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# **Design, Fabrication and Characterization of Inductance-Matched Current Sensing dc-SQUIDs for Metallic Magnetic Calorimeter Readout**

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This thesis describes the design and development of a current-sensor dc-SQUID, optimized for the readout of metallic magnetic calorimeter based particle detectors. Maximizing the energy resolution of the detector is achieved by minimizing the intrinsic noise of the SQUID and by matching the input inductance of the SQUID with the inductance of the detector. The design developed here features a two-turn input coil to better match the pickup coil inductance  $L_p = 6.65\text{ nH}$  of the maXs100 detector. To mitigate resonant structures in the SQUID dynamics, two new damping techniques are employed. On the one hand, we inductively damp the feed lines on the SQUID chip with a galvanically isolated gold layer. On the other, we fabricated the input coil as a Au/Nb bilayer, forming a lossy microstrip line. Successful damping was achieved by the former as it resulted in substantially smoother current-voltage characteristics. The noise measurements of the new SQUID with various combinations of the damping scheme were performed at a temperature of  $10\text{ mK}$ . These yield results comparable to previous low-noise SQUIDs from this working group with noise contributions as low as  $\sqrt{S_{\Phi_s,w}} = 0.22\text{ }\mu\Phi_0/\sqrt{\text{Hz}}$  for the white noise and  $\sqrt{S_{\Phi_s,1/f}} = 2.0\text{ }\mu\Phi_0/\sqrt{\text{Hz}}$  for the  $1/f$  noise contribution at  $1\text{ Hz}$ . The achieved input inductance of  $L_i = 6.4\text{ nH}$  leads to a larger coupling of  $\Delta\Phi_s/\Delta\Phi = 2.25\%$ , which exceeds the coupling achieved with  $L_i = 1.27\text{ nH}$  by 15 %. The new design thus represents an improved current-sensor SQUID for the maXs100 detector readout.

## Design, Herstellung und Charakterisierung von Stromsensor-dc-SQUIDs mit Induktivitätsanpassung zur Auslese von Metallischen Magnetischen Kalorimetern

In der vorliegenden Arbeit wird das Design und die Entwicklung eines Stromsensor-dc-SQUIDs beschrieben, welches hinsichtlich der Auslese von Teilchendetektoren auf Basis metallischer magnetischer Kalorimeter optimiert wurde. Die Maximierung der Energieauflösung des Detektors wird durch Minimierung des intrinsischen Flussrauschen des SQUIDs und durch die Anpassung der Eingangsinduktivität des SQUIDs an die Induktivität des Detektors erreicht. Das hier entwickelte Design verwendet eine Einkoppelpulpe mit zwei Windungen, um die Eingangsinduktivität an die Detektorinduktivität  $L_p = 6.65 \text{ nH}$  des maxS100-Detektors anzugeleichen. Um Resonanzstrukturen in den SQUID-Kennlinien zu minimieren, werden zwei neue Dämpfungstechniken eingesetzt. Einerseits werden die Zuleitungen auf dem SQUID-chip durch eine galvanisch getrennte Goldschicht induktiv gedämpft, andererseits wird die Einkoppelpulpe als Au/Nb Zweischichtstruktur strukturiert, wobei letzteres eine verlustreiche Mikrostreifenleitung darstellt. Mit Erstem wurde eine erfolgreiche Dämpfung erreicht, was durch wesentlich glattere Strom-Spannungs-Kennlinien demonstriert werden konnte. Die Rauschmessungen wurden für das neue SQUID mit verschiedenen Kombinationen der Dämpfungsmethoden bei einer Temperatur von  $10 \text{ mK}$  durchgeführt. Diese ergaben bis zu  $\sqrt{S_{\Phi_s,w}} = 0.22 \mu\Phi_0/\sqrt{\text{Hz}}$  für den frequenzunabhängigen Beitrag, während für die  $1/f$ -Komponente ein Wert von  $\sqrt{S_{\Phi_s,1/f}} = 2.0 \mu\Phi_0/\sqrt{\text{Hz}}$  bei  $1 \text{ Hz}$  erreicht werden konnte, was vergleichbar mit früheren rauscharmen SQUIDS aus dieser Arbeitsgruppe und mit dem Stand der Technik weltweit ist. Die erzielte Eingangsinduktivität von  $L_i = 6.4 \text{ nH}$  führt zu einer verbesserten Kopplung  $\Delta\Phi_s/\Delta\Phi = 2.25 \%$ , welche  $15 \%$  über der bisher mit  $L_i = 1.27 \text{ nH}$  erreichten liegt. Damit stellt das neue Design ein verbessertes Stromsensor-SQUID zur Auslese des maxS100-Detektors dar.



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# 1. Introduction

Particle detectors providing a high energy resolution combined with a large bandwidth have been successfully realized in the form of cryogenic metallic magnetic calorimeters (MMC) [Ens05]. These high-precision low-temperature detectors are used in a broad range of applications, such as for mass spectrometry of heavy particles [Gam19], spectroscopy of highly charged heavy ions [Hen17] or the investigation of the mass of the electron-neutrino [Gas17].

The working principle is based on the conversion of the energy of an incoming particle into a magnetic flux change. This is realized by the use of a paramagnetic temperature sensor in a weak magnetic field, which experiences a temperature change upon the absorption of an incoming particle within an absorber that is thermally coupled to the sensor. The temperature increase is accompanied by a change in magnetization of the paramagnet, which creates a flux change that can be read out by a direct current superconducting quantum interference device (dc-SQUID). SQUIDs are highly sensitive state-of-the-art magnetometers with a broad bandwidth. The SQUID can couple to the MMC via a flux transformer setup, where its input coil is connected to the superconducting pickup coil of the MMC experiencing the initial flux change  $\Delta\Phi$ . The induced current change in the input coil creates a flux change  $\Delta\Phi_s$  in the superconducting SQUID loop. The latter is a superconducting loop that is intersected by two Josephson junctions. A current-biased dc-SQUID produces a finite voltage change upon a magnetic flux change. The resulting non-linear current-voltage-characteristics require a broadband FLL feedback to linearize the output voltage. To reduce the noise contribution of the feedback electronics, a secondary SQUID is typically added to act as a low temperature preamplifier. The apparent noise in the current-sensor SQUID, however, adds to the intrinsic noise of the detector.

For the detectors developed in this working group, this added noise dominates the spectral power density above a few kHz, whereas  $1/f$  noise of the sensor material and thermodynamic energy fluctuations are mostly responsible for low frequency noise [Kem18]. The energy resolution  $\Delta E_{\text{FWHM}}$ , which depends on the signal to noise ratio, is, therefore, negatively affected by the SQUID readout chain. To quantify this effect, we use the extrinsic energy sensitivity  $\epsilon_p$  given by the intrinsic noise of the SQUID  $S_{\Phi_s}$  and the inverse of the flux-to-flux coupling  $\Delta\Phi_s/\Delta\Phi$ .

The objective of this thesis was to optimize the coupling between the SQUID and the X-ray detector maXs100 developed in this working group [All24]. This was realized by adding a second turn to the input coil of the detector SQUID to better match its inductance with the one of the pickup coil, as this maximizes the coupling

$\Delta\Phi_s/\Delta\Phi$ . Additionally, two new damping strategies based on resistive gold layers were employed to mitigate possible resonances in the IVCs. The resonance and noise behavior were investigated to assess that the new design leads to an overall reduction of the extrinsic energy sensitivity  $\epsilon_p$ .

In chapter 2 we introduce the theoretical background of Josephson junctions that is needed to describe the working principle of a dc-SQUID. As such, we briefly cover macroscopic quantum effects such as the Josephson effect and flux quantization. These explain the characteristic properties of Josephson junctions. To motivate the use of SQUIDs, we discuss the behavior of the junction in an external magnetic field. Following this, we cover the general aspects of dc-SQUIDs including their characteristics and noise behavior. Lastly, we discuss parasitic  $LC$  resonances arising from geometric properties of the SQUID circuit.

Chapter 3 covers the realization of practical SQUIDs more in depth and how their fundamental properties affect the expected performance. We demonstrate the importance of using a second-stage SQUID as a low temperature preamplifier, to significantly improve the signal to noise ratio. The last part of the chapter gives a short introduction to MMCs, summarizing their core features. We follow with a brief overview of the extrinsic energy sensitivity regarding the SQUID-based readout, which motivates the adaption of the input coil to the pickup coil of the detector, thereby improving the coupling between SQUID and detector.

In chapter 4 we present a new SQUID design based on the SQUID developed in [Bau22], where a second turn has been added to the input coil with the premise to better match the pickup coil inductance  $L_p = 6.65 \text{ nH}$  of the maXs100 detector. Further optimal design parameters are discussed which ensure minimal flux noise as well as smooth current-voltage-characteristics (IVCs). To mitigate the influence of resonances, various damping techniques are applied, such as shunting the  $LC$  circuits with resistors. We discuss new damping approaches involving inductively coupled gold layers distributed over the feed lines as well as placing a gold layer beneath the input coil, which represents a lossy microstrip line.

The measured characteristic properties of the new SQUID design are presented in chapter 5 including the input coil inductance and compare it to the expected value. To gain insight into the resonance behavior, we measured the IVC's of various new SQUID variants containing different combinations of the damping schemes described in chapter 3. Additionally, a SQUID without input coil has been tested for comparison to help identify input coil related resonances. The chapter ends with the discussion of noise measurements.

## 2. Theoretical Background

This chapter provides a short introduction into Josephson junctions and their role in dc-SQUIDs (**d**irect **c**urrent **S**uperconducting **Q**Uantum **I**nterference **D**evice), which will be the main focus of this thesis. We start with a brief overview on macroscopic quantum phenomena such as the Josephson effect and explain the general working principle of superconductor-isolator-superconductor (SIS) tunnel contacts, followed by a summary of their basic properties. This theoretical description of Josephson tunnel junctions enables us to form the corresponding framework for SQUIDs, which are developed in this group and optimized within the scope of this thesis. Lastly, we will take a closer look into their parasitic resonance behavior and investigate different methods to mitigate the quality factors. We will closely follow the derivations from the textbooks [Cla04] and [Gro16].

### 2.1 Josephson junctions

The *Josephson junction*, named after Brian D. Josephson, consists of two identical superconductors weakly coupled to each other. In the case of the junctions produced in this working group, such coupling is realized through a few nm of a thin insulating layer between the superconducting electrodes. Consequently, they are also referred to as SIS junctions. This is realized with a trilayer structure whereby in the framework of this thesis we use a Nb/Al-AlO<sub>x</sub>/Nb trilayer, with niobium being used for the superconductors and the insulating layer being provided by the aluminum oxide. The schematic structure of a SIS-type junction is shown in figure 2.1. By connecting the tunnel junction to a current source they exhibit a non-trivial current-voltage behavior, which will be covered in the following.

#### 2.1.1 Josephson effect

According to the BCS theory developed by Bardeen, Cooper and Schrieffer in 1957 [Bar57], electrons in a superconductor form pairs below a material dependent critical temperature  $T_c$ . These composite particles are also referred to as *Cooper pairs* and they represent the superconducting charge carriers with twice the mass and charge of a single electron. Within the framework of the BCS-theory the dissipationless current flow in superconductors can be explained by the formation of the energy gap in the density of states, suppressing the scattering of Cooper pairs and other particles in the solid body. Alongside the supercurrent with zero dc-resistance, the Meissner-

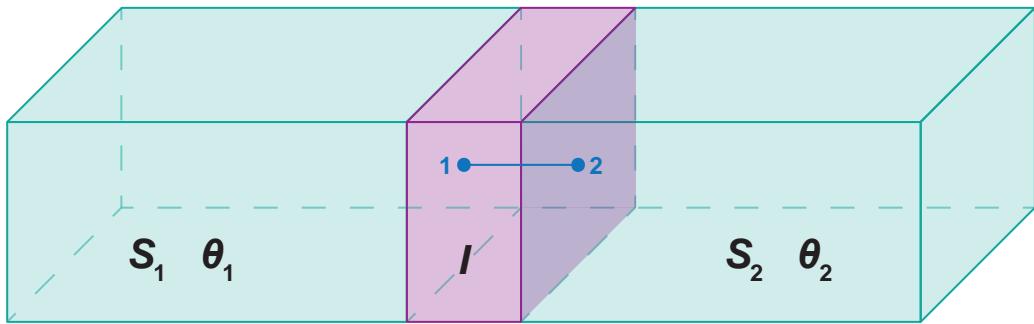


Figure 2.1: Schematic of a Josephson (SIS) junction. Both superconducting electrodes  $S_1$  and  $S_2$  are weakly coupled with each other through a thin tunnel barrier  $I$ .  $\theta_1$  and  $\theta_2$  represent the macroscopic phases of each superconductor.

Ochsenfeld effect [Mei33] is the most characteristic feature of a superconductor. The latter describes the magnetic field expulsion provided the external magnetic field is smaller than a critical field  $B_c$ . A detailed description of the microscopic theory of superconductivity can be found in [dG64].

For temperatures below the critical temperature of niobium  $T_c = 9.3\text{ K}$  [Ina80] Cooper pairs are able to tunnel across a Nb/Al-AlO<sub>x</sub>/Nb Josephson junction despite the insulating barrier. In fact, their tunneling behavior resembles that of a single electron, which can be motivated by the *Macroscopic Quantum Model* that was formulated by Fritz London in 1953. This model heavily focuses on the quantum mechanical phase  $\theta$  of ensemble of Cooper pair, whose macroscopic nature can be understood by the following arguments. On one hand, the distance between both electrons in a Cooper pair is approximately 10 to 1000 nm which is significantly larger than the spacing between Cooper pairs, resulting in strongly overlapping wave functions. On the other hand, Cooper pairs have to obey Bose-Einstein statistics. Thus, all Cooper pairs share the same ground state, and as a consequence, the energies and temporal evolutions of the phases are equal. These two effects lead to what is known as *phase-lock* [Gro16]. The phases of neighboring pairs synchronize such that this quantum mechanical property now holds on a macroscopic scale. This gives rise to the phase-dependent macroscopic wave function

$$\Psi(\mathbf{r}, t) = \Psi_0(\mathbf{r}, t)e^{i\theta(\mathbf{r}, t)} , \quad (2.1)$$

which describes all charge carriers of a bulk superconductor. Here, the charge carrier density is given by  $|\Psi_0(\mathbf{r}, t)|^2 = n_s$ . The phase of the Cooper pair ensemble depends on the time  $t$  and the position  $\mathbf{r}$ . As a result of sharing the same phase, both electrons of a Cooper pair consequently possess the same tunneling probability as

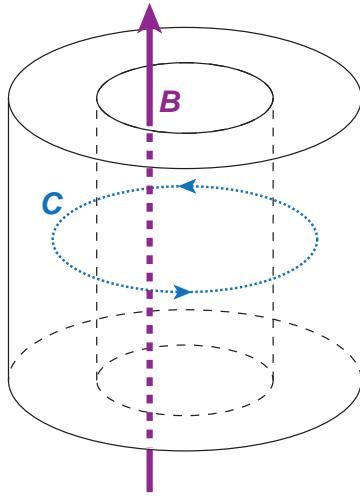


Figure 2.2: Superconducting ring-shaped cylinder threaded by an external magnetic field. By applying the field at low temperatures, shielding currents arise to expel the field from the superconductor. Upon turning off the external field the shielding currents will remain due to the lack of resistance, causing magnetic flux to be trapped. The dotted blue path  $C$  is situated at the center of the cylinder wall, which we assume to be current-free due to the London penetration depth  $\lambda_L$  being much smaller than the thickness of the cylinder wall.

an individual electron, enabling the supercurrent. This coherence phenomenon is referred to as the *Josephson effect* [Jos62]. Another significant consequence of the macroscopic quantum model is flux quantization.

This phenomenon can be explained through the capture of an external magnetic flux within a superconducting cylinder (see figure 2.2). The wave function must remain unchanged after circumnavigating the cylinder due to  $e^{i\theta} = e^{i\theta+2\pi n}$ . As a result, upon integrating along the current-free center of the cylinder wall (path  $C$ ), the following equation holds for the captured flux [Dea61]

$$\Phi = \frac{h}{q_s} n = \frac{h}{2e} n \equiv \Phi_0 n . \quad (2.2)$$

Here,  $n \in \mathbb{N}$  and  $\Phi_0 = 2.07 \times 10^{-15} \text{ T m}^2$  [Tie21] represents the so-called magnetic flux quantum. The captured flux is thus quantized, a consequence solely arising from the macroscopic nature of the phase. This quantity plays a crucial role in the theoretical description of Josephson junctions.

The current and voltage behavior in a SIS junction is described by the *Josephson equations*. Crucial to this description is a critical current  $I_c$ , which marks the boundary between two operational modes; the zero-voltage state and the voltage state. Additionally, due to the macroscopic nature of the superconducting phase factor, the supercurrent  $I_s$  oscillates with the gauge-invariant phase difference between the two macroscopic wavefunctions of the superconducting electrodes  $\varphi$ , leading to the **first Josephson equation** [Jos65]

$$I_s = I_c \sin(\varphi) . \quad (2.3)$$

The critical current  $I_c$  is proportional to the coupling strength  $\kappa$ , which describes the overlap of the wave functions  $\Psi_1$  and  $\Psi_2$  within the insulating layer. The relationship is given by

$$I_c = \frac{4e\kappa V n_s}{\hbar} , \quad (2.4)$$

where  $V$  represents the volume of the superconducting electrode and  $e$  denotes the elementary charge of an electron. We assume that the Cooper pair density  $n_s$  of the two superconductors  $S_1$  and  $S_2$  is identical, meaning  $n_{s1} = n_{s2} = n_s$ .

The gauge-invariant phase difference refers to the phases  $\theta_1$  and  $\theta_2$  of the respective electrodes at the boundary of the insulating layer (positions 1 and 2, see figure 2.1). Taking into account possible external electromagnetic fields within the barrier, the general form using the vector potential  $\mathbf{A}$  is given by [Gro16]

$$\varphi(\mathbf{r}, t) = \theta_2(\mathbf{r}, t) - \theta_1(\mathbf{r}, t) - \frac{2\pi}{\Phi_0} \int_1^2 \mathbf{A}(\mathbf{r}, t) \cdot d\mathbf{l} . \quad (2.5)$$

Assuming a constant supercurrent density  $J_s$  across the junction, taking the time derivative of equation (2.5) yields the **second Josephson equation** [Jos65]

$$\frac{\partial \varphi}{\partial t} = \frac{2\pi}{\Phi_0} V . \quad (2.6)$$

The first of the two above-mentioned operating modes describes the zero-voltage state, i.e.  $I < I_c$ . Here, the entire injected current is carried by Cooper pairs, so  $I = I_s = \text{const}$ . As a result,  $\varphi$  is constant over time, which, according to equation (2.6), leads to  $V = 0$ . This is known as the *dc Josephson effect*.

For  $I > I_c$  however, Cooper pairs begin to break up into quasiparticles which carry that portion of the current leading to a voltage drop  $V$ . According to the second Josephson equation, the phase  $\varphi$  becomes time dependent, and after integration one obtains

$$\varphi = \frac{2\pi}{\Phi_0} V t + \varphi_0 = w_J t + \varphi_0 \quad \text{with} \quad w_J = \frac{2\pi}{\Phi_0} V . \quad (2.7)$$

Thus, if we insert equation (2.7) into equation (2.3), we observe that the current  $I_s$  oscillates with the *Josephson frequency*  $\frac{f_J}{V} = \frac{w_J}{2\pi V} = \frac{1}{\Phi_0} \approx 483.6 \text{ MHz}/\mu\text{V}$ . Accordingly, this phenomenon is referred to as the *ac Josephson effect*.

### 2.1.2 Josephson Junctions in a Magnetic Field

To motivate the structure of a dc-SQUID, it is essential to first investigate the current behavior of an extended Josephson junction in the presence of an external magnetic field. So far, all previous formulae apply for point-like junctions, assuming a spatially constant phase difference  $\varphi$  and Josephson current density  $J_s$  across the junction area. This is not the case for three-dimensional (extended) junctions with a length  $L$  and width  $W$ . The *Josephson penetration depth*  $\lambda_J$  is a quantity used to classify an extended junction as short ( $W, L \leq \lambda_J$ ) or long ( $W, L \geq \lambda_J$ ) and is defined as [Wei69]

$$\lambda_J = \sqrt{\frac{\Phi_0}{2\pi\mu_0 J_{\text{c}} t_B}} . \quad (2.8)$$

Here, the magnetic thickness is defined as  $t_B = d + \lambda_{L,1} + \lambda_{L,2}$ , where  $d$  is the

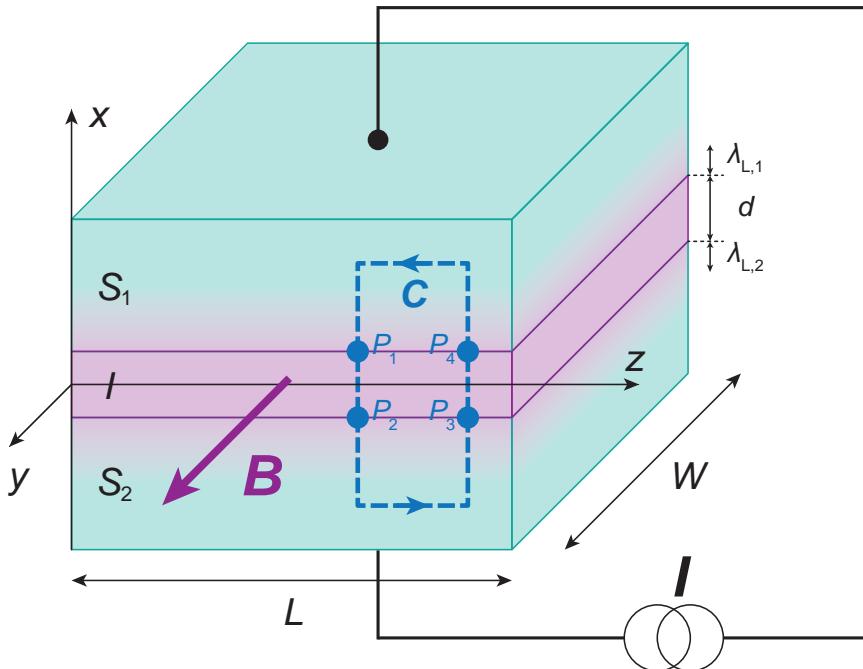


Figure 2.3: Short Josephson junction connected to a current source in the presence of an external  $B$ -field in  $y$ -direction, parallel to the junction area. Inside the electrodes the magnetic field decays exponentially according to the London penetration depths  $\lambda_{L,1}$  and  $\lambda_{L,2}$ , visually shown by the purple color gradient. The closed contour  $C$  is used to derive expressions for the spatially dependent phase difference  $\varphi$  and current density  $J_s$ .

geometric thickness of the isolator and  $\lambda_L$  the London penetration depth of respective superconducting electrode. It describes how far an external magnetic field penetrates both superconducting electrodes if applied parallel to the junction area, as depicted in figure 2.3. The respective London penetration depths are  $\lambda_{L,1}$  and  $\lambda_{L,2}$  and  $J_c = \frac{I_c}{WL}$  is the critical current density. This distinction is needed to determine whether the magnetic self-field generated by the supercurrent is negligible in comparison to the external field (short junctions) or not (long junctions). Within the scope of this thesis, we only use short junctions.

To analyze the current and phase distribution of such a junction we consider the setup shown in figure 2.3. A short junction is connected to a current source and is penetrated by an external  $B$ -field in  $y$ -direction, parallel to the junction area. Now, obtaining an expression for the phase requires a similar approach as the calculation for the quantized flux, where we assumed that the phase changes by  $2\pi n$  around a closed loop. Here, we again integrate over a closed contour  $C$ , with the points  $P_1 - P_4$  marking the transitions between superconductor and isolator. Using equation 2.5, we find

$$\frac{\partial \varphi}{\partial z} = \frac{2\pi}{\Phi_0} B_y t_B \quad \text{and} \quad \frac{\partial \varphi}{\partial y} = -\frac{2\pi}{\Phi_0} B_z t_B . \quad (2.9)$$

The magnetic field points in  $y$ -direction only, meaning  $\varphi$  will only vary along the  $z$ -axis. Integrating the first expression in equation 2.9 then leads to

$$\varphi(z) = \frac{2\pi}{\Phi_0} B_y t_B z + \varphi_0 . \quad (2.10)$$

Here, the integration constant  $\varphi_0$  represents the phase difference for the case  $z = 0$ . Inserting equation 2.10 into the first Josephson equation and using  $J_s = \frac{I_s}{WL}$  gives

$$J_s(y, z, t) = J_c(y, z) \sin(kz + \varphi_0) \quad \text{with} \quad k = \frac{2\pi}{\Phi_0} B_y t_B . \quad (2.11)$$

If we now assume the critical current density  $J_c$  to be constant across the junction area, we can integrate equation 2.11 to get a flux-dependent maximum Josephson current

$$I_s^m(\Phi) = I_c \left| \frac{\sin\left(\frac{kL}{2}\right)}{\frac{kL}{2}} \right| = I_c \left| \frac{\sin\left(\frac{\pi\Phi}{\Phi_0}\right)}{\frac{\pi\Phi}{\Phi_0}} \right| , \quad (2.12)$$

where  $\Phi = B_y t_B L$  is the total flux threading through the junction. This expression describes the so-called Fraunhofer diffraction pattern, shown in figure 2.4. The result

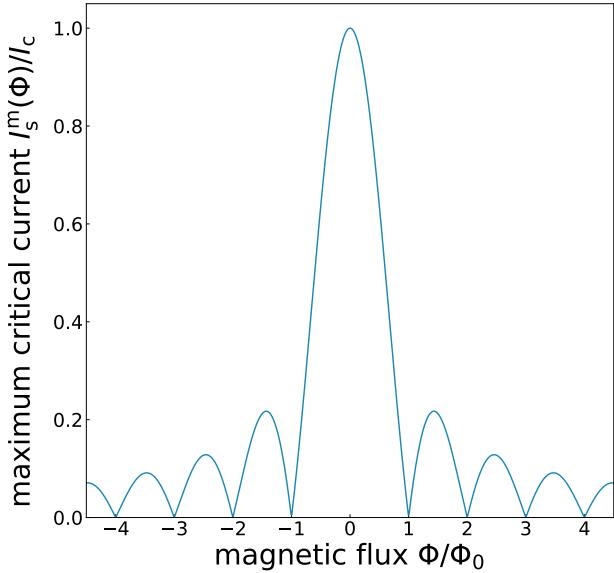


Figure 2.4: Normalized flux-dependent maximum Josephson current  $I_s^m(\Phi)$  showing a Fraunhofer pattern. It modulates with the flux quantum  $\Phi_0$ , peaking at  $\Phi = 0$  with subsequent maxima at  $\Phi = \pm(\frac{3}{2} + n)\Phi_0$  with  $n \in \mathbb{N}_0$ . For  $\Phi = \pm(n + 1)\Phi_0$  the total net current is zero.

resembles the single slit experiment, where the same pattern is found for the light intensity behind the slit. Here, the analogy works by considering the integral of the critical current density  $J_c$  as a transmission function which is constant inside the junction and zero outside.

### 2.1.3 RCSJ Model

The Fraunhofer pattern describes the flux-dependent current for the case of  $I < I_c$ , staying in the so-called zero-voltage state. In this regime, only the dc Josephson effect applies as discussed in subsection 2.1.1. Switching to the voltage stage, i.e.  $I > I_c$ , Cooper pairs start breaking up into quasiparticles if the electric energy  $eV$  exceeds the sum of both electrodes' gap energies  $\Delta_1(T) + \Delta_2(T)$  [Bar57]. Consequently, at the *gap-voltage*

$$V_g = \frac{\Delta_1(T) + \Delta_2(T)}{e} \quad (2.13)$$

quasiparticles start to cross the tunnel barrier resulting in a steep rise of a resistive normal current  $I_n$ . This process also occurs at finite temperatures for  $k_B T > \Delta_1(T) + \Delta_2(T)$ , leading to a reduction of  $I_c$  as well as  $V_g$ . Under a dc current source, the condition  $I = I_s + I_n$  must be constantly fulfilled. This results in an oscillating normal current and therefore voltage, since  $I_s$  oscillates with  $f_J$  according to the ac Josephson effect. According to the second Josephson equation (2.6) the oscillating

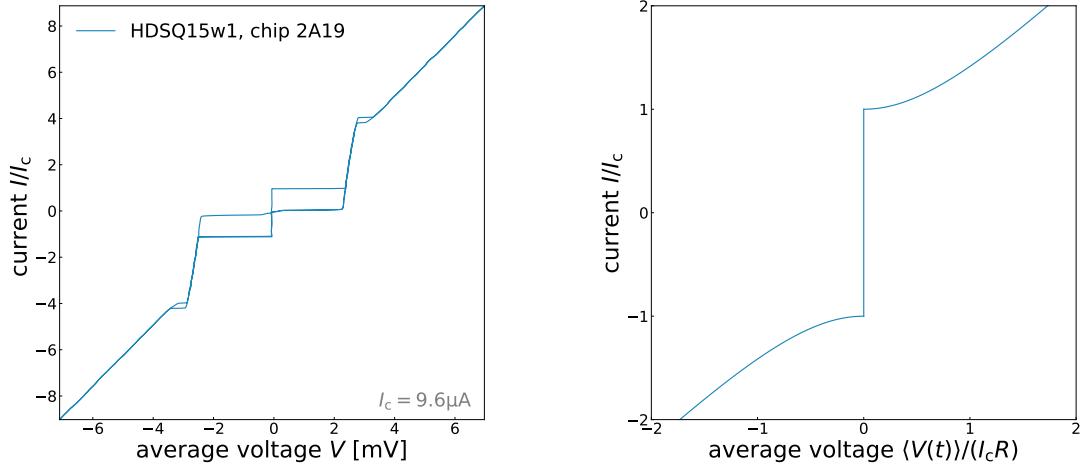


Figure 2.5: Left: Measured IVC of an underdamped junction manufactured in this working group, showing the typical hysteresis. Right: Theoretical IVC of an overdamped junction with a current-voltage shape that is independent of the current sweep direction.

voltage thus causes the term  $\frac{d\varphi}{dt}$  to vary sinusoidally, causing both  $I_s$  and  $I_n$  and in turn the resulting voltage to oscillate in a complex manner. As a voltage with such a high frequency cannot be measured, only the time-averaged voltage will be considered in the following discussion.

Now, further increasing the energy of the quasiparticles ( $T > T_c$  and/or  $V > V_g$ ) leads to a transition into normal-conducting electrons, which exhibit an ohmic dependence. This behavior can be seen in the typical current-voltage-characteristic (IVC) depicted in figure 2.5.

For real junctions, however, one needs to take into account that they are comprised of two electrodes separated by a thin insulating layer, which represent a parallel plate capacitor with the Al-AlO<sub>x</sub> layer being the dielectric material. Therefore, a junction capacitance  $C$  needs to be taken into account. A displacement current  $I_d$  will flow as a consequence, given we are in the voltage state. Lastly, thermal and 1/f noise cause a small fluctuating current  $I_f$ . All these current channels were defined in the so-called Resistively and Capacitively Shunted Junction (RCSJ) model [McC68], [Ste68], which models the total current of a lumped (0-dimensional) junction to a sufficiently high accuracy. A schematic of an effective circuit diagram is shown in figure 2.6 (left). Combining every current channel utilizing Kirchhoff's law leads to the *Basic Junction Equation*, which is defined as [Gro16]

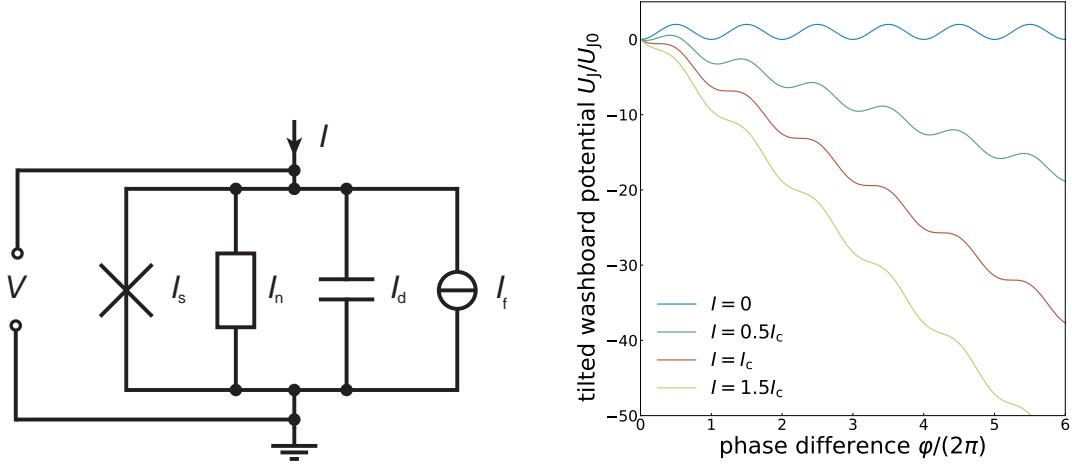


Figure 2.6: Left: Schematic circuit of a lumped Josephson junction with all four current channels connected in parallel, according to the RCSJ model. The junction is represented by the cross symbol on the left, marking the supercurrent  $I_s$ . The normal current  $I_n$  is realized with a resistance  $R$ , while the displacement current  $I_d$  and the noise  $I_f$  are attributed to a capacitor  $C$  and a current source, respectively. Right: Tilted washboard potential for different currents, ranging from 0 to  $1.5I_c$ . The tilt increases with the injected current  $I$ .

$$I = I_s + I_n + I_d + I_f = I_c \sin(\varphi) + \frac{1}{R(V)} \frac{\Phi_0}{2\pi} \frac{d\varphi}{dt} + C \frac{\Phi_0}{2\pi} \frac{d^2\varphi}{dt^2} + I_f . \quad (2.14)$$

By defining the Josephson coupling energy  $U_{J0} = \frac{\hbar I_c}{2e}$  and the normalized currents  $i = \frac{I}{I_c}$  and  $i_f(t) = \frac{I_f(t)}{I_c}$ , equation 2.14 can be rewritten to

$$\left(\frac{\hbar}{2e}\right)^2 C \frac{d^2\varphi}{dt^2} + \left(\frac{\hbar}{2e}\right)^2 \frac{1}{R(V)} \frac{d\varphi}{dt} + \frac{d}{d\varphi} \{U_{J0} [1 - \cos \varphi - i\varphi + i_f(t)\varphi]\} = 0 . \quad (2.15)$$

The expression in the curly brackets represents the potential energy in the system  $U_J$ , allowing equation 2.14 to be compared to

$$M \frac{d^2x}{dt^2} + \eta \frac{dx}{dt} + \nabla U = 0 , \quad (2.16)$$

which describes a particle with mass  $M$  and damping  $\eta$  moving inside the potential  $U$ . This mechanical analogue therefore allows us to interpret as the equation of motion

for a *phase particle*, where its motion corresponds to a change of the gauge-invariant phase difference  $\varphi$  within a potential  $U_J$  [Cla04]. Consequently, it is attributed with a mass  $M = \left(\frac{\hbar}{2e}\right)^2 C$  and damping  $\eta = \left(\frac{\hbar}{2e}\right)^2 \frac{1}{R(V)}$ . Figure 2.6 (right) visualizes how this phase particle behaves for different currents  $I$ . Given the shape of  $U_J(\varphi)$ , the potential is referred to as the *tilted washboard potential*.

For  $I = 0$ , the phase particle will remain within one of the potential minima. As the current increases, however, the potential starts to tilt such that the depth of the minima reduces until it vanishes for  $I = I_c$ , thus becoming a saddle point. Up until this point, the phase particle can't overcome the potential barrier to move downward, which agrees with the second Josephson equation as the phase difference  $\varphi$  should remain constant on average for  $I < I_c$ . Further increasing the current and therefore the tilt of the potential causes the phase particle to fall along the potential, resulting in a voltage drop across the junction ( $\frac{\partial\varphi}{\partial t} > 0$ ).

Reversing the current sweep showcases the importance of the particle's mass  $M$  and damping  $\eta$ , as they determine if the return path equals the current shape described above or not. For the case of a small mass (small  $C$ ) and large damping (small  $R$ ), the phase particle will, due to a lack of momentum and strong damping, come to a halt as soon as minima reappear in the washboard potential by reducing the current below  $I_c$ . The current path will therefore remain unchanged as  $I$  is reduced back to 0, as shown in figure 2.5 (right). Such a junction is consequently called an *overdamped* junction.

The other case describes an *underdamped* junction (figure 2.5 (left)) and involves a large mass (large  $C$ ) and small damping (large  $R$ ). This allows the phase particle to continue to move downward as it now carries enough momentum to overcome the arising maxima and minima. The finite voltage drop despite the current being below  $I_c$  is displayed as the steep quasiparticle current curve, which ends with a return current  $I_R$  that arises with the recapture of the particle in a minimum. This leads to a hysteretic IVC, as depicted in figure 2.5 (left).  $I_R$  can be calculated via [Lik86]

$$I_R = \frac{4}{\pi\sqrt{\beta_C}} I_c , \quad (2.17)$$

with  $\beta_C$  being the dimensionless Stewart-McCumber parameter, that is used to quantitatively distinguish between both junction types. It is given by

$$\beta_C = \frac{2\pi}{\Phi_0} I_c R^2 C \quad (2.18)$$

with  $\beta_C \gg 1$  corresponding to a strongly underdamped junction, whereas  $\beta_C \ll 1$  represents a strongly overdamped junction. The junctions developed and produced

within the scope of this thesis aim to be overdamped, which is why we take a closer look on the time-averaged voltage for  $I > I_c$  in the case of  $\beta_C \ll 1$ . Neglecting the noise in equation 2.15, as well as assuming the resistance to be linear below and above the gap voltage  $V_g$ , i.e.  $R(V) = R$ , the time-averaged voltage can be derived to [Cla04]

$$\langle V(t) \rangle = I_c R \sqrt{\left(\frac{I}{I_c}\right)^2 - 1} \quad \text{for} \quad \frac{I}{I_c} > 1 . \quad (2.19)$$

This equation will be crucial to determine the voltage drop of a dc-SQUID, as its derivation is analogous to that of a single junction, which will be covered in the next section.

## 2.2 dc-SQUIDS

We have now covered the theoretical framework necessary to understand the working principle of a dc-SQUID, which consists of a superconducting ring intersected by two identical Josephson junctions with critical Josephson currents  $I_c$ , as depicted in figure 2.7. Both junctions are shunted with shunt resistors  $R_s$  to avoid hysteretic behavior in the respective IVCs. If the SQUID is biased with a bias current  $I_b$ , it is possible to convert small flux variations inside the loop into a measurable voltage change. dc-SQUIDs are, therefore, used as highly sensitive flux-to-voltage transducers.

### 2.2.1 Zero Voltage State

In order to fully understand the working principle of a dc-SQUID it is again necessary to first cover the zero voltage stage as we did for a single junction. The parallel connection of the two junctions allows the bias current with identical critical currents, i.e.  $I_{c,1} = I_{c,2} = I_c$  to split into two supercurrents  $I_{s1}, I_{s2}$ . Here we assume  $I_b < 2I_c$  to ensure that no voltage drop across both junctions occurs ( $V_s = 0$ ). Applying Kirchhoff's law we then obtain the expression

$$I_b = I_s = I_c \sin \varphi_1 + I_c \sin \varphi_2 = 2I_c \cos\left(\frac{\varphi_1 - \varphi_2}{2}\right) \sin\left(\frac{\varphi_1 + \varphi_2}{2}\right) . \quad (2.20)$$

In chapter 2.1.2 we concluded that a magnetic flux  $\Phi$  causes the supercurrent to modulate with  $\Phi_0$ . A dc-SQUID can be considered as a single junction with a much larger effective area  $A_{\text{eff}}$  (loop area), that could be penetrated by an external magnetic flux. It is, therefore, reasonable to expect a similar behavior for a dc-SQUID.

The same approach as with a single junction is used to determine the flux dependence of the total supercurrent, where a closed loop integral is performed around the SQUID loop. The calculation leads to the relation [Gro16]

$$\varphi_2 - \varphi_1 = \frac{2\pi\Phi}{\Phi_0} , \quad (2.21)$$

which can be directly inserted into equation 2.20 to obtain

$$I_s = 2I_c \cos\left(\pi \frac{\Phi}{\Phi_0}\right) \sin\left(\varphi_1 + \pi \frac{\Phi}{\Phi_0}\right) . \quad (2.22)$$

In the most general case, however, one needs to take into account the inductance  $L_s$  of the SQUID loop and, therefore, a circulating current  $I_{\text{cir}} = (I_{s1} - I_{s2})/2$  that induces the additional flux  $\Phi_{\text{cir}} = L_s I_{\text{cir}}$ . With the external flux  $\Phi_e$  we can thus write for the total flux

$$\Phi = \Phi_e + \Phi_{\text{cir}} \quad (2.23)$$

$$= \Phi_e - L_s I_c \sin\left(\pi \frac{\Phi}{\Phi_0}\right) \cos\left(\varphi_1 + \pi \frac{\Phi}{\Phi_0}\right) \quad (2.24)$$

$$= \Phi_e - \frac{1}{2} \beta_L \Phi_0 \sin\left(\pi \frac{\Phi}{\Phi_0}\right) \cos\left(\varphi_1 + \pi \frac{\Phi}{\Phi_0}\right) . \quad (2.25)$$

Here, we introduced the dimensionless screening parameter  $\beta_L = \frac{2L_s I_c}{\Phi_0}$ , which relates the maximum possible flux  $\Phi_{\text{cir}}^{\max} = L_s I_{\text{cir}}^{\max} = L_s I_c$  produced by screening currents to  $\Phi_0/2$ . This quantity describes the influence the screening currents have on the total flux  $\Phi$ , which in turn affects  $I_s$  in equation 2.22. We will now simplify the

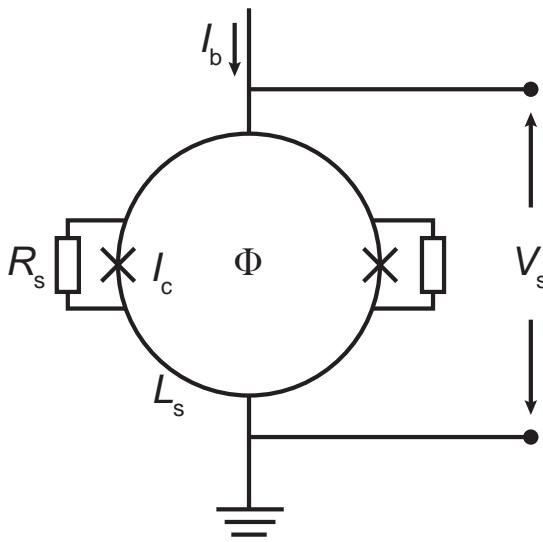


Figure 2.7: Schematic circuit diagram of a shunted dc-SQUID. A superconducting loop with inductance  $L_s$  is interrupted by two Josephson junctions such that they form a parallel connection. With bias current is  $I_b$  and external magnetic flux  $\Phi$ . To avoid hysteresis effects, shunt resistors  $R_s$  are connected in parallel to each junction.

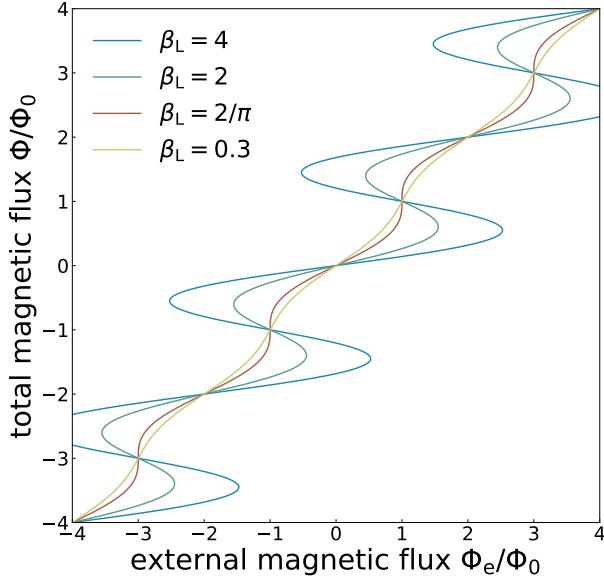


Figure 2.8: Normalized flux  $\Phi$  modulated by the external flux  $\Phi_e$ . The amplitude of the modulation depends on the screening parameter  $\beta_L$ , where  $\Phi(\Phi_e)$  remains a single-valued function for  $\beta_L \leq 2/\pi$ .

expression above by considering the limiting case for small currents, i.e.  $I_s \ll 2I_c$ . This condition implies that  $\sin \varphi_1 \approx -\sin \varphi_2$  and thus  $\varphi_1 \approx -\varphi_2$ , leading to a vanishing argument  $\varphi_1 + \pi \frac{\Phi}{\Phi_0} \approx 0$  in the cosine resulting to

$$\Phi = \Phi_e - \frac{1}{2}\beta_L\Phi_0 \sin\left(\pi \frac{\Phi}{\Phi_0}\right) . \quad (2.26)$$

Figure 2.8 showcases this relation for several values of  $\beta_L$ . High values ( $\beta_L > 2/\pi$ ) correspond to hysteretic characteristics, meaning there can be multiple values of total flux  $\Phi$  for the same applied flux  $\Phi_e$ . For practical dc-SQUIDs, it is, therefore, desirable to avoid this ambiguous behavior. The intersections of each curve represents the case for  $\Phi = n\Phi_0$ , such that the screening currents vanish and the total flux equals the external flux ( $\Phi = \Phi_e$ ). This is to be expected as the flux in a superconducting ring needs to be quantized (see equation 2.2). Consequently, the SQUID tries to maintain the total flux at integer values of  $\Phi_0$  for the limiting case of  $\beta_L \gg 1$ , where  $\Phi_{\text{cir}}$  dominates over any applied flux. This compensation is visualized by the strong modulation for high  $\beta_L$  in figure 2.8, where a wide range of  $\Phi_e$  values remain in the proximity of  $n\Phi_0$ . The other limiting case, i.e.  $\beta_L \ll 1$ , allows us to neglect the circulating currents such that we can write  $\Phi \approx \Phi_e$ . From equation 2.22 we then obtain the maximum possible supercurrent

$$I_s^m(\Phi_e) = 2I_c \left| \cos\left(\pi \frac{\Phi_e}{\Phi_0}\right) \right| . \quad (2.27)$$

The modulation of this current quickly diminishes for increasing  $\beta_L$ , as was derived in [Cla04] to

$$\frac{\Delta I_s^m(\Phi_e)}{2I_c} \approx 1 - \frac{2\Phi_e}{\Phi_0\beta_L} . \quad (2.28)$$

For the SQUIDs produced within the scope of this thesis, values of  $\beta_L \approx 1$  were considered optimal to minimize resonant behavior without reducing the SQUID inductance  $L_s$  too much.

### 2.2.2 Voltage State

To utilize dc-SQUIDs as sensitive magnetometers, it is necessary to operate them in the voltage state by applying a large enough bias current  $I_b$ , such that  $I_b > 2I_c$ . In the case of negligible screening ( $\beta_L \ll 1, \Phi \approx \Phi_e$ ) and strong damping ( $\beta_C \ll 1$ ), i.e. by choosing a small junction capacitance  $C$  and SQUID inductance  $L_s$ , it is possible to derive the flux dependence of the resulting voltage drop across the SQUID. Following the RCSJ model, we are only left with the supercurrent  $I_s$  and the resistive current  $I_n$ , such that by using equation 2.22 we can write for the bias current

$$I_b = 2I_c \cos\left(\pi \frac{\Phi_e}{\Phi_0}\right) \sin\left(\varphi_1 + \pi \frac{\Phi_e}{\Phi_0}\right) + 2 \frac{V_s}{R} , \quad (2.29)$$

where we again assumed identical junctions, each shunted by a small shunt resistor  $R_s \ll R_n$ . Here  $R_n$  denotes the normal resistance of a single, unshunted junction. Therefore, the total normal resistance  $R$  for each branch of the SQUID is approximately  $R \approx R_s$ . Additionally, we can define the phase  $\varphi = \varphi_1 + \pi \frac{\Phi}{\Phi_0}$  and with the maximum supercurrent from equation 2.27 we obtain a current-phase relation that resembles that of a single junction:

$$I_b = I_s^m(\Phi_e) \sin(\varphi) + \frac{2}{R_s} \frac{\Phi_0}{2\pi} \frac{\partial \varphi}{\partial t} . \quad (2.30)$$

This equivalence of a dc-SQUID and a single junction underlines the above-mentioned fact that the SQUID loop represents a single Josephson contact that provides a larger effective area external fields can penetrate. It is therefore possible to derive the voltage drop across the SQUID in the same manner as in subsection 2.1.3. With the critical current  $I_s^m(\Phi_e)$  being flux-dependent with a modulation period of  $\Phi_0$ , we can

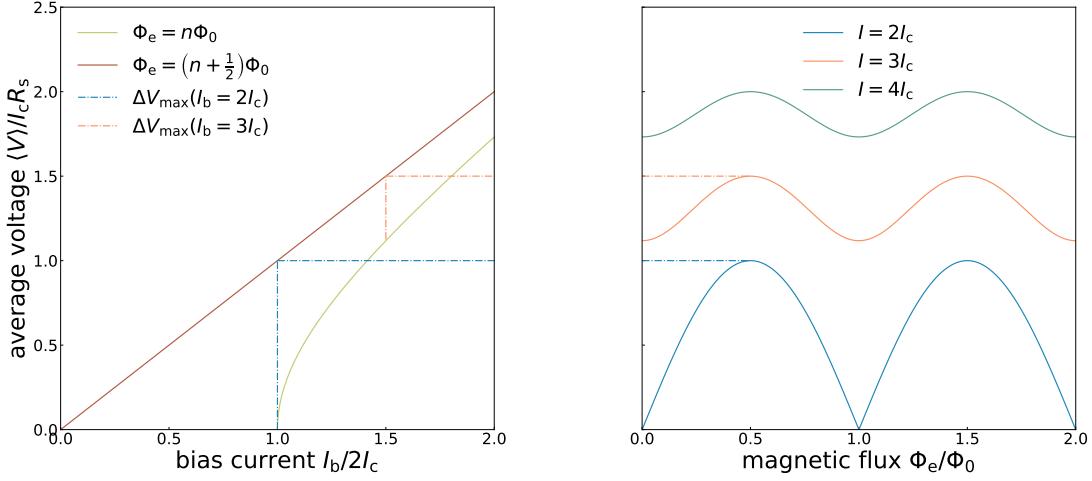


Figure 2.9: Left: IV-characteristics for the total flux  $\Phi \approx \Phi_e$  being an integer and half integer number of flux quanta, given that  $\beta_C \ll 1$  and  $\beta_L \ll 1$ . The maximum voltage swing  $\Delta V_{\text{max}}$  is approximately given at  $I_b \approx 2I_c$  and corresponds to  $I_c R_s$  for a resistively shunted dc-SQUID. Right: The projection of equation 2.31 onto the  $V\Phi$ -plane shows the flux dependence of the voltage at the bias current values  $I_b = 2I_c$ ,  $I_b = 3I_c$  and  $I_b = 4I_c$ . The amplitude of the modulation decreases for increasing  $I_b$ .

compare to equation 2.19 and obtain for the time averaged voltage [Cla04]

$$\langle V(t) \rangle = \frac{R_s}{2} \sqrt{I_b^2 - I_s^m(\Phi_e)^2} \quad (2.31)$$

$$= I_c R_s \sqrt{\left(\frac{I_b}{2I_c}\right)^2 - \left[\cos\left(\pi \frac{\Phi_e}{\Phi_0}\right)\right]^2}. \quad (2.32)$$

Evidently, both the critical current and the voltage are flux dependent and are modulated by  $\Phi_0$ . Figure 2.9 (left) showcases this behavior by considering the case for the minimum and maximum critical current, i.e. for  $\Phi_e = n\Phi_0$  and  $\Phi_e = (n + \frac{1}{2})\Phi_0$ , with  $n \in \mathbb{Z}$ . The current-voltage-characteristics at these flux values are particularly interesting, as they can be used to extract crucial SQUID parameters like the voltage swing  $\Delta V_{\text{max}}$ . This property describes how the voltage varies with the applied flux  $\Phi_e$ , at a given current  $I_b$ . It is maximal at  $I_b \approx 2I_c$ , as depicted in figure 2.9 (right).

It is, however, important to note that equation 2.31 doesn't hold for practical SQUIDS, as they are typically not fabricated to fulfill the limiting case of  $\beta_C \ll 1$  and  $\beta_L \ll 1$ . With few adjustments, the conclusions reached here will nevertheless

be applicable to practical SQUIDS.

### 2.2.3 Optimal Parameters

Negligible screening is not feasible, as it would require an extremely small SQUID inductance  $L_s$ , which in turn deteriorates the sensitivity to magnetic fields. The main reason to construct a dc-SQUID was to obtain a highly sensitive magnetometer by creating a large area for magnetic fields to penetrate. Also, the fabrication process doesn't allow to produce an arbitrarily small junction capacitance  $C$ . The parameter  $\beta_C$  will, therefore, reach a lower limit as well, since also decreasing  $R_s$  too much reduces the voltage swing  $\Delta V_{\max}$  and increases the energy sensitivity  $\epsilon(f)$ , as we will see in subsection 2.2.4. Taking into account displacement and fluctuation currents, the current and voltage expressions for the dc-SQUID become analytically unsolvable and therefore have to be solved numerically. In [Tes77] such numerical simulations lead to optimal values of  $\beta_C \approx 1$  and  $\beta_L \approx 1$  to minimize the energy sensitivity.

To further fine-tune the SQUID parameters, e.g. loop inductance and critical current, it is essential to look at how dc-SQUIDs are typically operated to achieve the highest possible flux sensitivity. Here, we distinguish between a current and a voltage bias, where the former was assumed in figure 2.9. Maximizing sensitivity in this mode is done by keeping the flux constant through a feedback loop at the steepest point in the  $V\Phi$ -curve, which is referred to as the working point. This allows for the largest possible voltage change  $\Delta V$  at a given flux change  $\Delta\Phi$ . Similarly, at a voltage bias the working point (WP) will mark the steepest point in the  $I\Phi$ -curve. To quantify this, we introduce the transfer coefficients at the working point

$$V_\Phi \equiv \left| \frac{\partial V}{\partial \Phi} \right|_{\text{WP}} \quad (2.33)$$

$$I_\Phi \equiv \left| \frac{\partial I}{\partial \Phi} \right|_{\text{WP}} . \quad (2.34)$$

As mentioned above, at  $I_b \approx 2I_c$  (current bias) the amplitude of the voltage modulation is maximal. This is modified for practical SQUIDS, where thermal fluctuations can't be neglected. The resulting thermal current  $I_{\text{th}}$  causes a rounding of the edge at  $I_b = 2I_c$  (figure 2.9 (left)), thereby reducing  $\Delta V_{\max}$  and  $V_\Phi$  [Iva68]. To minimize this effect, numerical simulations were made that lead to the condition [Cla88]

$$\frac{I_c}{5} \geq I_{\text{th}} \equiv \frac{2\pi k_B T}{\Phi_0} . \quad (2.35)$$

A lower bound for  $I_c$  at  $T = 4.2\text{ K}$  will, therefore, be approximately  $1\text{ }\mu\text{A}$ . This effect shifts the current  $I_{b,\max}$ , at which the voltage swing is maximal, according to [Dru96b] by a temperature correction factor leading to

$$I_{b,\max} \approx 2I_c(1 - \sqrt{\Gamma/\pi}) , \quad (2.36)$$

where  $\Gamma$  is defined as  $\Gamma = I_{\text{th}}/I_c$ . Lastly, the thermal current can also be used to set an upper limit to the SQUID inductance. We can define a thermal inductance  $L_{\text{th}} = \Phi_0/(2I_{\text{th}})$  for the thermal current inducing half a flux quantum. This should be significantly larger than the SQUID inductance  $L_s$  to minimize the impact of these thermal fluctuations. Again, simulations provide a constraint for optimization, giving the relation [Cla88]

$$5L_s \leq L_{\text{th}} \equiv \frac{\Phi_0}{2I_{\text{th}}} = \frac{\Phi_0^2}{4\pi k_B T} . \quad (2.37)$$

For  $T = 4.2\text{ K}$  we would obtain  $L_s \leq 1\text{ nH}$ , which is typically fulfilled for practical dc-SQUIDs.

#### 2.2.4 Noise

The above-mentioned energy sensitivity, also called spectral noise energy density or energy resolution, is defined as the flux noise per SQUID inductance  $L_s$  and is typically expressed through a power spectral density as

$$\epsilon(f) = \frac{S_\Phi(f)}{2L_s} . \quad (2.38)$$

This conveniently allows to compare noise properties from SQUIDs with different loop inductances [FS24]. The flux noise power spectral density  $S_\Phi(f)$  is typically calculated from the voltage noise using the transfer coefficient introduced above:

$$S_\Phi(f) = \frac{S_V(f)}{V_\Phi^2} . \quad (2.39)$$

The flux noise in SQUIDs is typically separated into a frequency-independent white noise at higher frequencies and a low frequency  $1/f$ -noise component [Koc07]. To derive an expression for  $S_V(f)$  we consider the white noise only, limiting ourselves to higher frequencies to avoid any significant influence of  $1/f$ -noise. For this we need to distinguish between the noise stemming from the bias current and the circulating current inside the SQUID-loop. The former is affected by the total resistance of the

SQUID which corresponds to  $R_s/2$  for  $I \gg I_{b,\max}$  and to the dynamic resistance  $R_{\text{dyn}} = \partial V / \partial I$  for operation at the working point. The circulating current, however, is affected by two resistances in series, i.e.  $2R_s$ . By taking into account noise inducing resonances due to the simulated optimal values of  $\beta_C \approx 1$  and  $\beta_L \approx 1$  that exhibit hysteretic behavior, both noise currents can be used to obtain the voltage noise given by [Tes77], [Bru82]

$$S_V(f) = \frac{4k_B T}{R_s} \left[ 2R_{\text{dyn}}^2 + \frac{L_s^2 V_\Phi^2}{2} \right] . \quad (2.40)$$

Here, we used the approximation  $V_\Phi \approx \frac{I_c R_s}{\Phi_0/2} \approx \frac{R_s}{L_s}$  valid for  $\beta_L \approx 1$ . This approximation leads to the expression  $R_{\text{dyn}} \approx \sqrt{2}R_s$  for the dynamical resistance, which can be used to rewrite equation 2.40 to

$$S_V(f) = 18k_B T R_s . \quad (2.41)$$

Inserting this into equations 2.39 and 2.38 leads to the expression

$$\epsilon(f) \approx 16k_B T \sqrt{\frac{L_s C}{\beta_C}} \approx 16k_B T \sqrt{L_s C} \quad \text{for } \beta_C \approx 1 . \quad (2.42)$$

For a moderately damped dc-SQUID ( $\beta_C \approx 1$ ) at a temperature  $T = 100 \text{ mK}$  and with a suitably small inductance  $L_s = 100 \text{ pH}$  we would, therefore, obtain a theoretical value of  $\epsilon \approx 0.32 \text{ h}$ .

Finally we will look at the total flux noise spectrum by considering the  $1/f$  component, which is added to the above derived white noise dominating in the low frequency regime. The origin of this component is still unclear [Kem16]. It is parameterized by a power density at 1 Hz, giving a total noise power density of

$$S_\Phi(f) = S_{\Phi,w} + \frac{S_{\Phi,1/\text{f}}(1 \text{ Hz})}{f^\alpha} , \quad (2.43)$$

with  $S_{\Phi,w}$  being the white noise component and  $\alpha$  ranging from 0.5 to 1 [Dru11], which can be determined by experiment.

### 2.2.5 Parasitic Resonances

We have seen that optimizing properties of dc-SQUIDs requires careful fine-tuning of various parameters to ensure high sensitivity and low noise at the same time. Another

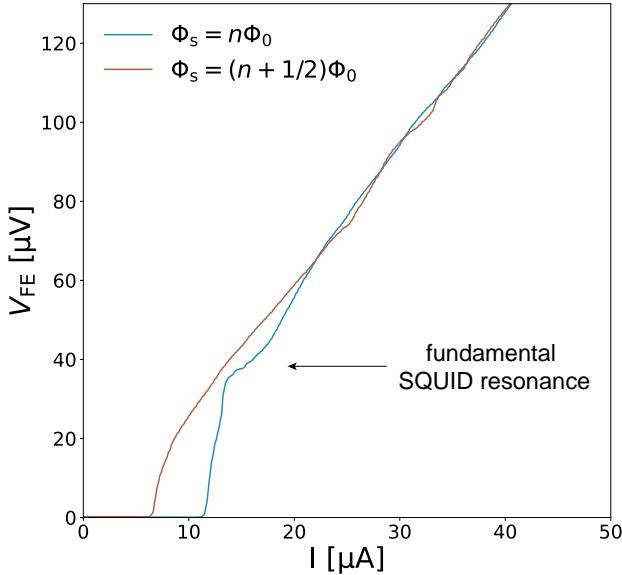


Figure 2.10: Measured IV-characteristic of a dc-SQUID developed within the scope of this thesis. The first current step at  $V_s \approx 40 \mu\text{V}$  corresponds to the fundamental SQUID resonance for  $L_s = 119 \text{ pH}$  and  $C = 0.95 \text{ pF}$ .

constraint in this regard involves the presence of resonances in the system that we need to take into account. As discussed in subsection 2.1.3, a Josephson contact also represents a parallel plate capacitor with capacitance  $C$ . Here we take into account the total capacitance of  $C/2$ , due to the series connection of the two junctions within the SQUID loop. The SQUID with its loop inductance  $L_s$  will, therefore, form an  $LC$  resonator that can be excited if the Josephson currents oscillate with the resonance frequency  $f_{L_sC} = 1/(2\pi\sqrt{L_sC/2})$ . This is fulfilled for the condition

$$\frac{V_s}{\Phi_0} = \frac{1}{2\pi\sqrt{L_sC/2}} , \quad (2.44)$$

where  $V_s$  denotes the voltage drop across the SQUID that is associated with the Josephson frequency  $f_J = V_s/\Phi_0$  resulting from the ac Josephson effect. With optimal  $\beta_C$  and  $\beta_L$  values the voltage corresponding to this resonance frequency will move towards the vicinity of the working point, thus becoming relevant for the SQUID's performance. This so-called *fundamental SQUID resonance* negatively affects the operation range by manifesting itself through a current step in the IVC, as shown in figure 2.10. As a result, the IV curves for integer and half integer flux quanta intersect as a result multiple times, due to higher harmonics of the resonance causing additional current steps [Cla04]. The voltage swing  $\Delta V$  and consequently  $V_\Phi$  are limited by this resonance, which favors the choice of small values for  $L_s$  and  $C$ .

Resonance inducing structures are, however, necessary for practical reasons. Those

include the need for effectively coupling external flux changes into the SQUID in order to take advantage of its high sensitivity. For this an input coil with inductance  $L_i$  is usually fabricated on top of the SQUID loop, separated by an insulating layer to ensure the coupling to be solely inductive (see section 3.1). The SQUID can, therefore, be used as a current sensor by converting small current signals in the input coil into small magnetic flux changes in the SQUID. This additional coil, however, provides a parasitic capacitance  $C_p$  resulting from its fabrication on top of the SQUID loop. Consequently, rf currents can couple from one system into the other, causing further parasitic effects. Particularly, problematic  $LC$  resonances arise from the added inductance  $L_i$  and capacitance  $C_p$ , namely the  $f_{L_s C_p}$  and  $f_{L_i C_p}$  resonance. The latter can be excited for  $f_J = f_{L_i C_p}$ , since the Josephson currents are able to couple into the input coil. The parasitic capacitance  $C_p$  and consequently both resonance frequencies depend on geometrical factors such as the length of the input coil, as well as the widths and heights of the individual components that form the resonator. It has been shown, that these resonances lead to the energy sensitivity increasing proportional to  $\sqrt{1 + 2C_p/C}$ , as long as  $C_p/C \leq 2$ . For  $C_p/C \geq 2$  the energy sensitivity saturates due to the resonance frequencies shifting below the working point [Ryh92]. It is, therefore, desirable to minimize  $C_p$  as much as possible in order to reduce the resulting voltage noise.

The fabrication of the input coil on top of the SQUID loop also represents a microstrip transmission line, consisting of a conductor carrying the signal (input coil) and a ground plane (SQUID loop), separated by a dielectric layer. This forms a waveguide allowing electromagnetic waves to propagate alongside it, which undergo reflections where impedance mismatches occur. These would arise whenever the input coil leaves the underlying ground plane. Standing waves occur if the length  $l$  of the signal carrying line above the ground plane corresponds to an integer multiple of half the wavelength of the Josephson frequency. This is consequently also called the  $\lambda/2$  resonance. The corresponding resonance frequency

$$f_l = \frac{mc_{\text{str}}}{2l} , \quad (2.45)$$

where  $m \in \mathbb{N}$ , depends on  $l$  and the material-dependent wave propagation velocity  $c_{\text{str}}$ . This resonance behavior also emerges if we consider the SQUID loop to be the signal carrying line with the input coil acting as the corresponding ground plane. Here it is again possible to move the resonance further away from the operation frequency at the working point by choosing adequate geometric proportions, e.g. by varying the length of the input coil or the SQUID loop. These methods to mitigate the influence of resonances can be complemented by direct damping through attenuators, which will be discussed in section 4.2. In chapter 4 we will cover how

various parameters are chosen in the SQUID design to suppress and avoid possible resonances, given the constraints we derived in subsection 2.2.3 and 2.2.4.



### 3. Experimental Methods

So far we discussed general aspects of dc-SQUIDs and how their working principle allows for highly sensitive measurements of magnetic flux changes. As already briefly seen in subsection 2.2.5, when it comes to practical SQUIDs many theoretical considerations regarding parameter optimization need to be reevaluated to account for usability in practical experiments. We begin this chapter with general concepts of a practical SQUID design and introduce a typical low-noise setup with a room temperature readout electronics. In this working group, SQUIDs are mainly developed for the readout of *Metallic Magnetic Calorimeters (MMCs)* (see section 3.3). We will see in the following how those SQUIDs need to be designed to optimize their coupling to these detectors.

#### 3.1 Practical dc-SQUIDs

The SQUIDs developed in this working group are primarily used as current sensors for the MMC readout by coupling the detected signals from the *pickup coil* of the MMC to the parallel input coil of the SQUID. The need for this input coil entails a parasitic capacitance resulting in several resonances, as discussed in subsection 2.2.5. To achieve high inductive coupling between input coil and SQUID loop with inductances  $L_i$  and  $L_s$ , respectively, it is necessary to fabricate them closely on top of each other, only separated by a thin insulating layer. The coupling strength is given by the dimensionless parameter

$$k_{is} = \frac{M_{is}}{\sqrt{L_i L_s}} \quad , \quad (3.1)$$

where  $M_{is} = \Delta\Phi_s / \Delta I_i$  is the mutual inductance, describing how much flux  $\Delta\Phi_s$  is generated in the SQUID loop for a current change  $\Delta I_i$  in the input coil. This allows us to define the so-called coupled energy sensitivity  $\epsilon_c(f)$  with respect to the input coil, which by using equation 2.38 is given as

$$\epsilon_c(f) = \frac{\epsilon(f)}{k_{is}^2} = \frac{L_i S_{I,i}}{2} \quad . \quad (3.2)$$

This expression refers to the apparent current noise  $S_{I,i} = S_{\Phi_s} / M_{is}^2$ , which is generated by the flux noise from the SQUID loop through the coupling  $M_{is}$ . A strong

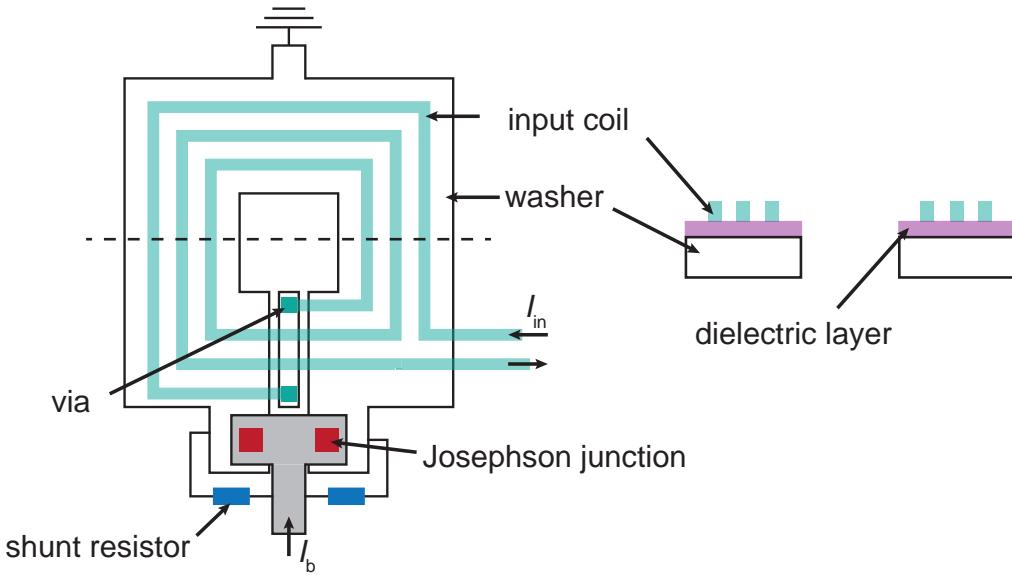


Figure 3.1: Schematic drawing of a typical planar thin-film dc-SQUID. The SQUID loop is realized as a square *washer*-geometry interrupted by a narrow slit, only connected at the junction area. A thin insulating layer separates the washer from the planar multi-turn input coil above. Left: View from the top. Right: Cross section marked by the dashed line.

coupling can be achieved by the commonly used square *washer*-geometry with a planar input coil [Jay81], as shown in figure 3.1.

Here, the SQUID loop is represented by the washer, whereas each turn of the input coil is located on top of it to maximize the coupling between both coils. A cross section of this setup is depicted in figure 3.1 (right), showing the insulating dielectric layer separating each coil. The washer is intersected by a slit, which starts at the square hole in the middle and ends at the remotely situated junction area that connects each side of the loop. The total inductance of the SQUID loop can be calculated by adding the dominating washer hole inductance  $L_h$ , the slit inductance  $L_{sl} \approx 0.3 \text{ pH}/\mu\text{m}$  given for small slit dimensions and the much smaller parasitic inductance  $L_j$  associated with the junction area, giving [Ket91]

$$L_s = L_h + L_{sl} + L_j . \quad (3.3)$$

The latter is referred to as parasitic due to its position outside of the input coil, thus not contributing to the coupling. By neglecting  $L_{sl}$  and  $L_j$ , we can approximate the washer inductance in the limit of  $d \ll w$  to  $L_s \approx L_h \approx 1.25\mu_0 d$ , where  $d$  and  $w$  are the inner and outer side lengths, respectively [Jay81]. This is a reasonable result considering that the supercurrent will only flow along the inner edge of the washer

[Ket82], thereby being independent of the outer side length  $w$ . The effective area  $A_{\text{eff}}$  of the SQUID loop has been calculated to  $A_{\text{eff}} \approx dw$  [Ket85], showing that this geometry allows for high sensitivity while keeping the SQUID inductance small. The input coil inductance on the other hand can be approximated by  $L_i = L_{\text{str}} + n^2 L_s$ , where  $L_{\text{str}}$  is the stripline inductance (see section 5.2) and  $n$  is the number of input coil turns [Jay81]. The dc-SQUID designs used in this working group, however, are too complex to provide such analytical expressions and therefore need to be calculated numerically using simulation softwares such as *InductEX*<sup>1</sup>.

### 3.1.1 Gradiometer

The high flux sensitivity of a SQUID makes it prone to detect unwanted magnetic bias fields and/or gradients that may be present during its operation. Typical SQUIDs are, therefore, built in a gradiometric design to counteract this effect [Ket78]. A first order gradiometer consists of two identical conducting loops connected in series or parallel, with opposing orientation as shown in figure 3.2 (left, middle). Under the presence of a homogeneous bias field  $\mathbf{B}$  in  $x$ -direction (perpendicular to the gradiometer plane), this configuration produces a zero net current after a field change  $\Delta B_x$ , due to the opposing currents induced in each turn. To also achieve the same effect for a field gradient  $\frac{\partial \mathbf{B}}{\partial z}$  or  $\frac{\partial \mathbf{B}}{\partial y}$ , a second order gradiometer composed of four loops in series or parallel is required, see figure 3.2 (right), where only the currents induced in the upper loops are drawn for the sake of clarity. In order to incorporate this into a practical SQUID, the input coil and the SQUID loop will consist of four serial and parallel turns, respectively. This configuration enables to combine a small SQUID inductance with a large input coil inductance while maintaining a strong coupling between the two, as each turn of both coils can be produced with similar dimensions. The low SQUID inductance results from the reciprocal summation over each loop inductance  $L_l$  due to the parallel connection, giving

$$L_s = \frac{L_l}{4} . \quad (3.4)$$

Whereas a serial gradiometer gives

$$L_i = 4L_l \quad (3.5)$$

for the input coil, provided it consists of only one turn. This gradiometric setup allows for adapting the input coil to the pickup coil of an MMC by choosing a large enough inductance  $L_i$ , which will be discussed in chapter 4.

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<sup>1</sup>SUNMagnetics, 15 De Beer Rd, Stellenbosch Central, Stellenbosch, 7600, South Africa

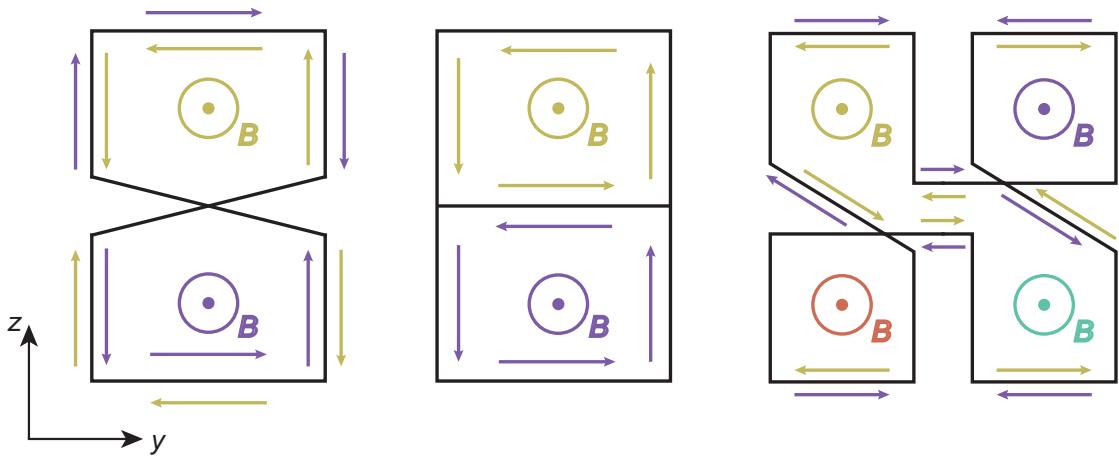


Figure 3.2: Schematic examples of a gradiometric dc-SQUID configuration threaded by a homogeneous magnetic field  $\mathbf{B}$ . A first order gradiometer can be realized by either connecting two loops in series (left) or in parallel (middle). A magnetic field change  $\Delta B_x$  induces two opposing currents that cancel each other out. Right: Second order gradiometer consisting of four loops connected in series. This geometry results in a net zero current also for an applied field gradient  $\frac{\partial \mathbf{B}}{\partial z}$ . For the sake of clarity, the currents induced in the two bottom loops are omitted.

### 3.2 Operation of a dc-SQUID

The SQUIDs developed within the scope of this thesis are produced in the institute's cleanroom and then tested both in a single- and a two-stage setup (see subsection 3.2.2). The measurements described in chapter 5 are carried out at either  $T = 4.2\text{ K}$  in a liquid helium transport vessel or in a dilution cryostat BF-LD250 from BlueFors<sup>2</sup> with a base temperature of  $T = 10\text{ mK}$ . The former submerges the SQUIDs in liquid helium via a dipstick, which provides a sample holder for PCBs. The SQUIDs are glued onto those PCBs and are electrically connected to them via aluminum bond wires, by utilizing the bond pads shown in figure 4.4. The sample holder is equipped with both a superconducting (niobium) and a soft-magnetic Cryoperm shield to suppress external magnetic fields. The read-out is done by a broadband SQUID electronics of the type XXF-1 (see subsection 3.2.1) to both supply the necessary bias and ramp current signals to the SQUID as well to provide the FLL feedback. The SQUID electronics is controlled via software and the voltage output observed on a Keysight InfiniiVision<sup>3</sup> oscilloscope.

<sup>2</sup>BlueFors Cryogenics Oy, Arinatie, 00370 Helsinki, Finnland

<sup>3</sup>Keysight Technologies Deutschland GmbH, Herrenberger Straße 130, 71034 Böblingen

### 3.2.1 Flux-Locked Loop

We have seen in section 2.2.2 that the periodic  $V\Phi$ -characteristic provides an approximately linear dependence at  $\Phi = (2n + 1)\Phi_0/4$ , which only holds for  $\Delta\Phi \approx \Phi_0/4$ . This restricts the dynamic range greatly, as the linearity vanishes for larger flux changes and for  $\Delta\Phi > \Phi_0/2$  the voltage even becomes ambiguous.

The standard readout method involves a flux feedback circuit to maintain the operation at the working point independently of the flux [Dru02]. This so-called flux-locked loop (FLL) readout technique first amplifies the output signal of the SQUID  $V_s$  with a differential amplifier operated at room temperature, where the voltage  $V_b$  corresponding to the working point is provided by a voltage source on the second amplifier input. This voltage compensation at the working point ensures that only variations  $\Delta V = V_s - V_b$  that correspond to the flux change  $\Delta\Phi$  are amplified. The signal is then fed into an integrator, which integrates it over time and thus creates a changing output voltage  $V_{\text{out}}$ . By connecting a feedback resistance  $R_{\text{fb}}$  to the output circuit, a feedback current  $I_{\text{fb}}$  is generated that flows to a feedback coil with inductance  $L_{\text{fb}}$ . This coil is coupled to the SQUID analogous to the input coil (see chapter 4), but with the opposite orientation. A compensation flux  $-\Delta\Phi$  is generated until the initial flux change is fully canceled out, i.e.  $V_s - V_b \rightarrow 0$ . The integrator will, therefore, approach a constant value due to the vanishing voltage at the input circuit. This voltage signal is proportional to the current that is needed to completely compensate for the input signal  $\Delta\Phi$ , leading to the relation

$$V_{\text{out}} = \frac{R_{\text{fb}}}{M_{\text{fb}}} \Delta\Phi . \quad (3.6)$$

A schematic for this readout process is shown in figure 3.3. With this setup the SQUID is used as a null-detector that allows for the linearization of the quantity of interest, while also providing a large dynamic range. A state-of-the-art, low-noise SQUID readout electronics by the company Magnicon<sup>4</sup> of the type XXF-1, which is used in this working group, provides the necessary current and voltage sources, as well as the room temperature amplifiers within the FLL circuit described above. The amplifier of this SQUID electronics exhibits an intrinsic voltage noise of  $\sqrt{S_{V,\text{el}}} \approx 0.33 \text{ nV}/\sqrt{\text{Hz}}$  and intrinsic current noise of  $\sqrt{S_{I,\text{el}}} \approx 2.6 \text{ pA}/\sqrt{\text{Hz}}$  [Dru06]. The included current source for the feedback coil couples current noise into the SQUID. In the work of [Kaa20], this contribution was determined to  $\sqrt{S_{I,\Phi}} = 2.4 \text{ pA}/\sqrt{\text{Hz}}$  for the white noise and  $\sqrt{S_{I,\Phi}} = 10 \text{ pA}/\sqrt{\text{Hz}}$  for the  $1/f$  component at 1 Hz. Furthermore, a large bandwidth of 6 MHz is provided to ensure high sensitivity for short signal rise times. Adding these terms to the intrinsic flux noise  $S_{\Phi_s}$  leads to the total

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<sup>4</sup>Magnicon GmbH, Barkhausenweg 11, 22339 Hamburg

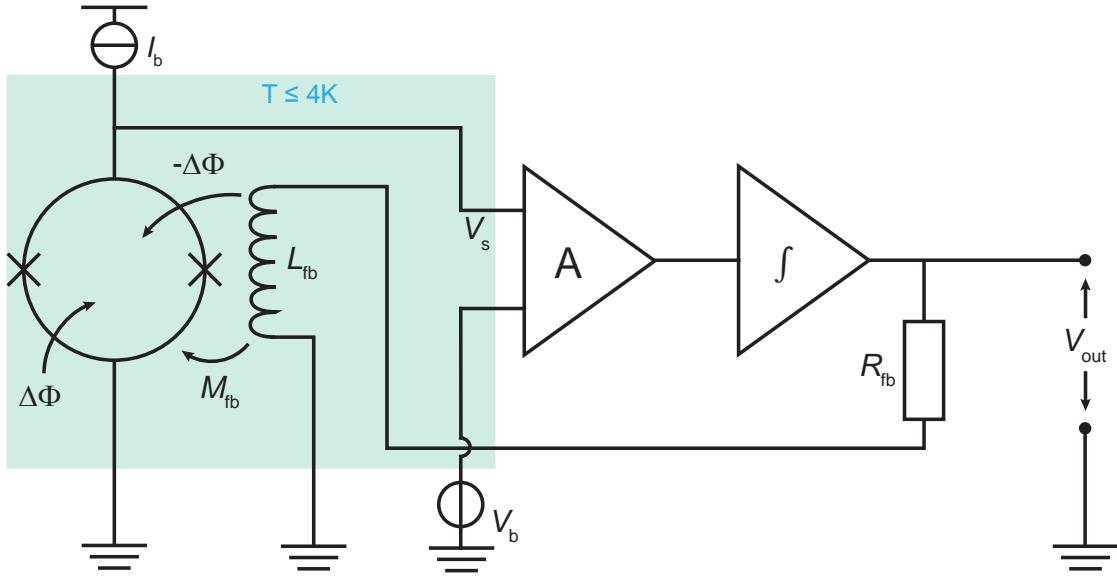


Figure 3.3: Schematic circuit diagram of a flux-locked loop dc-SQUID readout. The amplified and integrated SQUID voltage signal  $V_s$ , which is caused by a detected flux change  $\Delta\Phi$ , is fed back to a feedback coil with inductance  $L_{\text{fb}}$ , creating a compensating flux  $-\Delta\Phi$ . This enables the operation at the working point, while flux changes far greater than  $\Phi_0/4$  can be linearized.

apparent flux noise in the SQUID, which is expressed as the spectral power density

$$S_{\Phi_s, \text{SQ}} = S_{\Phi_s} + \frac{S_{V, \text{el}}}{V_{\Phi_s}^2} + \frac{S_{I, \text{el}}}{I_{\Phi_s}^2} + S_{I, \Phi} M_{\text{fb}}^2 . \quad (3.7)$$

Typical values for the transfer coefficients of SQUIDs produced within the scope of this thesis, are  $V_{\Phi_s} = 80 \mu\text{V}/\Phi_0$  and  $I_{\Phi_s} = 20 \mu\text{A}/\Phi_0$ , leading to the SQUID electronics having a total noise contribution of  $4.13 \mu\Phi_0/\sqrt{\text{Hz}}$ . The intrinsic noise of the SQUIDs produced in this working group, however, typically reaches values of  $\sqrt{S_{\Phi_s}} \leq 1 \mu\Phi_0/\sqrt{\text{Hz}}$ , depending on frequency and temperature. The amplifier noise, therefore, dominates the noise level, thereby deteriorating the signal to noise ratio. To avoid this effect, a second SQUID is typically added to act as a low temperature amplifier [Wel93]. This method significantly reduces the apparent flux noise in the detector SQUID, which is crucial for MMC readout, as the readout noise should not be larger than the intrinsic noise of the MMC.

### 3.2.2 Two-Stage Configuration

Implementing a low temperature amplifier can be realized through a second stage SQUID, situated between the first stage (detector) SQUID and the room temperature amplifier, as depicted in figure 3.4. Here, the subscripts 's' and 'x' refer to the first and second stage, respectively. The first stage SQUID, also referred to as a front-end SQUID, is typically operated in a voltage bias for this two-stage setup. This can be achieved by connecting a gain resistor  $R_g$  in parallel to the series connection of the front-end and the input coil of the amplifier SQUID. If a bias current  $I_{b,s}$  is injected into the circuit, all the current will flow through the Front-End, as long as it stays superconducting. Once it becomes normal conducting by further increasing  $I_{b,s}$ , the current will start shifting to  $R_g$ , whose resistance is chosen to be much smaller than the dynamic resistance  $R_{dyn}$  of the first stage SQUID, until most of the current flows through  $R_g$ . At this point, the resulting voltage across both components becomes approximately independent of the flux in the front-end SQUID. This behavior can be visualized through a loadline created by the parallel resistance, which intersects the IVC of the Front-End. The loadline voltage  $V_s$  between both extremal IV curves will then remain nearly constant, as the slope is given by the small gain resistance  $R_g$ . If a detector signal is now coupled into the Front-End through the input coil with mutual inductance  $M_{is}$ , the current in the SQUID will move along the loadline in the  $I\Phi$ -plane, corresponding to the externally induced flux  $\Phi_s$ . The attached input coil of the second stage SQUID would experience these current changes, hence creating a flux change  $\Delta\Phi_x$  in the amplifier SQUID, which is operated in a current bias. To maximize the amplification, the second stage SQUID is typically realized as a  $N$ -SQUID series array consisting of  $N$  serially connected identical SQUID cells. This results in a large voltage drop across the array, given by  $V_{array} = NV_{cell}$ . Analogous to the single stage readout, the signal will then be amplified at room temperature and fed back to the front-end SQUID via a feedback coil with mutual inductance  $M_{fs}$  to compensate for the initial flux change  $\Delta\Phi_s$ . An additional feedback coil with mutual inductance  $M_{fx}$ , spanning identically across every SQUID cell, allows to add an additional flux offset to adjust the relative phase of front-end and array SQUID. The resulting two-stage  $V_x\Phi_s$ -characteristic will strongly depend on the flux gain defined as

$$G_\Phi = \frac{\partial\Phi_x}{\partial\Phi_s} = \frac{M_{ix}}{R_g + R_{dyn}} V_{\Phi_s} \approx \frac{M_{ix}}{R_{dyn}} V_{\Phi_s} , \quad (3.8)$$

which relates the flux change induced in the second stage SQUID with a given flux change in the front-end. For  $\Delta\Phi_x = \Delta I_s M_{ix} > \Phi_0/2$ , additional minima and maxima emerge in the  $V_x\Phi_s$ -characteristic. These start to overlap for  $\Delta\Phi_x > \Phi_0$ , thereby creating multiple working points that prevent a practical FLL operation. This sets

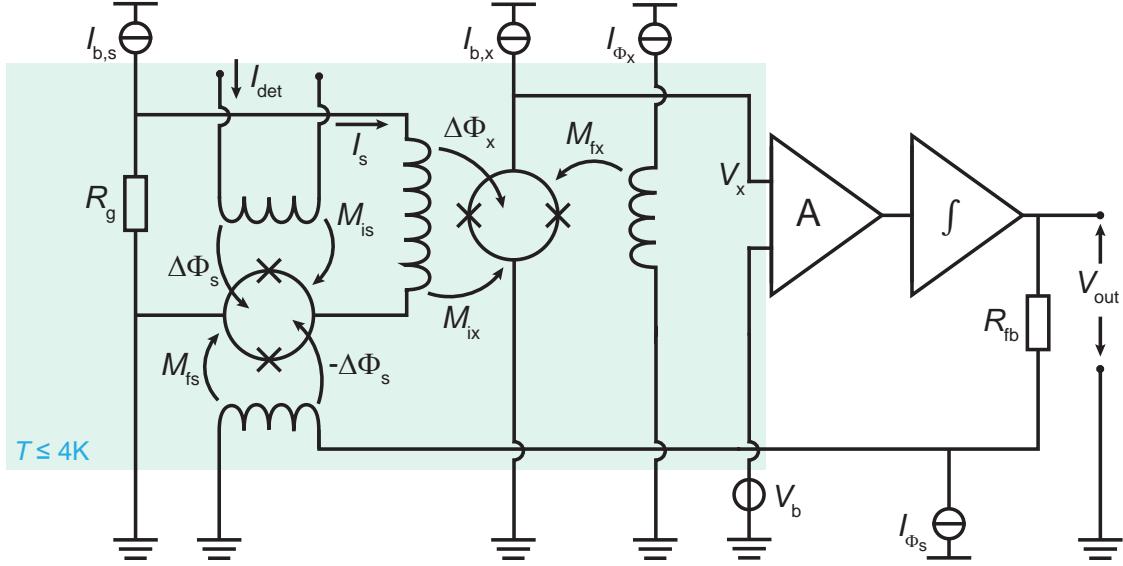


Figure 3.4: Schematic circuit diagram of a two-stage dc-SQUID readout. The first stage consisting of the detector SQUID is connected in parallel with a gain resistor  $R_g$ . The detector signal  $I_{\text{det}}$  induces a current change  $\Delta I_s$  in the front-end and, therefore, a flux change  $\Delta\Phi_x$  in the second stage SQUID, which acts as a low temperature preamplifier. In FLL mode, the signal will be fed back to compensate for the initial flux change  $\Delta\Phi_s$ , thereby linearizing the detector signal. Choosing an  $N$ -SQUID series array amplifies the voltage swing and transfer coefficient, thus significantly reducing the overall apparent flux noise in the detector SQUID.

an upper limit for the flux gain, however, it should be chosen as large as possible to reduce the apparent flux noise of the Front-End SQUID. An optimal flux gain has been calculated to  $G_\Phi \approx \pi$ , corresponding to  $\Delta\Phi_x \approx \Phi_0/2$  [Dru96a].

The two-stage setup contributes additional noise sources to equation 3.7, namely the gain resistor and the amplifier SQUID. However, the resulting conversion to the flux  $\Phi_s$  of the front-end SQUID significantly reduces the influence of the room temperature amplifier, which in turn strongly improves the overall signal to noise ratio. The total apparent flux noise of the Front-End then reads [Dru96a]

$$S_{\Phi_s, \text{SQ}} = S_{\Phi_s} + \frac{4k_B T R_g}{G_\Phi^2 (R_g + R_{\text{dyn}})^2} M_{\text{ix}}^2 + \frac{S_{\Phi_x}}{G_\Phi^2} + \frac{S_{V,\text{el}}}{G_\Phi^2 V_{\Phi_x}^2} + \frac{S_{I,\text{el}}}{G_\Phi^2 I_{\Phi_x}^2} + \frac{S_{I,\Phi_x} M_{\text{fx}}^2}{G_\Phi^2} . \quad (3.9)$$

The second term describes the Nyquist current noise caused by the gain resistor, which becomes negligible with a voltage biased front-end SQUID where  $R_g \ll R_{\text{dyn}}$ . The low temperature amplifier (array), the feedback coil current source as well as the room temperature amplifier contributions are reduced by the flux gain, which can't

be chosen arbitrarily large as mentioned above. However, using a SQUID array for the second stage increases the voltage swing and thus the transfer coefficient by an  $N$ -fold, i.e.  $V_{\Phi_x} = NV_{\Phi_{\text{cell}}}$ , where the subscript 'cell' refers to a single array SQUID cell. Consequently, the total noise level can be further reduced by choosing a high number  $N$  of SQUID cells. Here it is noteworthy, however, that these considerations only account for the magnetic flux in a single cell, as otherwise the transfer coefficient would remain constant [Sta93, Fog93]. Using equation 2.39 we obtain for the SQUID array flux noise

$$\sqrt{S_{\Phi_x}} = \frac{\sqrt{S_{V_x}}}{V_{\Phi_x}} = \frac{\sqrt{NS_{V_{\text{cell}}}}}{NV_{\Phi_{\text{cell}}}} = \frac{1}{\sqrt{N}} \sqrt{S_{\Phi_{\text{cell}}}} , \quad (3.10)$$

hence the intrinsic noise of the second stage gets reduced by a factor of  $\frac{1}{\sqrt{N}}$  [Sta93]. This also has a consequence for the coupled energy sensitivity of the SQUID array, which is calculated by summing the array flux noise over all  $N$  cells, giving

$$\epsilon_{c,x} = N \frac{S_{\Phi_x}}{2L_{\text{cell}} k_{i,\text{cell}}} = \frac{S_{\Phi_{\text{cell}}}}{2L_{\text{cell}} k_{i,\text{cell}}} . \quad (3.11)$$

Where the parameter  $k_{i,\text{cell}}$  denotes the coupling of a cell with inductance  $L_{\text{cell}}$  to its respective input coil segment. Connecting  $N$  SQUIDs in series does, therefore, not affect the energy sensitivity, provided that  $k_{i,\text{cell}}$  remains constant across the array. The arrays produced in this working group either contain 16 [Kem15] or 18 [Kra23] cells. Applying this to equation 3.9, we would reduce the above-mentioned contribution of the room temperature amplifier flux noise of  $4.13 \mu\Phi_0/\sqrt{\text{Hz}}$  to  $0.29 \mu\Phi_0/\sqrt{\text{Hz}}$  and  $0.26 \mu\Phi_0/\sqrt{\text{Hz}}$  for  $N = 16$  and  $N = 18$ , respectively. Provided an optimal flux gain of  $G \approx \pi$ , the noise contribution would further decrease to  $0.09 \mu\Phi_0/\sqrt{\text{Hz}}$  or  $0.08 \mu\Phi_0/\sqrt{\text{Hz}}$ , well below the typical noise level of a front-end SQUID.

Besides the improvement of the noise behavior, a major advantage of a two-stage setup consists of the low power dissipation produced by the front-end SQUID. As opposed to the current bias operation in a single-stage setup, the Front-End is held at a constant, low voltage, such that the dissipated power given by  $P = UI$  only consists of approximately 1 nW [Dru07].

### 3.3 Metallic Magnetic Calorimeters

Low temperature detectors used in this working group exhibit detector arrays consisting of metallic magnetic calorimeters, which are able to absorb and detect photons (X- or Gamma-rays) or heavy particles. The focus of this thesis is to improve the dc-SQUID based MMC readout, which is why we briefly cover the working principle

of such calorimeters.

A general overview of the detector setup can be seen in figure 3.5 [Fle05]. It consists of a particle absorber, which, following the law of energy conservation, converts the energy  $E$  of an incoming particle into a change in temperature  $\Delta T$  in the absorber. A paramagnetic sensor is directly attached, placed in a weak magnetic field. Strong thermal coupling allows for an effective temperature increase in the sensor, which is given by the total heat capacity  $C_{\text{tot}}$  of both the sensor and the absorber, resulting in

$$\Delta T \approx \frac{E}{C_{\text{tot}}} . \quad (3.12)$$

Not shown in figure 3.5 is a thermal bath weakly linked to the sensor, which reverts the temperature change by an exponential decay of the signal towards the equilibrium temperature. Due to its paramagnetic nature, the sensor changes its magnetization  $M(T)$  upon a temperature rise, according to the Curie law  $M \propto \frac{1}{T}$ . This induces a magnetic flux change  $\Delta\Phi$  proportional to  $\Delta M$ , which is detected by a meander-shaped pickup coil that is strongly coupled to the sensor. With equation 3.12 we can write for the induced flux in the meander-shaped pick-up coil

$$\Delta\Phi \propto \Delta M \approx \frac{\partial M}{\partial T} \frac{E}{C_{\text{tot}}} , \quad (3.13)$$

provided the deposited energy  $E$  is small. Evidently, it is desirable to both reduce the heat capacity and increase the magnetization gradient in order to maximize the flux change  $\Delta\Phi$ . Another important aspect are short signal rise times that can be realized by a high thermal conductivity of the sensor and the absorber, which is particularly crucial for coincidence measurements. All these material-dependent requirements are fulfilled for the normal conducting metal gold as an absorber material and gold or silver for the sensor. The latter needs to be doped with a few hundred ppm of the rare-earth metal erbium in order to obtain paramagnetic properties. State-of-the-art MMCs of this kind are successfully developed in this working group, providing an exceptional energy resolution of 1.6 eV at 6 keV, a large energy bandwidth and fast signal rise times well below 100 ns [Kem18].

Typically two meander-shaped pickup coils, each with an inductance  $L_m$ , are connected as a parallel gradiometer and are located beneath their respective sensor pixel, with a connection to the input coil of the SQUID. A parasitic inductance  $L_{\text{par}}$  needs to be taken into account due to the bonding wires between the SQUID and the detector (figure 3.5, right). A constant bias current  $I_0$  is injected into the pickup coils to provide the weak bias field needed for the paramagnet. If a particle gets detected by one of the absorbers, a flux change  $\Delta\Phi$  and thus a current is induced

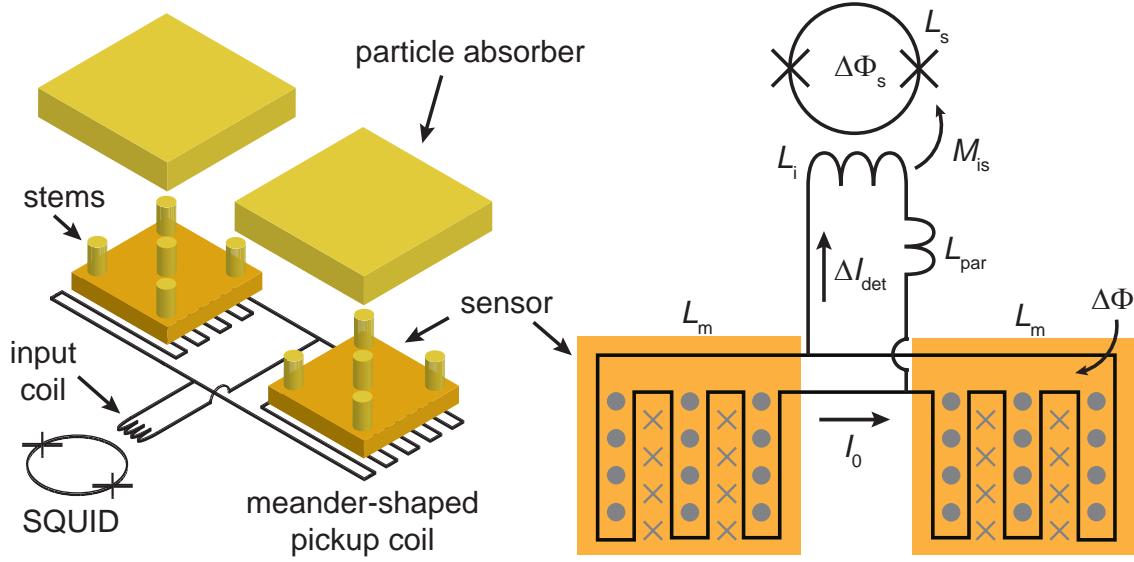


Figure 3.5: Schematic depiction (left) and circuit diagram (right) of a common MMC channel consisting of two pixels. The planar particle absorber is fabricated on top of the paramagnetic sensor, separated by cylinder-shaped stems to prevent energy losses through athermal phonons [Fle09], while still providing good thermal coupling. Beneath the sensors two meander-shaped pickup coils connected in parallel detect the flux change  $\Delta\Phi$ , which can be precisely measured with a SQUID. The grey dots and crosses within the meander mark a magnetic field induced by a bias current  $I_0$ . The figure was adapted from [Bau22].

in the meander. By applying Kirchhoff's first law and knowing that magnetic flux in a closed superconducting loop is conserved [Lan60], we can calculate the resulting current signal in the input coil of the SQUID [Bur04]

$$\Delta I_{\text{det}} = \frac{\Delta\Phi}{L_m + 2(L_i + L_{\text{par}})} . \quad (3.14)$$

A significant advantage of this setup lies in the fact, that only one SQUID is needed to readout two pixels. The gradiometric geometry ensures, that depending on which absorber gets hit, the current  $\Delta I_{\text{det}}$  will flow in one direction or the other. Thus, every pixel can be assigned a polarity of the resulting flux in the SQUID loop, making them distinguishable. Together with the mutual inductance  $M_{\text{is}} = \Delta\Phi / \Delta I_{\text{det}}$  between the input coil and the SQUID loop, we can now introduce the flux-to-flux coupling

$$\frac{\Delta\Phi_s}{\Delta\Phi} = \frac{M_{\text{is}}}{L_m + 2(L_i + L_{\text{par}})} , \quad (3.15)$$

which by considering equation 3.1 is maximal for  $L_i = L_m/2 + L_{\text{par}}$ . This crucial

parameter is a measure for the flux transfer and gives, therefore, information about the quality of the readout setup. Consequently, it is desirable to choose the optimal input coil inductance  $L_i$  for a given pickup coil inductance  $L_p = L_m/2$ .

### 3.3.1 Extrinsic Energy Sensitivity

In this regard it is reasonable to discuss the noise behavior for such a readout setup. Since noise can couple in both directions, the intrinsic flux noise of the SQUID will also couple into the pickup coil, creating the apparent flux noise  $\sqrt{S_{\Phi_s,p}}$  in the detector. To quantify this noise influence, we define the extrinsic energy sensitivity with respect to a given pickup coil inductance  $L_p$  analogous to equation 2.38 as [Knu88]

$$\epsilon_p = \frac{S_{\Phi_s,p}}{2L_p} . \quad (3.16)$$

Here, it is necessary to point out that the description of the detector as a single coil with inductance  $L_p$  adds a factor of 2 to equations 3.14 and 3.15, such that the flux-to-flux coupling reads

$$\frac{\Delta\Phi_s}{\Delta\Phi_p} = \frac{M_{is}}{L_p + L_i + L_{par}} . \quad (3.17)$$

This advantage in coupling, however, does not outweigh the above-mentioned benefit to readout two pixels with one read-out electronic, as the gradiometric design also cancels out unwanted signals through changes in external magnetic field or substrate temperature [Fle05]. In the following we consider the case for a single pickup coil with inductance  $L_p = \frac{L_m}{2}$ . The presence of a flux transformer is accompanied with a shielding effect with respect to the SQUID loop, such that the geometric inductance is reduced to an effective inductance given by [Gro16]

$$L'_s = L_s(1 - k_{is}^2 s_i) , \quad (3.18)$$

where  $s_i = L_i/(L_i + L_p + L_{par})$  denotes the screening factor [Cla04], which reaches the value of 0.5 for the condition  $L_i = L_p + L_{par}$  that maximizes the flux-to-flux coupling. By using the relation  $\Delta\Phi_s/\Delta\Phi_p = \sqrt{S_{\Phi_s}}/\sqrt{S_{\Phi_s,p}}$ , we obtain with equations 3.16 to 3.18 the extrinsic energy sensitivity [Knu88]

$$\epsilon_p = \frac{S_{\Phi_s}(L_i + L_{par} + L_p)^2}{2L_p M_{is}^2} = \frac{S_{\Phi_s}(L_i + L_{par} + L_m/2)}{L_s k_{is}^2 s_i L_m/2} . \quad (3.19)$$

This parameter is minimized by maximizing the flux-to-flux coupling  $\Delta\Phi_s/\Delta\Phi_p$  as well as minimizing the intrinsic flux noise of the SQUID  $S_{\Phi_s}$  given by the expressions derived in subsection 2.2.4. Finding the minimum for the latter provides optimal values for the shunt resistors  $R_s$  and the SQUID loop inductance  $L_s$ , as we will see in the following.



## 4. dc-SQUID Design

The main objective for this thesis was the optimization of an existing front-end SQUID design for an improved coupling to the maXs100 detector developed in this working group. As we have seen in the previous section 3.3, adjusting  $L_i$  to the detector coil ensures the maximization of the flux-to-flux coupling and, therefore, minimizes the extrinsic energy sensitivity. The parasitic inductance  $L_{\text{par}}$ , that arises from the aluminum bonds between the SQUID and the detector, has been estimated to 0.5 nH [Hen17]. The previous SQUID design exhibits a design value of  $L_i = 1.64$  for the input coil inductance [Bau22], which, therefore, fulfills the condition  $L_i = L_p + L_{\text{par}}$  for the pickup coil inductance of the ECHo-100k detector of  $L_p = 1.14$  nH [Man21]. Other MMCs from this working group such as the 4k-pixel molecule camera MOCCA and the X-ray detector maXs100 require higher input inductances, as their pickup coil inductances are  $L_p = 8.8$  nH and  $L_p = 6.65$  nH, respectively. In [Bau22] SQUIDs with matching input inductances for the MOCCA and maXs100 detector were developed for the first time using an intermediary coupling transformer. These improved the calculated energy resolution  $\Delta E_{\text{FWHM}}$  of the detectors, although the effect was minimal for the latter. Specifically for the maXs100 detector, a different approach was, therefore, followed in the framework of this thesis to achieve a better coupling while avoiding a significant increase of the detector noise.

### 4.1 dc-SQUID with a Two-Turn Input Coil

Increasing the input coil inductance can be realized either by changing the geometry of the coil itself, or by implementing an intermediary flux transformer, with the benefit of easily adapting the inductance independently of the SQUID design. The latter, however, was accompanied with a reduction of the effective coupling constant  $k'_{\text{is}}$  in the work of [Bau22] regarding the maXs100 detector readout, which led to a lower flux-to-flux coupling despite the higher input inductance  $L'_i = 5.47$  nH. Only the white noise reduction of the SQUID, which was caused by the shielding effects of the added flux transformer, resulted in a calculated small overall improvement of  $\epsilon_p$  and  $\Delta E_{\text{FWHM}}$ .

In this work we designed a new detector SQUID with *window-type* Josephson junctions, which is based on the design developed in [Bau22]. A drawing of this new SQUID is shown in figure 4.1. Four large oval loops form the second order gradiometer described in subsection 3.1.1, where the lower niobium layer (Nb1) contains the SQUID loop as a parallel gradiometer. The feed line coming from the top connects

to the input coil, which consists of four loops in series. It was fabricated in a second niobium layer (Nb2) on top of the SQUID loop, only separated by an insulating  $\text{SiO}_2$  layer. This geometry allows to combine a small SQUID loop inductance  $L_s = L_l/4$  with a large input inductance  $L_i = 4L_l$ . Also in the Nb2 layer, the feedback coil is located below the input coil as another second order serial gradiometer with feed lines on the bottom left. This coil has a design inductance of  $L_f = 336 \text{ pH}$  with a line width of  $3 \mu\text{m}$ . Both coils exhibit the same geometry as the corresponding underlying washer strip to maximize the overlap and, therefore, the coupling. At the same time, the coupling between input and feedback coil is kept small to mitigate cross talk between the two coils. The feed lines of the SQUID loop at the bottom center lead to the junction area, which is shown in the close-up. The window-type junctions are realized by structuring window-shaped vias on top of the square junction area, which connects the junction with the niobium layer above. Both square-shaped Nb/Al-AlO<sub>x</sub>/Nb junctions are realized with the dimensions  $4.5 \mu\text{m} \times 4.5 \mu\text{m}$  and a targeted critical current of  $I_c = 6 \mu\text{A}$ , leading to a critical current density of  $j_c = 30 \text{ A/cm}^2$ . Two AuPd junction shunt resistors  $R_s$  are located on the left and right side of the junction area, respectively. Both are attached to a large heat sink made of two gold layers, a thick galvanized layer on top of a sputtered, thin one. These so-called *cooling fins* provide a better electron-phonon coupling thanks to their large volume, which reduces the electron temperature of the normal-conducting shunts and thus mitigates the corresponding thermal noise [Maz24]. A third AuPd resistor  $R_d$  with the same dimensions as the shunt resistors is placed above the junction area and connected in parallel with the washer loop. This so-called *washer shunt* provides damping properties to reduce quality factors of parasitic resonances, as will be discussed in section 4.2.

As opposed to the previous design, the input coil is realized with two turns instead of one. Neglecting the stripline inductance  $L_{\text{str}}$  (see section 5.3), the input inductance becomes approximately proportional to the number of turns squared  $n^2$  [Ket81, Jay81], giving an expected value of  $L_i^{\text{theo}} \approx 2^2 \cdot 1.64 \text{ nH} \approx 6.56 \text{ nH}$ . To implement the second turn a wider washer loop line width of  $w_s = 10 \mu\text{m}$  was needed. The line width of the input coil  $w_i$  with two turns remained at  $3 \mu\text{m}$ , which would not have been possible to fabricate on top of the previous washer width of  $w_s = 5 \mu\text{m}$  without sacrificing coupling strength. A general increase in  $w_i$ , however, entails the risk of capturing noise inducing flux vortices. These form if the B-field pointing perpendicular to the SQUID plane exceeds a critical field given by [Kui08]

$$B_{v,\text{crit}} = 1.65 \frac{\Phi_0}{w_s^2} . \quad (4.1)$$

At the earths surface, its magnetic field reaches a maximum amplitude of  $65 \mu\text{T}$ ,

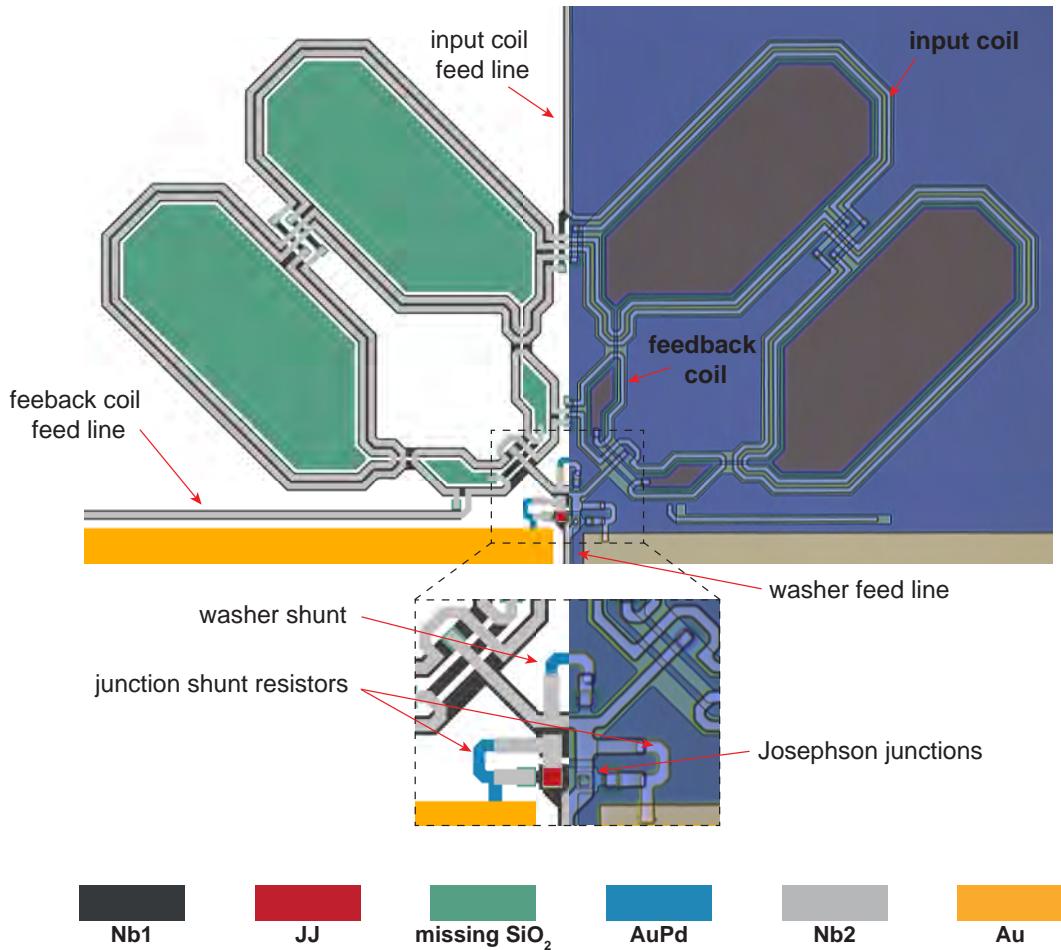


Figure 4.1: New dc-SQUID design with a two-turn input coil (top), including a zoom into the junction area (bottom). The schematic drawing is on the left, whereas a mirrored image with the use of an optical microscope is on the right. Four large oval-shaped loops follow the microstrip transmission line structure given by the washer SQUID loop in the lower niobium layer (Nb1) and the input coil in the upper niobium layer (Nb2). Fabricated in the same manner is a feedback coil between the junction area and the input coil. Both coils are free of sputter-deposited SiO<sub>2</sub>, visualized through the green areas that indicate missing SiO<sub>2</sub>.

which would give a threshold width of  $w_s = 7.2 \mu\text{m}$ . The dilution refrigerators used to cool down both the SQUIDs and the detectors typically provide magnetic shielding, such that we consider a width of  $w_s = 10 \mu\text{m}$  to have a still negligible impact on the flux noise attributed to flux vortices. As the input inductance was the only parameter necessary to adjust, we attempted to keep the other design parameters unaltered. The widened washer width would, therefore, need to be compensated with a larger washer hole circumference in order to maintain the same SQUID loop inductance.

The necessary increase was estimated by modeling the rather complicated oval washer loop geometry as a ring-shaped structure whose inductance can be calculated with the relation  $L = \mu_0 R \left( \ln \left( \frac{8R}{a} \right) - 2 \right)$ , where  $R$  denotes the loop radius and  $a$  the radius of the wire [Den16]. This resulted in an estimated increase of 15 %, necessary to compensate for the widened washer loop. For the sake of safety, the circumference was increased only by 10 % to prevent possible hysteretic behavior, that can occur if the SQUID loop inductance  $L_s$  and, therefore, the screening parameter  $\beta_L$  grows too large. In addition to these geometric adjustments, an extensive (sputter) deposition of the  $\text{SiO}_2$  layer in the area of the washer loops was also omitted, allowing the interior to remain free from the insulating layer. This prevents potential flux noise induced by unavoidable magnetic impurities as well as due to dangling bonds in the  $\text{SiO}_2$ . The same method was applied to the smaller loop areas formed by the feedback coil. The absence of insulation within the loops is visualized by the green-colored areas in figure 4.1.

The design values for  $R_s$  and  $L_s$  are chosen such that the extrinsic energy sensitivity given in equation 3.19 is minimized, which requires the maximization of the flux-to-flux coupling ( $L_i = L_p + L_{\text{par}}$ ) and the minimization of the intrinsic white noise of the SQUID. In [Bau22] this numerical calculation was done with the constraints  $\beta_C \leq 0.7$  and  $\beta_L \leq 1$  to avoid hysteretic behavior, which led to the optimal parameters  $\beta_C = 0.7$  and  $\beta_L = 0.86$ . Consequently, additional noise through voltage jumps caused by hysteretic IVCs as well as Nyquist noise from higher harmonics of the Josephson frequencies [Cla96] have been neglected for this minimization, which was not the case for the derivation of equation 2.40. The intrinsic white noise of the SQUID used for the numerical calculation is, therefore, given by the adjusted expression [Knu88]

$$S_{\Phi_s} = 2k_B T \frac{L_s^2}{R_s} \left[ (1 - k_{is}^2 s_{\text{in}})^2 + \frac{\sqrt{2}(1 + \beta_L)^2}{\beta_L^2} \right] . \quad (4.2)$$

To achieve the targeted critical current of  $I_c = 6 \mu\text{A}$  with the given junction dimension of  $4.5 \mu\text{m} \times 4.5 \mu\text{m}$ , we aim for a critical current density of  $j_c = 29.63 \text{ A/cm}^2$ . The current density can be used to calculate the junction capacitance  $C$  by using the empirical relation  $1/C' = p_1 + p_2 \log_{10} j_c$ , where  $p_1$  and  $p_2$  are constant fit parameters [Mae95]. Here, the intrinsic capacitance  $C'$  excludes any parasitic capacitances arising from the window-type fabrication technique. For simplicity reasons, we assume  $C \approx C'$  and thus obtain  $C = 0.95 \text{ pF}$ . The optimal Stewart McCumber and screening parameter then provide the values  $R_s = 6.3 \Omega$  and  $L_s = 147 \text{ pH}$ , respectively. The designed shunt resistor in both the previous and the new design was rounded to  $6 \Omega$ , which results with Ohm's circuit law in a normal resistance of  $R_n = 3 \Omega$  for the whole SQUID. This consequently corresponds to a slightly lower damping parameter of  $\beta_C = 0.62$ . The coupling constant was set to an upper limit of  $k_{is} = 0.75$ , which

is typically the highest achievable value for the SQUIDs produced in this working group. Under the assumption of  $k_{\text{is}}$  being maximal and  $L_i = L_p + L_{\text{par}} = 7.15 \text{ nH}$  for the maXs100 detector read-out, the theoretically obtainable flux-to-flux coupling regarding a single pickup coil is  $\Delta\Phi_s/\Delta\Phi_p = 5.38\%$ . As expected, this value succeeds the value obtained for the previous design with a single-turn input coil, which yielded  $\Delta\Phi_s/\Delta\Phi_p = 3.9\%$  [Bau22]. Lastly, for the extrinsic energy sensitivity we would obtain with equation 4.2  $\epsilon_p = 0.53 \text{ h}$ , assuming a detector operation temperature of  $T = 20 \text{ mK}$ .

## 4.2 Damping Methods

As discussed in subsection 2.2.5, several SQUID parameters can be optimized to mitigate the influence of various resonances in the circuit. However, we would like to choose these parameters accordingly to the minimization of the extrinsic energy sensitivity as discussed in section 4.1. This choice imposes substantial limitations on the extent to which modifications can be implemented. For instance, increasing the length of the input coil  $l_i$  would move the corresponding microstrip resonance given by equation 2.45 away from the operation frequency, while on the other hand leading to a larger input inductance, thus impeding the maximization of the flux-to-flux coupling. Furthermore, even resonances far away from the operation frequency can prove problematic as thermally activated transitions between different states increase the noise level [Sep87]. This motivates a more practical approach to suppress  $LC$  resonances, which can be realized through damping with attenuators, such as the damping resistor  $R_d$  shown in the top of the junction area in figure 4.1. These are typically connected in parallel to the resonant circuit, as this reduces the quality factor  $Q$  of the corresponding  $LC$  resonance given an appropriate dimensioning of the resistor. Consequently, the  $L_sC_p$  and  $L_sC$  resonances, that arise due to the parasitic capacitance  $C_p$  and the junction capacitance  $C$ , can be damped by this *washer shunt*  $R_d$  [Ono97, Ryh92]. Although the crossing of the IVCs caused by the  $L_sC$  resonance could not be fully eliminated in previous works of this group, they showed a significant smoothing of the curves as the accompanying step structures were reduced [Bau18]. The current noise introduced through this resistor, on the other hand, deteriorates the energy sensitivity and thus limits the damping benefit. However, this effect is minimal for the condition  $R_d \approx R_s$  with  $\beta_L = 1$  [Enp86, Ryh92], which is why we choose  $R_d = 6 \Omega$ . The input circuit is shunted with an  $R_xC_x$  attenuator to damp the  $L_iC_p$  resonance, where the added capacitance  $C_x$  blocks low frequency current noise [Sep87]. Both damping techniques to suppress  $C_p$ -related resonances were proven to be effective for various SQUID designs [Knu87, Enp86, Can91, Bau18]. As for the  $\lambda/2$  resonances, both the  $R_d$  and the  $R_xC_x$  attenuator

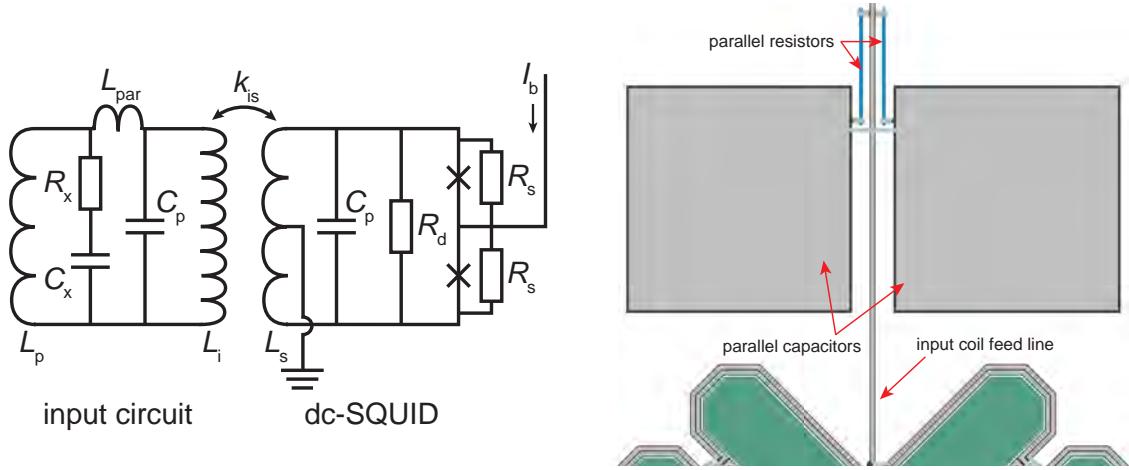


Figure 4.2: Left: Schematic circuit of a coupled dc-SQUID with damping components  $R_d$  and  $R_xC_x$ . The input coil forms a flux transformer with a pickup coil of inductance  $L_p$ . The parasitic capacitance between the input circuit and the SQUID loop is connected in parallel to both  $L_i$  and  $L_s$ . Adapted from [Bau22]. Right: Schematic drawing of the  $R_xC_x$  attenuator located at the input coil of the dc-SQUID. Both devices,  $R_x$  and  $C_x$  are each realized as two parallel components.

provide good damping as well [Can91], since they terminate the microstrip lines and thus avoid impedance mismatches given a suitable dimensioning of the resistors (see section 5.3). A schematic of the resulting circuit diagram of the coupled dc-SQUID with all damping components is depicted in figure 4.2 (left). Shown on the right is the design of the  $R_xC_x$  shunt, which has been adapted from [Bau22]. The capacitance  $C_x = 10 \text{ pF}$  is divided into two parallelly connected square-shaped planar capacitors with capacitance  $C_x/2$ , each. The top and bottom plate of each capacitor are fabricated in the Nb2 and Nb1 layer, respectively. The shunt  $R_x$  is split as well into two parallel AuPd resistors with a resistance of  $2R_x$  each. Optimal values for these components will be discussed in chapter 5.

#### 4.2.1 Lossy Input Coil

Although the above-mentioned damping methods should provide significant improvement regarding the resonance behavior, further suppression of additional resonances can be, depending on the SQUID design, favourable or even necessary. For this reason, we apply two additional damping techniques, which require resistive gold layers electrically or inductively coupled to the circuit. The former will be covered in the following, whereas the latter is explained in subsection 4.2.2.

Several approaches to reduce  $Q$  values of  $\lambda/2$  resonances associated with microstrip

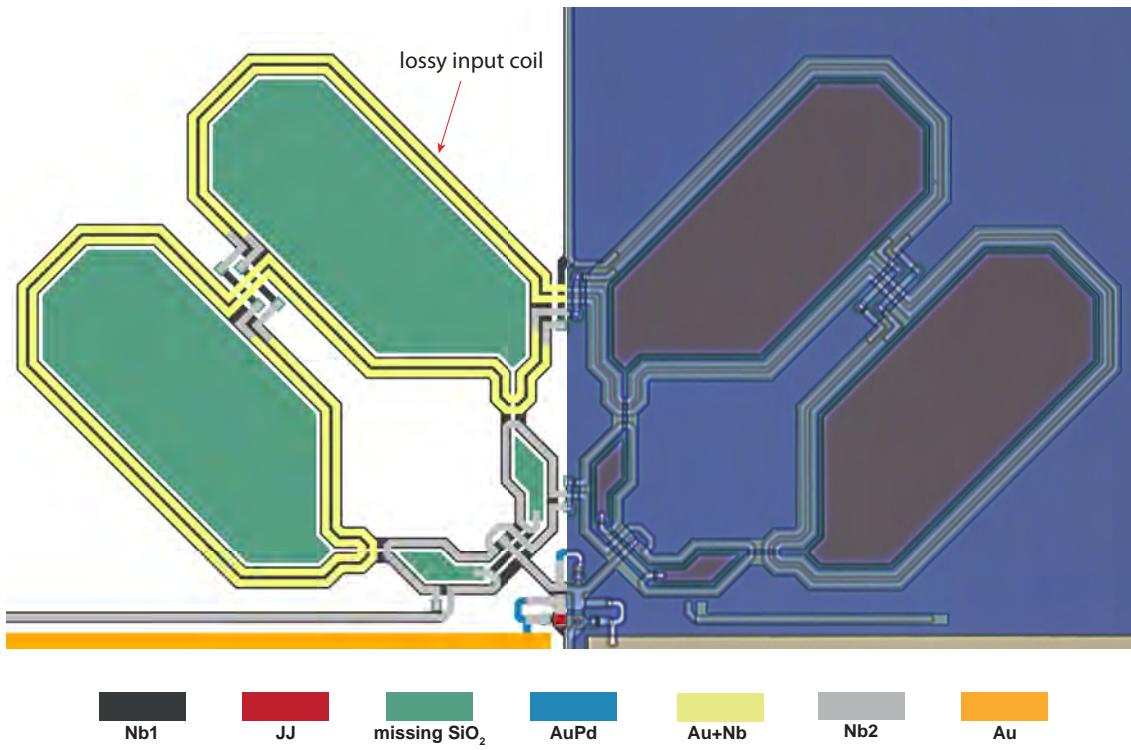


Figure 4.3: dc-SQUID design with a two-turn input coil realized as a lossy microstrip line. The schematic drawing is on the left, whereas a mirrored image with the use of an optical microscope is on the right. A gold layer is structured between the insulating  $\text{SiO}_2$  and an upper niobium layer and has been fabricated in a single step as a bilayer. To ensure the superconductivity of vias, they are contacted in a separate niobium layer (Nb2).

lines have been investigated in [Boy22]. The experimental setup consists of two parallel meanders with each having a length of 6.5 mm and an inductance of 2 nH. They are fabricated on top of each other with insulating  $\text{SiO}_2$  between them and in a direct-coupled MMC setup, where as opposed to the flux transformer setup in section 3.3, the SQUID is directly coupled to the sensor. The upper meander thereby represents the SQUID loop, whereas the lower carries a magnetizing current of approximately 100 mA. This setup produced high  $Q$  resonances at integer and half-integer wavelengths. Whereas placing an individual resistor in parallel to one meander only damped the half-integer modes, a more distributed damping scheme in the form of an insulated gold layer between both meanders provided strong damping of all microstrip resonances while maintaining a low detector noise level. The  $Q$  values further reduced by structuring the Au layer with the same geometry as the microstrip lines and electrically connecting it to one of the meanders, thereby preventing large, normal-conducting loops that would induce additional Johnson noise.

The attenuation  $\alpha$  of the microstrip line was measured to further quantify the damping. Generally,  $\alpha$  represents the real part of the complex wave propagation constant  $\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$ , where  $\beta$  is the imaginary part and  $j$  the complex number. The parameters  $R$ ,  $L$ ,  $G$  and  $C$  represent the series resistance, the inductance, the conductance and the capacitance of the transmission line, respectively [Poz11]. The attenuation constant  $\alpha$  of this lossy microstrip line increases with frequency and thickness of the Au layer, while the noise stayed insensitive upon thickness changes.

Within the scope of this thesis, such damping techniques were implemented and tested on the dc-SQUID described in section 4.1. For this, we sputtered a gold layer between  $\text{SiO}_2$  and Nb2 with the same geometry as the input coil running above the SQUID loop. This gold layer was omitted around the vias between the washer loops to avoid additional normal resistances. Gold and niobium were structured together as a bilayer in the same microfabrication step by sputtering the upper niobium layer directly after the gold. A schematic of our front-end design with this lossy input coil, including an image obtained with an optical microscope, is shown in figure 4.3. Our front-end SQUIDs with this gold layer beneath the input coil will be referred to as 'lossy' for the upcoming discussions.

#### 4.2.2 Inductive Damping

Together with the attempt of suppressing resonances through a lossy microstrip line, we introduce a second damping technique denoted as inductive damping. This method is based on the principle of magnetic damping, where a change in magnetic flux creates eddy currents in a nearby conductor, which following Lenz's law induces a flux trying to compensate the initial one. The flux change is, therefore, damped by effectively transferring part of the magnetic energy to the kinetic energy of the induced current, which in the case of a normal conductor emits heat. This phenomenon also causes the reduction of the geometric inductance of the SQUID loop due to the shielding effect of the flux transformer (see subsection 3.3.1). A strong indicator that this mechanism can be applied to SQUIDs shows the first mentioned experiment in [Boy22], where a square gold layer representing the MMC sensor was placed at a height of 300 nm above an isolated meander. This lead to a significant reduction of the high  $Q$  values of the meander modes. The concept is now applied to the feed lines on our SQUID chip, where large gold pads have been placed across the feedlines of both, SQUID washer and feedback coil, as shown in figure 4.4. The sharp voltage spikes associated with high  $Q$  resonances, that can be present within those lines would, therefore, be damped by partly converting their energy into heat in the normal conducting gold layer. The gold pads are fabricated in the same layers as the heat sinks for the shunt resistors and are consequently sputtered first before being

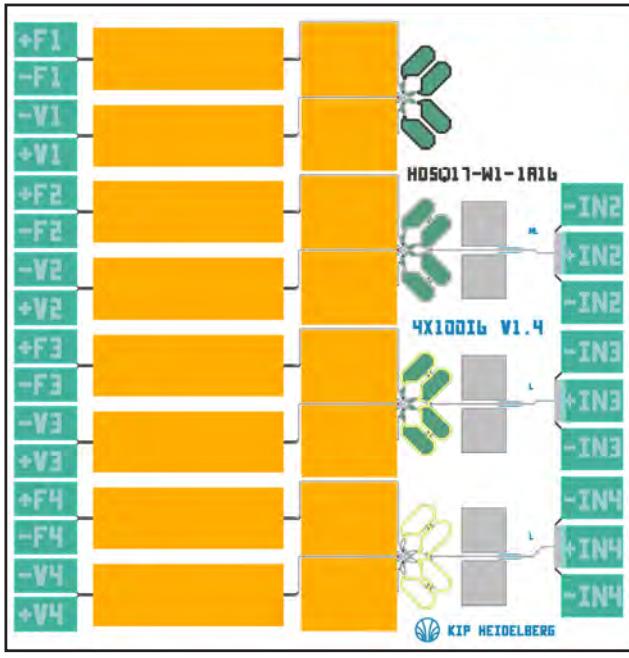


Figure 4.4: Inductive damping scheme on the front-end SQUID chip of the type 4x100i6 v1.4. All feedback coil and SQUID loop feed lines are covered with large, insulated gold layers. Each SQUID channel is realized with a different SQUID variant. Channel 1: front-end without an input coil. Channel 2: Non-lossy SQUID design presented in section 4.1. Channel 3: Lossy SQUID design from section 4.2.1. Channel 4: Lossy design with  $\text{SiO}_2$  inside the washer loop.

electroplated. The latter step significantly increases their volume, which allows for larger and more effective eddy currents. Additionally, the generated heat is expected to better dissipate into the chip substrate as the electron-phonon coupling increases with volume [Wel94]. All feed lines are typically realized as microstrip lines, but underneath the Au plane they are routed in a coplanar structure to prevent the need of another insulation layer after Nb<sub>2</sub>. Figure 4.4 shows the chip design consisting of four distinct front-end variants, each provided with the inductive damping scheme. The first channel at the top is realized without an input coil in order to better allocate possible resonance structures visible in the SQUID's IVCs. The design introduced in section 4.1, also referred to as 'non-lossy', is represented in channel 2, followed by the lossy variant in channel 3. The last channel contains a lossy front-end as well, however, the washer loop interiors were not kept  $\text{SiO}_2$ -free. This allows to investigate the influence of possible magnetic moments in the insulation material and to assess whether it can be regarded as negligible or not. The same chip design has been produced with and without gold pads on the feed lines, resulting in 8 different front-end SQUIDs that were developed and tested within the scope of this thesis.



## 5. Experimental Results

This chapter provides an overview over the general performance of the dc-SQUIDs developed within the framework of this thesis. Particularly, the resonance behavior and the noise spectra were investigated. We begin with a summary of characteristic parameters obtained by our measurements and compare them with the target values. The noise measurement has been conducted with the two-stage setup at  $T = 10\text{ mK}$  in the cryostat, whereas the single-stage measurements to obtain characteristic front-end properties have been done both in the helium vessel and the cryostat.

### 5.1 Characteristic dc-SQUID Parameters

Despite all distinctions between the front-end variants mentioned in chapter 4, most characteristic parameters such as the SQUID loop inductance  $L_s$  or shunt resistance  $R_s$  are unaffected by these variations. We, therefore, consider the following measurements to be representative for all variants and assume possible variations to be caused from fabrication-related variations across the wafer.

Figure 5.1 shows the current-voltage as well as the voltage-flux characteristics of one of the measured SQUIDs. This SQUID (chip no. HDSQ17-W1-3C16) is of the type 'non-lossy' with inductive damping and has been measured at a cryostat temperature of  $T = 10\text{ mK}$ . First, we measured the slope of the  $IV$  curves across the ohmic regime, which allows to determine the normal resistance  $R_n$  of the SQUID. In total 10 SQUIDs from the wafer HDSQ17-W1 covering all variants we measured. The median normal resistance taken from these SQUIDs yields the value  $R_n = 3.0\Omega$ , with the largest deviation being 13.3 %. This corresponds to a shunt resistance of  $R_s = 2R_n = 6.0\Omega$ , which exactly corresponds to the targeted value.

The maximum voltage swing as seen in figure 5.1 (right) is obtained by applying a bias current  $I_b^{\max}$  supplied by the SQUID electronics, while driving a ramp signal through one of the coupled coils to provide a varying external flux. This yielded median values of  $\Delta V_{\max} = 29.95\mu\text{V}$  and  $I_b^{\max} = 11.62\mu\text{A}$ . For  $T = 0\text{ K}$ ,  $I_b^{\max}$  should correspond to twice the critical current  $I_c$  (compare figure 2.9). Due to the finite temperature we obtain noticeable thermal smoothening visible in the IVC of figure 5.1 (left), such that the critical current is better approximated by [Dru96b]

$$I_c \approx \frac{I_b^{\max}}{2} + \frac{k_B T}{\Phi_0} \left( 1 + \sqrt{1 + \frac{I_b^{\max} \Phi_0}{k_B T}} \right) . \quad (5.1)$$

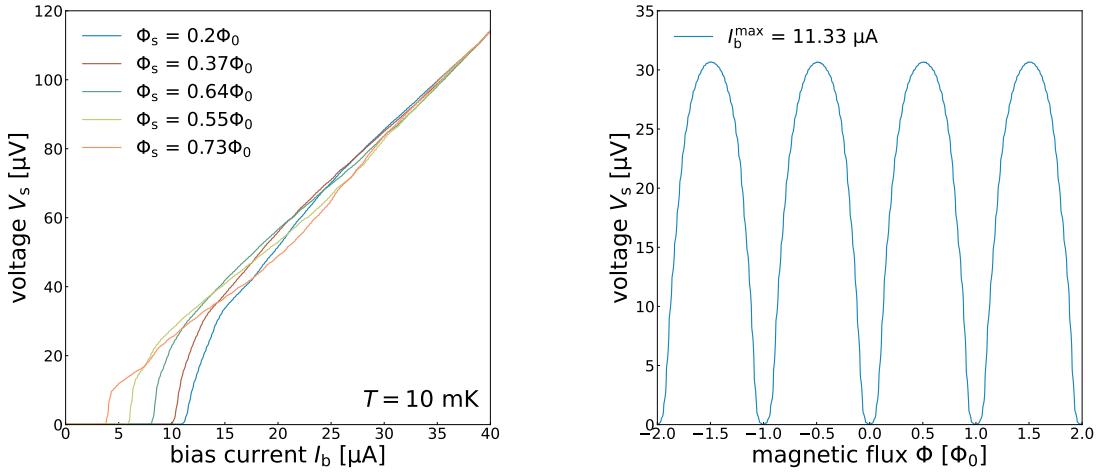


Figure 5.1: Current-voltage (left) and voltage-flux characteristics (right) of a non-lossy SQUID with inductive damping (chip no. HDSQ17-W1-3C16). The extremal IV curve with the lower critical current  $I_{c,1}$  does not correspond exactly to  $\Phi = (n + \Phi_0/2)$ , which is a consequence of an asymmetric current injection. The  $V\Phi$  curve was measured at  $I_b^{\max} = 11.33 \mu\text{A}$ , resulting in a maximal voltage swing  $\Delta V_{\max} = 30.7 \mu\text{V}$ . Note that both plots have different scaling on the  $V$ -axis.

We, therefore, calculate the median critical current to  $I_c = 5.84 \mu\text{A}$ . With a junction size of  $4.5 \mu\text{m} \times 4.5 \mu\text{m}$ , this corresponds to a critical current density of  $j_c = 28.8 \text{ A/cm}^2$ , such that both values only deviate 2.7 % from the design values  $I_c = 6 \mu\text{A}$  and  $j_c = 30 \text{ A/cm}^2$ . The junction capacitance thus becomes slightly smaller than the design value according to the empirical relation introduced in section 4.1. Using the empirical dependence  $1/C' = 0.132 - 0.053 \log_{10} j_c$  for the intrinsic junction capacitance  $C'$  obtained in [Bau22] and by again assuming  $C' = C$ , we get the capacitance  $C = 0.948 \text{ pF}$ , which is in excellent agreement with the target value of  $0.95 \text{ pF}$ . The McCumber parameter can now be calculated to  $\beta_C = 0.61$ , also fitting well with the design value of 0.62.

As for the transfer coefficients, different values for the positive and negative slopes of the  $I\Phi$  curve are expected due to the asymmetric current injection, which is realized by differing SQUID loop inductances between each arm of the loop, ultimately leading to a unilaterally larger transfer coefficient  $I_\Phi$  [Fer15]. The measured values are  $I_{\Phi,+} = 10.2 \mu\text{A}/\Phi_0$  and  $I_{\Phi,-} = 23.1 \mu\text{A}/\Phi_0$  for the positive and negative slope, respectively. The voltage transfer coefficient is approximately symmetric (compare figure 5.1) and yields  $V_{\Phi,+} = 94.0 \mu\text{V}/\Phi_0$  and  $V_{\Phi,-} = 93.0 \mu\text{V}/\Phi_0$ . The variance of these transfer parameters is rather large, with deviations from the median of up to

39 %. These results are nevertheless comparable to previous SQUIDs produced in this working group [Ric17].

**Screening parameter  $\beta_L$**  The screening parameter is typically derived by using the  $\beta_L$ -dependent normalized current swing  $\Delta I_{\max}/2I_c$  at  $V = 0$ , which has been numerically simulated in [Tes77]. We estimated the current swing by extrapolating both extremal IV curves for the theoretical case of  $T = 0$ , thus neglecting the thermal rounding and obtaining a minimal and maximal critical current  $I_{c,1}$  and  $I_{c,2}$ . The median of the maximal current swing  $\Delta I_{\max}$  yields  $7.18 \mu\text{A}$ , ranging from  $6.09 \mu\text{A}$  to  $7.73 \mu\text{A}$  for the lowest and highest measured value. The value for  $\Delta I_{\max}$ , which approximately corresponds to the current swing of the voltage-biased SQUID in a two-stage configuration, indicates that the SQUID is well adapted to the arrays produced in this working group. These arrays have a typical reciprocal mutual inductance of  $1/M_{ix} = 11.7 \mu\text{A}/\Phi_0$ , such that the flux  $\Delta\Phi_x$  coupled to the array is  $\Delta\Phi_x = M_{ix}\Delta I_{\max} = 0.61 \Phi_0$ , which is close to the optimal value of  $\Phi_0/2$  to achieve a flux gain of  $G_\Phi \approx 3$  (see section 3.2.2). The current swing of each SQUID allows us to calculate the median screening parameter to  $\beta_L = 0.65$ , which is rather low compared to the design value of  $\beta_L = 0.86$ , indicating an overestimation of the SQUID loop inductance  $L_s$ . It is noteworthy, however, that the variations of both the measured critical currents and the current swings were rather large with up to 11 % and 15 %, respectively. This of course provides an uncertainty for  $\beta_L$  that needs to be taken into account. Such variances have also been observed in [Bau22], where it was shown that SQUIDs with *cross-type* junctions exhibit a smaller variance across the wafer, making them more reliable than window-type junctions regarding  $I_c$ .

According to the obtained  $\beta_L$  and  $\beta_C$ , hysteretic behavior should not occur and thus be absent in the IVCs. The intersections and current steps seen in figure 5.1, therefore, likely stem from resonances, which will be discussed in section 5.3.

**SQUID loop inductance  $L_s$**  With the above-determined  $I_c$  and  $\beta_L$  values we obtain a median SQUID loop inductance of  $L_s = 108 \text{ pH}$ . This represents a significant deviation of 27 % from the target value of  $L_s = 147 \text{ pH}$  discussed in section 4.1, which can be partly explained by the conservative estimation for the geometric adjustment of the washer loop size, resulting in a smaller inductance. On the wafer HDSQ17-W1 we included for comparison the original design [Bau22] with a single turn input coil, where we measured two channels of the chip HDSQ17-W1-3A09. These provided the respective values  $L_s = 128 \text{ pH}$  ( $\beta_L = 0.80$ ) and  $L_s = 138 \text{ pH}$  ( $\beta_L = 0.87$ ), which is in accordance with the value of  $135 \text{ pH}$  that was measured in [Bau22]. This confirms that the washer loops were designed too small, resulting in a smaller SQUID loop inductance than the targeted  $147 \text{ pH}$ . The measured inductance  $L_s$

can, however, also deviate from the real value, which could be explained by the asymmetric  $I\Phi$  characteristic as this aspect has not been included in the numerical simulation for  $\Delta I(\beta_L)$  in [Tes77]. Additionally, the uncertainty of  $L_s$  is dominated by the measurement uncertainty regarding the maximal current swing, leading to large deviations of up to 23 % from the median value of 108 pH. The measured value of 108 pH, however, fits well with the simulated value of 106 pH, which was simulated with InductEx after the SQUIDs were already produced. The reduction of  $L_s$  reduces the coupling  $M_{is}$  between input coil and SQUID loop, which in turn reduces the flux-to-flux coupling  $\Delta\Phi_s/\Delta\Phi$  and thus the extrinsic energy sensitivity  $\epsilon_p$  given by equations 3.17 and 3.19, respectively.

**Mutual inductances** The mutual inductances of the input and feedback coil were obtained by sending a current ramp to the corresponding coils. The input coil was only connected to the SQUID electronics during the measurements in the helium vessel at 4.2 K, where  $M_{is}$  was measured from 22 front-end channels from 9 chips, covering all non-lossy and lossy variants. The measured median values are  $M_{is} = 611 \text{ pF}$  and  $M_{fs} = 51 \text{ pF}$ , where the latter coincides well with the measured value of 50 pF from [Bau22], as it is expected because the geometry of the feedback coil was not changed. The input coil current sensitivity is almost twice as large as the value determined with the previous design with a single-turn input coil, which yielded  $M_{is} = 328 \text{ pF}$  [Bau22]. This result is expected from the general linear dependence of the mutual inductance on the number of turns  $n$  of the input coil and the SQUID loop inductance, i.e.  $M_{is} \approx nL_s$  [Ket81].

The fact that  $M_{is}(n=2)/M_{is}(n=1) < 2$  can be explained by two reasons. On the one hand, the inductance  $L_s$  has been reduced by approximately 19 % as compared to the previous design, which according to equation 3.1 leads to a decreased current sensitivity  $M_{is}$ . On the other hand, the geometric two-turn input coil structure might deteriorate the coupling constant  $k_{is}$ , as compared to a single turn design.

Taking into account the input inductance calculated in section 5.2 and the other parameters determined above, we can calculate the coupling constant via equation 3.1 to  $k_{is} = 0.73$ , which is close to the target value 0.75 that has been set as a realistic upper limit for the minimization of the extrinsic energy sensitivity  $\epsilon_p$ . This suggests that the reduction of the input coil current sensitivity  $M_{is}$  is mostly due to the smaller SQUID loop inductance  $L_s$ .

## 5.2 Input Coil Inductance

To determine whether the new SQUID design with increased input inductance exhibits better coupling to the maXs100 detector, it is essential to measure the input

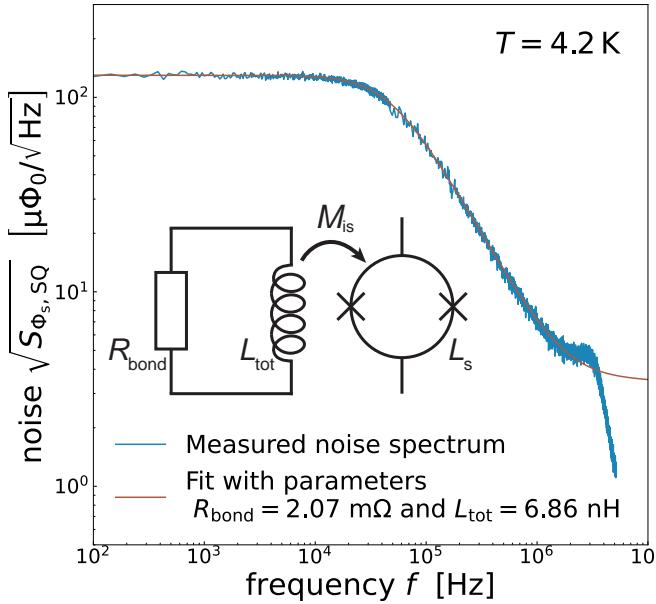


Figure 5.2: Noise spectrum of the input coil inductance  $L_i$  measurement. The input coil is resistively shorted through aluminum bond wires with resistance  $R_{\text{bond}}$ , which results in thermal noise that couples into the SQUID. The added noise has a cut-off frequency  $f_c = R_{\text{bond}}/(2\pi L_{\text{tot}})$  that allows to extract the added white noise component by a numerical fit. The input inductance can be derived by subtracting the parasitic inductances from  $L_{\text{tot}}$ , which is a fit parameter alongside  $R_{\text{bond}}$ .

coil inductance  $L_i$ . For this measurement, the input coil is electrically shorted via its bond pads with aluminum bond wires while the front-end is in a two-stage setup at 4.2 K. In this special case, the detector SQUID was part of an integrated chip that contains both the new non-lossy front-end design as well as an array for the second stage, which has been developed and tested in [Kra23]. These integrated chips were produced on the wafer HDSQ16-W1 and were designed with the same material thicknesses, such that we expect the measurement to be representative for the non-lossy SQUIDs from HDSQ17-W1. The aluminum bond wires are normal conducting at the temperature of 4.2 K and thus provide a resistance  $R_{\text{bond}}$ . The closed loop couples frequency-independent thermal noise from the resistive wires into the SQUID loop via the mutual inductance  $M_{\text{is}}$ , which adds to the apparent noise of the first stage SQUID. The resistance  $R_{\text{bond}}$  forms an  $RL$  lowpass filter together with the total inductance  $L_{\text{tot}}$  consisting of the input coil, its feed line and the wire inductance. The attributed cut-off frequency damps higher noise frequencies up to the point where this noise contribution becomes negligible, as can be seen in the measured noise spectrum of figure 5.2. The second cut-off at a frequency around 5 MHz represents the lowpass characteristic of the SQUID electronics, as it provides a limited bandwidth of up to 6 MHz. The total apparent noise of the Front-End SQUID is then given by

$$S_{\Phi_s,\text{SQ}} = M_{\text{is}}^2 \frac{4k_B T}{R_{\text{bond}}} \left[ \frac{1}{1 + (\frac{2\pi f L_{\text{tot}}}{R_{\text{bond}}})^2} \right] + S_{\Phi_s,\text{w}} , \quad (5.2)$$

where the first term describes the additional current noise of the shorted circuit,

which is schematically shown in the inset of figure 5.2. The second term represents the apparent white noise of the SQUID. This expression is numerically fitted to the measured data to obtain both the resistance of the bonding wires and the total inductance  $L_{\text{tot}}$ . Two measurements from separate chips (2C14 and 2C23) were carried out, where the one from 2C14 is depicted in figure 5.2. The fits result in  $R_{\text{bond}}^{2\text{C}14} = 2.07 \text{ m}\Omega$  and  $L_{\text{tot}}^{2\text{C}14} = 6.86 \text{ nH}$  for the chip 2C14, whereas  $R_{\text{bond}}^{2\text{C}23} = 1.86 \text{ m}\Omega$  and  $L_{\text{tot}}^{2\text{C}23} = 6.74 \text{ nH}$  for the chip 2C23. For the aluminum wires it has been shown in [Hen17] that they, depending on their arrangement, exhibit an inductance of  $L_{\text{bond}} \approx 0.14 \text{ nH/m}\Omega$ , which consequently results in  $L_{\text{bond}}^{2\text{C}14} = 0.29 \text{ nH}$  and  $L_{\text{bond}}^{2\text{C}23} = 0.26 \text{ nH}$ . The parasitic inductance from the input coil feed lines can be estimated with the microstrip inductance per unit length [Cha79]

$$L_{\text{str}} = \mu_0 \frac{h + 2\lambda}{w_i + 2h + 4\lambda} , \quad (5.3)$$

with  $\lambda$  being the London penetration depth of the superconductor,  $w_i$  the width of the upper line and  $h$  the thickness of the insulating layer. The length of the feed lines is approximately 724  $\mu\text{m}$ , the width 3  $\mu\text{m}$  and  $\text{SiO}_2$  has been fabricated as a 375 nm thick layer. We also assume  $\lambda = 90 \text{ nm}$ , which is a typical value for the fabricated niobium in this working group. This yields a small contribution of 0.12 nH for the feed line inductance. Thus, we finally obtain the input inductances  $L_i^{2\text{C}14} = 6.45 \text{ nH}$  and  $L_i^{2\text{C}23} = 6.36 \text{ nH}$ . These are only slightly smaller than the originally targeted value of  $L_i^{\text{theo}} \approx n^2 1.64 \text{ nH} \approx 6.56 \text{ nH}$  mentioned in section 4.1. However, two effects overestimate the expected value. On the one hand, the measured inductance of the previous design only yielded 1.27 nH. On the other hand, the approximation does not take the additional microstrip inductance  $L_{\text{str}}$  into account (see section 3.1), which grows only linearly with  $n$  rather than  $n^2$  [Ket91]. This effect would reduce the increase of  $L_i$  upon adding a second turn. In contrast to that, a third effect leads to an underestimation of the target value. The enlargement of the washer loop has a strong influence on  $L_i$  due to the series connection of the input coil, where the resulting inductance increase per washer loop is multiplied by 4. Taking all effects into account, the enlargement of the washer loops should lead to additional 0.48 nH per turn, which after multiplying by  $n^2$  results in the average input coil inductance  $L_i = 6.40 \text{ nH}$  obtained from the two measurements. With InductEx we obtained a simulated value of  $L_i = 5.6 \text{ nH}$ , which deviates 13 % and 12 % from  $L_i^{2\text{C}14}$  and  $L_i^{2\text{C}23}$ , respectively. These moderate deviations are in accordance with previous inductance simulations in several works of this working group [Fer15, Bau22]. For the following discussions, we will use the average  $L_i = 6.40 \text{ nH}$  of the two measurements.

An overview of the most relevant measured parameters compared to their respective design values is given by table 5.1, showing an overall good agreement.

Parameter	$R_s$ [ $\Omega$ ]	$I_c$ [ $\mu\text{A}$ ]	$M_{is}$ [pH]	$M_{fs}$ [pH]	$L_s$ [pH]	$L_i$ [nH]	$\beta_L$	$\beta_C$	$k_{is}$
Measured	6	5.84	611	51	108	6.40	0.65	0.61	0.73
Design	6	6	-	50	147	6.56	0.86	0.62	0.75

Table 5.1: Summary of measured characteristic parameters of the new dc-SQUID design with a two-turn input coil, which are compared with the corresponding target values.

### 5.3 Resonance Behavior

In this section we will investigate how various  $LC$  resonances might affect the IVCs of the SQUIDs and whether the applied damping techniques proved to be effective or not. Figure 5.3 showcases the IVCs of 8 distinct SQUIDs from the 10 chosen for the characterization. The order from top to bottom corresponds to the order shown in the inductively damped chip with various SQUID types in figure 4.4. These damped SQUIDs are on the right side (plots e) through h), while the corresponding SQUID types without inductive damping are on the left side (plots a) through d) ) in the same order. There are a few differences between certain SQUID variants, however, they collectively share a prominent feature at  $V_s \approx 40 \mu\text{V}$ , especially for the non-lossy SQUIDs. This artifact shows a pronounced current step with corresponding intersections between the IV curves. The fact that it is also present in the SQUIDs without input coil (d) and h)) rules out any input coil related resonances to be the cause. Regarding the feedback coil, due to the small serially connected inductance its resonance  $f_{L_f C_p} = 77 \text{ GHz}$  lies far above the operation frequency  $f_{op} \approx 5.1 \text{ GHz}$ , as we will see later on. Consequently, we attribute the pronounced current step at approximately  $40 \mu\text{V}$  to the fundamental SQUID resonance  $f_{L_s C}$ , which by using equation 2.44 is calculated to  $f_{L_s C} = 22.2 \text{ GHz} = 46 \mu\text{V}/\Phi_0$ , using the measured values of  $L_s = 108 \text{ pH}$  and  $C = 0.95 \text{ pF}$ . This value does not exactly correspond to the observed  $40 \mu\text{V}$ , which would indicate an underestimation of either the SQUID loop inductance  $L_s$ , junction capacitance  $C$  or both. As discussed in section 5.1, the parasitic capacitance arising from the insulation window structure of the junction is typically neglected, i.e.  $C = C'$ . The real junction capacitance should, therefore, be slightly larger, leading to a decreased resonance frequency  $f_{L_s C}$ . Furthermore, the measurement uncertainty of  $L_s$  can also be quite large, as discussed in section 5.1.

The comparison of each side of figure 5.3 demonstrates a significant smoothening of all IVCs of the inductively damped SQUIDs, as compared to their non-damped counterparts. We can, therefore, conclude that damping with insulated gold pads proved to effectively suppress all visible resonances to some extent, regardless of the SQUID design. Consequently, one would expect the resulting intrinsic flux noise of

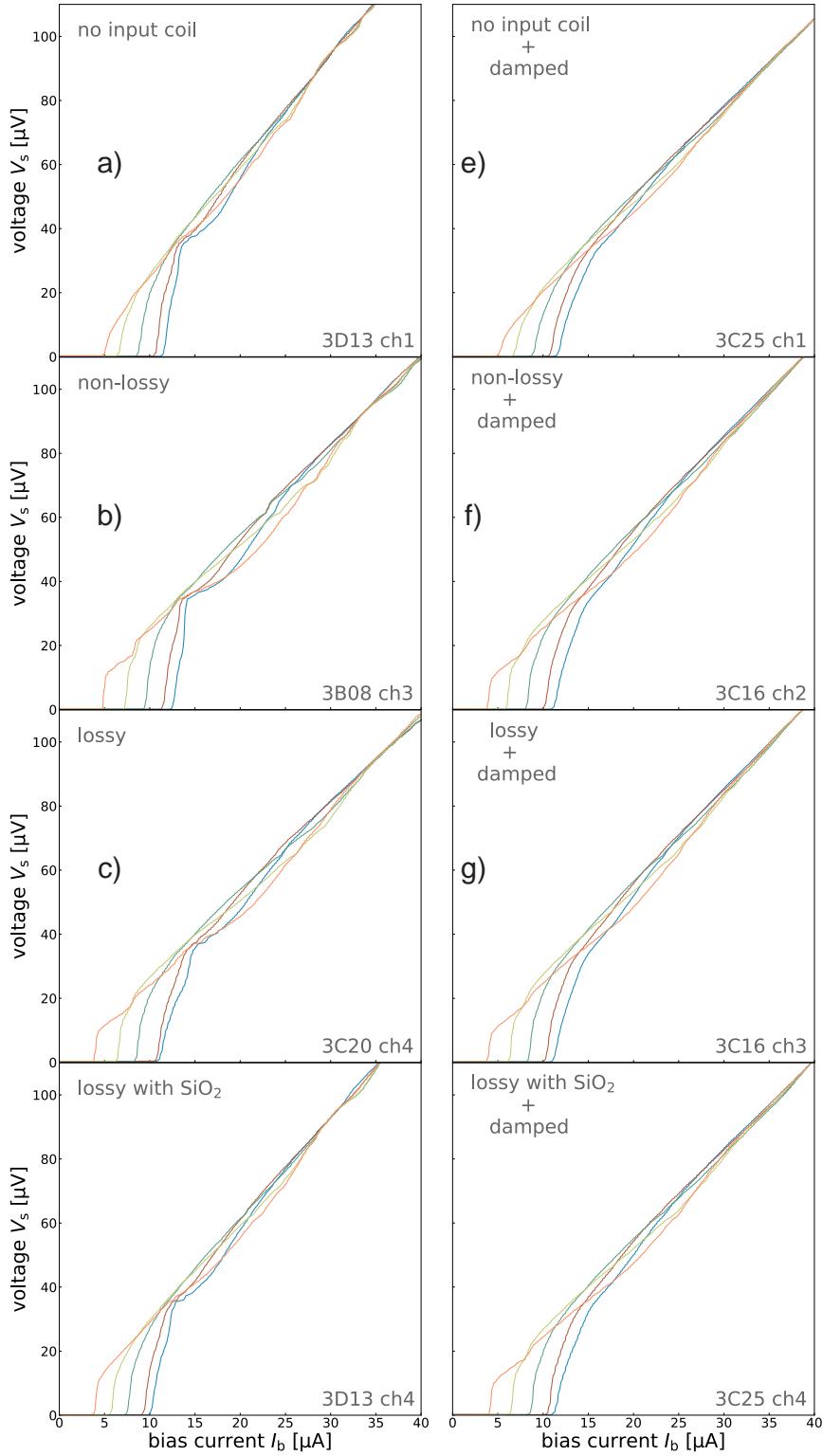


Figure 5.3: a) - d): Measured IVCs of the SQUID types presented in figure 4.4 in the same order, but without inductive damping. e) - h): The same measurement for the respective SQUID types but with inductive damping. Measured at  $T = 10\text{ mK}$ .

the damped and current-biased SQUIDs to be lower. Whether the various step structures apart from the large step at 40  $\mu\text{V}$  can be assigned to the different microstrip and  $LC$  resonances described in subsection 2.2.5 will be the subject of the following discussion.

A second prominent step structure seen in the IVCs of figure 5.3 occurs at approximately 15  $\mu\text{V}$  for the extremal curve with the smallest critical current  $I_{c,1}$ . This feature is absent in the IVCs of the SQUIDs without input coil (plots a) and e) ) and was also not visible for the design with a single-turn input coil, which was measured alongside the SQUIDs discussed in this section. Consequently, the second turn seems to cause an additional resonance effect, which could not be identified with the resonances described in the following. It is also unclear why this artefact does not appear in plot d) of figure 5.3.

Interestingly, there are no considerable differences between the characteristics of lossy and non-lossy variants. It is, however, worth noting that the input coils were not shorted or connected to a pickup coil during the experiments described in this chapter, which might result in an altered resonance behavior. The resonance as well as the noise properties of the new SQUID design with a shorted input coil should, therefore, be investigated in future works.

**$C_p$ -related resonances** For the  $C_p$ -related resonances, we need to take into account that the SQUID loops are connected in parallel and the input coil in series. The parasitic capacitance  $C_p^{\text{loop}}$  per washer loop, therefore, has to be multiplied by a factor of 4 or 1/4 respectively to obtain the total parasitic capacitance  $C_p$ . A theoretical value has been calculated to [Enp91, Enp92]

$$C_p^{\text{loop}} = \frac{l_{i,1/4} C_{\text{str}}}{8} = \frac{l_{i,1/4}}{8} \frac{\epsilon_0 \epsilon_r w_i K_f(w_i, t_i, h)}{h} , \quad (5.4)$$

where  $l_{i,1/4}$  denotes the input coil length on a single washer loop and  $C_{\text{str}}$  the microstrip line capacitance per unit length. The latter depends on the vacuum and relative permittivity  $\epsilon_0$  and  $\epsilon_r$ , as well as the width of the input coil  $w_i$ , the fringing factor  $K_f(w_i, t_i, h)$  and the insulation thickness  $h$ . The fringing factor also depends on  $w_i$ ,  $h$  and the thickness of the input coil denoted as  $t_i$  [Cha79]. With  $h = 375 \text{ nm}$ ,  $w_i = 3 \mu\text{m}$  and a sputtered Nb2 layer with  $t_i = 400 \text{ nm}$  we obtain the value  $K_f = 1.45$ . In case of the lossy SQUID variants, the niobium thickness of the input coil was fabricated to be  $t_i = 200 \text{ nm}$ , leading to  $K_f = 1.42$ . This has a small effect on the resulting resonance frequencies, as they only deviate approximately 2 % from the respective frequencies obtained for the non-lossy type. For the sake of clarity, we will, therefore, only consider the non-lossy SQUID in the following discussion, which will qualitatively represent the lossy type as well.

With a fringing factor of  $K_f = 1.45$ , we obtain a microstrip capacitance of  $C_{\text{str}} = 0.4 \text{ nF/m}$ . The input coil length per washer loop is approximately given by  $l_{i,1/4} \approx l_i/4 \approx 860 \mu\text{m}$ , with the total length of the input coil being  $l_i \approx 3.44 \text{ mm}$ . Consequently, we obtain for the parasitic capacitance per washer loop  $C_p^{\text{loop}} = 0.043 \text{ pF}$ . The total parasitic capacitance regarding the washer is, therefore,  $C_p = 4C_p^{\text{loop}} = 0.17 \text{ pF}$ , which is still small compared to the junction capacitance  $C = 0.95 \text{ pF}$ , i.e.  $C_p/C \ll 2$ . The influence of this parameter on the energy sensitivity should consequently be small, as explained in subsection 2.2.5. The  $L_s C_p$  resonance can now be calculated to  $f_{L_s C_p} = 1/(2\pi\sqrt{L_s 4C_p^{\text{loop}}}) = 34 \text{ GHz}$ , where we used the measured SQUID loop inductance  $L_s = 108 \text{ pH}$ . This corresponds to a voltage drop of  $V_s = 70.2 \mu\text{V}$  for the condition  $f_J = f_{L_s C_p}$ . Mostly the plots a) through c) from figure 5.3 showcase small distinctive step-like structures around  $75 \mu\text{V}$ . These could, however, also stem from a higher harmonic of the fundamental SQUID resonance, which should be the case for the SQUID without input coil (a)), as there is no corresponding parasitic capacitance. The value of  $f_{L_s C_p} = 34 \text{ GHz}$  is well above the optimal operation frequency for a current-biased SQUID, given by  $f_{\text{op}} \approx 0.3f_c \approx 5.1 \text{ GHz}$  ( $V_s = 10.5 \mu\text{V}$ ), with  $f_c = I_c R_s / \Phi_0$  [Can96] and is, therefore, not relevant in typical SQUID operation.

To discuss the  $f_{L_i C_p}$  resonance we need to take the  $R_x C_x$  attenuator into account. The capacitances  $C_x$  and  $C_p$  can approximately be added up to the total capacitance  $C_{\text{tot}} = C_x + C_p^{\text{loop}}/4$  due to the serial connection of the input coil. The geometric input coil inductance is shielded analogously to  $L_s$  (compare equation 3.18), giving the relation  $L'_i = (1 - k_{is}^2 s_s) L_i$  [Can96]. Here, the screening factor was derived to  $s_s = \beta_L s_i k_{is}^2 / (6 + 2\beta_L + \beta_L s_i k_{is}^2)$ . The input inductance of the SQUID and pickup coil inductance of the detector also form a parallel connection, resulting in  $L_{\text{tot}} = L'_i L_p / (L'_i + L_p)$ . The resonance frequency then reads

$$f_{L_i C_p} \approx \frac{1}{2\pi\sqrt{C_{\text{tot}} L_{\text{tot}}}} , \quad (5.5)$$

which by using our measured and calculated parameters yields  $0.9 \text{ GHz}$ . The corresponding voltage drop is  $V_s = 1.8 \mu\text{V}$ , which is well below the above-mentioned operation voltage and is, therefore, not visible in the IVCs from figure 5.3. The capacitance  $C_x$  is chosen small to minimize the attributed  $Q_{L_i C_p}$  value, which for an RCL parallel circuit is given by

$$Q_{L_i C_p} \approx R_x \sqrt{\frac{C_{\text{tot}}}{L_{\text{tot}}}} \approx \frac{R_x}{f_{L_i C_p} L_{\text{tot}}} . \quad (5.6)$$

Evidently,  $C_x$  cannot be chosen arbitrarily small as it would shift the resonance to the vicinity of the working point. An optimal value has been found to be  $Q_{L_i C_p} \approx 2$

[Can96]. The resistive component of the attenuator is typically dimensioned such that it corresponds to the nominal impedance  $Z_0$  of the microstrip line in order to mitigate wave reflections occurring where the input coil leaves the SQUID loop. This impedance can be expressed as [Enp91]

$$Z_0 = \sqrt{\frac{L_{\text{str}}}{C_{\text{str}}}} . \quad (5.7)$$

With the produced geometric proportions of the microstrip, the optimal value yields  $R_x = 20.6 \Omega$ , which was chosen as the design value. Therefore, with equation 5.6 we obtain an optimal capacitance of  $C_x = 31 \text{ pF}$ . However, for this first design, the capacitance of  $10 \text{ pF}$  chosen for the previous design in [Bau22] has been adapted to be on the safe side regarding the quality factor. In our case we obtain  $Q \approx 1.13$ , however, the corresponding resonance frequency is still small enough to consider the effect of this resonance as negligible.

We can also further neglect the  $C_p$ -related resonance with respect to the feedback coil, as its small design inductance of  $L_f = 336 \text{ pH}$  leads to a resonance frequency of  $f_{L_f C_p} = \frac{160 \mu\text{V}}{\Phi_0} = 77 \text{ GHz}$ , well above  $f_{\text{op}}$ .

**Microstrip line resonances** Regarding the  $\lambda/2$  resonances a distinction is made whether the inpt coil or the SQUID loop act as the signal carrying line. They will consequently be referred to as the input coil or washer resonance with the resonance frequencies  $f_{l_i}$  or  $f_{l_w}$ , respectively. These should ideally fulfill the condition  $4f_{l_i} < f_{\text{op}} < f_{l_w}/4$  to mitigate their noise inducing influence at the working point [Can91]. The wave propagation velocity  $c_{\text{str}}$  depends on  $L_{\text{str}}$  and  $C_{\text{str}}$ , such that equation 2.45 becomes [Enp92]

$$f_{l_i} = \frac{c_{\text{str}}}{2l_i} = \frac{1}{2l_i \sqrt{L_{\text{str}} C_{\text{str}}}} . \quad (5.8)$$

With the length of the input coil  $l_i$ , we obtain for the input coil resonance  $f_{l_i} = 17.6 \text{ GHz} = 36.4 \mu\text{V}/\Phi_0$ . However, we need to include the length  $l_i^{\text{feed}} \approx 620 \mu\text{m}$  of the input coil feed lines which are also structured as a microstrip transmission line. This results in  $14.9 \text{ GHz}$  and accordingly  $V_s = 30.9 \mu\text{V}$ . Even though the condition  $4f_{l_i} < f_{\text{op}}$  is not fulfilled, there are no visible step structures in the vicinity of this frequency, which suggests successful damping of this microstrip resonance. The effectiveness of the  $R_x C_x$  attenuator has also been demonstrated in [Bau22], where prominent step structures at  $f_{l_i} = 54 \mu\text{V}/\Phi_0$  were eliminated entirely after adding the  $R_x C_x$  component to the SQUID design with  $L_i = 1.27 \text{ nH}$ .

The washer resonance is more difficult to determine for the given SQUID loop design, as the complex washer structure surrounding the junction area affects the effective

length  $l_w$  of the signal carrying line. A first approximation can be done by adding the SQUID loop feed line length to the circumference of a single washer loop due to the parallel connection to estimate  $l_w$ . We thus obtain  $f_{l_w} \approx 26.9 \text{ GHz} \approx 50.3 \mu\text{V}/\Phi_0$ , which is more than four times the operation frequency. Just like for  $f_{l_i}$ , there aren't any artifacts in the SQUID IVCs that could be assigned to this resonance. This indicates that the washer shunt exhibits good damping properties, as has already been demonstrated in [Ric17] and [Bau18].

## 5.4 Noise Performance

The noise measurement was conducted at approximately 10 mK in the dilution cryostat, using the two-stage setup described in subsection 3.2.2, where the second stage SQUID is provided by a separate SQUID array chip. Figure 5.4 depicts the two-stage noise measurement of two lossy SQUIDs with  $\text{SiO}_2$  inside of the SQUID loop, one without and one with inductive damping from the chips 3A13 and 2A16, respectively. In the following we will use the name of the front-end chip to represent the corresponding two-stage setup. The flux noise  $\sqrt{S_\Phi}$  is given as the square root of the power spectral density in and is plotted against the frequency, ranging from approximately 30 mHz to 25 kHz. The measured data is shown as black dots and consists of a frequency-independent white noise and a low frequency  $1/f$  noise component, as described by equation 2.43. The fit of the total apparent flux noise are represented by a blue curve. The corresponding fit values are shown in table 5.2.

Chip	Contribution	$\sqrt{S_{\Phi_{s,w}}}$ [ $\mu\Phi_0/\sqrt{\text{Hz}}$ ]	$\sqrt{S_{\Phi_{s,1/f}}}(1 \text{ Hz})$ [ $\mu\Phi_0/\sqrt{\text{Hz}}$ ]	$\alpha$
3A13	total	0.41	3.90	0.87
	array	0.24	1.50	0.41
	front-end	0.23	3.52	0.85
2A16	total	0.57	4.50	0.84
	array	0.31	2.44	0.66
	front-end	0.30	3.54	0.82

Table 5.2: Overview of the determined noise contributions of the two-stage measurements with 3A13 and 2A16.

These values are, apart from the white noise of 3A13, relatively large which stems from larger contributions from the first and second stage SQUID, as discussed in the following. Both plots additionally contain the individual noise contributions from equation 3.9. To disentangle contributions it is necessary to also perform a noise measurement of the isolated SQUID array within the two-stage setup, which

is achieved by driving no currents through either the front-end or its coupled coils. The noise measured by this method corresponds to the apparent flux noise in the array, which is calculated by equation 3.9 for  $G_\Phi = 1$ ,  $S_{\Phi_s} = 0$  and  $R_{\text{dyn}} = 0$ . For the array chip in this operating mode we assumed a temperature of  $T = 150 \text{ mK}$ , which is caused by self-heating.

For the shunt resistors determined in section 5.1, we obtain a sheet resistance of  $1.23 \Omega/\square$ . The gain resistance  $R_g$  with a length of 0.3 squares thus yields  $0.37 \Omega$ . The transfer coefficients  $V_{\Phi_x}$  and  $I_{\Phi_x}$  are obtained prior to the noise measurement and yield  $V_{\Phi_x} = 602.2 \mu\text{V}/\Phi_0$  and  $I_{\Phi_x} = 11.71 \mu\text{A}/\Phi_0$  for 3A13, whereas for 2A16  $V_{\Phi_x} = 762.3 \mu\text{V}/\Phi_0$  and  $I_{\Phi_x} = 9.02 \mu\text{A}/\Phi_0$ . The intrinsic flux noise of the array  $S_{\Phi_x}$  can, therefore, be determined by subtracting the noise terms due to the gain resistor, current source and room temperature amplifier from the measured total noise. For the two-stage setup, the array noise contribution is reduced by the flux gain squared, as explained in subsection 3.2.2. For the measurements displayed in figure 5.4, we obtained  $G_\Phi = 2.34$  for the setup with the 3A13 chip and  $G_\Phi = 1.31$  for 2A16. The resulting array noise contribution is shown in table 5.2 for both measurements. The higher noise contribution of the array in the 2A16 setup as well as the lower  $G_\Phi$  value can be explained by dephasing of the array. This might be prevented in future experiments with better shielding of the experimental setup. Insufficient shielding allows external magnetic flux to be trapped inside the individual SQUID cells during the cooldown of the cryostat. If this added flux bias varies across the cells, the corresponding  $V\Phi$ -characteristics of the cells will be out-of-phase, resulting in a deteriorated  $V\Phi$ -characteristic of the array, which reduces  $\Delta V_{\max}$  and  $V_\Phi$ . For the two-stage measurement we assumed a larger array chip temperature of  $500 \text{ mK}$  due to the current flowing through  $R_g$ , resulting in considerable more dissipated heat. However, comparing with the array only measurement, the contribution of the gain resistor (yellow curve) is strongly reduced due to  $G_\Phi > 1$  and  $R_{\text{dyn}} \gg R_g$ , thus becoming negligible. The noise contribution of the current source for the array flux bias has been calculated with the values given in [Kaa20] and is shown by the pink curve in figure 5.4, which is also negligible. The green curve represents the room temperature amplifier noise, consisting of the current and voltage noise as described in equation 3.9. It is calculated with the values given in [Dru06] (see subsection 3.2.1). In the white noise regime it is comparable to the array contribution (brown curve), while for low frequencies the contribution of array noise is larger. The total noise, however, is dominated by the intrinsic noise of the first-stage SQUID (orange curve), which is obtained analogously to the intrinsic array noise by subtracting all other contributions from the total noise. The corresponding fit according to equation 3.9 provides the parameters shown in table 5.2. The white noise levels of both front-ends are comparable to previous low-noise SQUIDs developed in this working group [Fer15], however, the  $1/f$  noise is only slightly larger than the best

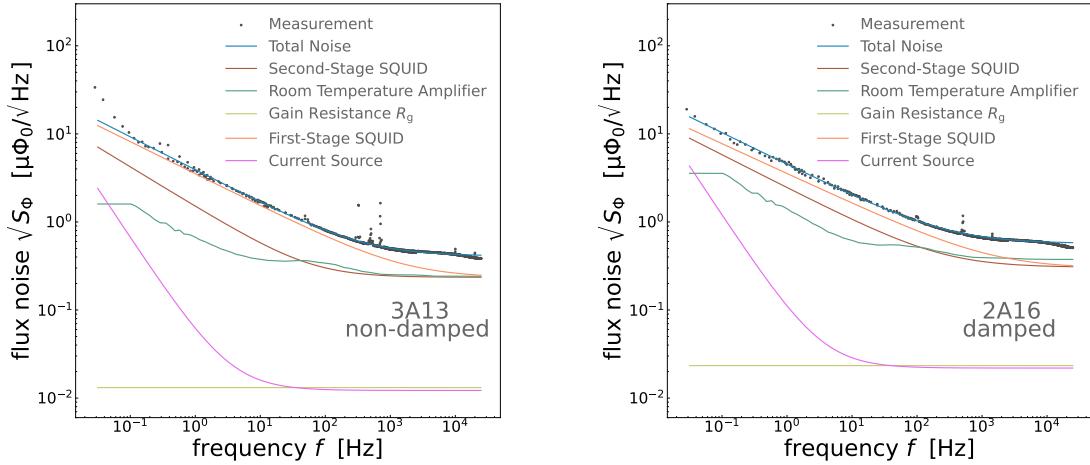


Figure 5.4: Measured noise spectra at  $T = 10\text{ mK}$  using a two-stage setup, where the low temperature amplifier is provided by a separate SQUID array chip. The blue curve represents the numerical fit of the measured total noise, whereas the remaining curves are the individual contributions of the first- and second-stage SQUID as well as the SQUID electronics consisting of the room temperature amplifier and the current source. Left: Lossy variant with  $\text{SiO}_2$  from the chip 3A13. Right: Lossy variant with  $\text{SiO}_2$ , including inductively damped feed lines (2A16).

devices produced in the working group, which are around  $1\text{-}3 \mu\Phi_0/\sqrt{\text{Hz}}$  [Kem16]. Surprisingly, the intrinsic noise of the first-stage SQUID from 3A13 is slightly lower than the inductively damped SQUID from 2A16. The smoother IVCs obtained by the inductive damping scheme should have a positive effect on the flux noise level. The extent to which this effect occurs is, however, unclear. The difference of the white noise level between both SQUIDs can also stem from differing SQUID loop inductances  $L_s$  or varying chip temperatures  $T$ , as the noise depends on these parameters according to equation 4.2. It is thus crucial to further investigate the noise spectra of all SQUID variants produced within the scope of this thesis to provide a more conclusive comparison.

Two other noise measurements involving the new SQUID design, but with different setups, were conducted in this working group and are now briefly compared to the aforementioned results. The first setup involves the integrated two-stage chip HDSQ16-W1-3A15 developed and tested in the work of [Kra23]. This chip contains a modified version of the array that was used for our two-stage measurements, which exhibits 18 SQUID cells instead of 16. Located next to it is the first-stage SQUID, where the non-lossy variant described in section 4.1 has been chosen. The noise mea-

surement was performed with the same two-stage configuration as described above. All noise contributions were determined the same way as well, which resulted in  $\sqrt{S_{\Phi,w}} = 0.42 \mu\Phi_0/\sqrt{\text{Hz}}$ ,  $\sqrt{S_{\Phi,1/f}}(1 \text{ Hz}) = 2.0 \mu\Phi_0/\sqrt{\text{Hz}}$  and  $\alpha = 0.60$  for the intrinsic noise of the front-end. The larger white noise contribution could be explained by the increased self-heating due to the proximity of the array to the front-end. Interestingly, in contrast, the low-frequency noise and the corresponding  $1/f$  exponent are reduced to values, which are comparable with the best SQUIDs produced in this working group. Further noise measurements of the non-lossy variant will allow to assess whether these values are attributed to the integrated two-stage setup, the Front-End design itself or just represent a statistical variation.

Lastly, a different type of noise measurement has been conducted by [Maz24], which involved a cross-correlation setup at 10 mK in the cryostat. The setup utilizes two array channels of a single array chip that are both connected to the front-end via their input coils [Her23]. As this method allows to cancel out the array noise, it is possible to directly obtain the intrinsic noise of the detector SQUID. The measured front-end from the chip 2A02 is of the same type as the one from 3A13 used for our two-stage setup, i.e. lossy with SiO<sub>2</sub> inside of the SQUID loop. The noise was determined to be  $\sqrt{S_{\Phi,w}} = 0.22 \mu\Phi_0/\sqrt{\text{Hz}}$ ,  $\sqrt{S_{\Phi,1/f}}(1 \text{ Hz}) = 3.0 \mu\Phi_0/\sqrt{\text{Hz}}$  and  $\alpha = 0.85$ , which is comparable to the values for the SQUID from 3A13. This measurement was part of a series of temperature-dependent noise measurements, which allowed to determine the minimum temperature  $T_{\min}$  of the front-end chip. The thermal decoupling of the SQUID from the cryostat occurred at  $T_{\min} \approx 90 \text{ mK}$ , which we will use for the following discussion regarding the energy sensitivity.

## 5.5 Energy Sensitivity

With the obtained flux noise we are able to determine the intrinsic and extrinsic energy sensitivities introduced in subsections 2.2.4 and 3.3.1. The theoretical values are obtained by using equation 4.2 for the theoretical intrinsic white noise of the SQUID. For the sake of clarity, they will be compared only to the noise measurement of 3A13, whereas the values for the other chips are summarized in table 5.3. With the measured parameters discussed in section 5.1 and for the assumed chip temperature  $T \approx 90 \text{ mK}$  we get  $\sqrt{S_{\Phi_s,w}^{\text{theo}}} = 0.11 \mu\Phi_0/\sqrt{\text{Hz}}$ , which compared to the measured values is reduced by a factor of 2 to 4. This could be explained by added noise that is mixed down due to the resonance-induced hysteretic behavior shown in the IVCs [Kem15]. Additionally, the current noise from the washer shunt is not taken into account in equation 4.2. The measured intrinsic energy sensitivity  $\epsilon_{s,w}$  will, therefore, deviate accordingly, yielding  $\epsilon_{s,w} = 1.58 h$ , whereas the theoretical value is  $\epsilon_{s,w}^{\text{theo}} = 0.36 h$ .

Parameter	$\sqrt{S_{\Phi_{s,w}}}$ [ $\mu\Phi_0/\sqrt{\text{Hz}}$ ]	$\sqrt{S_{\Phi_{s,1/f}}(1 \text{ Hz})}$ [ $\mu\Phi_0/\sqrt{\text{Hz}}$ ]	$\alpha$	$\epsilon_{s,w}$ [ $h$ ]	$\epsilon_{s,1/f}$ [ $h$ ]	$\epsilon_{p,w}$ [ $h$ ]	$\epsilon_{p,1/f}$ [ $h$ ]
3A13 (L)	0.23	3.52	0.85	1.58	369.3	25.26	5916
2A16 (LD)	0.30	3.54	0.82	2.68	373.5	43.0	5983
integrated (NL)	0.42	2.0	0.60	5.26	119.2	84.2	1910
2A02 (L)	0.22	3.0	0.85	1.44	268.2	23.11	4297
transf. (CJJ)	0.38	3.33	0.70	3.45	265.0	114.4	8784

Table 5.3: Overview of the determined noise and energy sensitivities of the SQUIDs described in this section. The acronyms stand for lossy (L), lossy and inductively damped (LD), non-lossy (NL), and cross-type Josephson junctions (CJJ). The latter represents the 6 nH double flux transformer SQUIDs measured in [Bau22].

The measured value of  $\epsilon_{s,w}$  is surprisingly low compared to typical values achieved in this working group [Kem15], however, for the  $1/f$  noise the value of  $\epsilon_{s,1/f} = 369.3 \text{ } h$  is rather large, but still within the range of typically achieved values of approximately  $50 - 500 \text{ } h$  [Kem16].

As discussed in subsection 3.3.1, the main goal of the increased input inductance was to achieve a better extrinsic energy sensitivity when using the SQUID for the read-out of MMCs with inductances of  $L_p = 6.65 \text{ nH}$ . By considering the gradiometric geometry of the pickup coil of the maXs100 detector and by using the obtained values  $M_{is} = 611 \text{ pH}$  and  $L_i = 6.4 \text{ nH}$ , the flux-to-flux coupling is calculated to  $\Delta\Phi_s/\Delta\Phi = 2.25 \%$ . This represents an increase over the value obtained in [Bau22], where the use of an additional flux transformer even deteriorated the coupling from 1.95 % to 1.75 %. In [Bau22], the noise with the 6 nH double flux transformer setup with the previous design using both window-type and cross-type junctions, had been measured using the cross-correlation method. The window-type variant exhibited an unexplained increase of the targeted critical current by a factor of 2, which led to strong hysteretic behavior and higher noise. We will therefore restrict the comparison to the cross-type SQUID 'SQ-CJJ-4w1 Chip10 #1'. The measured noise yields  $\sqrt{S_{\Phi_{s,w}}} = 0.38 \mu\Phi_0/\sqrt{\text{Hz}}$  for the white noise,  $\sqrt{S_{\Phi_{s,1/f}}(1 \text{ Hz})} = 3.33 \mu\Phi_0/\sqrt{\text{Hz}}$  for the  $1/f$  component and  $\alpha = 0.70$ . The obtained white noise of our measurements, therefore, represent an improvement, which together with the coupling gain will result in an overall smaller extrinsic energy sensitivity. As for the intrinsic energy sensitivity  $\epsilon_{s,w}$ , using equation 3.19 the measured extrinsic energy sensitivity  $\epsilon_{p,w} = 25.26 \text{ } h$  exceeds the predicted value  $\epsilon_{p,w}^{\text{theo}} = 5.71 \text{ } h$ . The  $1/f$  contribution results accordingly also in a large value of  $\epsilon_{p,1/f} = 5916 \text{ } h$ . These values, however, represent a significant reduction as compared to the previous design with an intermediary coupling transformer. Given the aforementioned parameters, the cross-type version yields  $\epsilon_{p,w} = 114.4 \text{ } h$  and  $\epsilon_{p,1/f} = 8784 \text{ } h$ . It should be noted though, that the

white noise strongly depends on the chip temperature, which can vary from setup to setup. Furthermore,  $1/f$  noise typically exhibits large variations across the wafer. It is, therefore, necessary to collect a larger set of noise measurements to confirm whether the new design exhibits a flux noise comparable to previous SQUID designs.



## 6. Summary and Outlook

In this work a new dc-SQUID design was developed and optimized for the readout of the MMC-based X-ray detector maXs100, which is developed in this working group [All24]. For this, a design from the work of [Bau22] was adapted by increasing the input coil inductance to match the inductance of the detector pickup coil. Contrary to the intermediary flux transformer approach tested in [Bau22], the inductance was adjusted by adding a second turn to the input coil. This maximizes the flux-to-flux coupling  $\Delta\Phi_s/\Delta\Phi$ , which is a crucial parameter to maximize the signal-to-noise ratio, thereby increasing the energy resolution  $\Delta E_{FWHM}$  of the detector setup.

To assess whether the design changes lead to additional noise or to a deteriorated resonance behavior, the SQUID was tested both in a single-stage setup to investigate general properties including the current-voltage characteristics, and in a two-stage setup to measure the flux noise. Additionally, new damping techniques were implemented and tested to suppress various resonances originating from either *LC* resonant circuits or from microstrip structures. These involved the deposition of additional galvanically isolated gold structures onto the washer and feedback coil feed lines, as well as the fabrication of the input coil as a bilayer containing both gold and niobium. The new SQUID design was produced with different damping scheme configurations to examine the effectiveness of the employed damping approaches.

The single-stage experiments were conducted both in a liquid helium transport vessel at 4.2 K and in a dilution cryostat at 10 mK. In the case of the latter, the characteristic parameters of 10 SQUIDs were measured, which except for the SQUID loop inductance resulted to be close to their target values. The SQUID loop inductance was determined to  $L_s = 108 \text{ pH}$ , deviating 27 % from the target value of 147 pH. This can be explained by the conservative adjustment of the washer loops, which had to be enlarged to compensate for the inductance loss from widening the lines from 5  $\mu\text{m}$  to 10  $\mu\text{m}$  fit the second input coil turn. Additionally, with the previous SQUID design the inductance  $L_s$  had already been overestimated, yielding only 135 pH. This could be confirmed in this work with similar values. The input coil inductance was measured at 4.2 K and determined to  $L_i = 6.4 \text{ nH}$  as the average of two measurements. This represents a deviation of 14 % from the simulated value of 5.6 nH, which is in agreement with observations from [Fer15, Bau22].

The IVCs of eight SQUIDs were obtained at 10 mK in the cryostat, each with a different damping scheme to explore the influence of the damping structures on the resonances. The most effective method was the inductive damping of the feed lines, as it showcased significantly smoother IV curves as compared to the non-

damped counterparts. The implementation of the input coil as a lossy microstrip line, however, seemed to provide no considerable reduction of the observed resonance artefacts. It is noteworthy though, that the input coils were in an open loop and not shorted or connected to another coil. This might affect input coil related resonances, which should be investigated in future works.

The flux noise measurements were performed alongside the IVC measurements in the same cryostat and at the same temperature. Three front-end chips were measured and yielded white noise values below  $0.3 \mu\Phi_0/\sqrt{\text{Hz}}$ , while the  $1/f$  noise contribution was around  $3.5 \mu\Phi_0/\sqrt{\text{Hz}}$ . These values are similar to the noise of the previous design with an intermediary flux transformer setup measured in [Bau22]. The results are comparable to other low-noise SQUIDs developed in this working group, even though smaller values down to  $< 1 \mu\Phi_0/\sqrt{\text{Hz}}$  have been achieved in the low frequency regime [Fer15]. The noise measurements indicate, however, that the inductive damping technique did not significantly raise the overall noise levels, thus representing a technique worth pursuing in the future. The flux-to-flux coupling  $\Delta\Phi_s/\Delta\Phi$  regarding the maXs100 detector yielded 2.25 %, which represents an increase of 29 % over the value of 1.75 % obtained with the intermediary flux transformer in [Bau22]. Consequently, the extrinsic energy sensitivities are comparatively low, where the best value regarding the white noise yielded  $\epsilon_{p,w} = 23.11 \text{ h}$  and regarding the  $1/f$  noise  $\epsilon_{p,1/f} = 1910 \text{ h}$ .

A possible measure to further increase the flux-to-flux coupling while also reducing the white noise would be a better adjustment of the SQUID loop and input coil inductance, which can be realized by expanding the washer loop, such that the inductances are shifted towards their target values  $L_s = 147 \text{ pH}$  and  $L_i = 7.15 \text{ nH}$ , respectively. Furthermore, substituting window-type with cross-type junctions would allow the use of shunt resistances twice as large, leading to a significant reduction of the intrinsic white noise by a factor of 2. The four times smaller junction capacitance would also shift the fundamental SQUID resonance further away from the working point, resulting in better performance. Nevertheless, the new SQUID design represents an improved detector SQUID for MMCs with larger pick-up coils, as for example the maXs100 detector.

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Ich versichere, dass ich diese Arbeit selbstständig verfasst und keine anderen als die angegebenen Quellen und Hilfsmittel benutzt habe.

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