between 50 Ohm impedance and SIS junction at operating point. An entire on-chip layout was included for consideration.

Figure IV-7 Equivalent circuit representation of calculated design. P1 is waveguide RF port; P2 is SIS junction port; P3 is IF output port.

This fact confirms competitive noise performance, demonstrated by this design in ALMA band 9 production. Technically, that is achieved due to a very low parasitic capacitance, which is mainly dominated by the SIS junction area and the overlap area of the integrated tuning circuit. Avoiding parallel tuning structures like a stub+tuner leads to reduced parasitic capacitance for IF. Based on comparison of the current ALMA mixer FTS response and the model calculation, we conclude that the model is suitable for re-optimization of the chosen SIS technology.

IV.5 OPTIMIZATION FOR CHOSEN (GARD) TECHNOLOGY REQUIREMENT

This section summarizes the optimization process for requirements of GARD SIS technology. Optimization was limited to variations of the geometry of superconducting layers on top of a quartz substrate, and mainly the geometry integrated with superconducting microstrip matching and tuning structure. The junction area was found to be the most critical difference, and optimization was carried out for junction areas of $1 \mu\text{m}^2$ and $2 \mu\text{m}^2$, which are areas with a typical $R_n A$ value of $10 \text{ Ohm}/\mu\text{m}^2$ and a specific capacitance of $0.85 \text{ pF}/\mu\text{m}^2$.

IV.5.1 GARD SIS FABRICATION TECHNOLOGY REQUIREMENTS

The main difference between the requirements used for the band 9 SIS process and those used in GARD is in the junction area. Since optical lithography is used in the GARD SIS fabrication process (which is a requirement of SIS junction area, as follows from the GARD SIS design rules: min $3 \mu\text{m}^2$ as baseline and $2 \mu\text{m}^2$ on the best effort basis), further miniaturization of the SIS junction is likely possible; however, this is a subject of process development. The typical achievable R_nA values considered in this project are in the range of 10–20 $\Omega\mu\text{m}^2$. Other process parameters, such as layer thicknesses and layer geometry definitions, are very similar between laboratories because similar tools and methods are utilized for production.

IV.5.2 OPTIMIZATION PARAMETERS AND RESULTS

To provide cost effective and efficient demonstration of GARD SIS technology for ALMA band 9, we have decided to leave all waveguide dimensions constrained and only modify the geometrical dimensions of an on-chip integrated tuner. The design type of a single SIS junction with an end-loaded stub, followed by a two-stage microstrip transformer and a bow tie waveguide probe remain fixed as well. Since the main challenge of re-optimization is to increase junction area toward requirement of the technology, we decided to retain a single junction design, as dual junction would reduce the normal resistance even further compared to the already low normal resistance of a large current density single SIS device.

Single junction design also allows avoidance of SQUID-like effects with a very fine structure of SIS critical current vs. magnetic field dependence, which makes it difficult to suppress Josephson current in the SIS mixer. This suppression is a requirement for ALMA band 9 frequencies to achieve stable low noise SIS mixer operation. Other chip parameters remain the same as those used to evaluate a typical ALMA band 9 mixer design. The optimization was completed using full 3D EM analysis in the Microwave Studio package.

The 3D model includes: the model of a dielectric material with losses; the model of a superconductor metal structure, with kinetic inductance and losses included; the model of the SIS junction, represented by a lumped element port with the real part of SIS junction impedance; and a lumped element capacitor representing SIS junction capacitance. The geometrical parameters of all defined microstrip transformers, such as the waveguide probe and RF band rejecting filters and the waveguide structures, were included in the 3D model.

The absolute match $|S_{21}|$ between the waveguide input port and the SIS junction port was calculated over the required frequency range to evaluate performance. First, an initial set of parameters for nominal design was evaluated to check the model. Then, the SIS junction area was increased stepwise by $0.5 > 1 > 2 > 3 \text{ } \mu\text{m}^2$. In each step, a parametric search (gradient optimization) was performed to maximize the $|S_{21}|$ parameter in the required frequency range. The output from this set of parameters was plotted on graphs similar to those included in this thesis. For this optimization we did not change the current principle of junction matching, i.e., the end-loaded stub with a two-stage transformer between the SIS junction and the waveguide probe; this design also works well with a small junction area.

The junction capacitance is responsible for limiting IF performance in the absence of an external IF tuning circuit. If we added an external tuning circuit and/or an RF mechanical structure, it would lead to a complete redesign of the mixer, which was not our intention. Our goal was design of a drop-in replacement for our current mixer block. The MWS user manual can be referenced for underlying calculation principles. As discussed before, we aim to re-optimize the current successful ALMA band 9 SIS mixer design for a much larger SIS junction area. We completed optimization for $1 \text{ } \mu\text{m}^2$ and $2 \text{ } \mu\text{m}^2$ junction areas (process best-effort). We were able to find suitable layout geometries to achieve a good match throughout the RF band.

The optimized mixer layout for both junction areas is shown in Figure IV-8. The tuning circuit overlap area increases with increased SIS junction area, which we will see back in the IF performance of the designs. Figure IV-7 presents the equivalent circuit layout to the extent of the 3D distributed elements, which can be mapped to a lumped element/stripline representation. As in the scheme, the absolute RF coupling can be calculated as $|S_{21}|$ and the IF coupling as $|S_{23}|$, while the RF impedance and IF impedance are used for the port 3 reference impedance calculation.

Figure IV-8 SIS junction on-chip layout: a) central part of the chip optimized for a $1 \mu\text{m}^2$ junction area, b) central part of the chip optimized for a $3 \mu\text{m}^2$ junction area, and c) general layout of the circuit in the waveguide, which is similar to baseline ALMA band 9 design. Rose color is the top Nb layer, violet color is the bottom Nb layer. The quartz substrate is depicted in green.

Figure IV-9 SIS mixer input match calculated for the optimum layout optimized around a $2 \mu\text{m}^2$ junction area. Junction area was varied as indicated in the figure legend. The dielectric thickness is $0.25 \mu\text{m}$ and the SIS critical current density is defined as 20 kA/cm^2 .

The C_{sis} is calculated from the specific capacitance and the junction area. The RF impedance is calculated from the $R_n A$ product and the junction area. The other parameters are related in the 3D model. The calculated input match for layout configurations in Figure IV-8 are shown in Figure IV-10 and Figure IV-11.

Figure IV-10 SIS mixer input match calculated for the optimum layout optimized around a $1 \mu\text{m}^2$ junction area. Junction area was varied as indicated in the figure legend. The dielectric thickness is $0.25 \mu\text{m}$ and the SIS critical current density is defined as 20 kA/cm^2 .

Figure IV-11 SIS mixer input match calculated for the optimum layout optimized around a $3 \mu\text{m}^2$ junction area. Junction area was varied as indicated in the figure legend. The dielectric thickness is $0.25 \mu\text{m}$ and the SIS critical current density is defined as 20 kA/cm^2 .

Good band coverage can be achieved for both junction sizes. The input match was also calculated for $\pm 20\%$ of the junction area for a small junction and $\pm 10\%$ for a large junction. The linear dimension tolerances are independent of the junction area. Both design configurations are robust with respect to

junction area variation, which is the main uncertainty in the SIS junction production process. The overall matching curve shape is very similar for both junction areas. The calculated output IF match for both design configurations is shown in Figure IV-12.

Figure IV-12 SIS mixer output match calculated for optimum layouts: original 0.5 μm^2 area, 1 μm^2 junction area, 2 μm^2 junction area, and 3 μm^2 junction area.

Compared with the baseline ALMA band 9 design, IF performance is degraded in terms of the 3dB IF bandwidth and in terms of the absolute match figures. ALMA specifications on the 4–12 GHz IF bandwidth are still likely to be met with a sufficient margin, with the noise performance expected to degrade by 10 K for a smaller junction and by 20 K for a larger one, on top of typical 100 K DSB noise. These estimates were calculated for an amplifier noise temperature of 5 K and a mixer conversion gain of 3dB. The increased noise temperature still meets ALMA band 9 specifications.

The input noise temperature can be presented as a sum of the mixer noise temperature (T_{mix}) and an IF amplifier contribution ($T_{\text{amp}}/L_{\text{mix}}$), where T_{amp} is the IF amplifier noise temperature and L_{mix} is the mixer conversion gain. Although the exact value of L_{mix} can only be calculated from Tucker theory, which requires much more effort, the mixer output match can be estimated from microwave theory as the S23 parameter in our model. When the S23 parameter changes by 3dB, as is evident in the figures, the IF amplifier contribution will show a corresponding increase. The nominal design mixer conversion gain is estimated at -6dB, with an IF amplifier contribution of 20 K. If that gain is decreased by 3dB, the amplifier contribution increases by 10 K. Those values are estimates, taking into

account the assumptions given above and also dependent on IF and LO frequency. Table IV-3 summarizes the parasitic IF capacitance of the tuning structure compared with the capacitance of the SIS junction itself. As junction area increases, the parasitic IF capacitance also increases, limiting IF bandwidth according to the Fano-Bode theorem. This effect is clearly visible in the IF match calculation in Figure IV-12. The calculated SIS junction embedding impedance is shown in Figure IV-14. For optimization of both designs, the embedding impedance curve occupies the stable SIS operation region, which manifests absence of a potential negative dynamic resistance of the SIS junction at the operating point.

name	$A_{\text{sis}}, \mu\text{m}^2$	$A_{\text{idle}}, \mu\text{m}^2$	$C(\text{idle}-A_{\text{sis}}), \text{fF}$	$C_{\text{sis}}, \text{fF}$
TT 3	3	1198	160,828	255
TT 2	2	1160	155,848	170
TT 1	1	1578	212,239	85
b9 old	0,5	604,2	81,2483	30

Table IV-3 Wiring overlap and capacitance for different chip designs. The TTs are new designs and "b9 old" is one of the prior designs.

Figure IV-14 SIS mixer embedding impedance: a) optimized around a $1 \mu\text{m}^2$ junction area, b) optimized around a $3 \mu\text{m}^2$ junction area.

IV.6 WIDE BAND IF PERFORMANCE

To test the numerically estimated wide IF band, we ordered new cold low noise amplifiers (LNAs; Yebes Y214 LNA, 2–20 GHz of frequency range) and warm LNAs from B&Z (2–26.5 GHz of frequency range). We then performed a noise temperature test for a band 9 mixer at four different LO frequencies (621, 648, 675, and 702 GHz). In Figure IV-13, looking at the lowest parts of the standing wave pattern (where the coupling should be optimal), it seems that the useful band seen in the noise temperature extends up to approximately 24 GHz. These results were obtained even though the experiment was carried out with a cold LNA specified for up to 20 GHz. The LO frequency was confirmed to have been set correctly.

Figure IV-13 Mixer noise temperature with the Yebes 2–20 GHz cold LNA and the B&Z 2–26.5 warm LNA.

IV.7 CONCLUSION

We have created a model of an existing ALMA band 9 design produced at NOVA/TU Delft. The model includes the losses in Nb film and a full 3D structure. The modeling results were close to the measured FTS response of ALMA band 9 SIS mixers, which confirms its validity for further optimization. We have used the model to optimize mixer chip layout to accommodate the main GARD SIS process requirement: a junction area of $3 \mu\text{m}^2$. To show the design evolution, we performed optimization for $1 \mu\text{m}^2$ and $2 \mu\text{m}^2$ areas as well. All optimizations yielded layouts of SIS chips that perform sufficiently and demonstrate GARD SIS technology (see Table IV-2) in ALMA band 9.

Our interim conclusion is that the SIS devices have an IF coupling that extends up to 24 GHz, as predicted by the simulations. However, it is also clear that, without proper input matching of the LNAs, this range will still not be available for practical use. Another obstacle is the absence of a built-in bias tee, because this may well be largely responsible for the standing waves.

CHAPTER 5

V Flux-flow Josephson oscillator as the broadband tunable terahertz source to open space

The flux-flow oscillator (FFO) based on a long Josephson junction was implemented as a broadband tunable terahertz (THz) source to open space. For this purpose, the transmitting slot antenna was coupled to the oscillator. Additionally, an elliptical lens with a diameter of 10 mm was matched to the antenna, forming a narrow output beam of the THz emission. Two designs for the antenna, integrated with the oscillator and developed for operation at different frequency ranges of 0.32–0.55 THz and 0.4–0.7 THz, have been investigated. The FFO was phase locked to an external reference oscillator by utilizing a harmonic mixer. Its linewidth in the phase locking regime is determined by the phase noise of the reference oscillator and the number of harmonics used; linewidth as low as 0.1 MHz was measured by an external receiver (Superconducting Integrated receiver SIR). A free-running FFO linewidth from about 2 MHz to several MHz, depending on the operating point, was obtained. Output emission to open space was measured by a superconducting integrated spectrometer located in a separate cryostat. The FFO operation as an external source with the achieved emission power and spectral characteristics has demonstrated its applicability for different tasks and purposes where tunable THz sources are required.

V.1 INTRODUCTION

Wideband tunable sources of EM emission in the terahertz (THz) region are of great importance, as they are required in many fields, such as radio physics, astronomy and spectroscopy, medicine, and security systems. The FFO based on a long Josephson junction with a length much greater than the Josephson penetration depth ($l \gg \lambda_J$) is a promising THz source due to its excellent operating bandwidth, which can be up to 100% of the central frequency. The FFO was studied experimentally [26], [70]–[76] and theoretically [11], [77]–[82], and was finally utilized as the LO of the heterodyne THz receiver in the real flight mission, TELIS [83], [84] for atmospheric research. In this mission, as well as in other studies of the FFO and its applications, it was used as an on-chip planar source. It has never been used as a source to open space.

The output frequency f is strictly defined by the Josephson equation $hf = 2eV_{DC}$, where h is Planck's constant and V_{DC} is the voltage at the junction. An oscillator based on a Nb/AlN/NbN tri-layer with dimensions of $16 \times 400 \text{ } \mu\text{m}^2$ has an operating range from 250 to 750 GHz [26], corresponding to V_{DC} in the region from 0.52 mV to 1.55 mV. The typical free-running spectral linewidth of the oscillator is of the order of 1 MHz, and the line shape is Lorentzian.[72]–[74], [76]. For narrowing the actual spectral line down to a value less than 100 kHz, and for collecting in the peak up to 97% of the spectral power (this ratio is called the Spectral ratio), a phase-locking loop (PLL) to the reference oscillator is commonly used in Institute of Radiorngineering and Electronics of Russian academy of science (IRE RAS). The upper border of the operating frequency range is limited by the gap frequency $2\Delta/h$, which is about 700 GHz for all-Nb junctions, about 1.4 THz for all-NbN junctions, and 1.2 THz for all-NbTiN junctions. However, the actual frequency limit of Nb/NbN junctions is generally about 750–850 GHz [26], [75] due to the high surface resistance of NbN and therefore sharp increase in losses above 750 GHz. These surface losses lead to decreased emitted power due to a damping of the wave propagating along both the tunnel junction and the output transmitting line, as well as linewidth increase due to a higher differential resistance. For higher frequencies, NbTiN-based junctions are preferable [85] due to much lower losses at frequencies above 750 GHz, where the upper frequency limit for operation is predicted to be as high as 1–1.2 THz [86], [87]. It should be noted that the emission frequency range available for FFO applications is also defined by matching structures and other circuit elements with typically narrower bandwidths. The emission power of the FFO ranges from fractions of 1 μW to a few microwatts in magnitude [70]. This is not large, but it is quite enough for applications such as heterodyne receiving and spectroscopy. Furthermore, the power can be increased by using a larger current density in the junctions [75].

V.2 CONCEPT AND NUMERICAL SIMULATIONS

The main goal of this project is on-chip integration of the FFO with the transmitting double-slot lens antenna to provide FFO emission to open space by using matched transmitting lines. The concept is shown in Figure V-1. The FFO is coupled to the slot antenna via the output microstrip line, as illustrated in Figure V-1(a). Tuning the FFO frequency is equivalent to setting the FFO voltage V_{DC} with the coefficient 483.6 GHz/mV. To set V_{DC} , two currents are applied: the control line current I_{CL} is used for supplying the magnetic field and can typically be set from 0 mA up to 100 mA, and the bias current I_{BIAS} is applied via the FFO tunnel junction. The typical operation bias is from 10 to 30 mA for current densities J_c in the range 3–10 kA/cm^2 . An impedance

transformer is used to match the low output impedance of the FFO ($\sim 0.5 \Omega$) and the much higher input impedance of the slot antenna, which is on the order of tens of Ohms. Some part of the FFO power is branched to the harmonic mixer (HM), based on the SIS junction and used in the feedback loop with the oscillator for phase locking by means of the PLL system [72], [76].

Figure V-1 (a) Layout of the planar microcircuit containing FFO, double-slot antenna, harmonic mixer, and the coupling structures between the elements. Both tunnel junctions, the FFO, and the HM are to be fabricated using Nb/AlN/NbN or Nb/AlOx/Nb tri-layers in one technological process; top and bottom electrodes constitute the microstrip lines with an SiO_2 insulator layer. The full dimensions of the slot structure shown are $242 \times 90 \mu\text{m}^2$. (b) Scheme of the chip with the microcircuit shown in (a) placed at the far focus of the lens and inside the magnetic shield.

Another impedance transformer is used to couple the FFO emission to the HM, with an impedance of several ohms. Different FFO operating points at the same frequency (voltage) can be chosen to optimize the spectral characteristics and the pumping of the HM. The double-slot antenna is made using a thin film of Nb with a thickness of $\sim 200 \text{ nm}$, which is also used for

the bottom electrode of the FFO, the microstrip lines, and the HM. A SiO₂ insulator with a thickness of 400 nm is used between the top and bottom electrodes. The dimensions of the FFO are $400 \times 16 \mu\text{m}^2$, as was successfully used in previous studies [26], [83], [84], and the HM area is approximately $1\text{--}1.5 \mu\text{m}^2$. The integrated circuit is fabricated on a Si substrate with a thickness of $D_{\text{Si}} = 0.535 \text{ mm}$ and a dielectric constant of $\epsilon = 11.7$, so the wavelengths in the circuit are $(\epsilon)^{0.5} = 3.42$ times smaller than those in a vacuum. The length of the antenna slots depends on the frequency range and is about 0.36λ or $1.25\lambda_{\text{Si}}$, where λ and λ_{Si} are the wavelengths in a vacuum and in silicon, respectively, for the central frequency range. The distance between the slots is about 0.1λ or $0.4\lambda_{\text{Si}}$. In Figure V-1(a), the specific design for the 330–570 GHz range (central frequency 450 GHz, $\lambda_{\text{Si}} = 195 \mu\text{m}$) is shown. The length of the antenna dipoles is $242 \mu\text{m}$, the width of the slots is $15 \mu\text{m}$, and the distance between the centers of the slots is $75 \mu\text{m}$. This type of antenna has been used previously as the receiving antenna coupled to a SIS detector in the receiver, at 100 GHz, 246 GHz, and 500 GHz [88]. We use this double-slot antenna because of the possibility of locating it at the minimum distance from the FFO and fabricating it in the same technological cycle with the FFO and HM tunnel junctions. The chip with the FFO, antenna, and HM integrated circuit is mounted on the back surface of an extended semielliptical lens with a diameter of 10 mm [see Figure V-1(b)] and at a far focus distant from the geometrical center of the lens by $D = 1.529 \text{ mm}$. The extension of the lens is $D - D_{\text{Si}} = 0.994 \text{ mm}$ so that the center of the antenna is located exactly at the far focus of the lens. The lens and the substrate of the chip are both made of Si to avoid beam refractions on the chip-lens border. It was shown that this type of lens with a matched antenna mounted at its focus can be effectively used over a wide frequency range of 0.1–1 THz without changing the lens geometry [75]. Since the FFO is extremely sensitive to external magnetic fields, the circuit is located inside a magnetic shield made of permalloy and copper with a lead-plated layer. The shield is open on one side. The shield with the sample is then placed inside a liquid helium (LHe) cryostat with a bath temperature T_{bath} of 4.2 K and a Mylar output window at a room temperature stage. Additional Gortex windows are used at 4.2 K and 77 K stages to operate as IR filters, and both Mylar and Gortex windows are transparent in the THz region. For developing structure designs for different frequency ranges, numerical simulations were performed using specialized software for 3D microwave modeling and optimization. The main task for modeling is the coupling, at THz frequencies, of the oscillator with a low output impedance to the lens antenna with a higher input impedance and forming the required beam pattern for applications.

Figure V-2 Input impedance Z of antenna designs #1 and #2: the real part $\text{Re } Z$, the imaginary part $\text{Im } Z$, and the full impedance $|Z|$ are shown.

Figure V-3 FFO power absorbed by the harmonic mixer (dashed line), emitted to open space (solid line), and absorbed by the full output structure (dashed-dotted line), normalized to the total output power for (a) design #1 and (b) design #2.

A simultaneous modeling task is the coupling to the HM based on the SIS junction with some intermediate impedance (higher than the FFO's and lower than the antenna's) in the maximally similar frequency range as for the FFO-to-antenna coupling. Matching to the antenna should be as high as possible, while matching to the HM should be just enough for proper PLL operation and should not take away much power. Pumping of the HM at ~5%–10% of the current jump at the SIS-mixer gap voltage is often enough for HM and PLL operations [26], [73], [75]. Both couplings should be as wideband as possible for an antenna and transmitting lines of this type. To account for the superconducting state of the antenna and microstrips, the London penetration depth for thin Nb films $\lambda_L = 85$ nm is used, and the SiO_2 dielectric constant $\epsilon = 4.2$ was taken for thin films. Calculating the real FFO output impedance, which is frequency-dependent, is an extremely complicated task due to self-coupling and standing-wave effects [82], [89], so the impedance was taken to be constant and equal to 0.5Ω . It is known from antenna engineering and also verified during our simulations that the slot antenna bandwidth is usually not more than $\sim 0.5f_0$, where f_0 is the central frequency. At the same time, the FFO operation bandwidth is up to f_0 . Hence, we developed two designs to cover the operating range 0.4–0.7 THz. As a result, two designs were numerically simulated: design #1 for $f_0 = 450$ GHz [as depicted in Figure V-1(a)] and design #2 for $f_0 = 600$ GHz. The calculated input impedances of the antenna at the point of connection to the microstrip line for designs #1 and #2 are presented in Figure V-2. The full impedance dependence is rather flat: it varies between $\sim 30 \Omega$ and $\sim 40 \Omega$ in the target frequency regions. The numerical results for the emission to open space and the coupling with the HM for both designs are illustrated in Figure V-3. The calculations were performed as follows: the emitting edge of the FFO was simulated as the input port (FFO port) with an impedance of 0.5Ω , and the HM was simulated as a port (HM port) with an impedance equal to the normal resistance of the junction $R_n = 17.85 \Omega$, corresponding to an area of the junction $A_{HM} = 1.4 \mu\text{m}^2$, based on a tri-layer with a normal-state resistance area product $R_n \times A = 25 \Omega \mu\text{m}^2$. The capacitance of the HM was taken into account in the model by embedding a capacitor according to the specific capacitance of the tunnel tri-layer taken as $85 \text{ fF}/\mu\text{m}^2$. The microstrip line after the HM contains the rf short based on a radial stub, and then transforms to a CPW line for proper connection to the contact pads. The simulation program calculates the power absorbed by the HM port from the transmitting line PHM at each frequency (dashed lines in Figure V-3) and the power returned to the input FFO port Preturn. Hence, the power coupled to the whole integrated output structure is $P_{\text{couple}} = 1 - \text{Preturn}$ (dashed-dotted line) and the power emitted to open space is $P_{\text{emit}} = 1 - \text{Preturn} - \text{PHM} = P_{\text{couple}} - \text{PHM}$ (solid line). All the power values are normalized to the FFO output power. Additional computation of the far-field distribution simulated

by the program showed that all the constituents of the FFO power were taken into consideration; in particular, only about 1%–2% of the power at operating frequencies passed out via the rf short. According to the numerical simulation, the operating frequency ranges of the antenna, evaluated as the ranges with emission above 0.7 of FFO total power, are 320–550 GHz for design #1 and 400–700 GHz for design #2. Thus, the two designs cover the range 0.32–0.7 THz. The MWS CST allows export of Near-Fields and then calculation of Far-Fields. Figure V-4 shows the antenna chip radiation gains for different frequencies and designs.

Figure V-4 (a) Far-Field Gains of A-350 (test design #0), A-450 (design #1), and A-600 (design #2) antenna designs at the selected frequency and (b) a set of Far-Field Gains for the A-450 antenna design in the frequency range from 350 to 650 GHz with 50 GHz increments.

V.3 EXPERIMENT

V.3.1 SAMPLE PREPARATION

The technology used by our group for fabrication of samples containing integrated circuits with all-Nb and Nb/NbN tunnel junctions of micrometer and submicrometer dimensions has been described elsewhere [31], [64]. The key technological processes are electron-beam lithography for mask preparation, magnetron sputtering, and "lift-off" lithography. In this study, $4 \times 4\text{-mm}^2$ chips with designs #1 and #2 containing the FFO and HM based on Nb/AlN/NbN tri-layers were fabricated on a silicon substrate. According to DC testing, the tunnel junctions of the batch have a normal-state resistance-area product $R_n \times A \sim 18 \Omega \mu\text{m}^2$, which corresponds to a current density of $J_c \sim 12 \text{ kA/cm}^2$. The gap voltage V_g is about 3.45–3.55 mV and the quality factor, defined as the sub-gap resistance to normal-state resistance ratio R_j/R_n , is about 30, which is a good result for the NbN technique.

For testing at THz frequencies, the chip is stuck to the lens as shown in Figure V-1(b) and bonded to the bias plate by aluminum wires utilizing ultrasonic bonding. The six contact pads are used to fully control the FFO with a four-point biasing current-control scheme ($+I_{\text{BIAS}}$, $+V_{\text{BIAS}}$, $-I_{\text{BIAS}}$, $-V_{\text{BIAS}}$) and an additional two-point control line scheme for applying the magnetic field ($+I_{\text{CL}}$, I_{CL}). The HM is controlled via a four-point biasing voltage-control scheme. Two additional transmitting lines to the chip are used: one for supplying the reference rf signal of about 20 GHz to the HM and for measuring the output intermediate frequency (IF) signal, and the other for applying the feedback signal to the FFO from the PLL system. The final cryogenic module with the chip on the lens, bias plate with DC, and rf wiring inside the magnetic shield, is mounted inside the cryostat opposite the output window.

V.3.2 MEASUREMENT SETUP

A complex block diagram of the setup for measuring the FFO emission to open space is shown in Figure V-5. Two liquid helium (LHe) cryostats are used for the oscillator and the receiver: the left part of the diagram shows the "emitting" cryostat (#1) and the right part shows the "receiving" cryostat (#2). The schematics of the FFO microcircuit design are discussed in section V.2. Signals from the FFO and n^{th} harmonics from the synthesizer at 19–21 GHz are mixed by the HM. The IF signal from the HM in the range 0–800 MHz is amplified first by the cryo-amplifier based on a high electron mobility transistor (HEMT) and then by the room temperature amplifier. It is then branched to a PLL system and to spectrum analyzer (SA) #1 for direct reading of the FFO spectra.

Figure V-5 Block diagram of the experimental setup for studying the FFO emission to open space using a THz spectrometer based on a superconducting integrated receiver.

The main part of the FFO output power is coupled to the slot antenna and radiated to open space, then directed to the THz spectrometer based on the superconducting integrated receiver (SIR). Operation of the SIR (right side of diagram in Figure V-5) has been discussed previously [73], [90]. Two FFOs are used in the experimental setup: one (FFO #1) is the object of the study and the other (FFO #2) is used as the local oscillator (LO) of the SIR. The SIR IF range is 4–8 GHz and the FFO frequency is strictly defined by the Josephson voltage. Therefore, for the SIR to receive the emitting signal, the voltages of the FFOs should be at a distance from one another within the range of 8.3–16.5 μV . The backend of the SIR is SA #2, which immediately shows the receiving signal converted down to the IF range. Since the gains of the amplifiers are different in the IF chain of the HM in cryostat #1 and in the IF chain of the SIS mixer in cryostat #2, the spectra detected by spectrum analyzers #1 and #2 should not be compared by power. Nevertheless, they can be compared by shape, taking into account that the spectrum detected by SA #1 is the convolution of FFO #1 signal and the n^{th} harmonic of the synthesizer (assumed to be a δ -function), while the spectrum detected by SA #2 is the convolution of the two FFO signals, at least one of which (#2) should be phase locked during the measurements, forming a signal close to a δ -function.

V.3.3 RESULTS AND DISCUSSION

A set of the current-voltage curves (IVCs) of fabricated FFOs measured for different magnetic fields produced by I_{cl} in the range 20–90 mA, with a step size of 2 mA (36 curves altogether), is presented in Figure V-6(a). The bottom axis, frequency, strictly corresponds to the top axis, voltage. The set is quite similar to the IVCs previously obtained for Nb/AlN/NbN-based FFOs [26], except for the slope of the IVCs (high differential resistance) and therefore the absence of clearly expressed Fiske steps [26], [73], as well as

the gap in operating points at currents from 13 mA to 22 mA between ~ 400 GHz and ~ 490 GHz. Both features are the consequence of disadvantages in the present control line design and the superconductor difference gaps, which could be redesigned in the future.

Figure V-6 (a) IVCs of the FFO with dimensions $400 \times 16 \mu\text{m}^2$ measured for different magnetic fields produced by the control line current with an ICL step of 2 mA. (b) Experimental results of HM pumping by the FFO power and calculation results of the FFO power absorbed by the HM port. The HM I_{pump} is normalized to the "current jump" at the SIS gap voltage, and equal to $206 \mu\text{A}$ for the HM in design #1 and $199 \mu\text{A}$ for the HM in design #2. The HM absorption power is normalized to the total FFO output power.

In addition, the J_c of the batch turned out to be higher than expected: 12 kA/cm^2 instead of $7\text{--}8.5 \text{ kA/cm}^2$. This led to a stronger slope in the IVCs and degradation of Fiske steps. Nevertheless, neither the slope nor the lack of operating points in some regions of the IVCs affects the fundamental Josephson oscillation or the frequency characteristics of the antenna design and the transmission lines. The pumping of the SIS-based HM by the FFO

power leads to the appearance of the quasiparticle step I_{pump} on the IVC of the HM, which is measured directly over the whole frequency range of the FFO operation. In Figure V-6(b), the experimental results for the frequency dependence of the normalized HM pumping current are presented together with the calculation results for the absorbed power for designs #1 and #2 [shown earlier by dashed lines in Figure V-3(a) and Figure V-3(b)]. The axes of frequency in (a) and (b) are exactly one above the other, for clarity. There is good agreement between the experimental data and the numerical simulations. The spectral characteristics of the FFO were measured both by SA #1 (on-chip emission) and by SA #2 (open-space emission detected by the SIR).

Typical spectra are shown in Figure V-7 for design #2 at 590 GHz. Free-running and phase-locked spectral lines measured in the upper sideband are presented. The linewidths of free-running spectra are similar (red and magenta lines). The IF power levels of spectral lines are quite different due to different gains of IF circuits, as mentioned in section V.3.2. The phase-locked line measured by SA #2 (gray line) seems to be wider than that measured by SA #1 (black line) due to the different resolution bandwidths (RBW) of the analyzers. The linewidth of the phase-locked spectra in Figure V-7 is completely defined by the RBW of SA (see, for example, figures shown in previous publications [72], [74]) and proved to be about 100 kHz when $\text{RBW} = 100 \text{ kHz}$ is used during the measurements. The actual linewidth of the phase-locked spectra is defined by the phase noise of the synthesizer used multiplied by n^2 , where n is the number of harmonics [72]. The spectral ratio defined for a phase-locked line as the percentage of the power concentrated in the peak to the total power is a little less in the case of measurement by the SIR, and due to the higher RBW of SA #2.

The two samples, design #1 and design #2, were carefully studied in their operating frequency ranges. The measurements shown in Figure V-7 were carried out at many different frequencies, and the results are summarized in Table V-1. The linewidth at many of the studied operating points is not as good as may be desired — typically above 5 MHz — due to the high J_c that leads to a strong slope in the FFO IVCs (high differential resistance). This issue can be solved simply by fabricating another batch with more appropriate tri-layer parameters. Therefore, such values should not be considered representative of the FFO. In many previous papers, a linewidth of about 1–2 MHz was obtained for the Fiske steps IVC region (lower than $V_{\text{gap}}/3$) and of about 4–7 MHz for the true flux-flow regime (above $V_{\text{gap}}/3$) [72], [74], [76], [83].

Figure V-7 Power spectra of the FFO at a frequency of 590 GHz for the sample with design #2 measured using the on-chip HM (SA #1)—free-running (red line) and phase-locked (black line) spectra, and measured using the SIR (SA #2)—free-running (magenta line) and phase-locked (gray line) spectra. The bottom and top axes of frequency represent the different IF ranges of 0–800 MHz for the on-chip HM IF and 4–8 GHz for the SIR IF. The span for both axes is the same, 100 MHz, and all the spectra are measured in the upper sideband. An estimated linewidth for the free-running spectra and a spectral ratio for the phase-locked spectra are shown near the curves. The resolution bandwidths of the analyzers are different: 1 MHz for SA #1 and 1.8 MHz for SA #2. For the spectra measured by SIR (magenta and gray lines), the LO of the SIR, FFO #2, is phase locked.

We emphasize that the main results of this research are the detection of the FFO emission to open space and demonstration of similar spectral properties for on-chip radiation and open-space radiation, regardless of the measurement technique. Therefore, the best results obtained for the FFO as an on-chip oscillator can be repeated for the FFO as an open-space oscillator. As for the frequency operating ranges, which were calculated to be 320–550 GHz for design #1 and 400–700 GHz for design #2 (using the level of 0.7 of total FFO power), the experimental results confirmed the numerical simulations (see Table V-1). The exception to this confirmation is the case of frequencies outside the SIR operating range of 450–700 GHz. Thus, the operating frequencies 320–450 GHz for design #1 and 400–450 GHz for design #2 could not be studied accurately by the SIR. Nevertheless, even outside the SIR operating range, the emission at some frequencies was

registered without stabilization of the SIR LO (FFO #2), e.g., at 387 GHz and 392 GHz for design #1 and at 720 GHz, 730 GHz, and 735 GHz for design #2. In these cases, the FFO spectral linewidth could still be measured using the on-chip HM, provided the oscillator could be phase locked.

A beam pattern for the lens antenna was not measured accurately because a complex optical experimental setup is required, and the pattern of the receiver must also be taken into account. We made a rough estimate for the width of the central beam lobe of about a few degrees by manually moving and rotating cryostat #1 relative to cryostat #2. Enlarging the distance between the cryostat windows from 6 cm to 30 cm had almost no influence on the receiving power, but rotating the output window by 1° – 2° from the central position (defined as the maximum power position) led to a sharp decrease in receiving power. This result is as expected, because the elliptical silicon lens used for the oscillator is the same as that used for the SIR, where the lens is coupled to a double-dipole receiving antenna. It is the lens that mostly defines the beam pattern of such lens antenna systems. The beam diagram for the SIR optics has been presented previously [73], [83] (Fig. 13 and Fig. 3, respectively, therein) for the samples for the TELIS flight mission. All the spectral measurements in this paper were made by manually tuning the relative positions of the “emitting” and “receiving” cryostats to obtain the maximum receiving signal.

The concept of integrating the FFO with the antenna on a single chip turned out to be a very good perspective for developing a tunable THz source emitting to open space. The main issue for such a source is the need for a specific frequency or frequency range and the design of the antenna and the transmitting line structures between the FFO, the antenna, and the HM. From the point of view of producing devices for practical use, some directions for future development are planned. The first is the widening of the operation range in both the lower and upper frequencies. The lower frequency border of the present samples of the FFO is about 200–250 GHz, so the output structures are to be coupled to the oscillator at these frequencies. The upper frequency limit is defined by the gap frequency and can potentially reach 1–1.2 THz if NbTiN-based junctions and transmitting lines are used [86], [87]. For both upper and lower frequency ranges, another experimental setup based on the Golay cell or the Si bolometer should be used for detecting the THz emission, since such frequencies are outside the SIR operating region.

Sample	R_n of the HM (Ω)	Operating frequency of the FFO emission to open space (GHz)	The best free-running linewidth obtained at this frequency (MHz)	Signal-to-noise ratio (dB)	SIR LO phase locking ability (+ or -)
Sample, design #1	13.6	387	2.5	29.8	-
		392	3.2	25.2	-
		497	14.5	21.7	+
		505	4.0	27.4	+
		536	10.1	20.1	+
		540	2.7	29.6	+
		572	2.7	27.2	+
		592	3.4	23.3	+
		600	12.8	20.4	+
		650	13.6	17.6	+
Sample, design #2	14.8	518	5.9	28.0	+
		550	8.6	29.9	+
		560	4.3	30.1	+
		580	2.5	28.4	+
		590	3.6	30.0	+
		600	12.9	22.8	+
		620	29.1	15.2	+
		650	16.1	20.2	+
		670	10	22.3	+
		690	5.6	23.4	+
		700	7.0	23.1	+
		720	7.5	16.3	-
		730	10.6	13.0	-
		735	8.9	12.2	-

Table V-1 Main results of the FFO emission measurements to open space by the SIR.

The second direction is redesign of the FFO control line and fabrication of samples with a more appropriate current density to avoid the gap at the points of operation below $V_{\text{gap}}/3$ on the IVCs, and to provide continuous frequency tuning with a free-running spectral linewidth of not more than 5–

7 MHz over the whole range. Finally, the third direction is the accurate measurement of the calibrated emission power and the beam pattern of the oscillator. For calibrated power measurements, a Golay detector or THz bolometer could be used.

V.4 CONCLUSION

We elaborated upon the idea for the THz source based on the FFO; the devices that emit to open space at a tunable frequency in the ranges of 0.32–0.55 THz and 0.4–0.7 THz have been developed, which cover the 0.32–0.7 THz region. The slot antenna is used for providing the emission from the chip surface to open space. A small part of the FFO power ($\sim 10\%$ – 20%) is branched to the HM for phase locking of the oscillator, using the feedback locking loop. Numerical simulations for the sample designs were undertaken, and samples based on Nb/AlN/NbN Josephson junctions with a gap voltage of ~ 3.5 mV were fabricated and experimentally studied. Emission to open space was successfully obtained and measured using a superconducting spectrometer located in a separate cryostat. The spectral lines of the FFO emission with a signal-to-noise ratio of up to 30 dB and a free-running linewidth as low as a few megahertz were obtained. At the same time, the measured linewidth of the phase-locked oscillator is defined by the resolution bandwidth of the spectrum analyzer, down to a value of 100 kHz [72], [74], while the actual linewidth is determined by the phase noise of the reference oscillator multiplied by the square of the number of harmonics. The experimental results for the frequency ranges of the open-space emission and of the HM pumping by the FFO are in good agreement with the numerical simulations [91].

Thus, we have demonstrated and tested all necessary planar elements which are required to design a balanced and integrated superconductor receiver [91]–[100].

CHAPTER 6

VI The thesis conclusions and prospects

- A method for microwave superconducting circuit numeric calculation in the frequency range of 300–700–950 GHz, for receiving devices based on tunnel SIS junctions, has been developed, applied, and confirmed.
- A 211–275 GHz receiving DSB element was developed, manufactured, and measured: averaged over a 5–8 GHz IF point, the noise temperature (according to Callen & Welton) was $\sim 16 \pm 1.5$ K at a frequency of 255 GHz.
- A mixing element for the range of 700–950 GHz was developed, manufactured, and mounted on a telescope with an APEX DSB matrix receiver. A noise temperature of 120 K was demonstrated at a frequency of 725 GHz, exceeding the quantum limit by only a factor of 3.
- A source of THz frequencies based on a distributed tunnel junction with the possibility of smooth frequency tuning in the range of 400–600 GHz was developed, manufactured, and studied. The applicability of the FFO as an external radiation source for various applications that require weak tunable THz radiation sources with good spectral characteristics has been demonstrated.
- Technologies, such as FFO, HM, SIS, waveguide probes, and a planar antenna design, which are necessary for a planar receiver array development, have been demonstrated.
- A wide IF band was predicted by calculations and was confirmed by experimental data. These results demonstrate the possibility of extending an IF band up to 24 GHz.

SAMENVATTING

Sinds de oudheid heeft de mensheid geprobeerd om het universum te begrijpen door het vervaardigen van vaardigheden en het verbeteren van haar observatietechnieken. De observatie van zichtbaar licht was hierin het meest natuurlijk, aangezien het visuele systeem van de mens gevoelig is voor dit type licht. Echter kon men hiermee alleen de helderste bronnen, zoals sterren en planeten, waarnemen. In de koudste gebieden van het interstellair medium bevinden zich compacte clusters van interstellair stof en moleculair gas. De grootste hiervan staan bekend als Grote Moleculaire Wolken (GMWen). Deze wolken zijn zo koud dat ze geen zichtbaar licht uitzenden en het stellair stof in deze wolken blokkeert de achtergrondstraling van sterren.

Voor het bestuderen van de fysische processen die plaatsvinden in een GMW moet een moleculaire wolk worden waargenomen terwijl het belicht, verwarmd en geïoniseerd wordt door een nabije ster. Voor het in kaart brengen van de onderliggend fysische processen meet men spectrale emissielijnen. Voor koud moleculair gas liggen de langste golflengtes van het emissiespectrum in het millimeter- en submillimeter regime, welke overeenkomen met Giga Hertz (GHz) en Tera Hertz (THz) frequenties. Dit komt door de excitatie van rotatie- en vibratievrijheidsgraden van complexe chemische verbindingen. Koolmonoxide (CO) heeft bijvoorbeeld een spectrale emissielijn in het submillimeter bereik en komt vrij veel voor in interstellair gaswolken. Door het spectrale profiel van CO, H₂O, HCl, NH, CH, CO, OH en andere verbindingen te meten, is het mogelijk om de gassamenstelling en de onderliggende fysische processen van koude moleculaire wolken in kaart te brengen.

Het observeren van submillimeter golven vanaf de Aarde wordt bemoeilijkt door het feit dat het lichtsignaal erg zwak is en atmosferische waterdamp veel submillimeter licht absorbeert. Dit vereist de ontwikkeling van zeer gevoelige en ruisarme detectoren. Deze detectoren worden normaliter afgekoeld tot de temperatuur van vloeibaar helium, zodat de intrinsieke thermische fluctuaties (Johnson-Nyquist-ruis) in de ontvanger kleiner zijn dan de kwantumfluctuaties. Op deze manier werkt het detectiesysteem in de buurt van de fundamentele ruisgrens. Daarnaast worden deze detectoren

gemaakt van supergeleidend materiaal met sterk niet-lineaire eigenschappen, wat hen zeer geschikt maakt als ruisarme THz detectoren. Volgens de ruistheorie is de tijd die nodig is om zwakke lichtsignalen te registreren evenredig met het kwadraat van de ruistemperatuur. Daarom kunnen dergelijke detectoren de observatietijd aanzienlijk verkorten, wat van een enorm belang is voor de moderne Sterrenkunde.

In het algemeen zijn er twee type van millimeter- en sub-millimeter detectie: directe en heterodyne detectie. De laatste wordt gebruikt in dit onderzoek en in de meest geavanceerde instrumentatie voor Astronomie om een zo hoge mogelijke ruimtelijke en spectrale resolutie te verkrijgen.

In heterodyne detectie wordt het waargenomen signaal in niet-lineair element gemengd met een bekend en stabiele referentiesignaal, ook wel een lokale oscillator (LO). Dit proces leidt tot een transformatie van het oorspronkelijk inkomende spectrum (met een gegeven amplitude en fasen) naar een zogenaamd "intermediate frequency" (IF) spectrum. Met andere woorden, de IF is het frequentieverschil tussen de frequentie van het referentie signaal (de LO) en de frequenties van het waargenomen signaal. Vaak ligt het IF spectrum rond de 4-12 GHz, wat makkelijk te verwerken en analyseren is met commercieel verkrijgbare apparatuur.

Dit werk richt zich op het ontwerp en de ontwikkeling van supergeleidende golfpijpen op 211-275 GHz, 600-720 GHz en 780-950 GHz. Een onderdeel hiervan is het ontwikkelen van plenaire supergeleidende geïntegreerde schakelingen die gekoppeld zijn aan één vlakke Flux-Flow-Oscillator (FFO) op dezelfde microchip. Voor het ontwerpen en ontwikkelen van lage ruis heterodyne golfpijp mixers op basis van het Superconductor-Isolator-Superconductor (SIS) junctie principe, moet men tegelijkertijd aan de volgende voorwaarden voldoen: het afstemmen (matching) van de impedantie van het megelement met de ingangsimpedantie van de hoogfrequente (HF) golfpijp; matching van het megelement met uitgangsimpedantie op de IF; het verminderen van de HF signaalverliezen van de mixerchip; en om het megelement met een aanzienlijke "conversion gain" te laten werken. Dit proces vereist het ontwikkelen van rekenmodellen, benaderingswijzen en methoden, de belangrijkste hiervan worden beschreven in hoofdstuk 1. In de rest van de hoofdstukken worden de toepassingen beschreven per frequentiegebied. De Chipplay-outs die in dit werk worden beschouwd, zijn gebaseerd op dunne supergeleidende lagen en normale dunne metaalfilms met lage verliezen. De elektrische eigenschappen van deze dunne films hangen sterk af van de

fabricatiemethoden, waarbij het afschatten van hun eigenschappen erg belangrijk is. De elektrische eigenschappen van supergeleidende microstriplijnen zijn sterk afhankelijk van frequentie en de kwaliteit van het gebruikte materiaal. In hoofdstuk 2 worden de materiaaleigenschappen en de mixer analyse bij 780-950 GHz behandeld, en zijn geleverd aan de APEX telescoop.

In hoofdstuk 3 wordt het ontwerp van een mixer gepresenteerd en zijn presentatie geverifieerd. De strikte eis op een lage ruis heeft als gevolg dat het niet lineaire element (de SIS) een grote conversion gain heeft. Dit geeft aan dat het gebruik van AlOx-technologie essentieel is voor het ontwikkelen van SIS juncties en leidt tot een hoge kwaliteitseis op de matching. De vergelijking tussen het rekenmodel en de gemeten detector eigenschappen zijn gecompliceerder dan beschreven in het vorige hoofdstuk, als gevolg van de lager ruistemperaturen en hogere gain van de mixer. Verder worden de toegepaste methoden beschreven in dit hoofdstuk.

De ontwikkeling van de huidige mixers voor ALMA-band 9 (602-720 GHz) wordt behandeld in hoofdstuk 4. Aan de hand van drie dimensionale computer simulaties en een experiment werd aangetoond dat de IF-band van de ontvanger kan worden uitgebreid tot 24 GHz. Nieuwe mixerontwerpen zijn gemaakt die voldoen aan de verschillende eisen van de productiefaciliteit.

De aspecten van de supergeleidende FFO als afstembare GHz-bron worden behandeld in hoofdstuk 5. De FFO is een eenvoudig afstembare oscillator die samen met een harmonische mixer (HM) op de chip wordt aangebracht. Hierbij wordt de HM gebruikt om de frequentie en fase van de FFO te stabiliseren. Dit project toont aan dat het technische haalbaar is om alle vlakke componenten van een ontvanger array op één chip te ontwerpen en te vervaardigen, denk hierbij aan: 1) een LO met een stabilisatiesysteem; 2) een SIS-junctie die functioneert als niet-lineaire element; of 3) een antenne. In dit hoofdstuk worden ontwerpen van dit type gesimuleerd met een rekenmodel en worden de fabricageprocessen beschreven en aangetoond.

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