
Doctorate of National Interest
in
Space Science and Technology

Experimental and analytical study of
electrical crosstalk in CMB cold readout
systems and implications for power
spectra

Eugenia Di Giorgi

Supervisor
Prof. Giovanni Signorelli
Università di Pisa

Co-Supervisor
Prof. Luca Baldini
Università di Pisa

*While we were fearing it, it came
But came with less of fear
Because that fearing it so long
Had almost made it fair*
— Emily Dickinson

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Contents

Abstract	9
1 The Cosmic Microwave Background	11
1.1 An introduction to cosmology	11
1.1.1 Inflation	13
1.2 The Cosmic Microwave Background	15
1.3 CMB anisotropies	17
1.3.1 The dipole anisotropy	17
1.3.2 Primordial anisotropies	18
1.4 CMB polarization	21
1.4.1 E -modes and B -modes	22
1.4.2 A Closer Look at B -modes	25
1.5 Conclusions	27
2 The LiteBIRD experiment	28
2.1 The scientific goal	29
2.1.1 Other science outcomes	30
2.2 The instrument	31
2.3 The Transition Edge Sensors	35
2.3.1 The Noise Equivalent Power	40
2.3.2 The TESs for LiteBIRD	42
2.3.3 The TESs for Athena	43
2.4 The SQUIDs	45
2.4.1 Flux-Locked Loop	48
2.4.2 The SQUID gain-bandwidth product	49
2.5 The “warm” readout: the electronics of LiteBIRD	52
2.6 From the instrument to the CMB power spectra	53
2.7 Conclusions	55

3	The Frequency Domain Multiplexing (FDM) and the Electrical Crosstalk	56
3.1	Multiplexing readout	56
3.1.1	Lumped elements for Frequency Division Multiplexing	60
3.2	A systematic error for FDM: electrical crosstalk	63
3.2.1	Carrier leakage crosstalk	63
3.2.2	Mutual inductance crosstalk	64
3.2.3	Common impedance crosstalk	66
3.3	Conclusions	68
4	The design optimization of LC filters and multiplexing frequency scheduling	69
4.1	Multiplexing frequency scheduling	69
4.1.1	The algorithm applied to LF12	76
4.2	A LC filters chip design for LF12	78
4.3	Conclusions	83
5	Characterization and crosstalk measurements for an LC filters chip with parallel capacitors	87
5.1	A lithographed LC chip for the LSPE/SWIPE experiment	87
5.2	The LC filters geometry	88
5.3	The experimental setup	90
5.4	Measurements	92
5.4.1	Resonance frequencies, quality factor and parasitic resistance	92
5.4.2	Temperature scan	94
5.4.3	Frequency scan and common inductance evaluation	95
5.5	Electrical crosstalk measurements	98
5.5.1	Pulse shape modulation	99
5.5.2	Analytical model and measurements: a comparison	101
5.6	Sine-wave modulation	104
5.7	Conclusions	106
6	Characterization and crosstalk measurements for an end-to-end readout chain of LiteBIRD	107
6.1	The experimental set-up	107
6.2	Measurements and tests	109
6.2.1	SQUID properties	109

6.2.2	Network analysis	112
6.2.3	Preliminary assessment of Electrical Crosstalk	120
6.2.4	Magnitude and in phase signals for crosstalk analysis	125
6.3	Conclusions	125
7	Propagation of crosstalk on the CMB maps and spectra	127
7.1	Pipeline to simulate the propagation of crosstalk on the maps	127
7.2	Simulations and results	128
7.2.1	Detectors on the same pixel with 90° polarization separation and identical frequency channel	129
7.2.2	Detectors on different pixels with 45° polarization separation and identical frequency channel	133
7.2.3	Detectors on different pixels with 90° polarization separation and different frequency channels	135
7.2.4	Detectors on different pixels non adjacent in the multiplexing frequency sequence	137
7.3	Mitigation of Electrical Crosstalk	139
7.4	Conclusions	140
8	Conclusions	141
	References	143

Abstract

The Cosmic Microwave Background Radiation (CMB) originated in the early phases of the the life of the Universe, during the *recombination epoch*, when protons and electrons combined to form hydrogen, emitting photons that free-streamed to the present epoch.

The CMB frequency and power spectra have been measured and characterized by groundbreaking space missions such as COBE, which first detected the CMB anisotropies, WMAP, which provided improved angular resolution, and Planck, which achieved the most precise measurements to date. Future missions, such as the LiteBIRD (space-based) and LSPE/SWIPE (balloon-borne), aim to measure the polarization of the CMB, in particular the so called B-modes, to provide evidence for the current inflationary models.

The new generation of CMB experiments employ superconducting devices to improve the sensitivity of the observations. The detection chain of both LiteBIRD and SWIPE is based on Transition Edge Sensors (TESs), which exploit the superconductors' strong variation of resistance with temperature, thereby converting the detector heating produced by the sky signal into an electrical signal. Due to the small size of the signal Superconducting QUantum Interference Devices (SQUIDs) are used as cryogenic amplifiers. To limit the thermal load on the cryogenic stages, Frequency Division Multiplexing (FDM) is employed to reduce the number of SQUIDs and readout cables. In this approach, each TES is connected in series with a band-pass LC filter and biased at a specific frequency, while the combined output signals from multiple detectors are read out through a single SQUID.

FDM is most commonly implemented using fully lithographed LC filters, consisting of spiral inductors and planar parallel-plate or interdigitated capacitors. Nonetheless it carries instrumental systematic effects, among which electrical crosstalk is one of the most critical. This is related to the presence of the filters themselves which can create signal leakage among channels. There are three main sources of electrical crosstalk that were identified: carrier leakage crosstalk, associated with

the bandwidth of the filters; mutual inductance crosstalk, arising from possible electromagnetic coupling between spiral inductors; and common impedance crosstalk, caused by the presence of a non-zero inductance in series with all the multiplexing branches. The LC filters chip design, together with the sequence of the multiplexing frequencies, affects the overall crosstalk level. Measuring and understanding the impact of crosstalk on the readout chain of these experiments is of utmost importance. The work presented in this thesis focuses on the cold readout chain for those CMB experiments, addressing both simulation and experimental aspects. Optimization studies were carried out for the design of an LC filter chip entirely fabricated using lithographic techniques and composed of spiral inductors and interdigitated capacitors. In particular, an algorithm was developed to determine the optimal configuration that minimizes electrical crosstalk, based on the focal plane layout. LC assignments, linking the sequence of multiplexing frequencies to that of the TESs, have a significant impact on the crosstalk level.

Analytical methods for estimating crosstalk were applied to the LiteBIRD and SWIPE readout chains. The results show that the expected level of crosstalk depends strongly on the separation between multiplexing frequencies and on the value of the stray inductance associated with the SQUID input coil and the wiring.

Finally, experimental measurements on LC chips representative of both experiments were performed to validate the developed models, showing a good match. In the case of LiteBIRD readout, tests on a complete end-to-end channel, including the warm electronics, were carried out. The impact of electrical crosstalk on the observational outcomes was eventually assessed by propagating its effects to the CMB power spectra, evaluating the potential influence on the LiteBIRD results.

Keywords

[Multiplexing, electrical crosstalk, Transition Edge Sensors, CMB]

Chapter 1

The Cosmic Microwave Background

A general introduction to cosmology is presented in this chapter, followed by a more detailed description of the temperature and polarization properties of the Cosmic Microwave Background.

1.1 An introduction to cosmology

The Λ CDM (*Lambda cold dark matter*) model currently provides the most accurate and comprehensive description of our Universe. This is based on three major constituents: ordinary matter, cold dark matter (CDM) and a cosmological constant (Λ) quantifying the dark energy. In the Λ CDM model, the large-scale Universe is assumed to be homogeneous and isotropic, and can be mathematically described by the Friedmann–Lemaître–Robertson–Walker (FLRW) metric:

$$ds^2 = -dt^2 + a^2(t) \left[\frac{dr^2}{1 - Kr^2} + r^2 (d\theta^2 + \sin^2 \theta d\phi^2) \right]. \quad (1.1)$$

where $a(t)$ is the *scale factor* and K determines the curvature of the Universe. The quantities r , θ and ϕ are the spherical coordinates. The FLRW line element corresponds to the following metric tensor $g_{\mu\nu}$:

$$g_{\mu\nu} = \begin{pmatrix} -1 & 0 & 0 & 0 \\ 0 & \frac{a^2(t)}{1 - Kr^2} & 0 & 0 \\ 0 & 0 & a^2(t) r^2 & 0 \\ 0 & 0 & 0 & a^2(t) r^2 \sin^2 \theta \end{pmatrix} \quad (1.2)$$

Based on this, the metric equation can be rewritten as:

$$ds^2 = g_{\mu\nu} dx^\mu dx^\nu \quad (1.3)$$

The evolution of $a(t)$ is modeled by the *Friedmann equations*, a pair of equations describing the variation of the Universe scale in terms of the pressure p and the energy density ρ of a perfect fluid, as shown in the following equations.

$$\frac{3\ddot{a}}{a} = -4\pi G(3p + \rho) \quad (1.4)$$

$$\left(\frac{\dot{a}}{a}\right)^2 + \frac{K}{a^2} = \frac{8\pi G}{3}\rho \quad (1.5)$$

where G is the gravitational constant $G = 6.6743 \cdot 10^{-11} m^3 Kg^{-1} s^{-2}$. The energy density ρ includes three components: the non relativistic matter (ρ_M) (which includes barionic and dark matter), the ultra-relativistic radiation (ρ_R) and the vacuum energy (ρ_Λ).

Eq. 1.5 shows how the *Hubble* parameter $H = \dot{a}(t)/a(t)$ depends on the total energy density ρ . A positive energy density is linked to an expanding Universe through the relation $\rho \propto a^{-3(1+w)}$ where $w = p/\rho$. The rate of expansion varies over the dominant components, radiation, matter or dark energy, since each component scale differently with t .

- non relativistic fluid: $a(t) \propto t^{\frac{2}{3}}$
- ultra relativistic fluid: $a(t) \propto t^{\frac{1}{2}}$
- vacuum energy: $a(t) \propto e^{Ht}$

The relative contributions of matter, radiation, and dark energy determine whether the Universe is open, closed, or spatially flat. By comparing the total density of the Universe to a critical value, $\rho_c \equiv \frac{3H^2(t)}{8\pi G}$, defined for a flat Universe with $\Lambda=0$, one can introduce the density parameter $\Omega_{x,0}$, which characterize the present-day relative density for each component ρ_x . This parameter is defined as $\Omega_{x,0} = \rho_x(t)/\rho_c$, where x specifies the different energy density components. These include $\Omega_{M,0}$ for matter, $\Omega_{R,0}$ for radiation, $\Omega_{K,0}$ for curvature and $\Omega_{\Lambda,0}$ for dark energy.

The FLRW equation describes the evolution of the Universe as a function of these density parameters:

$$H^2 = H_0^2 \left(\frac{\Omega_{M,0}}{a^3} + \frac{\Omega_{R,0}}{a^4} + \Omega_{\Lambda,0} \right) \quad (1.6)$$

where $H_0 = \dot{a}(t_0)/a(t_0)$ is the Hubble parameter evaluated at the present epoch. The Λ CDM model predicts:

$$\Omega_{tot} = \Omega_{M,0} + \Omega_{R,0} + \Omega_{\Lambda,0} \approx 1 \quad (1.7)$$

The curvature term in the FLRW metric vanishes, corresponding to a spatially flat Universe ($K = 0$). This theoretical prediction is supported by observations of Type Ia supernovae and the Cosmic Microwave Background, which indicate that the Universe shows no measurable curvature on large scales. The data of the last space missions observing the CMB, *Planck*, show [1]:

$$\begin{cases} \Omega_R = 0 \\ \Omega_K = 0.0007 \pm 0.0019 \\ \Omega_M = 0.3147 \pm 0.0074 \\ \Omega_\Lambda = 0.6889 \pm 0.0056 \end{cases} \quad (1.8)$$

1.1.1 Inflation

Given the definition of the main parameters in the Λ CDM model, some uncertainties regarding the initial conditions still remain. Cosmic inflation is introduced to address these uncertainties by introducing an early phase of accelerated expansion ($\ddot{a} \gg 0$) and the generation of primordial fluctuations, that gave origin to the large-scale structures of today's Universe.

The existence of the inflationary period would solve three major problems that the

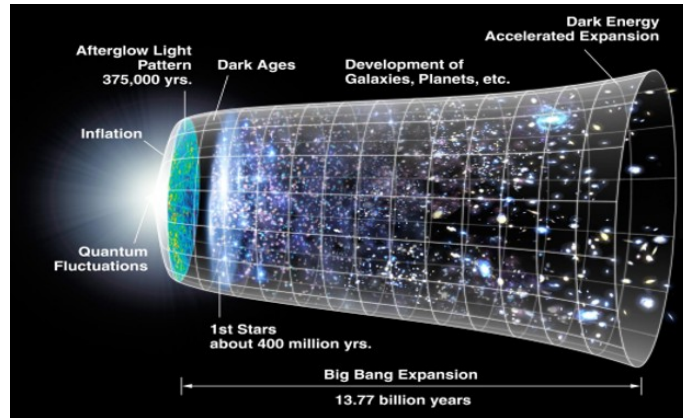


Figure 1.1: Scheme of the expansion phases of the Universe life.

Λ CDM model fails to describe:

1. The flatness problem arises because, in order to obtain the present-day value of $\Omega_{K,0}$, corresponding to a spatially flat Universe, the initial conditions of

the early Universe must be extremely fine-tuned. The variation of $\Omega_{K,0}$ from the Planck era to the matter dominated epoch must have followed a specific evolutionary behaviour, which implies the existence of a period of accelerated expansion, characterized by $\ddot{a} > 0$.

2. The horizon problem originates from the observation that regions of the Universe which were not causally connected at the last scattering surface exhibit a remarkable degree of homogeneity. The high level of isotropy observed in the CMB implies that these regions must have once been in causal contact long enough to reach uniformity. An exponential expansion in the early phase of the Universe life would account for this: the Universe, once condensed in a small and causally connected region, expanded rapidly, explaining the uniformity of the CMB that could not otherwise have been achieved.
3. The monopole problem concerns the non-observation of magnetic monopoles, whose existence was postulated by Grand Unified Theories (GUTs). Inflation is thought to have occurred at energies below the so-called GUT scale, that is, after the formation of monopoles. An exponential expansion of the Universe would naturally dilute their density to non-observable levels, thus providing a solution to this problem.

The dominant inflationary model today is the *slow-roll inflation* [2],[3]. Slow-roll inflation is described by a scalar “inflaton field” ϕ slowly rolling down a potential energy hill. From the energy-momentum tensor

$$T_{\mu\nu} = \partial_\mu\phi\partial_\nu\phi - g_{\mu\nu} \left[\frac{1}{2}\partial_z\phi\partial^z\phi + V(\phi) \right] \quad (1.9)$$

the energy and pressure of this field can be defined as:

$$\begin{cases} \rho_\phi = \frac{\dot{\phi}^2}{2} + V(\phi) \\ p_\phi = \frac{\dot{\phi}^2}{2} - V(\phi) \end{cases} \quad (1.10)$$

where $V(\phi)$ is the potential of the field. The time evolution of ϕ is given by:

$$\ddot{\phi} + 3H\dot{\phi} + V'(\phi) = 0 \quad (1.11)$$

where the prime indicates the derivative with respect to ϕ .

Here H behaves as a friction parameter and causes the “slow rolling” of the field. It is defined as:

$$H^2 = \frac{8\pi G}{3} \left(\frac{1}{2}\dot{\phi}^2 + V(\phi) \right) \quad (1.12)$$

Combining eq. 1.11 and eq. 1.12 we can calculate $\dot{H} = -4\pi G\dot{\phi}^2$. In this case, the Friedmann equation is:

$$\frac{\ddot{a}}{a} = -\frac{4\pi G}{3}(2\dot{\phi}^2 - 2V(\phi)) \quad (1.13)$$

Defining $\epsilon \equiv -\dot{H}/H$ as the *slow-roll parameter*, eq. 1.13 can be rewritten as:

$$\frac{\ddot{a}}{a} = \dot{H} + H^2 = (1 - \epsilon)H^2 \quad (1.14)$$

Based on this, the condition for the inflation to occur ($\ddot{a} \gg 0$) is to have $\epsilon \ll 1$. Defining N as the number of *e-foldings* for which $dN = H dt$, the slow-roll parameter can be rewritten as:

$$\epsilon = -\frac{d \ln H}{dN} \quad (1.15)$$

and can be demonstrated [3] that N can be determined by:

$$N \approx \sqrt{4\pi G} \int_{\phi_1}^{\phi_2} \frac{d\phi}{\sqrt{\epsilon}} \quad (1.16)$$

where ϕ_1 and ϕ_2 are the initial and final conditions of inflation. A slow-roll inflation lasting at least about $N \simeq 60$ –65 e-foldings expands the regions in causal contact enough to solve the horizon problem.

1.2 The Cosmic Microwave Background

The CMB was theorised in the 1940's but was measured for the first time in 1965 by Arno Allan Penzias and Robert Woodrow Wilson [4], while making some uncorrelated radio waves measurements. After removing all the known interference, a mysterious residual radiation at 3.5 K persisted in their results, coming from all the directions in the sky during both the day and the night hours. Teaming up with the Princeton researchers Dicke, Wilkinson and Peebles, Penzias and Wilson ended up discovering the Cosmic Microwave Background radiation, leading to important steps forward in the field of cosmology.

The CMB is a free streaming radiation which originated at the early stages of the life of the Universe. In particular, it was generated during the *epoch of recombination* [5]. Protons and electrons combined to form neutral atoms, leading to the emission of photons. From that moment, radiation decoupled from matter and began to travel freely through space. This occurred when the effects of scattering between electrons and photons became negligible, a moment referred to as the *last scattering*. The timing of this phenomenon can be calculated by comparing the Universe expansion

rate (H) with the interaction rate between the photons γ and the free electrons (Γ). The decoupling could take place when the former was larger than the latter. By equalling $H = \Gamma$, the redshift and the temperature of last scattering z_L and T_L can be defined: the numerical values are reported below ([6]).

$$\begin{cases} z_L \simeq 1100 \\ T_L \simeq 3000K \end{cases} \quad (1.17)$$

The surface located at the appropriate epoch, from which the CMB is considered to have been emitted, is called the *last scattering surface*.

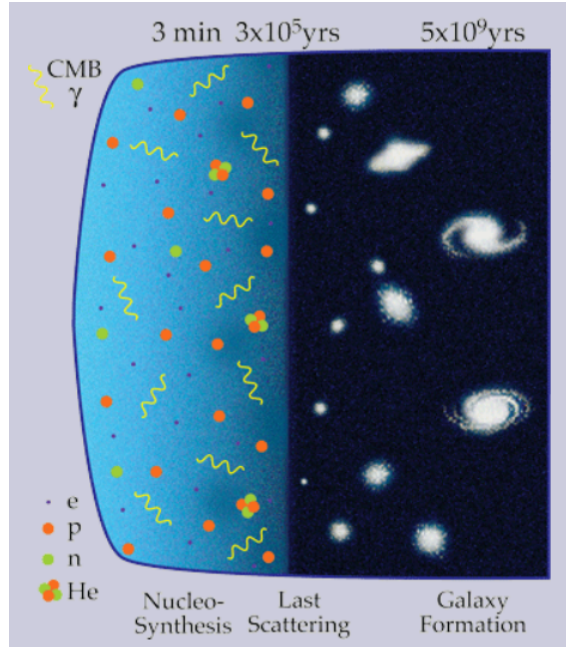


Figure 1.2: Visual representation of the periods before and after the CMB emission [7].

The CMB decoupled as an isotropic radiation in thermal equilibrium with matter. The number density distribution of photons in equilibrium at a temperature T , in the frequency range between ν and $\nu + d\nu$, has the shape of a *black-body spectrum* [8].

$$n_T(\nu)d\nu = \frac{8\pi\nu^2 d\nu}{e^{\frac{h\nu}{kT}} - 1} \quad (1.18)$$

Once decoupled and freely propagating, the CMB radiation retained the shape of a black-body spectrum. Nowadays, this spectrum peaks at a temperature of $T_{CMB} = 2.725 \pm 0.002$ K [9], having cooled due to the expansion of the Universe.

The first precise measurement of the black-body spectrum of the CMB was made by the FIRAS radiometer carried by the Cosmic Background Explorer Satellite (COBE)

[10] in the 1990s. It was shown that the cosmic background radiation actually fitted the black-body spectrum (Figure 1.3).

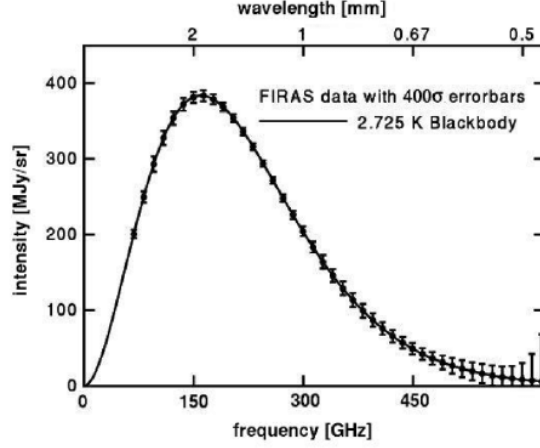


Figure 1.3: Intensity of radiation observed by FIRAS radiometer carried by COBE (dots). The continuous black line is a black-body spectrum fit with temperature $T = 2.728K$ ([10]). On the y -axis the intensity is presented in MegaJansky per steradian as a function of frequency ($c=1$). The errorbars are multiplied by 400.

1.3 CMB anisotropies

The Cosmic Microwave Background is isotropic and homogeneous with a good approximation. However, by increasing the resolution of the observations, variations in temperature can be observed.

1.3.1 The dipole anisotropy

Studying deviations from perfect isotropy provides fundamental insights into the evolution of the Universe. The first type of anisotropy originates from the motion of the Earth and the Galaxy with respect to the CMB.

Considering the photon density in a system at rest relative to the CMB and its Lorentz transform in the Earth's frame of reference (with a velocity $v = 369.82 \pm 0.11$ km/s and $\beta = v/c = (1.23357 \pm 0.00036) \times 10^{-3}$ [9]), we can estimate the transformed photon density and the corresponding temperature.

$$\begin{cases} N'_\gamma(p') = \frac{1}{h^3} \frac{1}{\exp(\frac{|p'|c}{kT}) - 1} \\ T' = \frac{1}{\gamma(1 + \beta \cos \theta)} \end{cases} \quad (1.19)$$

T' increases when $\cos \theta = -1$, that is, when the photons arrive from the direction toward which the Earth is moving, and decreases when $\cos \theta = 1$, corresponding to

photons coming from the direction opposite to the Earth's motion.

The difference in temperature between a moving system and a system at rest on a spherical surface can be described using Legendre polynomials [8]:

$$\Delta T = T' - T = T \left[-\frac{\beta^2}{6} - \beta P_1(\cos \theta) + \frac{2\beta^2}{3} P_2(\cos \theta) + \dots \right] \quad (1.20)$$

The main contribution to 1.20 is the dipole term with a minor contribution of the “kinematic quadrupole”. The contribution to the CMB anisotropy coming from the earth kinematics is of the order $\Delta T/T \sim 10^{-3}$.

1.3.2 Primordial anisotropies

Beyond the dipole anisotropy caused by the motion of the Earth with respect to the CMB rest frame, smaller-scale variations reveal the presence of primordial anisotropies. These were first detected by the COBE-DMR experiment. Following these measurements, additional observations were conducted either from the ground or from a balloon. Two satellites successors of COBE were launched: the *Wilkinson Microwave Anisotropy Probe* (WMAP) [11] and *Planck*. In figure 1.4 the improvement in the instrumental resolution can be observed.

The quantity ΔT in a specific spatial direction $\hat{n}$ can be defined as ([8]):

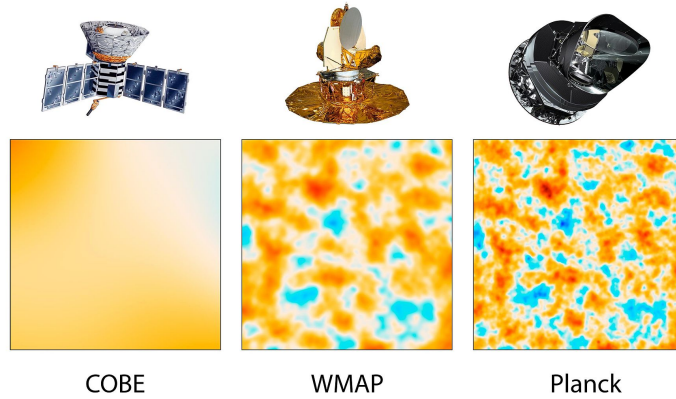


Figure 1.4: Comparison among COBE, WMAP and Planck instrumental resolution on CMB anisotropies observations ([12]).

$$\Delta T(\hat{n}) = T(\hat{n}) - T_0 \quad (1.21)$$

where $T_0 = 2.725K$ is the average CMB temperature.

The fluctuations can be decomposed into spherical harmonics as:

$$\frac{\Delta T}{T_0}(\hat{n}) = \sum_{\ell=0}^{\infty} \sum_{m=-\ell}^{\ell} a_{\ell m} Y_{\ell}^m(\hat{n}) \quad (1.22)$$

where the coefficients $a_{\ell m}$ quantify the intensity of the fluctuation in that particular direction and ℓ is the angular momentum labelling the order of multipole.

As we state that CMB is isotropic at large scales, eq. 1.23 sets a constraint on $\langle \Delta T \rangle$.

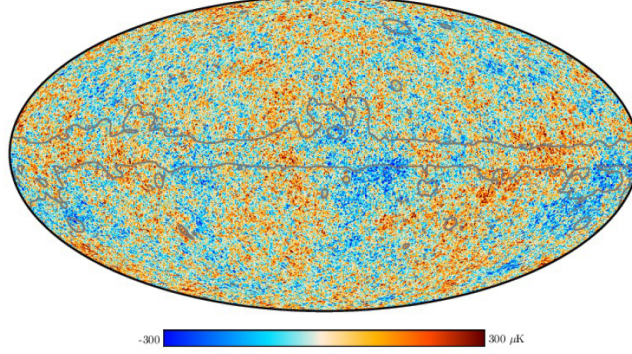


Figure 1.5: Temperature anisotropies map [9]. The Galaxy signal is masked.

$$\langle \Delta T \rangle = \int \Delta T(\hat{n}) d^2 \hat{n} = 0 \quad (1.23)$$

but the correlation between the fluctuations in two different directions, $\hat{n}$ and $\hat{n}'$, $C(\theta) \equiv \langle \frac{\Delta T(\hat{n})}{T_0} \frac{\Delta T(\hat{n}')}{T_0} \rangle$ is not negligible.

The equation for the correlation quantity $C(\theta)$ is:

$$C(\theta) = \sum_{\ell=2}^{\infty} C_{\ell} \frac{2\ell+1}{4\pi} P_{\ell}(\cos\theta) \quad (1.24)$$

where ℓ starts from the *quadrupole* (the dipole term was already included in the large scale fluctuations), P_{ℓ} are the *Legendre polynomials* and θ is the angle between $\hat{n}$ and $\hat{n}'$. The coefficients C_{ℓ} are defined through the relation

$$\langle a_{\ell m} a_{\ell' m'}^* \rangle = \delta_{\ell \ell'} \delta_{m m'} C_{\ell} \quad (1.25)$$

and describe the angular power spectrum of the temperature fluctuations. The multipole ℓ corresponds approximately to an angular scale π/θ , and the coefficients C_{ℓ} quantify the amplitude of the temperature fluctuations at that scale.

Inverting Eq.1.24 and applying the definition of $C(\theta)$ one obtains the relation used to calculate the coefficients:

$$C_{\ell} = \frac{1}{4\pi} \int d^2 \hat{n} d^2 \hat{n}' P_{\ell}(\hat{n} \cdot \hat{n}') \langle \Delta T(\hat{n}) \Delta T(\hat{n}') \rangle \quad (1.26)$$

The CMB temperature anisotropies map and the power distribution are shown by the data from *Planck* in Figure 1.5 and Figure 1.6 [9]. For $\ell > 1000$, the fluctuations

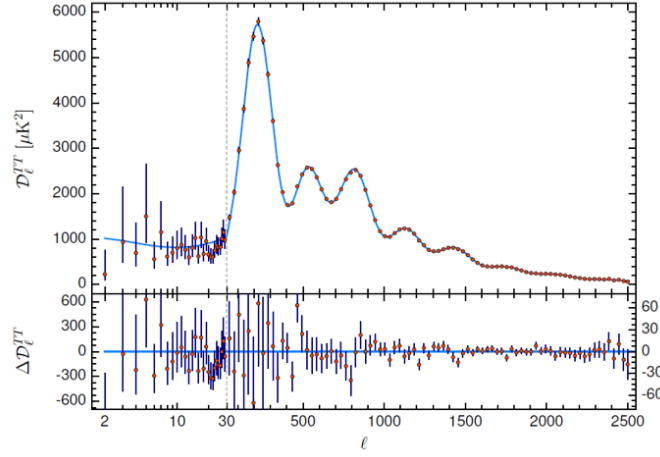


Figure 1.6: Power spectrum of the Cosmic Microwave Background in function of ℓ from [9]. Solid line shows the theoretical expectation.

are suppressed, and most of the correlations are concentrated around $\ell \simeq 200$.

The angular peaks observed in the power spectrum, within the multipole range $\ell \sim 100 - 1000$, are caused by *acoustic oscillations*. Before recombination, the baryonic matter density was subject to fluctuations driven by acoustic waves in the primordial plasma. The radiation pressure of the photons counteracted gravity, generating oscillations in the baryon-photon fluid.

As the plasma cooled and became neutral, these oscillations stopped and the resulting density variations froze. These oscillations amplified temperature fluctuations, and the modes that were frozen at the extrema of the oscillations now appear as the peaks observed in the CMB power spectrum.

The first peak, or fundamental mode, corresponds to regions that had reached maximum compression at the time of recombination, and its angular scale provides a measurement of the angular diameter distance to the surface of last scattering. The second peak represents regions that underwent both compression and rarefaction, reaching a minimal density. The third peak arises from regions oscillating three times as fast as those of the fundamental mode, which had sufficient time to undergo a second maximum compression by the time of CMB decoupling. These and the higher-order modes combine to form an harmonic series. The shape of the acoustic peaks is sensitive to key cosmological parameters, such as the densities of dark and baryonic matter.

At high multipoles, corresponding to small angular scales, the CMB power spectrum exhibits a decrease in temperature fluctuations. This phenomenon, known as *damping*, is caused by photon diffusion during recombination, as the distances traveled

by the photons exceed the wavelengths of the oscillations.

1.4 CMB polarization

In addition to temperature anisotropies, observations also reveal the angular variations in the *polarization* power spectrum of the CMB radiation. This property was first detected in 2002 by the *Degree Angular Scale Interferometer* (DASI) [13].

Thomson scattering of photons by free electrons at the last scattering surface initially generated the polarization of the CMB. Only quadrupole anisotropies, that is intensity variations of the unpolarized radiation along two perpendicular directions, forming a quadrupolar pattern, can produce a net linear polarization, as illustrated in Figure 1.7. This occurs because the differential intensity along these directions induces a preferred orientation in the scattered radiation.

During the early epochs, frequent Thomson scattering continuously randomized the photon directions, preventing the development of quadrupole anisotropies. Only near the last scattering surface the photon mean free path became sufficiently long to allow quadrupole temperature variations to grow, giving rise to the observed CMB polarization. Thus, the linear polarization pattern observed today reflects the temperature distribution at the time of recombination. However, since Thomson

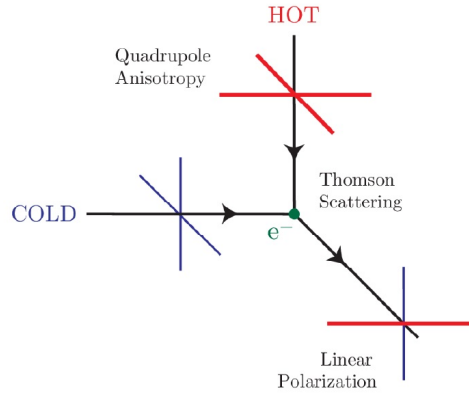


Figure 1.7: Quadrupole anisotropies in the radiation generate a net linear polarization in the scattered photons [3]. Different intensities along two perpendicular directions induce a preferred orientation of the electric field in the outgoing light.

scattering generates polarization only in the presence of quadrupole anisotropies in the radiation field, which were relatively small at that epoch, only about ten percent of the CMB radiation is polarized.

In general, there are three sources of quadrupole anisotropies ($\ell=2$) that can impact

on the polarization, with $m=0,\pm 1,\pm 2$ [7]:

1. $m=0$: scalars, coming from acoustic density perturbations, i.e. the compression and rarefaction.
2. $m=\pm 1$: vectors, coming from the Doppler shift associated with velocity. This quantity is negligible at the recombination.
3. $m=\pm 2$: tensors, related to gravitational waves.

The impact on the CMB polarization of 1. and 3. is described by specific coefficient that will be introduced in the next sections.

In general, the description of an electromagnetic wave polarization is made by using the *Stokes parameters*. The electric field for a wave propagating in the z direction can be written in the form [14]:

$$\vec{E}(\vec{x}, t) = \vec{E}_0 e^{i(\omega t - \vec{k}\vec{x})} \quad (1.27)$$

with

$$\vec{E}_0 \equiv \begin{pmatrix} a_1 e^{i\theta_1} \\ a_2 e^{i\theta_2} \\ 0 \end{pmatrix} \quad (1.28)$$

where a_1 and a_2 represents the amplitude and θ_1 and θ_2 the phase of each component. The Stokes parameters are a combination of these coefficients.

$$I \equiv \langle a_1^2 \rangle + \langle a_2^2 \rangle \quad (1.29)$$

$$Q \equiv \langle a_1^2 \rangle - \langle a_2^2 \rangle \quad (1.30)$$

$$U \equiv 2 \langle a_1 a_2 \cos(\theta_1 - \theta_2) \rangle \quad (1.31)$$

$$V \equiv 2 \langle a_1 a_2 \sin(\theta_1 - \theta_2) \rangle \quad (1.32)$$

where I is the total intensity, Q and U describe the linear polarization and V the circular polarization.

1.4.1 E -modes and B -modes

The Stokes parameters depend on the coordinates system in the sky. To obtain a coordinate-independent description, Q and U are combined into pseudo-scalar quantities, called E -modes and B -modes. The polarization is decomposed in a curl-free component (E) and a divergence-free component (B) (see Figure 1.8). The names refer to the same properties related to the electric (E) and magnetic (B) fields.

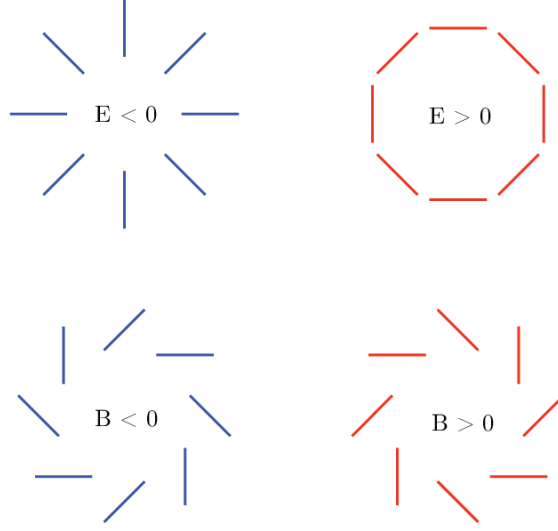


Figure 1.8: E-modes and B-modes pattern of polarization. E-modes are tangential around hot spots and radial around cold spots. B-modes change the orientation for hot and cold spots.

The rotational invariant quantity $(Q \pm iU)(\hat{n})$ is expanded in spin-2 spherical harmonics.

$$(Q \pm iU)(\hat{n}) = \sum_{\ell m} a_{\pm 2, \ell m} \pm 2 Y_{\ell m}(\hat{n}) \quad (1.33)$$

Defining the quantities

$$a_{\ell m}^E = -\frac{1}{2}(a_{2, \ell m} + a_{-2, \ell m}) \quad (1.34)$$

$$a_{\ell m}^B = -\frac{1}{2}(a_{2, \ell m} - a_{-2, \ell m}) \quad (1.35)$$

eq. 1.33 can be rewritten as:

$$(Q \pm iU)(\hat{n}) = \sum_{\ell m} (a_{\ell m}^E \pm i a_{\ell m}^B) \pm 2 Y_{\ell m}(\hat{n}) \quad (1.36)$$

The E - and B -modes coefficients $a_{\ell m}^E$ and $a_{\ell m}^B$ are used to compute the corresponding power spectra, exactly as in eq. 1.25.

$$C_\ell^{EE} = \frac{1}{2\ell + 1} \sum_{m=-\ell}^{\ell} |a_{\ell m}^E|^2 \quad (1.37)$$

$$C_\ell^{BB} = \frac{1}{2\ell + 1} \sum_{m=-\ell}^{\ell} |a_{\ell m}^B|^2 \quad (1.38)$$

These coefficients quantify the polarization in terms of multipoles and angular scale. In addition to these spectra, cross-correlation between temperature and polarization can be considered. C_ℓ^{TE} provides informations about the correlation between temperature and the E-modes polarization. In the standard cosmological model, parity

symmetry ensures that E-modes are even, B-modes are odd, and as a result, the TB and EB cross-correlations are expected to be zero.

An illustration of the power spectrum of the CMB, showing all the non-zero components is shown in Figure 1.9. In these plots $D_\ell = C_\ell \ell(\ell + 1)/2$. The C_ℓ^{TT} , C_ℓ^{EE} and C_ℓ^{BB} theoretical lines are displayed as dashed lines, along with measurements from the main CMB experiments over the years [9].

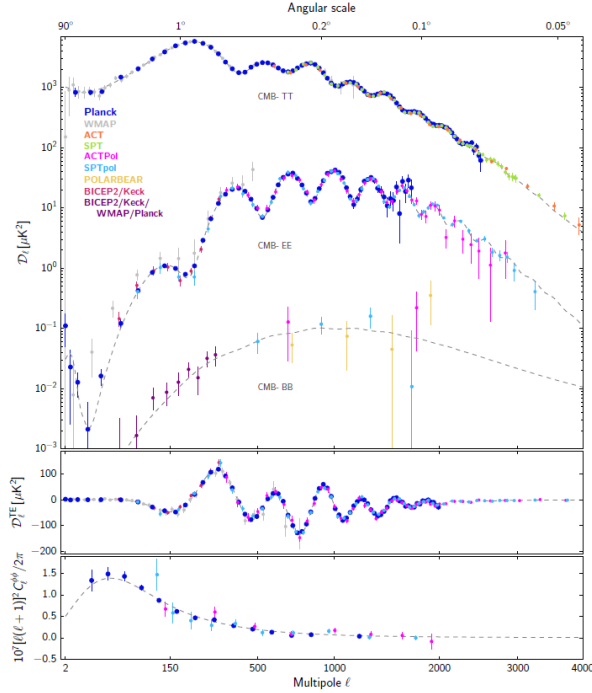


Figure 1.9: Power spectra for C_ℓ^{TT} , C_ℓ^{EE} and C_ℓ^{BB} , C_ℓ^{TE} theoretical lines and data [9].

Recalling the main sources of quadrupole anisotropies enumerated earlier, the E -modes are generated by density fluctuations, and exhibit highly symmetric pattern. In contrast, the B-modes arise exclusively by tensor components, and do not display a symmetric structure, or from gravitational lensing of the E -modes. In the latter case, large-scale structures deflect the photons along their path from the last scattering surface to the observer, converting part of the E -mode polarization into B -modes.

A representation of the polarization projected onto the spherical coordinates for both E - and B -modes is presented in Figure 1.10 [7].

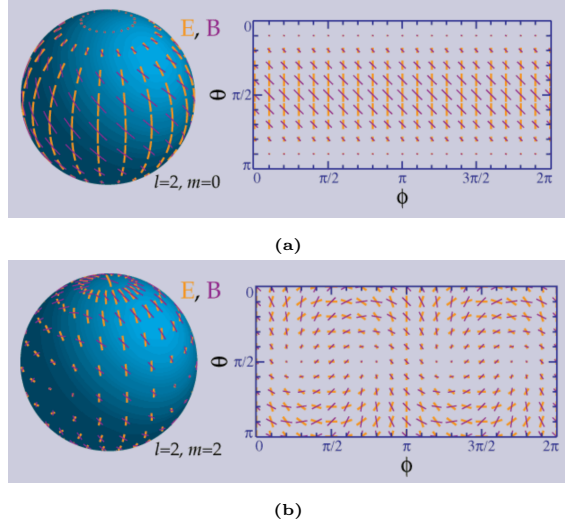


Figure 1.10: E - and B -modes pattern projected over spherical coordinates [7].

1.4.2 A Closer Look at B -modes

The investigation of B -mode polarization is a primary objective of current CMB experiments, since its detection would constitute a confirmation of the inflationary epoch in the early Universe.

During inflation, quantum fluctuations of the gravitational field generated perturbations in the space-time metric, known as *primordial gravitational waves*. These are tensor perturbations, described by small fluctuations $h_{\mu\nu}(t, \vec{x})$ in the metric tensor described by equation 1.2 [14]:

$$g_{00}(t, \vec{x}) = -1 \quad (1.39)$$

$$g_{0i}(t, \vec{x}) = 0 \quad (1.40)$$

$$g_{ij}(t, \vec{x}) = a^2(t)[\delta_{ij} + h_{ij}^{TT}(t, \vec{x})] \quad (1.41)$$

The term h_{ij}^{TT} represents the transverse and traceless components of the perturbation, which can be written in the form of matrix as

$$h_{ij}^{TT} = \begin{pmatrix} h_+ & h_\times & 0 \\ h_\times & -h_+ & 0 \\ 0 & 0 & 0 \end{pmatrix} \quad (1.42)$$

where h_+ and $h_\times$ are the two state of polarization of the gravitational wave. Einstein's equations give:

$$h_t'' + 2\frac{a'}{a}h_t' + k^2h_t = 0 \quad (1.43)$$

where $t = +, \times$ and the prime indicates the derivative over the conformal time η . This can be generally solved as a standard wave:

$$h_t(\vec{x}, \eta) = \int \frac{d^3k}{(2\pi)^3} e^{i\vec{k}\cdot\vec{x}} [A(\eta)e^{ik\eta} + B(\eta)e^{-ik\eta}] \quad (1.44)$$

This equation describes gravitational waves propagating in space-time. Figure 1.11 illustrates the distortion caused by gravitational waves.

These tensor perturbations leave a specific imprint on the polarization of the Cos-

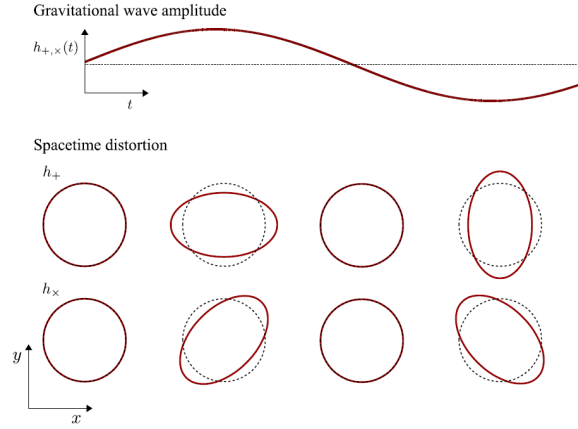


Figure 1.11: Illustration of the space-time distortion due to the gravitational waves representing the tensor perturbation of the metric ([14]).

mic Microwave Background. They induce divergence-free patterns. The detection of B-modes thus provides a direct observational signature of primordial gravitational waves and, consequently, of the inflationary epoch itself.

The *tensor-to-scalar ratio*, denoted by r , quantifies the relative amplitude of primordial tensor perturbations (gravitational waves) with respect to scalar perturbations (density fluctuations). This quantity provides a direct probe of the energy scale of inflation and the physics of the very early Universe.

The parameter r is defined by:

$$r = \frac{4P_h(k)}{P_{\mathcal{R}}(k)} \quad (1.45)$$

where k is a scale factor and $P_h(k, \eta)$ and $P_{\mathcal{R}}$ are respectively the spectra of tensor and scalar perturbations, which are described by the following equations.

$$P_h(k, \eta) = \frac{8\pi G H^2}{k^3} \Big|_{k|\eta|=1} \quad (1.46)$$

$$P_{\mathcal{R}} = \frac{8\pi G H^2}{9k^3 \epsilon_{sr}} \Big|_{aH=k} \quad (1.47)$$

where ϵ_{sr} is the slow-roll parameter. Since $\epsilon_{sr} = \frac{1}{16\pi G} \left(\frac{\partial_\phi V}{V} \right)$, the tensor-to-scalar ratio can be rewritten as:

$$r = 16\epsilon_{sr} \quad (1.48)$$

The relation between the energy scale of inflation V and the tensor-to-scalar ratio is given by:

$$V^{1/4} \sim \left(\frac{r}{0.01} \right)^{1/4} 10^{16} \text{ GeV} \quad (1.49)$$

An upper limit for r was set as $r < 0.044$ (95%CL) by combining the data from Planck and the BICEP-Keck Array [15]. Figure 1.12 shows the power spectra of the BB component in dependence of different values of r .

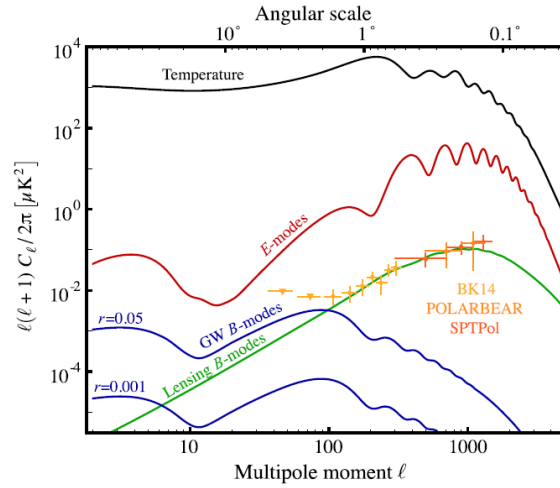


Figure 1.12: Power spectra showing the C_ℓ^{BB} coefficient as a function of r . The C_ℓ^{TT} and C_ℓ^{EE} are also shown. Data from BK14, POLARBEAR and SPTPol are shown [14].

1.5 Conclusions

In this chapter a summary of cosmology was presented, along with a more detailed description of the Cosmic Microwave Background characteristics. In particular, the CMB polarization was described, with particular emphasis on the B -modes and the important implications that their observation and discovery would have for fundamental physics.

Chapter 2

The LiteBIRD experiment

LiteBIRD (the Lite (Light) satellite for the study of B-mode polarization and Inflation from cosmic background Radiation Detection) was proposed as a JAXA (Japan Aerospace Exploration Agency) large class (L-class) mission in 2015 and was selected in 2019, after passing an initial downselection and a two year long pre-phase-A2 of *concept development*. LiteBIRD can be considered to be the successor of the Planck mission [9], already mentioned in Chapter 1. It will be launched by the next generation Jaxa H3 rocket and will be a 3-year all sky survey, operated in the Sun-Earth Lagrangian point L2.

LiteBIRD will observe the sky in 15 frequency bands, in the range 40, 402 GHz¹ in order to separate the CMB polarized signal from the more intense signal coming from other polarized sources in the Galaxy (thermal dust emission, synchrotron radiation...), with effective polarization sensitivity of 2 $\mu\text{K}/\text{arcmin}$ and angular resolution of 31 arcmin (at 140 GHz) [16].

One of the key features of the satellite will be the high sensitivity to low multipoles and large angular scale, in particular below $\ell \simeq 10$. This is possible thanks to the rapid polarization modulation (given by the presence of an half-wave plate, HWP), the position in the second Lagrangian point, which offers a large field of view and the screening of the Sun light, and the minimization of the systematic errors. LiteBIRD will be able to observe both the recombination and reionization bumps in the power spectrum, as visible in Figure 2.1, where the current available data are shown, together with a simulation of the sensitivity of LiteBIRD.

¹At the time of writing, LiteBIRD is under rescope studies to consolidate the mission's feasibility with the same scientific objectives. The rescope intends to simplify the mission, reduce the cost and improve manageability.

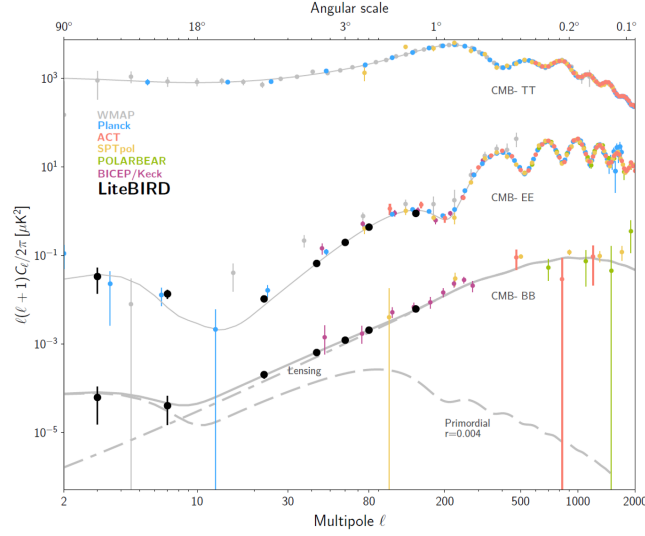


Figure 2.1: Simulation of the LiteBIRD sensitivity (black dots). The TT, EE and BB power spectra are reported. The solid lines are the best-fit for the Λ CDM model with a scalar-invariant tensor perturbation equal to $r=0.004$ ([16]).

2.1 The scientific goal

The detection of the polarization B-modes of the Cosmic Microwave Background, as mentioned in Chapter 1, would be a major breakthrough in understanding the early stages of the Universe's life. LiteBIRD is designed to detect B-mode signals with a sensitivity matching the amplitude predicted by current theoretical models. Moreover, the observation of B-modes, and thus of primordial gravitational waves, would provide evidence for the energy scale at which inflation occurred, which is expected to be close to the grand unification scale. Consequently, it would have a great impact on fundamental physics theories.

From cosmology, and in particular from inflationary models, we learn that the level of primordial gravitational waves, represented as tensor perturbations, is quantified by the tensor-to-scalar ratio, as discussed in eq. 1.48 from Chapter 1.

Different inflationary theories predict different values for r . Slow-roll inflationary models, coupled with the Friedmann equation and the amplitude of the scalar perturbations, define the energy scale of the inflation in terms of r as described in Eq. 2.1.

$$V^{1/2} = 1.04 \times 10^{16} \text{ GeV} \left(\frac{r}{0.01} \right)^{1/4} \quad (2.1)$$

As discussed in [16] and mentioned earlier in this chapter, the detection of the B-mode signal would provide a measurement of the tensor-to-scalar ratio, which in turn corresponds to an energy scale that could support grand unification theories.

The primary mission requirements of LiteBIRD² are:

- for $r = 0$ a total uncertainty of $\delta r < 1 \cdot 10^{-3}$, which should include the total instrumental noise fluctuations and the systematic errors uncertainty, residual foregrounds, lensing B-modes and observer bias.
- for $r = 0.01$ construct full sky linear CMB polarization maps in the recombination peak and reionization bump area with a 5σ detection.

2.1.1 Other science outcomes

LiteBIRD aims at other scientific targets beyond B-modes detection.

- **Epoch of re-ionization:** the scattering of the CMB photons at re-ionization causes an additional linear polarization. The quantity involved in this effect is the *optical depth* τ of the free electron column along the line of sight (see Figure 2.2a). The effects due to reionization can be observed in the E-modes power spectrum in the range $2 < \ell < 200$. The amplitude of this bump can be used to estimate τ and to compare this value with the number of free electrons expected to come from stars and galaxies formation. See Figure 2.2b from [16], adapted from [17].
- **Neutrino masses:** the optical depth studies have impact on the contribution of LiteBIRD in the estimation or constraint to the quantities related to the masses of light relics, in particular the sum of neutrino masses $\sum m_\nu$. Massive neutrinos slow down the formation of structures and we can get an evaluation of the masses from the comparison between the amplitude of the fluctuations at a low-redshift and at the last-scattering surface, described by the cosmological parameter A_s . This quantity can be estimated through the knowledge of τ . Setting the goal described in the previous bullet, LiteBIRD would increase the precision in the neutrino mass values compared to the *Planck* results. This comparison is shown in Figure 2.3a, adapted from [18].
- **Constraints on cosmic birefringence:** assuming that dark matter and dark energy behave as pseudo-scalar fields ϕ , the space would be considered like a birefringent material. The polarization of the CMB is sensitive to the rotation angle β of the plane of linear polarization. This would induce C_ℓ^{TB} and C_ℓ^{EB} components different from 0 on the power spectra and an additional

²See footnote (1)

B-modes signal, as can be seen in Figure 2.3b. The step forward produced by LiteBIRD would involve the reduction of the error bar on a global angle β and produce a map of β to be tested for cosmic birefringence.

- **Sunyaev-Zel’dovich effect:** the hot electrons do inverse Compton scattering on the CMB photons, distorting the black body spectrum. The LiteBIRD sensitivity and the frequency coverage can improve mapping of hot electrons in the Universe for $\ell > 20$ (see Figure 2.4a from [19]).
- **Anisotropic CMB Spectral Distortions:** effects causing the spatial distortion of the CMB power spectrum, other than the SZ effect, will be detected thanks to the LiteBIRD sensitivity. Among the contributor to this effect we can mention: the Rayleigh scattering of the CMB photons to the neutral hydrogen atoms formed at recombination, the “ μ distortion”, induced by non-Gaussian fluctuations during inflation and axion decay. This refers to the conversion of axions into photons caused by the Galactic magnetic field, which can produce polarized spectral distortions.
- **Primordial magnetic field:** PMFs were supposedly induced during cosmic inflation and can affect CMB in different ways: gravitational effects with magnetically induced perturbations, influence on the ionization of atoms during the history of the Universe, Faraday rotation of the CMB polarization, non-Gaussianity over the CMB anisotropies. LiteBIRD will have the sensitivity to detect all this effects, thanks to the measurements of the CMB polarization. Figure 2.4b from [20] shows the simulated impact of the PMFs on the BB power spectrum.

In addition to the secondary scientific case studies, LiteBIRD will clarify some anomalies found in the data of *WMAP* and *Planck* which cause tension with the Λ CDM model. Some Galactic science will be performed as well, involving interstellar turbulence, dust composition, anomalous microwave emission (AME) and so on.

2.2 The instrument

LiteBIRD will be a satellite working in the second Lagrangian point L2. It will observe the sky in 15 frequency bands from 40 GHz to 402 GHz, primarily to perform

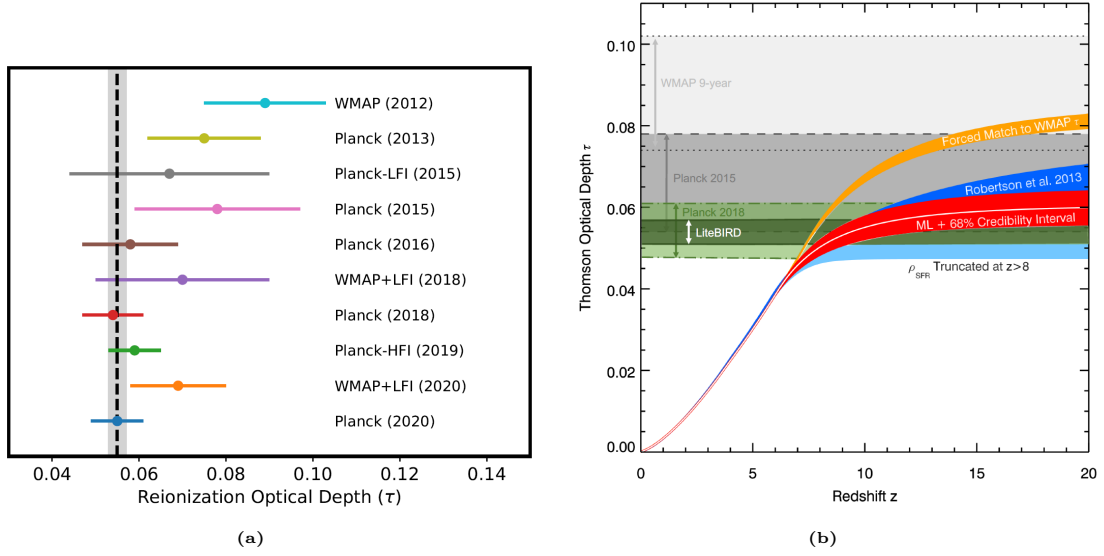


Figure 2.2: (a) Optical depth values from CMB power spectra from previous experiments. The grey colour indicates the minimum uncertainty around the fiducial value of $\tau = 0.054$ of LiteBIRD [16]. (b) Prediction for τ from various models. The green part indicates the range of sensitivity of LiteBIRD with a 65% and a 95% of confidence level around $\tau = 0.054$ ([16], adapted from [17]).

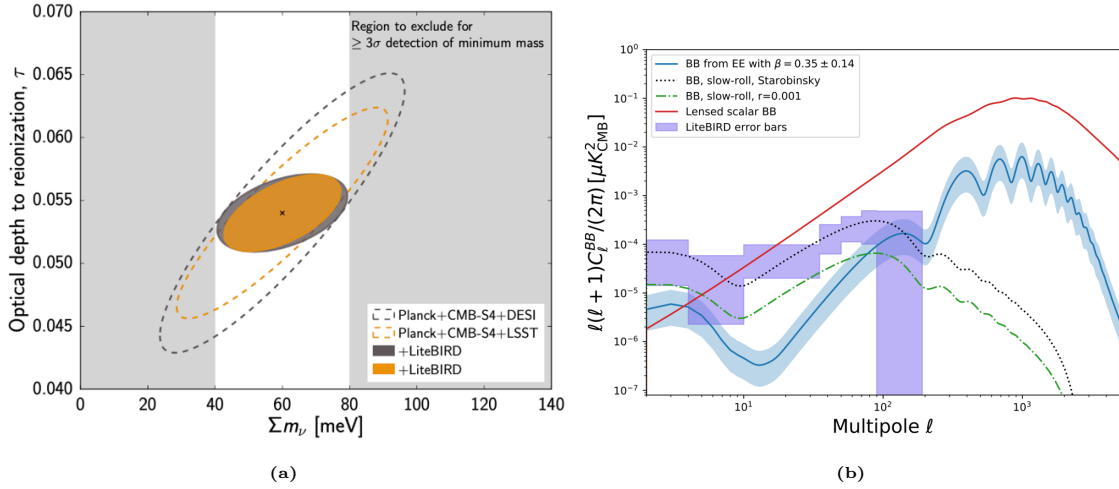


Figure 2.3: (a): Constrained area for τ and the sum of the neutrino masses considering CMB experiments and large class data, adapted from [18]. (b): The BB power spectrum from cosmic birefringence (blue solid line) for $\beta = 0.35 \pm 0.14$, estimated from the Planck data. The purple area relates to the expected constraints of LiteBIRD [16].

component separation and, consequently, foreground cleaning. This is a fundamental step to disentangle the CMB polarized signal from the “polarized foregrounds” coming from the galaxy. The two main foregrounds are **synchrotron radiation** and **thermal dust emission** which dominate the spectrum respectively below and above 80 GHz. Figures 2.5 and 2.6 illustrate these contributions, as well as the expected “layers” of emission that can obscure the observation of B -modes, making

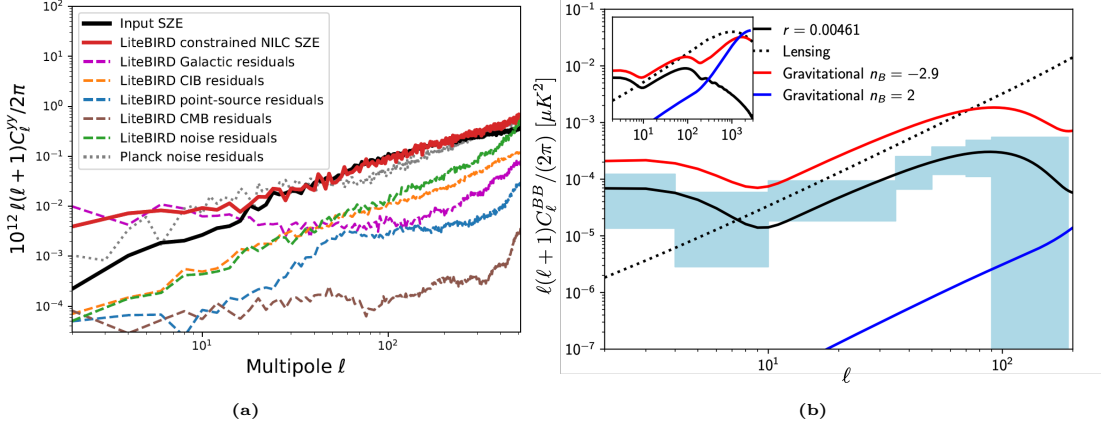


Figure 2.4: (a): Reconstructed power spectrum of the thermal SZ effect obtained with LiteBIRD simulations. The green dashed line indicates the noise power spectrum of LiteBIRD, to be compared to the grey one, related to *Planck* ([19]). (b): B-mode power spectrum from the gravitational effect from a PMF generated during inflation. The blue shaded area refers to the LiteBIRD constraints.

their detection more challenging.

Synchrotron and dust radiation can be described by the following equations.

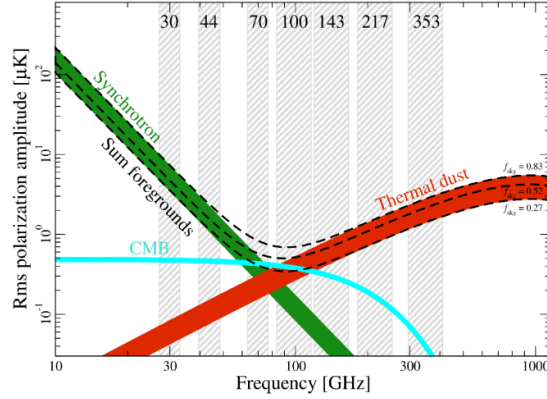


Figure 2.5: Spectrum showing synchrotron and dust signals over the CMB in polarization ([9]). The vertical grey bands correspond to the frequency channels of *Planck*.

$$[Q_s, U_s](\hat{n}, \nu) = [Q_s, U_s](\hat{n}, \nu_*) \left(\frac{\nu}{\nu_*} \right)^{\beta_s(\hat{n})} \quad \text{synchrotron radiation} \quad (2.2)$$

$$[Q_d, U_d](\hat{n}, \nu) = [Q_d, U_d](\hat{n}, \nu_*) \left(\frac{\nu}{\nu_*} \right)^{\beta_d(\hat{n})-2} \frac{B[\nu, T_d(\hat{n})]}{B[\nu_*, T_d(\hat{n})]} \quad \text{thermal dust radiation} \quad (2.3)$$

where $[Q_s, U_s](\hat{n}, \nu)$ and $[Q_d, U_d](\hat{n}, \nu)$ are respectively the Stokes parameters of synchrotron and thermal dust, on a given line of sight $\hat{n}$ at a frequency ν . The quantity ν_* is a reference frequency and $\beta_s(\hat{n})$ and $\beta_d(\hat{n})$ are the synchrotron and dust spectral indices. $T_d(\hat{n})$ is the effective temperature of the dust, $B[\nu_*, T_d(\hat{n})]$ is

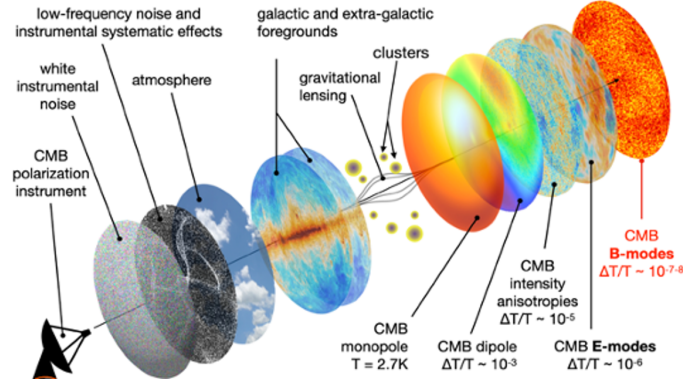


Figure 2.6: Representation of the expected challenges in the CMB B-modes observation. Image credits to Josquin Errard.

a blackbody spectrum.

In Figure 2.7 the simulated C_ℓ^{BB} power spectrum assuming $r = 0.00461$ and $\tau = 0.0544$ ($D_\ell = \ell(\ell + 1)C_\ell$) is shown, alongside the foreground cleaning residuals and the LiteBIRD error-bars.

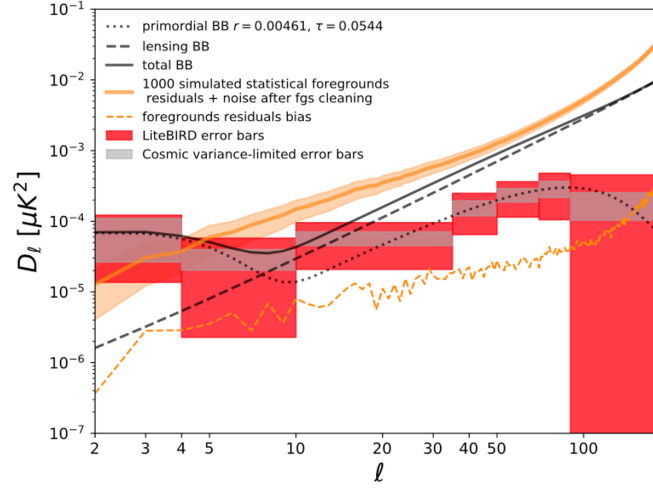


Figure 2.7: D_ℓ spectrum with $r = 0.00461$ and $\tau = 0.0544$. CMB primordial and gravitational lensing generated BB-modes (black solid, dot and dashed lines), the LiteBIRD error-bars (cosmic variance in grey and total error-bars in red), the average total residuals (including statistical and systematic foreground residuals, as well as noise after component separation (solid orange line) are displayed.

We note that at the time of writing LiteBIRD is undergoing rescope studies to consolidate the mission's feasibility with the same scientific objectives.

The optical part of the LiteBIRD instrument needs to be shaped in order to achieve the very high levels of sensitivity to reach the scientific goals. This is why a continuously rotating **half-wave plate** will be the first optical element [16], with the role of separating the polarized instrumental signals from the CMB radiation by shifting

the latter and modulating it at $4f_{HWP}$. This is also useful to diminish the impact of $1/f$ noise.

The first instrumental configuration, as shown in Table 2.8 from [16], published in 2022, was based on the mechanical construction organized in three telescopes: the *low (LFT)*, *medium (MFT)* and *high (HFT) frequency telescopes*, observing the sky in different frequency ranges.

Each telescope hosts a focal plane, with large FOVs, populated by multichroic pixels: every pixel will host multiple bolometers observing the sky in more than one frequency channels (see section 2.3.2 later in this chapter), with each detector viewing a single *antenna*. The focal planes are further divided in *focal plane sub-assemblies*,

Telescope	Detector type	Module	Frequency [GHz]	Pixel size [mm]	Module count	Pixel count	Detector count
LFT	Lenslet/Sinusoidal	LF12	40/60/78	32	4	24	144
			50/68/89	32		12	72
		LF34	68/89/119	16	4	72	432
			78/100/140	16		72	432
				Total LFT	8	180	1080
MFT	Lenslet/Sinusoidal	MF1	100/140/195	12	3	183	1098
		MF2	119/166	12	4	244	976
				Total MFT	7	607	2074
HFT	Horn/OMT	HF1	195/280	7	1	127	508
		HF2	235/337	7	1	127	508
		HF3	402	6.1	1	169	338
				Total HFT	3	423	1354
				Total mission	18	1210	4508

Figure 2.8: Configuration of the focal planes of the 3 instruments of LiteBIRD.

denoted as FPSA, hosting subsets of pixels. The modules are shown in Table 2.8. LFT will be divided in LF12 and LF34, MFT in MF1 and MF2, HFT in HF1, HF2, and HF3. This is important because, starting from this layout, the multiplexing and the SQUIDs assignment is planned. This will be presented in Chapter 4 in more detail.

The most important part of the instrument, the **detection chain**, is based on the use of superconducting Transition Edge Sensors (TESs) coupled to readout amplifiers called Superconducting QUantum Interference Devices (SQUIDs).

2.3 The Transition Edge Sensors

Transition Edge Sensors (TESs) are radiation detectors that exploit the steep relation between resistance and temperature of superconducting thin films [21]. They

are usually coupled to a bath reservoir with a temperature T_b through a weak link which guarantees a thermal conductance G . This also defines the characteristic time τ_0 over which the power deposited on the TES is dissipated through the thermal bath, with the relation $\tau_0 = C/G$, where C is the thermal capacity of the sensor. The sky signal, described in Figure 2.9 by the quantity P_{opt} , hits the TES and

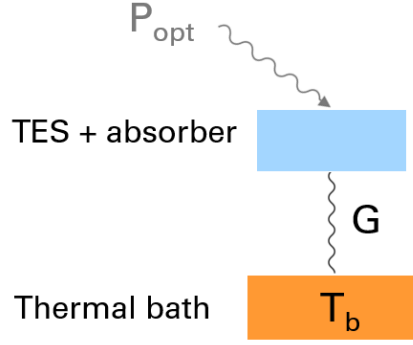


Figure 2.9: A scheme of a TES and the link to the thermal bath. P_{opt} is the optical power deposited on the TES, generally coupled to an absorber.

the coupled absorber, and warms it up. The power deposited is partly dissipated through the reservoir, partly warming the detector:

$$P_{opt} = C \frac{dT}{dt} + G(T - T_b) \quad (2.4)$$

The increasing temperature leads to an increase in the TES resistance when TESs are operated in the superconducting transition, and this variation can be measured as an electrical signal. Superconductors behave as normal metals above a critical temperature T_c , which is determined by the properties of the material (or materials, in case of a multi-layer TES). Below this temperature, the resistance of the material rapidly vanishes. The transition is very steep: small variations in temperature correspond to large changes in the resistance. This is described by the following relation.

$$\alpha \equiv \frac{d \ln R}{d \ln T} = \frac{T}{R} \frac{dR}{dT} \quad (2.5)$$

where α is called *logarithmic temperature sensitivity* and highlights the sensitivity of the resistance with temperature.

Operating the TES in a current bias configuration can lead to a thermal runaway: the increase of the temperature cause an increasing in the resistance and drives the TES outside its superconducting transition to the normal stage. The additional power $P = I^2 R$ is dissipated on the TES, an increase in the current corresponds to a grow of the dissipated power and an increase of the resistance. Based on this,

a voltage bias is preferred. The TES is connected in parallel to a *shunt resistor* R_{shunt} , as shown in Figure 2.10. In this way, the **Negative Electro-thermal**

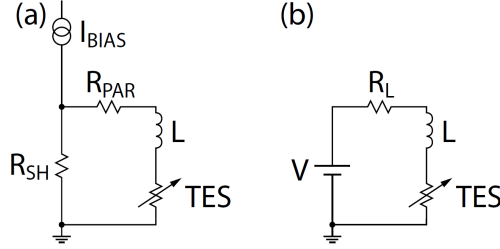


Figure 2.10: (a) The TES in a shunt circuit to perform voltage bias and reach thermal stability. The bias current I_{bias} is applied to the R_{shunt} in parallel with parasitic resistance R_{par} and a stray inductance L . (b) The equivalent Thevenin circuit where $R_L = R_{shunt} + R_{par}$ is shown. A bias voltage $V = I_{bias}R_{shunt}$ is applied ([21]).

Feedback (NEFT) is achieved. As the temperature and the resistance increase, the dissipated power at a fixed bias voltage, $P_J = V^2/R$, decreases and the system reaches equilibrium. Eq. 2.4 changes, since the additional term P_J is included in the left-hand side. The behavior of the TES in the voltage bias configuration showed in Figure 2.10-(b), is described by two differential equations, reported in Eq. 2.6 and 2.7, describing the thermal and electrical responses.

$$C \frac{dT}{dt} = -P_{bath} + P_J + P_{opt} \quad \text{thermal} \quad (2.6)$$

$$L \frac{dI}{dt} = V - IR_L - IR(T, I) \quad \text{electrical} \quad (2.7)$$

where P_{bath} is derived from $dP_{bath}/dT = G$ and is the power flowing from the film to the bath and P_J is the Joule power dissipation, I is the electrical current through the TES and $R(T, I)$ is the resistance of the TES [21]. Since:

$$dP_J = \frac{dP}{dR} \frac{dR}{dT} dT = -\frac{P_J \alpha}{T} dT \quad (2.8)$$

eq. 2.6 can be expanded around an equilibrium temperature T_e :

$$P + P_J - P_J \frac{\alpha}{T_e} (T - T_e) = C \frac{d(T - T_e)}{dt} + G(T - T_e - T_b) \quad (2.9)$$

The conditions of equilibrium are [21]:

$$P_{opt} + P_J = -GT_b \quad (2.10)$$

$$C \frac{d(T - T_e)}{dt} = -G \left(1 + \frac{P_J \alpha}{T_e G} \right) (T - T_e) \quad (2.11)$$

This expression differs from the case without ETF by the replacement of G with $G \left(1 + \frac{P_J \alpha}{T_e G} \right)$. Accordingly, the time constant τ_0 previously defined, is replaced by

an *effective* value τ_{eff} . A fundamental quantity to be defined to describe the TES behavior in the negative ETF regime is the **loop gain** [21].

$$\mathcal{L}_I \equiv \frac{P_{J_0} \alpha_I}{GT_0} \quad (2.12)$$

It depends on intrinsic quantities of the device. Given this definition, if $\mathcal{L}_I \gg 1$ the response of the TES is very fast and efficient. The time constant depends on $\mathcal{L}_I$ and is equal to $\tau_{eff} = \tau_0/(1 + \mathcal{L}_I)$.

Equations 2.6 and 2.7 can be solved by assuming to have small variations around the steady state values R_0, I_0, T_0 . For P_{bath} we assume a power law dependency [21]:

$$P_{bath} = K(T^n - T_{bath}^n) \quad (2.13)$$

where $K = G/n(T^{n-1})$ and $n = \beta + 1$, with β defined as a thermal conductance exponent. Being $G = dP_{bath}/dT$, eq. 2.13 can be rewritten for small signals around T_0 :

$$P_{bath} \approx P_{bath,0} + G\delta T \quad (2.14)$$

where $G = nKT^{n-1}$ and $\delta T \equiv T - T_0$. K and n are function of the link to the thermal bath. In the steady-state case, with $P_{bath,0} = P_{J_0} + P_0$ and a constant $P_{opt} = P_0$ the result for the temperature is:

$$T \simeq T_0 + \frac{P_0}{G} \quad (2.15)$$

The resistance of the TES is determined by expanding the solutions around R_0, I_0 and T_0 up to the first order.

$$R(T, I) \approx R_0 + \left. \frac{\partial R}{\partial T} \right|_{I_0} \delta T + \left. \frac{\partial R}{\partial I} \right|_{T_0} \delta I \quad (2.16)$$

where $\delta I \equiv I - I_0$. Substituting the coefficients:

$$\alpha_i \equiv \left. \frac{\partial \log R}{\partial \log T} \right|_{I_0} = \frac{T_0}{R_0} \left. \frac{\partial R}{\partial T} \right|_{I_0} \quad (2.17)$$

$$\beta_i \equiv \left. \frac{\partial \log R}{\partial \log I} \right|_{T_0} = \frac{I_0}{R_0} \left. \frac{\partial R}{\partial I} \right|_{T_0} \quad (2.18)$$

into Eq. 2.16, it becomes:

$$R(T, I) \approx R_0 + \alpha_i \frac{R_0}{T_0} \delta T + \beta_i \frac{R_0}{I_0} \delta I \quad (2.19)$$

By solving Eq. 2.6 and 2.7 in the steady-state assumption, we can study the output I as a function of the voltage bias V . The I-V curve describes the behavior of the

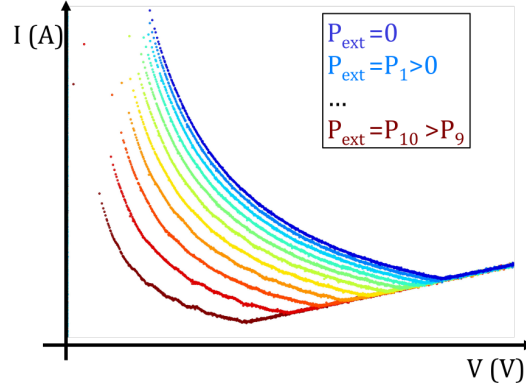


Figure 2.11: IV curve of a TES bolometer, figure from [22].

TESs, as shown in Figure 6.14. When a high bias voltage is applied to a TES in the normal regime and then gradually decreased, the current decreases linearly with the voltage ($I = V/R$), following the Ohm's law. When the current decreases below the critical point, the TES enters the transition from normal to superconducting. As $P_J = IV$ is kept constant, the bias voltage continues to decrease and the current increases until the TES resistance vanishes.

TESs can be operated in either a bolometric or calorimetric regime, depending on the the origin of the radiation to be detected.

- **bolometer:** the measured quantity is an external signal P_{opt} which varies more slowly than the response time of the TES. The sensor stabilizes around its working point. In the steady state the dissipated power is balanced by the power flowing to the thermal bath. The TESs are used in this configuration when measuring slowly modulated signal, coming for example from the CMB radiation
- **calorimeter:** the measured quantity is the energy coming from a transient signal, such as a single photon event. The P_{opt} contribution varies as $\sim E\delta T$ and the TES responds rapidly warming up and exponentially cooling down to its original equilibrium temperature. This is the configuration in which TESs are used for detecting high energy photons and to perform, for example, X-Ray Astronomy.

The TESs, in both configurations, need to be connected to a low noise amplifier. This is driven by the necessity of amplifying the very small current signals at the TESs output. To do so, these detectors are connected to superconducting quantum

interference devices (**SQUIDS**), which are cryogenic current amplifier with a very low impedance input [21]. SQUIDS also allow to implement the multiplexing configurations which are needed to decrease the heat load on the cryogenic stages, as will be shown later in this work.

2.3.1 The Noise Equivalent Power

For TESs employed in a bolometric configuration the quantity to measure is the power dissipated on the detector. The Signal-to-Noise ratio is estimated by the **Noise Equivalent Power (NEP)**. It is crucial to estimate the detector sensitivity and to characterize the noise sources. The Noise Equivalent Power is defined as the input power needed to have a S/N ratio of 1 on a 1 Hz bandwidth in output, which means the input power necessary for the signal to match the noise level of the detector.

The so-called *photon noise* is an unavoidable contribution arising from the variance of the photons arrival at the detector. A brief description follows.

Let's imagine to have a black-body source of photons. We can describe its brightness as a function of frequency and temperature as:

$$B(\nu, T) d\nu = \frac{2h\nu^3 d\nu}{c^2 [e^{h\nu/kT} - 1]} \quad [\text{W/m}^2 \text{ sr}] \quad (2.20)$$

where ν is the frequency and T is the temperature of the source. The number of photons measured by a detector with efficiency $\eta(\nu)$ per second per Herz is:

$$n = \frac{\eta}{\exp(h\nu/kT) - 1} = \left[\frac{\text{photons}}{\text{sHz}} \right] \quad (2.21)$$

The variance on the number of photons is:

$$\langle \Delta n^2 \rangle = n + n^2 \quad (2.22)$$

The energy in photons is $E_\nu = nh\nu$ and the variance in energy is calculated as $\langle E_\nu^2 \rangle = (h\nu)^2 \langle \Delta n^2 \rangle$. Combining this with eq. 2.22 the variance can be written as:

$$\langle E_\nu^2 \rangle = n(2h\nu)^2 + n^2(2h\nu)^2 \quad (2.23)$$

where the factor of 2 comes from the two photon polarizations.

Integrating over the frequency to get the power variance in the single bands, we can define the NEP related to the photon noise like:

$$(NEP_\gamma)^2 = \int 4n(h\nu)^2 d\nu + \int 4n^2(h\nu)^2 d\nu \quad [W^2/\text{Hz}] \quad (2.24)$$

Rewriting this equation in terms of photon power we get:

$$(NEP_\gamma)^2 = \int 2h\nu P_\nu d\nu + \int P_\nu^2 d\nu \quad (2.25)$$

If we assume that the power is uniform over the single frequency bands, we can define $P_{opt} = P_{\nu_0} \Delta\nu$ and

$$(NEP_\gamma)^2 \approx 2h\nu_0 P_{opt} + \frac{P_{opt}^2}{\Delta\nu} \quad (2.26)$$

where the first term is the *shot noise*, arising from the statistical uncertainty of the Poisson distributed photon arrivals, while the second term is the *bunching noise*, associated with the excess variance introduced by the correlated arrival of photons. CMB experiments appear to be dominated by the photon noise.

Other relevant noise sources are:

- the *internal thermal fluctuation* noise, generated by thermal fluctuation on the TES and absorber level, described by the power spectral density $S_{th} = 4k_B T^2 G$ ([23])
- the *Johnson-Nyquist* noise related to the resistive behaviour of the TES in the transition. To describe this contributor we can replace the noisy resistance with an equivalent circuit composed of a noiseless resistor plus a Johnson noise voltage generator with an RMS voltage fluctuation $V = \sqrt{4k_B T R \zeta(I)}$ [23], where $\zeta(I)$ describes the noise originating from the TES.
- the *phonon* noise, generated by the statistical energy fluctuations in the link between the TES-absorber pair and the heat bath. It is outlined by the power spectral density $S_{pn} = 4k_B T^2 G_{bath} F(T_0, T_{bath})$, where $F(T_0, T_{bath}) = ((T_{bath}/T)^{n+1} + 1)/2 \sim 0.5$ and n is the thermal-process dependent coefficient [23].
- the *readout* noise, generated by the elements composing the electronics: amplifiers, DACs, ADCs etc. It is identified through the noise current, $NEP_{readout} = V_{bias} \times NEI$. NEI stands for **Noise Equivalent Current** and describes the flux density necessary to match the noise level. The readout NEP can be minimized by reducing the current noise coming from the electronics or reducing the V_{bias} by decreasing the working point resistance of the TES.

The NEP must be converted into a temperature-related quantity, to be compared to the CMB signal, expressed in terms of temperature. The **NET (Noise Equivalent**

Temperature) is defined as:

$$NET = \frac{NEP}{\sqrt{2}dP/dT} \quad (2.27)$$

which expresses the relation between the temperature fluctuations and the power variations on the detector. The $\sqrt{2}$ term derives from the two polarizations. In the specific case of LiteBIRD [16]:

$$\frac{dP}{dT_{CMB}} = \int_{\nu_1}^{\nu_2} \left[k_B \eta(\nu) \left(\frac{h\nu}{k_B T_{CMB} (e^{h\nu/k_B T_{CMB}} - 1)} \right)^2 e^{h\nu/k_B T_{CMB}} \right] d\nu \quad (2.28)$$

where ν_1 and ν_2 represent the upper and lower edge of the frequency band. This expression is included in Eq. 2.27 to define the NET level.

2.3.2 The TESs for LiteBIRD

LiteBIRD will use TESs in a bolometric configuration, clustered into “pixels” and coupled to dual-polarization antennas.

The antennas are connected to the TESs with on-chip microstrips through bandpass filters. These elements split the signal from the antennas in different frequency channels to different TESs. The orientation of the antenna select the polarization state detected by each TES.

The orientation of the antenna is defined by its geometry, as it is illustrated in Figure 2.12. Each pixel will host at the same time a 0° and 90° or a 45° and -45° oriented antennas. This is necessary because it is impossible to reconstruct the

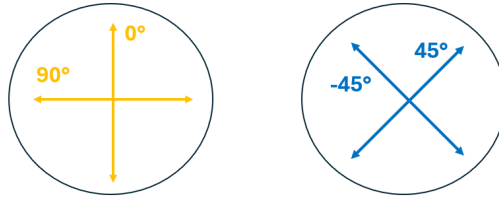


Figure 2.12: Sketch of antennas orientation for the Stokes parameter reconstruction. The black circle represents the pixel, while the arrows the antennas.

Stokes parameters Q and U (for polarization description) with a single direction. Let’s define $\theta(\nu)$ as the angle of the incident light polarization depending on the frequency. The power received by the bolometer connected to the single antenna

can be defined as [24]:

$$P_0 = \int \eta(\nu) [E(\nu) \cos \theta(\nu)]^2 d\nu \quad (2.29)$$

$$P_{90} = \int \eta(\nu) [E(\nu) \sin \theta(\nu)]^2 d\nu \quad (2.30)$$

$$P_{45} = \int \eta(\nu) \left[E(\nu) \cos \left(\frac{\pi}{4} - \theta(\nu) \right) \right]^2 d\nu \quad (2.31)$$

$$P_{-45} = \int \eta(\nu) \left[E(\nu) \sin \left(\frac{\pi}{4} - \theta(\nu) \right) \right]^2 d\nu \quad (2.32)$$

where $\eta(\nu)$ is the frequency-dependent efficiency of the detector and $E(\nu)$ is the amplitude of the electrical field in that direction as a function of the frequency. From eq. 2.29-2.32 the Stokes parameters are defined [24] combining measurable quantities.

$$Q = P_0 - P_{90} = \cos(2\theta) \int \eta(\nu) E^2(\nu) d\nu \quad (2.33)$$

$$U = P_{45} - P_{-45} = \sin(2\theta) \int \eta(\nu) E^2(\nu) d\nu \quad (2.34)$$

The TESs of LiteBIRD will be arranged as follows: on the same pixel, for each frequency channel, two detectors will be connected to two antennas oriented with a relative angle of 90 degrees, as shown in Figure 2.13. Moreover, each pixel hosts more than one frequency channel and is called **multichroic**. In detail, multiple traces originate from each antenna and are split through bandpass filters, which select the frequency channels before reaching single detectors, as shown in 2.14 from [25], where the LF12 subassembly is presented. Each green circle correspond to a pixel. It can be easily noticed that the antennas belonging to the same pixel are oriented with a 90 degrees relative angle, but some pixel are rotated by 45 degrees, to replicate the scheme shown in Figure 2.12.

At the time of writing, the TESs of LiteBIRD are of two types: those assembled in sinuous-antenna, lenslet-coupled pixels (LFT and MFT), and those assembled in OMT (orthomode transducer) horn-coupled pixels (LFT).

The TESs are made of a Al/Mn belayer film coupled to a Pd/Au thermal ballast with an expected critical temperature around 170 mK and normal resistance of 1 Ω [26].

2.3.3 The TESs for Athena

In this section, the structure and setup of an X-ray mission that employs TESs as microcalorimeters will be presented. Since we worked with the TESs of Athena to

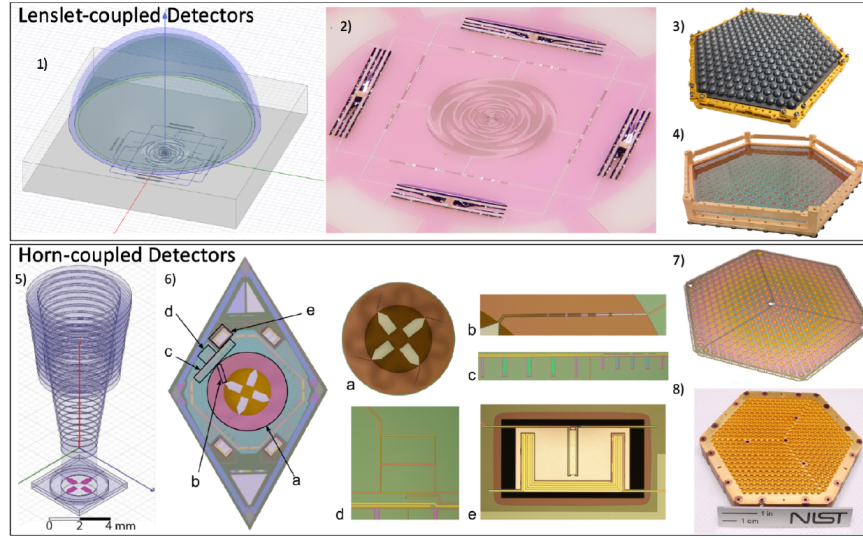


Figure 2.13: The LiteBIRD focal planes are populated by lenslet-coupled sinuous antennas pixels for LFT and MFT (1) and by horn-coupled orthomode transducer pixels for HFT (5). Details of a single pixel (2) and of an array of pixels (3, 4) hosting sinuous antennas are shown in the figure. Single elements of an OMT horn-coupled antenna, together with an array arrangement, are displayed in 6, 7, 8 ([16]).

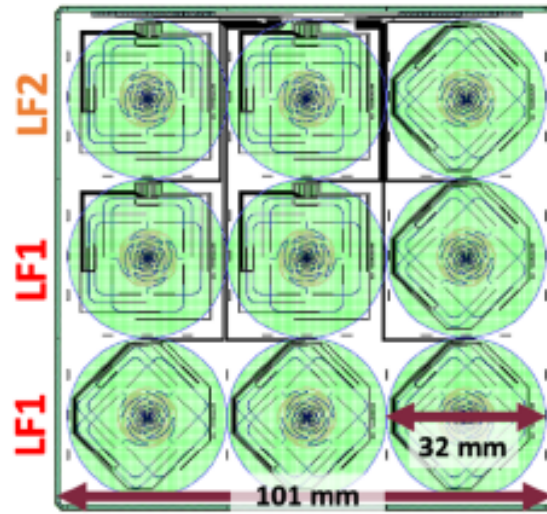


Figure 2.14: LF12 scheme, with highlights on the antennas, the microstrips and the antennas orientation [25].

test a prototype of cold readout (the work will be presented in the next chapters), we will briefly introduce the mission and the instrument to provide some context.

Athena is an ESA large class mission selected in 2014 [27]. From 2023 it was re-named **NewAthena** (New Advanced Telescope for High-ENergy Astrophysics) after undergoing a rescoping and it is now under mission designing and cost studying. The launch is planned for 2037 and the main scientific case to be explored will be X-ray astronomy. It will be the largest X-ray telescope ever built and will optimize the performance by increasing the quality of many technical aresas: collecting area,

speed, sensitivity and spectroscopy resolution. The main goal of the mission will be to observe strong X-ray sources, together with mapping the hot gases between galaxies and also observing transient events [27].

The payload will be composed of:

- an X-ray telescope with a 12 meters focal length
- an X-ray Integral Field Unit (**X-IFU**) for high-spectral resolution imaging
- a Wide Field Imager (**WFI**) for high count rate and a lower spectral resolution spectroscopy over a large field of view.

The X-IFU instrument will be equipped of ~ 1500 TES microcalorimeters in a cryogenic environment kept at a temperature of about 50 mK, ([28]). The microcalorimeters consist of an absorber, the TES itself used as thermometer and a thermal weak link to the bath ([23]).

The TESs of Athena are Mo/Au bilayers with a 90 mK transition temperature. The absorber is made of Bi/Au with a thickness of $\sim 6 \mu\text{m}$. They will be readout in a Time-Domain Multiplexing (see section 3.1) configuration [23]. Alternatively to the baseline cold readout, a backup Frequency Domain Readout (see section 3.1) has been studied. This configuration will be employed for the measurements reported in Chapter 5.

2.4 The SQUIDs

The TES detectors will be connected to SQUIDs (Superconducting QUantum Interference Devices) for a cryogenic first stage of amplification.

SQUIDs are composed of a superconducting loop interrupted by two Josephson junctions in parallel. The *Josephson effect* [29] arises when two superconductors are separated by a thin barrier made of an insulating layer (the Josephson junction): in this case the configuration is called S-I-S, for “superconductor-insulator-superconductor”. The current flow in the loop is due to the tunneling of the Cooper pairs and quasiparticles. The first Josephson equation is defined as:

$$I_s = I_c \sin \delta \quad (2.35)$$

where δ is the phase difference of the wave functions of the two superconducting halves and I_c is the critical current or maximum supercurrent. This quantity represents the maximum current that can flow through a Josephson junction without

generating a voltage, maintaining the superconducting state.

The second Josephson equation describes how the phase difference changes over time:

$$\dot{\delta} \equiv \frac{d\delta}{dt} = \frac{2e}{\hbar} U = \frac{2\pi}{\Phi_0} U \quad (2.36)$$

where $\Phi_0 = 2e/h = 0.207 \cdot 10^{-15}$ Wb. The magnetic flux $\Phi(B)$ is quantized in units of the flux quantum Φ_0 : $\Phi(B) = n\Phi_0$. U is the voltage applied across the junction when a phase difference is present. In a SQUID, two Josephson junctions are present (see

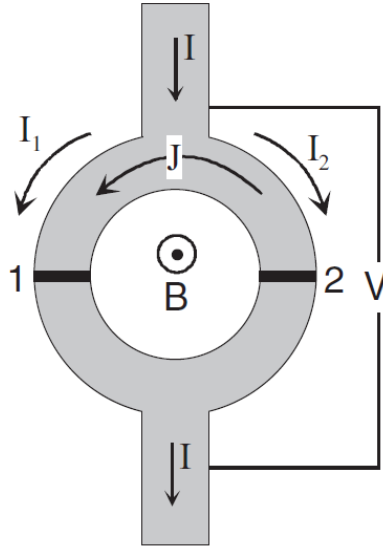


Figure 2.15: Schematic picture of a SQUID. The junctions (1 and 2) and the current flowing through each one of them (I_1 and I_2) are shown ([30]).

Figure 2.15): the current I is the sum of the currents I_1 and I_2 flowing through each junction. The phase difference induced by the two junctions (named here a and b) is described by the following equation.

$$\delta_a - \delta_b = \frac{2e}{\hbar} \left(\int_a \vec{A} \cdot d\vec{\ell} - \int_b \vec{A} \cdot d\vec{\ell} \right) = \frac{2e}{\hbar} \oint \vec{A} \cdot d\vec{\ell} = \frac{2e}{\hbar} \oint_S \vec{B} \cdot \hat{n} ds = \frac{2e}{\hbar} \Phi(B) \quad (2.37)$$

Based on this:

$$\delta_a - \delta_b = 2\pi \frac{\Phi(B)}{\Phi_0} \quad (2.38)$$

For the Josephson equation, the total current I can be written as follows.

$$I = I_{ac} \sin \delta_a + I_{bc} \sin \delta_b \quad (2.39)$$

where I_{ac} and I_{bc} are the critical currents of the two junctions. Assuming $I_{ac} = I_{bc} \equiv I_0$:

$$I = I_0 (\sin \delta_a + \sin \delta_b) = 2I_0 \sin \frac{\delta_a + \delta_b}{2} \cos \frac{\delta_a - \delta_b}{2} \quad (2.40)$$

We choose to define the phases to have:

$$\begin{cases} \delta_a = \delta_0 + \pi \frac{\Phi(B)}{\Phi_0} \\ \delta_b = \delta_0 - \pi \frac{\Phi(B)}{\Phi_0} \end{cases} \quad (2.41)$$

Eq. 2.40 becomes:

$$I = 2I_0 \sin \delta_0 \cos \left(\frac{\pi \Phi(B)}{\Phi_0} \right) \quad (2.42)$$

The SQUID behaves like a Josephson Junction with critical current modulated by the external flux $\Phi(B)$.

$$I_c = 2I_0 \left| \cos \left(\frac{\pi \Phi(B)}{\Phi_0} \right) \right| \quad (2.43)$$

The variation of the current as a function of the the magnetic flux is shown in Figure 2.16. The dependence of the current from the voltage is described by I-V

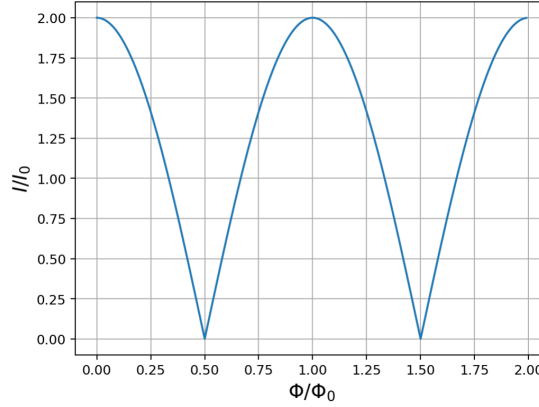


Figure 2.16: Dependence of the I/I_0 ratio to the quantity $\Phi(B)/\Phi_0$ [30].

curve. Using the RCSJ (resistively- and capacitively-shunted junction) model [31], the two junctions can be represented as shunted with a resistance-capacitor circuit and overdamped, as shown in Figure 2.17(a). The final relation is [32]:

$$V = R \sqrt{I^2 - 4I_0^2 \left(\cos \frac{\pi \Phi(B)}{\Phi_0} \right)^2} \quad (2.44)$$

From the I-V curve (Figure 2.17(b)) the variation of the voltage as a function of $\Phi(B)/\Phi_0$ (Figure 2.17(c)) with a fixed bias current I_b can be described. Based on this, the V- Φ curve, can be built. The performance of the SQUID is optimized when the chosen I_b corresponds to a V- Φ curve with the largest peak-to-peak amplitude. As previously mentioned, the SQUID is used in this configuration as an amplifier for TESs. The SQUID is paired to an input coil with inductance L_{in} connected in series to the detectors. The output current coming from the TESs is sent to the

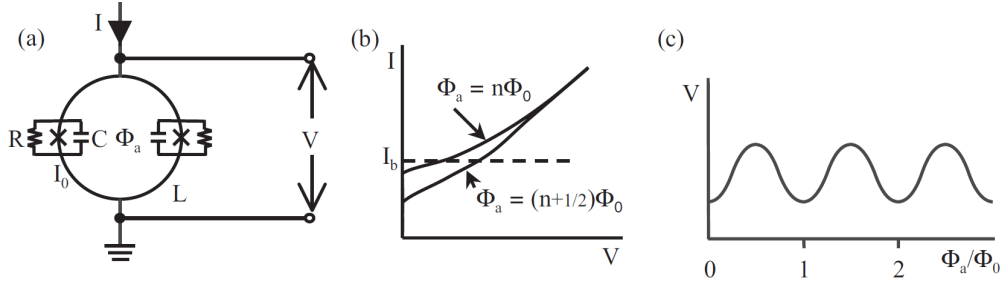


Figure 2.17: The circuitual scheme of the SQUID (a), I-V characteristics for integer and half-integer n coefficient of Φ_0 , $V-(\Phi(B)/\Phi_0)$ at current fixed (c).

input coil which induces a magnetic flux in the SQUID.

We define M_s and I_s respectively as the mutual inductance between the SQUID and the input coils and the current flowing through L_{in} . $\Phi_s = M_s I_s$ is the flux induced through the SQUID coil. The voltage for small flux variations $\delta\Phi$ is $\delta V = (dV/d\Phi) \cdot \delta\Phi$.

Consequently:

$$\delta\Phi_s = M_s \delta I_s \quad (2.45)$$

$$\delta V = (dV/d\Phi) M_s \delta I_s \quad (2.46)$$

More operative details on the SQUID will be enumerated in Chapter 6, when the measurements on a readout chain of LiteBIRD will be presented.

2.4.1 Flux-Locked Loop

The signal read and amplified with a SQUID is not-linear. To perform linearization some specific configurations must be implemented. One of these is the **Flux-Locked Loop (FLL)**. In this configuration, a feedback circuit is implemented as shown in Figure 2.18. It consists of a feedback resistance R_F and an additional coil coupled to the SQUID (M_f). The input flux Φ_s and the feedback flux Φ_F can be calculated:

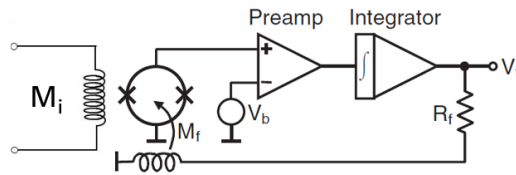


Figure 2.18: Feedback circuitual configuration to implement FLL scheme.

$$\Phi_s = M_{in} I_s \quad (2.47)$$

$$\Phi_F = M_F I_F = M_F \frac{V_{out}}{R_F} \quad (2.48)$$

To implement the FLL scheme, the signal to the SQUID is set to zero, $\Phi_F + \Phi_s = 0$. This relation is rewritten as follows:

$$I_s M_{in} = -M_F I_F = -M_F \frac{V_{out}}{R_F} \quad (2.49)$$

This quantity can be turned around to calculate V_{out} :

$$V_{out} = -R_F \cdot \frac{M_{in}}{M_F} I_s \quad (2.50)$$

Based on this, V_{out} is linearized with a factor of amplification equal to $R_F M_{in}/M_F$. This is the *stationary* description of the SQUID behavior. This description is valid if we consider an infinite gain and we assume this quantity to be constant with frequency. In this case, the feedback loop provides a linear relationship between the output voltage and the input current.

2.4.2 The SQUID gain-bandwidth product

Going into more detail, if we consider the intrinsic properties of a SQUID, displayed as an operational amplifier in series with an integrator, we need to take into consideration the **gain-bandwidth** figure of merit, which describes the frequency dependence of the gain. This quantity is constant, it means that if we increase the bandwidth, the gain is decreased and viceversa. In this case, we need to study the frequency dependency of the gain.

In Figure 2.19 a scheme of a generic amplification chain applied to the SQUID circuit is shown. The dependence of each quantity with the other is given by:

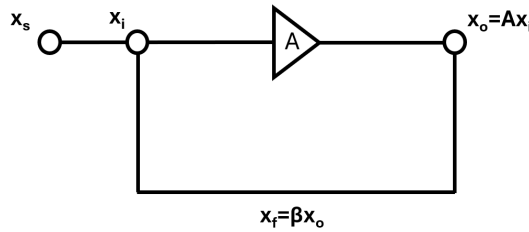


Figure 2.19: Scheme of the SQUID circuit. A is the open-loop SQUID gain, x_s the input signal and x_i the input to the SQUID. x_o is the output, amplified by the gain A and x_f is the feedback signal, scaled of a fraction of x_o , called β .

$$\begin{cases} x_i = x_s + x_f \\ x_i = x_s + \beta x_o = x_s + \beta A(\omega) x_i \rightarrow x_i(1 - \beta A(\omega)) = x_s \\ x_i = \frac{1}{1 - \beta A(\omega)} x_s \\ x_o = \frac{A(\omega)}{1 - \beta A(\omega)} x_s \end{cases} \quad (2.51)$$

where x_s is the input signal and x_i is the input to the SQUID, x_o is the output, A is the gain and x_f is the feedback signal, scaled of a fraction of x_o , called β . These parameters correspond to the physical quantities shown in Figure 2.20 we show the components of the signal. From 2.20 and eq. 2.52 the output voltage and the input

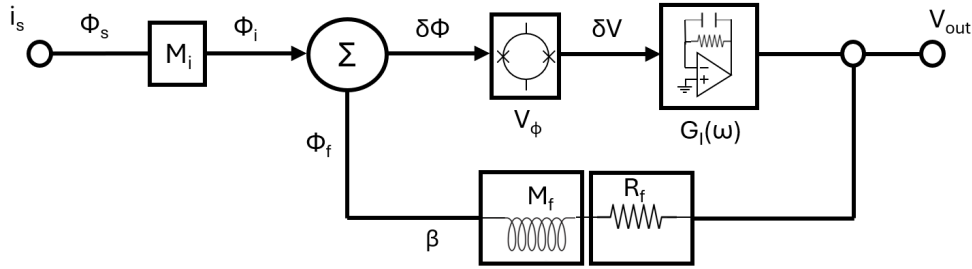


Figure 2.20: Scheme of the SQUID circuit signal components. The SQUID is represented as a integrator plus an operational amplifier. $\delta\Phi$ is the small variation applied to the magnetic flux generating the tension in the SQUID and V_ϕ is transimpedance. $G_I(\omega)$ is the gain of the integrator and β is the fraction of the output sent back to the SQUID through the feedback and summed to the input flux.

flux are defined.

$$\begin{cases} x_o = V_{out} \\ x_s = \Phi_s \end{cases} \quad (2.52)$$

The gain is determined as: $A(\omega) = G_I(\omega)V_\phi$, where $V_\phi = dV/d\Phi$ and $G_I(\omega) = \omega_1/(i\omega)$. From the definition of Φ_f the value of the fraction β is derived.

$$\Phi_f = -\frac{V_{out}}{R_f} M_f \rightarrow \beta = -\frac{M_f}{R_f} \quad (2.53)$$

Combining the last equation of 2.51 with eq. 2.52 and eq. 2.53 together:

$$V_{out} = \frac{A(\omega)}{1 - \beta A(\omega)} \Phi_s = \frac{\frac{\omega_1}{i\omega} V_\phi}{1 + \frac{M_f \omega_1}{R_f i\omega} V_\phi} M_i i_s = R_f i_s \frac{M_i}{M_f} \frac{1}{1 + \frac{R_f}{M_f} V_\phi \frac{i\omega}{\omega_1}} \quad (2.54)$$

Simplifying, V_{out} becomes:

$$V_{out} = R_f i_s \frac{M_i}{M_f} \frac{1}{1 + \frac{i\omega}{\omega_{cut}}} \quad (2.55)$$

where $\omega_{cut} = \omega_1 \frac{M_f}{R_f} V_\phi$.

In the gain-bandwidth dependence a “cut-off” frequency is introduced: above this

frequency the gain descends to zero, as shown in Figure 2.21. This is an important feature for the SQUIDS, which sets a limitation on the operation frequency range. In the presence of a delay a further element to the gain is added.

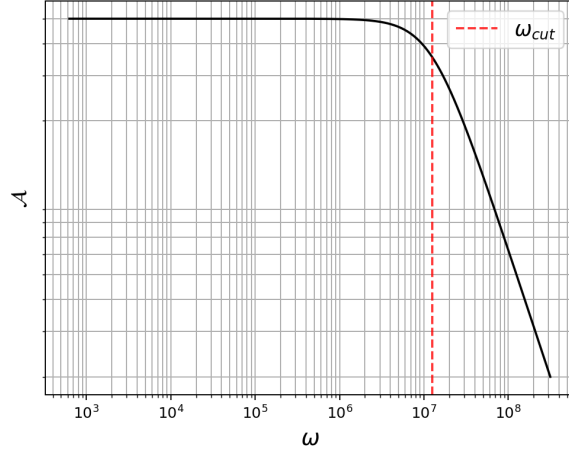


Figure 2.21: Gain-bandwidth product shown in the “cut” at ω_{cut} .

$$A(\omega) = \frac{1}{i\omega} V_\phi e^{-i\omega t_D} \quad (2.56)$$

where t_D is the time delay, responsible for additional phase shift $-\omega t_D$. Substituting this in 2.54:

$$A(\omega) = \frac{\frac{\omega_1 e^{-i\omega t_D}}{i\omega} V_\phi}{1 + \frac{M_f}{R_f} \frac{\omega_1}{i\omega} V_\phi e^{-i\omega t_D}} \quad (2.57)$$

Considering $e^{-i\omega t_D} = \cos(\omega t_D) - i \sin(\omega t_D)$:

$$A(\omega) = \frac{\omega_1}{i\omega} V_\phi \frac{\cos(\omega t_D) - i \sin(\omega t_D)}{1 + \frac{M_f}{R_f} V_\phi \frac{\omega_1}{\omega} [\cos(\omega t_D) - i \sin(\omega t_D)]} \quad (2.58)$$

Simplifying the expression, eq. 2.58 becomes:

$$\|A(\omega)^2\| \approx \frac{1}{\sqrt{1 + \frac{\omega}{\omega_2} \left(\frac{\omega}{\omega_2} - 2 \sin(\omega t) \right)}} \quad (2.59)$$

With the approximation $\sin(\omega t_D) \sim \omega t_D \sim \frac{\omega}{\omega_D}$ the following relationship is defined:

$$\|A(\omega)^2\| \approx \frac{1}{\sqrt{1 + \frac{\omega^2}{\omega_2^2} \left(\frac{\omega_D - 2\omega}{\omega_D} \right)}} \quad (2.60)$$

where ω_2 contains all the M_f , R_f , V_ϕ terms. The more the negligible the delay, the more the resulting gain will collapse into Figure 2.21. The delay should be considered in the context of the potential presence of long cables, which are known to introduce such effects.

2.5 The “warm” readout: the electronics of LiteBIRD

The readout chain of LiteBIRD consists of a cryogenic section and room temperature section. The “cold” part is designed to implement the *multiplexing*, a technique which will be described in detail in Section 3.1 of the next chapter. Multiplexing is used to optimize the efficiency of the detectors readout and to reduce the heat load on the cryogenic stages by decreasing the number of SQUIDs required, connecting a single SQUID to multiple detectors.

The cold section is connected to a “warm” electronics (WE), as shown in Figure 2.22, which is divided in three main parts:

1. the **SPA (Signal Processing Assembly)**, which generates the digital signals and communicates through the FPGA (field-programmable gate array).
2. the **DA (Digitizer Assembly)**, which converts the signal generated by the FPGA from digital to analog and transforms the tones from the SQUIDs back from analog to digital to be processed by the SPA
3. the **SCA (SQUID Controller Assembly)**, which includes the biasing electronics and preamplification stages for the output signals coming from the SQUIDs.

The number of WE elements is to be consolidated during the LiteBIRD reformation plans. The former configuration [16] included: 4 DAs for each SPA with 15 multiplexing modules.

In its current configuration, a single SCA is designed to readout 4 SQUIDs. As can be seen in Figure 2.22, the SPA and the four DAs are hosted in the **SPU (Signal Processing Unit)**. A yet to be fixed number of SPUs will be necessary to cover the complete readout of LiteBIRD.

The roles of the complete electronics include the management of the readout of the detectors and the SQUIDs, together with the signal processing, the temperatures control and the housekeeping. The coordination and centralization of these actions will be appointed to the **MDPUs (Mission Data-Processing Units)**.

In the configuration with two telescopes, the MDPUs would have been two: one for LFT and one for MHFT. The MDPUs are responsible for both the science signal processing and the cooling chain, including the 0.1 K, 0.3 K and the 1.8 K stages linked to the ADRs (Adiabatic Demagnetization Refrigerators), the Joule-Thomson

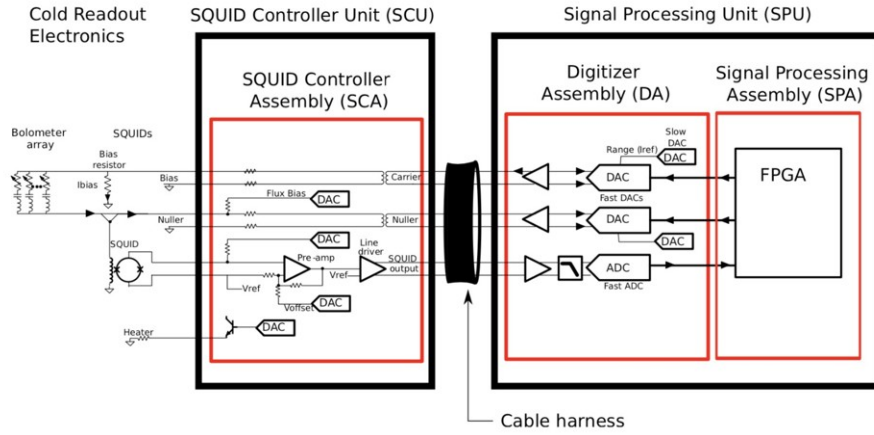


Figure 2.22: The scheme of the SPA-DA-SCA connected to a single SQUID as reported in [16]. The cold readout part, based on the Frequency Division Multiplexing will be discussed in the next chapter.

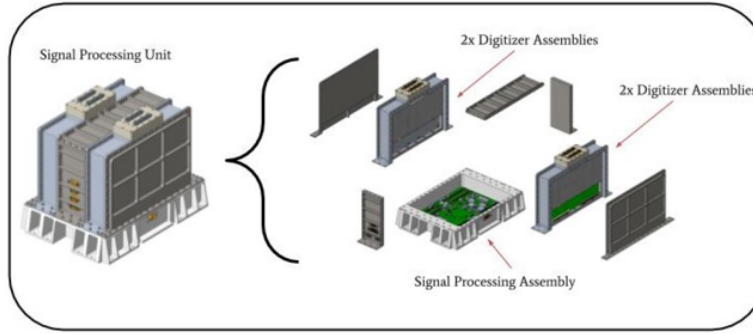


Figure 2.23: The Signal Processing Unit in all its parts: the four DAs and the SPA.

cooler driver and the 15 K Pulse Tube cooler driver for the “warmer” stages. A complete description of the LiteBIRD cryogenic system can be found in [16].

This concludes the description of the experimental aspects of LiteBIRD. The highly sensitive detectors and the complex readout chain are implemented to observe the CMB polarization, enabling the construction of sky maps and power spectra. The following section describes the data acquisition process and the map-making procedure.

2.6 From the instrument to the CMB power spectra

The data collected from LiteBIRD observations will be organized into time series known as **Time-Ordered Data (TODs)**, sampled at ≈ 19 Hz. This results in data

streams with a temporal spacing of approximately 0.05 seconds.

With regards to the TODs, they can be described by:

$$d_t = \sum_p A m_p + n_t \quad (2.61)$$

where t is the time, p is the sky patch pixel (the location of the sky observed by each detector), m_p is the sky signal, A is the pointing matrix (or mixing matrix) and n_t is the noise. A is shaped in this way: it is a $(N_t, 3N_p)$ sized matrix, where N_t is the number of samples in the TODs and N_p is the number of pixels. Each row represents an instant of time at which a sky pixel is observed. The columns represent the pixels: for each row, the number of columns is either 1 (for intensity measurements) or 3 (for polarization-sensitive measurements).

Suppose to have 2 sky patches observed with N_t samples and to measure intensity and polarization. This is the shape of the mixing matrix in this case:

$$A = \begin{pmatrix} 1 & \cos \Psi_1 & \sin \Psi_1 & 0 & \dots & 0 \\ 1 & \cos \Psi_2 & \sin \Psi_2 & 0 & \dots & 0 \\ 0 & 0 & 0 & 1 & \cos \Psi_3 & \sin \Psi_3 \\ \vdots & & & & & \end{pmatrix} \quad (2.62)$$

where Ψ_i is the angle of the detector with respect to the sky coordinates. If the sky patches were more than two, we would add additional columns. The three components per pixel represents the I , Q and U Stokes parameters. The TOD for each detector can be rewritten as:

$$d_{t,i} = I + Q \cos 2\Psi_i + U \sin 2\Psi_i \quad (2.63)$$

Various map-making techniques can be applied to the time-ordered data (TODs) to reconstruct the CMB maps. Map-making consists of estimating a sky quantity m from the TODs through a linear operation, $m = M \cdot d$. In the binning method, the data are averaged within the same pixel: $M = (A^T N A)^{-1} A^T N$, where N is the weight matrix that determines the noise properties of the resulting maps. One possible approach is to define $N = C_n^{-1}$, where C_n is the noise covariance matrix. In this way, M introduces additional weighting before projecting time-domain noise onto the sky [33].

In the case of purely white noise, the binned map-making approach implemented in HEALPix is suitable ([34], [33]). The most important quantity is the resolution N_{side} which selects the number of pixels $N_{pix} = 12 \times N_{side}^2$.

Whenever a low-frequency $1/f$ correlated noise is present, another procedure of map-making needs to be selected. A detailed discussion of these techniques, however, goes beyond the scope of this thesis and we refer the reader to [16] and [35].

2.7 Conclusions

This chapter has presented the LiteBIRD experiment by outlining its scientific objectives. The instrument has been described in some details, with a focus on the detectors (TESs) and the main components of the readout system (SQUIDs). The warm readout, consisting of the electronics, has been briefly introduced. The challenges associated with operating these extremely sensitive yet technically demanding devices have also been discussed. In the next chapter, the cold readout section will be extensively described.

Finally, the simplest map-making method for converting raw measurements into power spectrum maps has been introduced, providing the necessary background for the analysis developed later in this thesis. Further details on the calculation of the CMB power spectra following map-making will be presented in Chapter 4.

Chapter 3

The Frequency Domain Multiplexing (FDM) and the Electrical Crosstalk

In this chapter the issues associated with large numbers of detectors and SQUIDs will be discussed and the Frequency Division Multiplexing (FDM) technique implemented to address them will be described. The instrumental systematic effect known as electrical crosstalk will also be presented, together analytical models and calculations focused on evaluating its level.

3.1 Multiplexing readout

The presence of complex wiring systems linked to a large number of detectors at cryogenic temperature could cause extra heating of the cryogenic stage due to power dissipation. Specific readouts schemes are required to reduce the heat load and also the costs. *Multiplexing* consists in transmitting signals to a multi-channel system through a single line, and retrieving the output from the same system using a single return line.

There are several multiplexing configurations which can be implemented [36]. Among these, the most considered for space missions are the **time-division multiplexing** and the **frequency-division multiplexing**.

In Time Division Multiplexing (TDM) the detectors are read one at a time. TDM is performed by transmitting different signals on a common line and using *switches* to bias the designated SQUIDs. When the switch is closed, the input signal flows

through the SQUID, as shown in Figure 3.1. A specific implementation of TDM by *NIST* (National Institute of Standards and Technology) is shown in Figure 3.2 from [37]. A matrix of detectors (M rows and N columns) is connected to M current biased SQUIDS. The TESs are connected to a first stage SQUID (SQ1) and are sequentially readout one by one thanks to the switches. All the M SQUIDS are

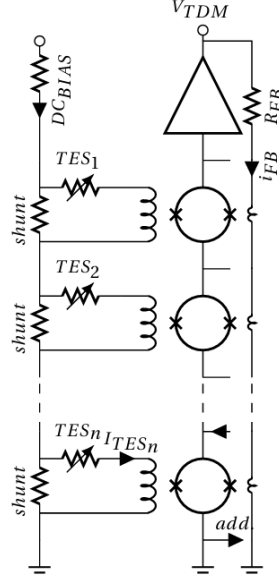


Figure 3.1: Scheme of the TDM configuration from [36]

readout in parallel. A DAC selects the time interval to “activate” the SQUIDS, with a duration sufficient to detect the signal. In this way, the signals from individual detectors can be identified during data processing by their characteristic “activation time”.

Experiments with different goals, like CMB-S4, observing the CMB polarization, and Athena, studying the X-rays, implement TDM.

Frequency Division Multiplexing (FDM) is obtained by dividing the total readout bandwidth in shorter frequency ranges for each detector. Each TES is connected in series to a bandpass filter in order to exploit the filter limited bandwidth. Filter composed by inductors and capacitors are most commonly used. The resonance frequency for this pair of elements is defined as follows.

$$f_{res} = \frac{1}{2\pi\sqrt{LC}} \quad (3.1)$$

In an FDM readout scheme, each *leg* consists of a Transition Edge Sensor (TES) connected in series with a band-pass LC filter. All the legs are connected in par-

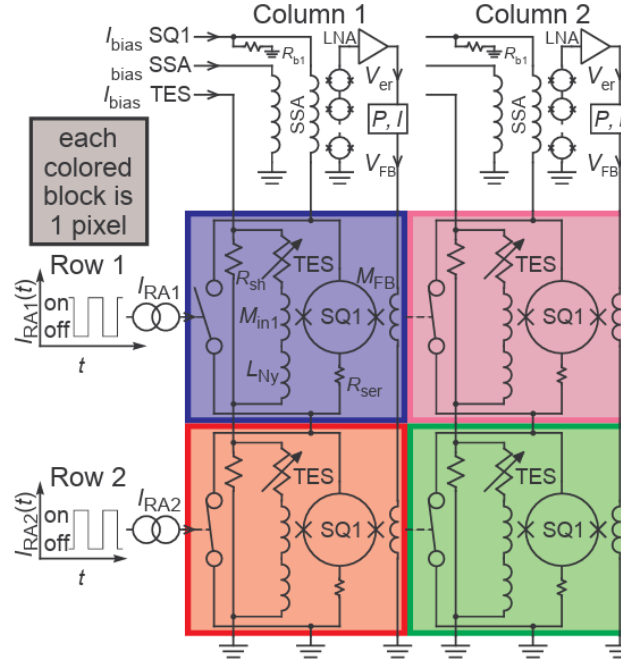


Figure 3.2: Schematic of a 2×2 matrix of TESs. Each detector is connected to a first stage SQUID (SQ1) and then to a SSA (SQUID Series Array). Each colored block shows the multiplexing of each TES and the switches turned on by a current I_{RA} ([37]).

allel, as illustrated in Figure 3.3. A “comb” of bias signals, each at the resonant LC frequency, is distributed to the individual branches through a common bias line. The signals from all branches are then summed and sent to the input coil of a single SQUID amplifier, providing a single readout channel that combines the information from multiple detectors simultaneously.

The presence of a small resistance R_{shunt} in parallel with the FDM legs provides a stable bias voltage to the detectors and the possibility for the Transition Edge Sensors to exploit negative ETF (see Chapter 2-Section 2.3 and Section 2.4).

The arrival of the sky signal hitting a TES will warm the detector’s absorber and will cause the increasing of the resistance of each TES. The ETF will drive the TES back in the operating bias voltage. As a consequence, the current at that specific multiplexing frequency is amplitude modulated. As will be discussed later in this work, the output signals at a specific frequency can be associated with the corresponding TES, allowing for the analysis of the physical sky signal at a later stage of the readout chain.

The combined signal is amplified by the SQUID and sent to the warm electronics, where it is further amplified and digitally demodulated. Digital demodulation separates the carrier tones, enabling the reconstruction of the individual TES responses.

Frequency Domain Multiplexing (FDM)

Frequency **transposition** of each input signal (S_x) is **summed**

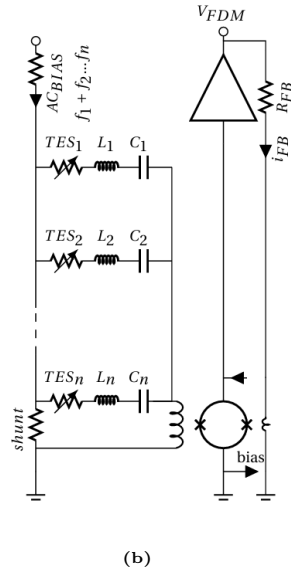
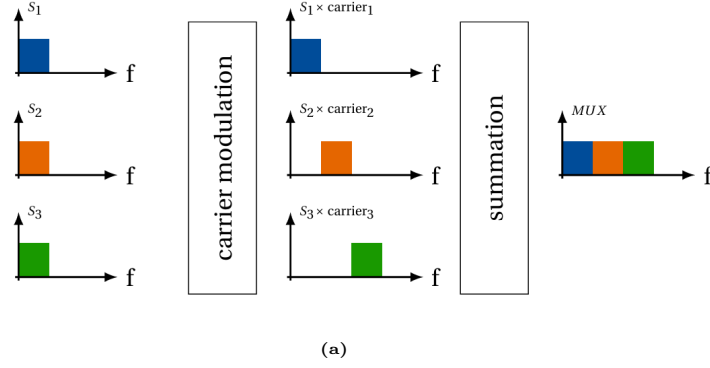


Figure 3.3: (a) A scheme of the FDM configuration. (b) Parallel legs hosting the TESs variable resistance, the inductor and the capacitor are shown. The presence of a bias resistance (R_{shunt}) and the SQUID input coil in series are represented as well.[36]

Experiments like *POLARBEAR* ([38]) and *SPT-3G* ([39]), observing CMB from ground, together with future experiments like LiteBIRD, implement FDM. The multiplexing frequency range needs to be tuned taking several circuitual constraints into account. As mentioned in Section 2.4 in Chapter 2, the delay of the signal through the SQUID introduces a cut-off in its amplification spectrum. FDM must account for this effect by allocating the multiplexing frequencies within a range below this cut-off frequency.

3.1.1 Lumped elements for Frequency Division Multiplexing

In this and the subsequent sections, the features and characteristic behavior of FDM configurations are discussed.

In many experiment, FDM is performed by connecting each TES to a completely lithographed LC filter.

For LiteBIRD, for example, the multiplexing frequency range, i.e. the resonant frequencies of the LC filters, will lay between 1.6 MHz and 5.5 MHz, with a spacing of around 90 kHz and different multiplexing factors for each subassembly.

The advantage of **lumped lithographic elements** mostly stands in the possibility of shaping the 2D design to precisely tune the capacitance and inductance values and to use the precision of optical lithography to reach the best design. High multiplexing factors and the subsequent increase of the number of filters, lead to the need of decreasing the parasitic resistance related to the components. Considering that the series impedance of ceramic elements increases with frequency, completely lithographic inductors and capacitors get closer to this goal. This drives the definition of quality factor or Q -factor of the LC filters, defined by the following equation:

$$Q = \frac{f_{res}}{\Delta f} \quad (3.2)$$

where f_{res} is the resonance frequency and Δf is the bandwidth of the filter, which is calculated as:

$$BW = \frac{R}{2\pi L} \quad (3.3)$$

where R and L are respectively the resistance and inductance values of the series. This quantity needs to be as small as possible, to properly isolate each TES from the neighbor and to minimize systematic effects like the one that will be discussed in the next section: electrical crosstalk. As it can be observed, the bandwidth gets smaller for small resistance. This leads to the requirement of minimizing the parasitic resistance, as previously reported. Maximizing the value of L represents another possible approach, but it presents some limitations since it is constrained by the physical size of the inductors. Therefore, a trade-off must be achieved to optimize the configuration.

Another advantage of lithographed filters is related to the possibility of packing a large number of LCs on a smaller chip. This is crucial for cryogenic experiments.

Two valid options for cryogenic readout are based on lithographed plane parallel or interdigitated capacitors coupled to coil inductors. The area size of the first type

is smaller but the latter can be realized with monolayer lithography. Examples and tests involving these two types of LCs will be shown later in this work.

Square coils are frequently used as lithographed inductors. The inductance value of these elements can be approximated by the following formula [40].

$$L = 1.25 \mu_0 d N^2 \quad (3.4)$$

where $\mu_0 = 4\pi \cdot 10^{-7} \text{H/m}$ is the magnetic vacuum permeability, d is the inner square side and N is the number of turns of the coil, as empirically found in [40]. A simple drawing of these devices and the picture of a single coil inductor are reported in Figure 3.4.

Different design variations can be implemented to optimize the performance. As

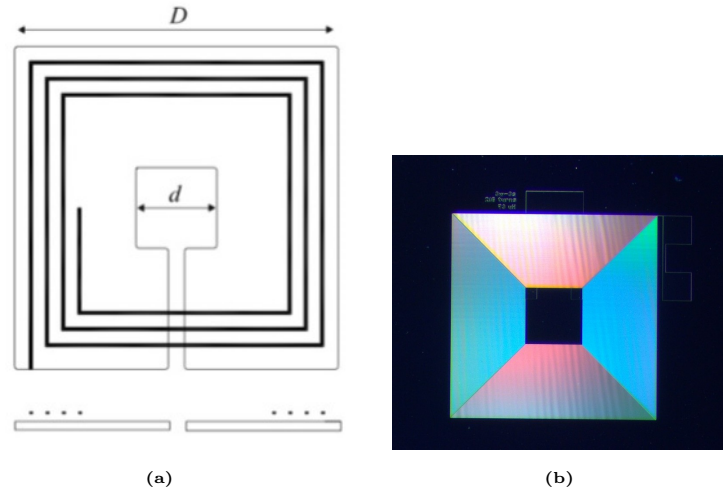


Figure 3.4: (a) Square coil inductors as schematized in [40]. (b) Picture of a lithographed spiral inductor. Dimensions are $3.8 \text{ mm} \times 3.8 \text{ mm}$.

illustrated in Figure 3.5 from [41], spiral inductors can be connected in series using a mirrored configuration (wound in opposite directions) to minimize mutual inductance. Adjacent multiplexing legs can be placed at a minimal distance without increasing the overall electromagnetic coupling. Lithographed parallel-capacitors have the well known behavior of the ceramic components. The capacitance depends on the geometry and it is expressed by the ratio between the charge Q on each conductor and the voltage V between them. From Gauss's law we can define the voltage. From this, the capacitance value depends on the dielectric coefficient of the insulating material ϵ between the plates, the area of the plates A and the distance d between them.

$$\begin{cases} V = \frac{Qd}{\epsilon A} \\ C = \frac{\epsilon A}{d} \end{cases} \quad (3.5)$$

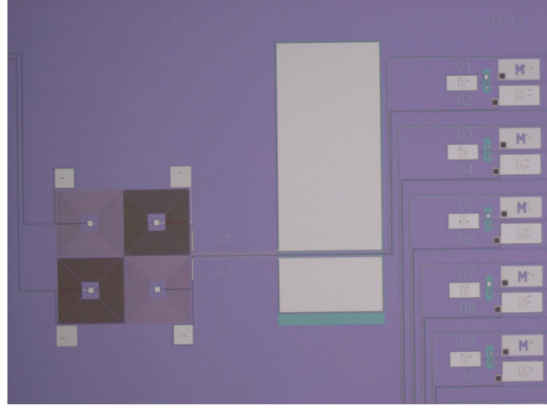


Figure 3.5: 2 coils inductor. The two grey shapes are the capacitors connected to two couples of inductors ([41]).

The capacitors are produced by multi-layer lithography, with the conductor plates and the dielectric in between. Each size is shaped to reach the required capacitance value.

On the other hand, interdigitated capacitors are designed as “multifinger” periodic structures which take advantage of the capacitance across the narrow gap between thin-film conductors, as shown in Figure 3.6(a). Through semi-empirical simulations, it has been shown in [42] that the relation between the capacitance and the different shapes of the “meanders” is the following.

$$C = (\epsilon_r + 1) \ell [(N - 3) 4.409 \cdot 10^{-6} + 9.92 \cdot 10^{-6}] \quad (pF) \quad (3.6)$$

where ϵ_r is the relative dielectric coefficient of the substrate, ℓ is the finger length and N is the number of fingers. The picture of an interdigitated capacitor is shown in Figure 3.6(b). Size needs to be smaller than the operating wavelength, which leads to $\ell < \lambda/4$. In both cases it can be noticed that it is possible to shape the

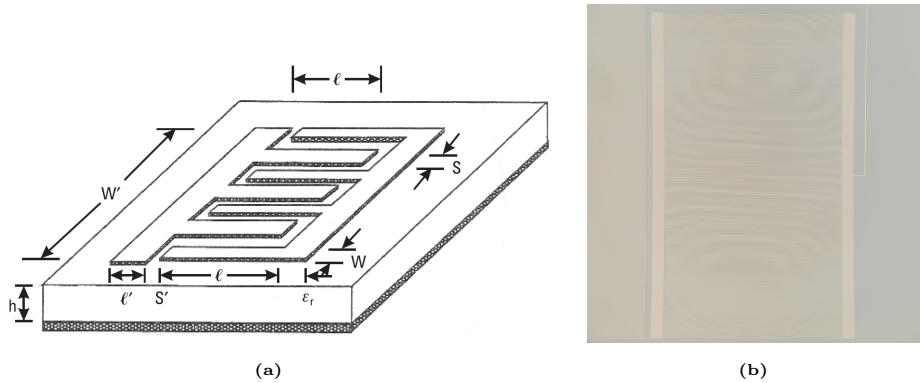


Figure 3.6: (a) Representation of interdigitated capacitor ([42]), (b) picture of an interdigitated capacitor. Dimensions are 3.4 mm×5.4 mm.

capacitors and the inductors to obtain the specific resonant frequency we are looking

for. This, together with the minimization of the parasitics, is why superconducting readout chains are nowadays commonly used.

3.2 A systematic error for FDM: electrical crosstalk

The employment of FDM, while offering many benefits, also carries certain risks, such as systematic effects which, if not properly modeled and studied, could compromise the ultimate objective of the experiment itself. Among these, electrical crosstalk is one of the most important to be modeled.

The inclusion of LC filters is intended to isolate one leg from the others and select the signal from a single TES using the resonance frequency of the FDM. This configuration can be disrupted by electrical crosstalk, leading to signal leakage from the on-resonance to the off-resonance legs. Sky signals incident on one TES may be misinterpreted as being detected by neighboring detectors, potentially compromising the measurement.

The main contributors which will be described in the next sections are: carrier leakage crosstalk, mutual inductance crosstalk and common impedance crosstalk.

This effects must be distinguished from *thermal crosstalk*, which is solely related to the dissipation of heat from the signal hitting a single TES into the thermal bath, which can subsequently influence neighboring detectors, and *optical crosstalk* which is related to signal leakage happening at the level of the optical chain. In this work, we will address only electrical crosstalk, omitting the analysis of other crosstalk effects.

3.2.1 Carrier leakage crosstalk

The *Carrier leakage* is an unavoidable form of electrical crosstalk and it is strongly related to the bandwidth of the single LC filter. It can be defined as a current leakage due to the non-infinite impedance of the off resonance legs.

When the on-resonance current flowing through the leg is modulated due to the sky signal hitting the detector, an off-resonance current is induced in the neighbors and modulated as well. Only frequency neighbors are affected by this, while spatial proximity does not play a role.

This type of crosstalk can be modeled as follows. We calculate the admittance for the i -th leg at the ω_i frequency considering the circuital elements. We include the bias impedance and the SQUID input coil in series with the entire multiplexing

scheme, as shown in Figure 3.3.

$$I_i(\omega_i) = I_0(\omega_i) \frac{1}{1 + \left(\frac{Z_{TES,i}(\omega_i) + Z_{LC,i}(\omega_i) + Z_{stray}(\omega_i)}{R_{bias}} \right)} \quad (3.7)$$

where $I_0(\omega_i)$ is the input current, $Z_{TES,i}$ and $Z_{LC,i}$ are the impedance of the TES and the LC filter, Z_{stray} represents the impedance of the SQUID input coil and associated wiring in series with the branches, R_{bias} is the shunt resistance.

This formula can be written for the off-resonance j -th legs.

$$I_j(\omega_i) = I_0(\omega_i) \frac{1}{1 + \left(\frac{Z_{TES,j}(\omega_i) + Z_{LC,j}(\omega_i) + Z_{stray}(\omega_i)}{R_{bias}} \right)} \quad (3.8)$$

where $j \neq i$.

In the latter, we are describing the current at frequency ω_i flowing across the neighbor legs. In this case the impedance should be infinite, but due to the finite bandwidth of the LC filters, this is not the case. As a consequence, the current flowing in the i -th leg induces a modulation of the nonzero current in the j -th leg when the i -th TES resistance is modulated by the sky signal.

Carrier leakage crosstalk can be quantified as the ratio between the modulation of the induced current and the variation of the on-resonance current, as defined in Eq. 3.9.

$$\text{xtalk}_{cl} = \frac{dI_j(\omega_i)/dR_{TES,i}}{dI_i(\omega_i)/dR_{TES,i}} \quad (3.9)$$

Considering the relations $[j\omega_i L + 1/(j\omega_i C_j)] \simeq j2L\Delta\omega$ and $[j\omega_i L + 1/(j\omega_i C_i)] \simeq 0$ and assuming $R_{TES,i}$ being larger than R_{bias} and Z_{stray} , eq. 3.9 is approximated by

$$dI_j(\omega_i)/dR_{TES,i} \simeq \frac{V_{bias}}{(R_{TES,i} + j2\Delta\omega L)^2} \quad (3.10)$$

$$dI_i(\omega_i)/dR_{TES,i} \simeq -\frac{V_{bias}}{R_{TES,i}^2} \quad (3.11)$$

where $V_{bias}(\omega_i) = I_0(\omega_i)R_1$ is the voltage bias generated by driving the input current through a large resistor R_1 . The crosstalk ratio becomes:

$$\text{xtalk}_{cl} \simeq \frac{R_{TES,i}^2}{(R_{TES,i} + j2\Delta\omega L)^2} \quad (3.12)$$

3.2.2 Mutual inductance crosstalk

The *Mutual inductance* crosstalk originates from the magnetic coupling among the inductors in the circuit. Coupling between a pair of coils can induce a voltage across

the inductors of the off-resonance legs, generating a current flow. This current will be in turn modulated when the TES resistance varies due to sky signal.

To investigate this mechanism in detail, two adjacent channels, labeled i and j are considered. The magnitude of this effect can be estimated by modeling the geometrical inductance as follows. The mutual inductance is defined as:

$$M_{ij} = \frac{N_j \Phi_{ij}}{I_i(\omega_i)} \quad (3.13)$$

where N_j is the number of coils of the “victim” inductor, Φ_{ij} is the magnetic field flux into the off-resonance coil and $I_i(\omega_i)$ is the current flowing through the “primary” inductor.

The current $I_i(\omega_i)$ will generate a magnetic field B_i , which in turn will induce a flux Φ_{ij} through the neighbor inductor spiral. The flux will be proportional to the current according to Eq. 3.13.

Since the current distribution is localized, we can use the approximate expression derived from the multipole expansion and valid in the far-field limit. Among the multipole terms, we consider the dipole contribution, defined as:

$$\vec{m} = N_i I_i A_i \hat{n} \quad (3.14)$$

where N_i and $A_i \hat{n}$ are respectively the number of turns and the oriented internal area (the one of the central hole) of i -th coil, which are defined by the Jaycox-Ketchen formula 3.4.

The flux through the j -th loop can be evaluated as:

$$\Phi_{ij} = - \int_{A_j} \frac{\mu_0}{4\pi} \frac{\vec{m} \cdot \hat{n}}{r^3} da_j \quad (3.15)$$

And using this result in equation 3.13:

$$M_{ij} = \frac{N \Phi_{ij}}{I_i} = - \frac{\mu_0}{4\pi} \frac{N^2}{r^3} A^2 \quad (3.16)$$

where we assumed $M_{ij} = M_{ji} = M$ and $L_i = L_j = L$. The **coupling constant** k is:

$$k = M/L \quad (3.17)$$

The mutual inductance will induce a voltage across the “victim” inductor and therefore a current flowing in the off-resonance legs, as shown in Eq 3.18.

$$\begin{cases} V_j = i\omega_i M_{ij} I_i(\omega_i) \\ I_j(\omega_i) = \frac{V_j}{R_{TES,j} + i\omega_i L + \frac{1}{i\omega_i C_j}} = \frac{iM_{ij}\omega_i I_i}{R_{TES,j} + i\omega_i L + \frac{1}{i\omega_i C_j}} \end{cases} \quad (3.18)$$

and substituting the value of M_{ij} :

$$I_j(\omega_i) = \frac{i\omega_i}{i\omega_i L} \frac{M_{ij} I_i(\omega_i)}{1 + \frac{R_{TES,j}}{i\omega_i L} - \frac{1}{\omega_i^2 C_j L}} = \frac{1}{L} \frac{k L I_i(\omega_i)}{\left(1 + \frac{R_{TES,j}}{i\omega_i L} - \frac{1}{\omega_i^2 C_j L}\right)} \quad (3.19)$$

Based on this, the mutual inductance crosstalk can be defined as the variation of current $I_j(\omega_i)$ when the resistance R_i of the on-resonance leg varies normalized to the variation of the on-resonance current $I_i(\omega_i)$.

$$\text{xtalk}_{mut} = \frac{dI_j(\omega_i)/dR_{TES,i}}{dI_i(\omega_i)/dR_{TES,i}} \quad (3.20)$$

3.2.3 Common impedance crosstalk

The common impedance crosstalk is a very important contributor for FDM readouts. This arises from the presence of the SQUID input coil and the wiring inductance in series with all the parallel LC legs. This causes the system to be operated slightly outside the resonance frequency given solely by the LC filters. As mentioned, Frequency Division Multiplexing is operated by connecting each TES to a LC filter and the combined output signals coming are sent to a single low-impedance SQUID with a unique set of wires. For this reason, a large common impedance can have a large impact on the electrical crosstalk level.

The effect of this crosstalk contributor is described in the following.

The sky signal incident on the TES in the i -th leg causes a modulation of the current

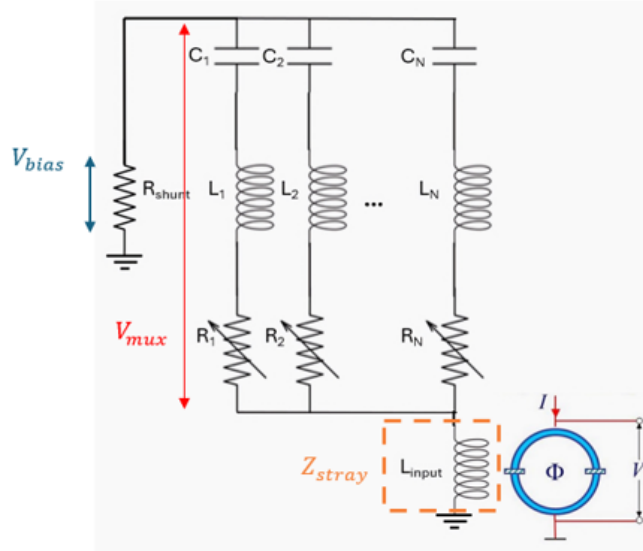


Figure 3.7: Schematic representation of what happens with common impedance crosstalk

flowing through the common impedance Z_{stray} [43].

A change in the voltage V_{stray} across the inductance L_{stray} produces a corresponding variation in V_{mux} , the voltage across the multiplexing branches. A scheme of this components is shown in Figure 3.7. Based on the Kirchhoff law, this can be re-wrote in the following way.

$$dV_{mux} = dV_{bias} - dV_{stray} \quad (3.21)$$

where V_{bias} is the voltage across R_{shunt} (also named R_{bias} in the previous section). By assuming V_{bias} to be fixed and explicitly expressing dV_{stray} as a function of the current modulation $dI_i(\omega_i)$, which is the variation of the current through the i -th leg at the on-resonance frequency ω_i , Eq. 3.21 becomes:

$$dV_{mux} = -dV_{stray} = -dI_i(\omega_i)Z_{stray} \simeq -dI_i(\omega_i)i\omega_i L_{stray} \quad (3.22)$$

where we assume that Z_{stray} is primarily defined by the common inductance. This voltage induces a current at the j -th frequency in the off-resonance leg $I_j(\omega_i)$. This current is therefore modulated when the current through Z_{stray} varies [43].

$$dI_j(\omega_i) = \frac{dV_{mux}}{dZ_j(\omega_i)} \simeq \frac{-dI_i(\omega_i)j\omega_i L_{stray}}{j2\Delta\omega L} \quad (3.23)$$

We can define the common impedance crosstalk as the ratio between the variation of the off-resonance current when the TES resistance R_i varies and the variation of the on-resonance current in the same case.

$$\text{xtalk}_{com} = \frac{dI_j(\omega_i)/dR_i}{dI_i(\omega_i)/dR_i} = -\frac{1}{2} \frac{\omega_i}{\omega_j - \omega_i} \frac{L_{stray}}{L} \quad (3.24)$$

This quantity linearly depends on L_{stray} and ω_i and inversely from the frequency distance between the two legs and the filter inductance value. The closer two legs are in the multiplexing frequency space, the more significant the impact of crosstalk.

Some comments on the electrical crosstalk follow. We calculated three contributions solely related to the electrical side of the circuit. As already mentioned during this chapter, we are not considering thermal or optical crosstalk.

Each type of crosstalk arises when a resistance change from the sky signal modulates the current in a given leg, inducing currents in neighboring off-resonance legs for different physical reasons.

For each type, we specified the frequency of the induced current, noting that it corresponds to the on-resonance frequency of the branch of the detector affected by the sky signal. For example, if the i -th TES resistance R_i changes, a modulation in the adjacent current $I_j(\omega_i)$ is induced. It will be shown that this also affects the

on-resonance current of the neighboring legs (denoted $I_j(\omega_j)$). This occurs because variations in the TES resistance lead to changes in the total impedance (assuming, as done here, that the off-resonance impedance is finite), which in turn modulates the current at each frequency. When measuring the crosstalk level or evaluating the current flowing into the SQUID, the sum of the currents at the node is always considered. Detailed analyses of specific experimental examples will be analyzed in detail in the following chapters.

Additionally, some remarks can be made on the expected levels of crosstalk. Mutual inductance crosstalk can be significantly minimized by optimizing the spatial arrangement of the filters. Other methods have been implemented to achieve this goal, as will be presented in the next sections. For instance, the design of spiral inductors can be adjusted to reduce electromagnetic coupling. Carrier leakage crosstalk can be reduced by narrowing the filter bandwidth and increasing the spacing between frequencies, but it cannot be minimized indefinitely. Calculations using characteristic quantities from CMB experiments indicate that the level of carrier leakage crosstalk is around 10^{-4} or lower. Similarly, common-impedance crosstalk is determined by L_{stray} , which is primarily set by the SQUIDs manufacturing process and is typically around 10^{-9} H. It is inversely proportional to the inductance of the filters, typically on the order of 10^{-6} H. Common impedance crosstalk, based on the ratio between these two quantities, is expected to be around 10^{-3} .

3.3 Conclusions

This chapter described the FDM configuration, enumerating the motivations for its implementation and highlighting the advantages of including a fully lithographed design. A description of the fundamental components was included, exploiting the differences between different choices.

The contributors to electrical crosstalk were thoroughly enumerated, and for each of them an analytical model was formulated and described. A discussion of their applicability and accuracy was also initiated, we will focus on more details in the next chapters.

Chapter 4

The design optimization of LC filters and multiplexing frequency scheduling

In this chapter the methods developed to define the multiplexing frequency sequence for each group of detectors, here called *multiplexing frequency scheduling*, is described. In particular, the analytical model of the crosstalk impact on the CMB power spectra is presented. The design of an LC filter following these guidelines is shown.

4.1 Multiplexing frequency scheduling

Although electrical crosstalk is an instrumental effect, it affects the data analysis, particularly the reconstruction of the CMB power spectra from the observational data. Previous works [44] have shown that the assignment of multiplexing frequencies to the TESs, based on their positions on the focal plane, can significantly affect the system performance.

The crosstalk is included in the TODs as a mixing matrix among the detectors.

$$\tilde{d}^i = \sum_j X_{ij} d^j \quad (4.1)$$

where X_{ij} is the crosstalk element between the i -th and the j -th detectors in the crosstalk matrix and d^j is the time-order data array for the j -th detector without crosstalk. $\tilde{d}^i$ is the resulting TOD with crosstalk. The TODs are the data streams in time collected by each detector. They will be downsampled to obtain a data rate

of 19 Hz.

Expanding Eq. 2.61 from Chapter 2, the maps are calculated from the TODs as shown in eq. 4.2. More details on this can be found in Chapter 2.

$$\hat{m}_p = \sum_t (A^t N A)^{-1} A^t N d_t \quad (4.2)$$

where A is the pointing matrix and N determines the noise properties of the system. Each row corresponds to a time sample, while each column represents a pixel in the observed sky.

A can be rewritten as $A^i(t, \vec{r}) = e^{-is\Delta a_i} A(t, R_i^{-1} \vec{r})$, where Δa_i denotes the relative offset between the boresight direction¹ and the detector position, and s is the spin of the field (0 for temperature, 2 for polarization).

As shown in Chapter 1, the coefficient C_ℓ are described by the following relation:

$$C_\ell = \frac{1}{2\ell + 1} \sum_{\ell, m} |a_{\ell m}|^2 \quad (4.3)$$

In the presence of electrical crosstalk, this formula will be modified. In the remainder of this section the procedure used in [44] to inject crosstalk in the power spectra will be summarized.

When crosstalk between detectors i and j is included, the map transforms as follows:

$$\tilde{m}^i(\vec{r}) = \sum_j X_{ij} e^{-isa^{i \rightarrow j}} m^j(R_j R_i^{-1} \vec{r}) \quad (4.4)$$

where $a^{i \rightarrow j}$ is the rotation from the orientation of detector i to that of detector j .

If we assume that the resultant maps are the weighted sum of individual detector maps and that each detector has a uniform map coverage, we can define:

$$\mathcal{X}_{ij} \equiv \omega_i X_{ij} \quad \text{with} \quad \sum_i \omega_i = 1 \quad (4.5)$$

where ω_i is the weight.

Defining $\mathcal{X} = \text{tr} \mathcal{X}_{ij}$, the map equation can be rewritten as:

$$m'(\vec{r}) = \mathcal{X} m(\vec{r}) + \sum_{i, j \neq i} \mathcal{X} e^{-isa^{i \rightarrow j}} m^j(\vec{r}) \quad (4.6)$$

Under the *flat-sky* approximation and assuming a fixed focal plane orientation, the map affected by crosstalk can be expressed as the convolution of the original map m with the point spread function.

$$m'(\vec{r}) = (m * P)(\vec{r}) \quad \text{with} \quad P(\vec{r}) = \mathcal{X} \delta^2(\vec{r}) + \sum_{i, j \neq i} \mathcal{X}_{ij} e^{-isa^{i \rightarrow j}} \delta^2(\vec{r} + \vec{r}_j) \quad (4.7)$$

¹The boresight direction corresponds to the satellite axis exhibiting the highest gain.

Based on this, the fields transform under crosstalk in the following way:

$$T'(\ell) = {}_0P(\ell)T(\ell) \quad (4.8a)$$

$$(Q' + iU')(\ell) = {}_2P(\ell)(Q + iU)(\ell) \quad (4.8b)$$

where T is the temperature field intensity, Q and U are the Stokes parameter combined to obtain the polarization fields E and B (see Chapter 1).

The amplitude of fluctuations, denotes here as $\mathcal{A}(\ell)$, transforms as:

$$\mathcal{A}'(\ell) = \begin{pmatrix} {}_0P(\ell) & 0 & 0 \\ 0 & \text{Re}({}_2P(\ell)) & -\text{Im}({}_2P(\ell)) \\ 0 & \text{Im}({}_2P(\ell)) & \text{Re}({}_2P(\ell)) \end{pmatrix} \mathcal{A}(\ell) = [\mathcal{X} + \sum_{i,j \neq i} \mathcal{X}_{ij} \mathcal{J}_{ij}(\ell)] \mathcal{A}(\ell) \quad (4.9)$$

where

$$\mathcal{J}_\ell(\vec{r} \cdot \ell, a) = \begin{bmatrix} e^{i\vec{r} \cdot \ell} & 0 & 0 \\ 0 & \cos r \cdot \ell - 2a & -\sin r \cdot \ell - 2a \\ 0 & \sin r \cdot \ell - 2a & \cos r \cdot \ell - 2a \end{bmatrix} \quad (4.10)$$

The term on the diagonal indicates the “auto-correlations” of the T, Q, U components. The nonzero term outside the diagonal refer to the combination of the Q and U elements.

Consequently, being $C_\ell \equiv \langle \mathcal{A}(\ell) \mathcal{A}^\dagger(\ell) \rangle$:

$$C'_\ell = \langle \mathcal{A}'(\ell)^\dagger \mathcal{A}'(\ell) \rangle = \mathcal{X}^2 C_\ell + \mathcal{X} \sum_{i,j \neq i} \mathcal{X}_{ij} \langle \mathcal{J}_{ij}(\ell) C_\ell + h.c. \rangle \quad (4.11)$$

where $h.c.$ stands for the Hermitian conjugate and the prime indicates the elements affected by crosstalk.

Moving from the flat-sky approximation to a curved sky with isotropic orientation and map coverage, the point spread function must be replaced by the spherical harmonics, as described in Chapter 1.

Introducing the rotation of the detector orientation we can rewrite:

$${}_s\mathcal{Y}_{\ell m}(\phi, \theta, \psi) \equiv e^{-is\psi} {}_sY_{\ell m}(\theta, \phi) \quad (4.12)$$

where ψ is the polarization angle of the detector and ${}_sY_{\ell m}(\theta, \phi)$ are the spherical harmonics for the order s field. The effect of crosstalk reflects on the amplitude

coefficients $a_{\ell m}$ in the following way.

$${}_0a'_{\ell m} \equiv {}_0(T * {}_0P)_{\ell m} = {}_0a_{\ell m}(\mathcal{X} + \sum_{i,j \neq i} \mathcal{X}_{ij} \mathcal{P}_\ell(\cos \theta^{i \rightarrow j})) \quad (4.13a)$$

$${}_2a'_{\ell m} \equiv {}_2(\mathcal{Q} * {}_2P)_{\ell m} = {}_2a_{\ell m}(\mathcal{X} + \sum_{i,j \neq i} \mathcal{X}_{ij} d_{22}^\ell(\theta^{i \rightarrow j}) e^{-2ia^{i \rightarrow j}}) \quad (4.13b)$$

$$-{}_2a'_{\ell m} = -{}_2a_{\ell m}(\mathcal{X} + \sum_{i,j \neq i} \mathcal{X}_{ij} d_{22}^\ell(\theta^{i \rightarrow j}) e^{2ia^{i \rightarrow j}}) \quad (4.13c)$$

where d_{22} is an element of the Wigner matrix and $\mathcal{P}_\ell(\cos \theta)$ are the Legendre polynomials.

Introducing the modified matrix $\mathcal{J}_\ell$ the C_ℓ coefficients become:

$$\mathcal{J}_\ell(\theta, a) = \begin{bmatrix} \mathcal{P}_\ell(\cos \theta) & 0 & 0 \\ 0 & d_{22}^\ell(\theta) \cos 2a & d_{22}^\ell(\theta) \sin 2a \\ 0 & -d_{22}^\ell(\theta) \sin 2a & d_{22}^\ell(\theta) \cos 2a \end{bmatrix} \quad (4.14)$$

$$C'_\ell = \mathcal{X}^2 C_\ell + [\mathcal{X} \sum_{i,j \neq i} \mathcal{X}_{ij} \mathcal{J}_{ij,\ell} C_\ell] + \sum_{i,j \neq i, m, n \neq m} \mathcal{J}_{ij,\ell} C_\ell \mathcal{J}_{mn,\ell} \quad (4.15)$$

Starting from this, the terms (C_ℓ^{TT} , C_ℓ^{EE} , C_ℓ^{BB} , C_ℓ^{TE} , C_ℓ^{TB} , C_ℓ^{EB}) with different values of θ (distance on the sky between two detectors) and ψ (the angle between the polarization of the two detectors) can be calculated. If the detectors belong to the same *pixel*, $\theta = 0$.

In Chapter 2 the layout of the focal planes of LiteBIRD was described. As mentioned, the FPSA will host multichroic detectors, each one connected to a bandpass lumped filter to select the frequency channel and a polarization sensitive sinuous antenna to define the polarization direction of the signal. The sinuous antennas are a combination of spiral and log-periodic antennas, and present a structure with four arms, with a relative angle of 90 degrees with each other. This kind of antennas present a large bandwidth, thank to their geometry: the parts closest to the center detect short wavelengths, the parts toward the edge are sensitive to large wavelengths. The handedness determines whether the sinuous antenna is right-handed or left-handed and is fundamental for detecting circular polarized radiation and minimize polarization losses. In Figure 4.1 the antennas design is shown: the disposition (Q, U) and the handedness (A, B) are highlighted. These details can also be observed in Figure 4.4.

A key aspect in the design of this part of the telescope concerns the routing of connections from the detectors to the cold readout. The order in which the TESs

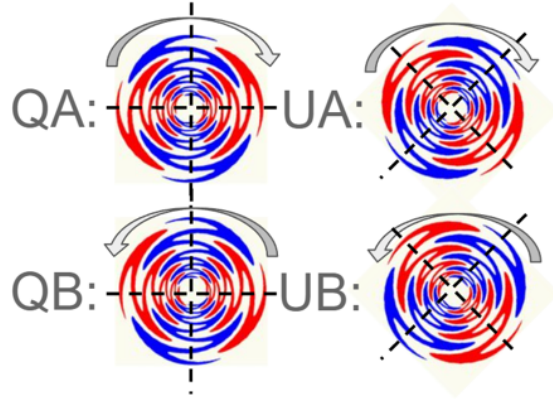


Figure 4.1: Orientation (Q, U) and handedness (A, B) of sinusoidal antennas [16].

are brought to the chip edge for connection to the readout chain is optimized to minimize wiring obstructions and to ensure a proper distribution of TESs among the SQUIDS.

The key question to be addressed is: which LC resonator should be connected to which TES? This assignment follows a set of rules developed to minimize crosstalk, starting from eq. 4.15.

Assuming to have detectors located in the same pixel ($\theta = 0$), the impact on the power spectra was evaluated for $\psi = 45^\circ$ and $\psi = 90^\circ$ [44]. First, it was simulated for two detectors and then for the entire focal plane sub-assembly, by changing configurations and studying the variation on the level of impact of crosstalk on the maps. The complete results can be found in [44].

The crosstalk matrix is structured as follows:

$$X = \begin{bmatrix} 1 & 2x \\ 2x & 1 \end{bmatrix} \quad (4.16)$$

where x is the crosstalk component between the two multiplexing legs. This is justified by the fact that the transformed matrix $\mathcal{X}_{ij}$ has trace equal to 1. As can be observed in Table 4.1, considering only first-order crosstalk terms (i.e. linear in x , excluding x^2 contributions), the impact on the power spectra at $a = 45^\circ$ is less significant, since $C_\ell^{EE} \gg C_\ell^{BB}$. C_ℓ^{TT} doesn't depend on a .

From these simulations, the following rules were set:

1. frequency band and pixel partners need to be neighbors in multiplexing frequency space
2. frequency bands need to be maximally separated in multiplexing frequency

$a = 45^\circ$	$a = 90^\circ$
$C_\ell'^{EE} = x C_\ell^{BB}$	$C_\ell'^{EE} = C_\ell^{EE} (1 - x)^2$
$C_\ell'^{BB} = C_\ell^{BB} + x^2 C_\ell^{EE}$	$C_\ell'^{BB} = C_\ell^{BB} (1 - x)^2$
$C_\ell'^{TE} = C_\ell^{TE} \left(\frac{x}{\pi} + 1 \right)$	$C_\ell'^{TE} = C_\ell^{TE} (1 - x^2)$
$C_\ell'^{TB} = -x C_\ell^{TE} (x + 1)$	$C_\ell'^{TB} = 0$
$C_\ell'^{EB} = x (C_\ell^{BB} - C_\ell^{EE})$	$C_\ell'^{EB} = 0$

Table 4.1: The power spectra components including crosstalk (x) and how the $C_\ell'^{XX}$ depend on the components without crosstalk.

space

- neighbor detectors in different pixels need to have a polarization angle difference of 45 or 135 degrees

We add a forth rule to minimize the inductive coupling among the coil inductors:

- on the LC chip layout, multiplexing frequency neighbor resonators should be separated in space

Each cluster of TESs is connected to a SQUID. Consequently, every SQUID on the same focal plane is associated with a different sequence of detectors. At the same time, a series of LC filters must be designed and assigned to the TESs. The assignment of the multiplexing frequencies to the detectors is optimized to minimize electrical crosstalk. For calibration purposes, optical detectors are accompanied by dark detectors (without antennas) and calibration resistors. Starting from a given sequence of TESs and a set of LC filters with different resonance frequencies, the goal is to determine the optimal pairing between each TES and its corresponding LC filter.

We developed a method and implemented a dedicated code to perform this task in the most accurate and efficient way. The main steps are:

- A random multiplexing frequency sequence is generated in a fixed range: for LiteBIRD 1.6-5.5 MHz
- Each multiplexing frequency is assigned to the optical and dark detectors, as well as to the calibration resistors, following the fixed order of the detectors and the randomly generated frequency sequence
- A “cost function” is defined: differently weighted “penalties” are assigned when the rules mentioned above are violated. The relative weight of the penalties reflects the assumed impact of each violation on the crosstalk level and is

intended to simulate the behavior of the algorithm when different weights are applied.

- penalty=500 when the band separation rule is violated
- penalty=100 when band/pixel partners are not neighbors in multiplexing frequency space
- penalty=10 when different pixel neighbors share the same polarization
- penalty=5 when multiplexing frequency neighbors are close in space
- The multiplexing frequency sequence, calibration resistor position, and selected dark bolometer are iteratively varied, with the cost function recalculated at each step.

This method is based on the so called *simulated annealing* algorithm, used to approximate the global minimum of a function. The name is inspired to the metallurgic process of heating a metal up and leaving it to slowly cool to reach a more stable structure with less energy.

Simulating all possible configurations would be prohibitively time consuming and computationally expensive. Thanks to this algorithm however, the probability of finding a high cost configuration (which means a configuration with many penalties applied) decreases as the number of iteration increases. Moreover, the goal is to avoid to fall into local minima and not converge to a better result.

In brief, the simulating annealing algorithm is performed by setting a number of steps N , the starting temperature $T_{initial}$ and the final temperature T_{final} , defining also the initial configuration and cost. At each iteration, the cost of a new configuration is evaluated and compared with that of the current configuration. If the new cost is lower, the configuration is accepted; if it is higher, it may still be accepted with a probability given by eq. 4.17:

$$P = e^{-\frac{\Delta C}{T}} \quad (4.17)$$

where C represents the cost and ΔC is the difference between the new and current costs. Allowing the acceptance of higher-cost configurations helps the algorithm to escape local minima.

The temperature T is gradually reduced at each iteration according to the factor $(T_{final}/T_{initial})^{1/N}$. As the temperature decreases, the probability of accepting higher-cost configurations diminishes, following eq. 4.17. The process continues until the predefined number of iterations is reached, at which point the configuration

with the lowest cost encountered is selected as the final solution.

The cost is determined by evaluating the current configuration and adding any applicable penalties. At each step, a random number n between 0 and 1 is generated, and the multiplexing frequency sequence is modified according to the following rules:

- $0 < n < 0.25$: two elements in the multiplexing frequency sequence are swapped.
- $0.25 < n < 0.5$: a subset of the sequence is shuffled.
- $0.5 < n < 0.75$: the multiplexing frequencies assigned to the calibration resistors are updated.
- $0.75 < n < 1$: if the TES sequence contains more than two dark bolometers, the subset of dark bolometers used is adjusted.

In summary, this routine produces a sequence of multiplexing frequencies designed to complement the TES ordering. This frequency map forms the basis for the placement and wiring strategy on the LC filter chip. To identify the most efficient configuration, the annealing algorithm is applied.

4.1.1 The algorithm applied to LF12

The LF12 FPSA is a matrix of 3×3 thrice dual polarization pixels and will host a total of 72 detectors divided in two groups: LF1 ([40, 60, 78] GHz) and LF2 ([50, 68, 89] GHz). Of these, 54 will be optical, i.e. connected to sinuous antennas, and are organized into 9 pixels².

The SQUID mapping for LF12 is arranged as shown in Figure 4.3. Pixels 0, 3, 6, 7 are connected to SQUID 0, pixels 1, 2, 4, 5, 8 to SQUID 1. Figures (a) and (b) in 4.4 show details of the TES chip layout [25]. Panel (a) shows the pixels arranged in 3 rows and 3 columns, with bondpads lines visible along the top and bottom edges of the chip. A design of the sinuous antenna is placed at the center of the pixels. Panel (b) presents a zoomed view of a single pixel. The top shows part of the sinuous antenna, whereas the TESs, the microstrip lines connecting the TESs to the bandpass filters, and the necessary crossovers are depicted along the sides.

In the case of LF12, the multiplexing frequency scheduling was performed to have

²At the time of writing, LiteBIRD is under rescope studies to consolidate the mission's feasibility with the same scientific objectives. The rescope intends to simplify the mission, reduce the cost and improve manageability.

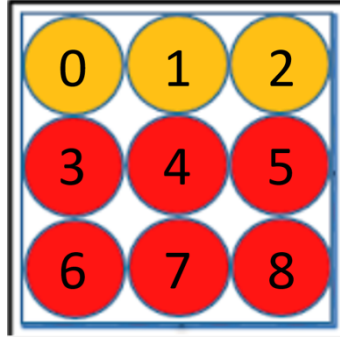


Figure 4.2: LF12 design: yellow circles represent the LF1 pixels, the red ones are the LF2 pixels.

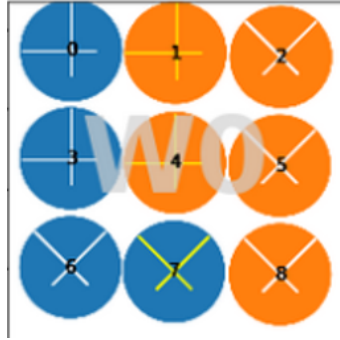


Figure 4.3: SQUID mapping for LF12 FPSA. Blue pixels are connected to SQUID 0, orange pixels to SQUID 1.

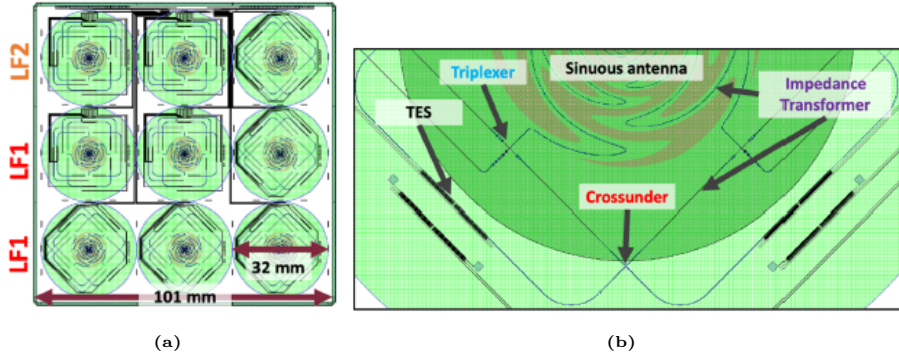


Figure 4.4: Detailed TESs chip design for LF12, see [25].

the same LC filters configuration for both SQUID 0 and SQUID 1. As mentioned, the pixels and TESs connected to the two SQUIDs differ in both frequency channels and polarization antennas. Moreover, the number of optical detectors is not the same: 24 for SQUID 0 and 30 for SQUID 1. To maintain a multiplexing factor equal to 34 for both, and accommodate two calibration resistors, 8 dark bolometers are included for SQUID 0, and 2 for SQUID 1. The number of dark bolometers and calibration resistors per SQUID was determined based on internal analyses of data handling and the expected on-board to ground data rate.

Consequently, the algorithm is required to identify a solution that simultaneously

minimizes the penalty functions and maximally satisfies the rules for both SQUID 0 and SQUID 1. This complicates the calculations but, at the same time, the simulated annealing method allows the code to converge even with a larger number of requirements.

Considering the multiplexing factor and the 1.6–5.5 MHz frequency range, the LC sequence that optimally satisfies the design requirements for LF12 is: [9, 7, 13, 15, 23, 21, 6, 11, 20, 25, 17, 12, 27, 3, 1, 31, 29, 18, 33, 28, 0, 5, 19, 16, 14, 24, 22, 8, 10, 26, 30, 32, 4, 2]. The numbers indicate the frequency ranks in ascending order: 0 corresponds to the lowest frequency (1.6 MHz), and 33 to the highest (5.5 MHz). The sequence matches the TES detector ordering, i.e., the first frequency value (9) is assigned to the first detector, and the last (2) to the last detector. The shape of the LC filters matrix was chosen to be 5×7 with an empty spot. A schematic representation is shown in Figure 4.5. Starting from this sequence we can proceed

bondpads side						
9	21	17	31	0	24	
7	6	12	29	5	22	30
13	11	27	18	19	8	32
15	20	3	33	16	10	4
23	25	1	28	14	26	2
bondpads side						

Figure 4.5: Scheme of the position of multiplexing frequencies on the LC chip. The sequence preceeds from top to bottom, from left to right.

by designing an LC chip for LF12 by placing the “real” filters in the scheme.

4.2 A LC filters chip design for LF12

The LF12 LC chip was designed using the Python *gdsapy* package, which allows the direct generation of a *gds* file from the design script. Some useful functions allow to draw custom sized components, the wiring and all the necessary components.

The resulting chip is sized 5×5 cm and, as mentioned, hosts 34 LC filters, properly sized to reach the requested frequency range.

The frequency range was chosen to meet the LiteBIRD requirements and to investigate the impact of the frequency sequence on electrical crosstalk. The multiplexing frequencies span from 1.6 MHz to 5.5 MHz. Two sequences were used to design the chips: one with linear spacing and one with a hybrid frequency arrangement, in

which the first half of the resonators are linearly spaced and the second half logarithmically spaced. This configuration provides larger separations for the highest-frequency resonators, which is recommended since higher frequencies generally have a greater impact on electrical crosstalk.

The inductance value is the same for each LC filter and the coil inductors are shaped to match $L = 60 \mu\text{H}$. This value is higher than in previous experiments from which LiteBIRD inherits its design and matches the normal resistance of the TESs $R_N \sim 1 \Omega$. This choice is made to reduce the bandwidth of each leg, defined by $BW = R/(2\pi L)$, resulting in $BW \sim 2 \text{ kHz}$. The companions are interdigitated capacitors, previously described in Chapter 3.

Setting the inductance value and the multiplexing frequency range, the capacitance values of the filters will be calculated accordingly to eq. 3.6 and the capacitors will properly shaped to match these quantities. As mentioned, the advantage of having lithographed capacitors stands in the possibility of reaching perfectly tuned capacitance and, consequently, resonance frequency values. The capacitance values result to be in the range (155.79, 13.13) pF.

In Tables 4.2 and 4.3 the values of the resonance frequencies for both designs are reported. Figure 4.6 presents a simulation of the current as a function of the frequency sequences in the linear and hybrid arrangements.

The inductors (Fig. 4.10-(b)) have a square spiral design and consist of 154 turns,

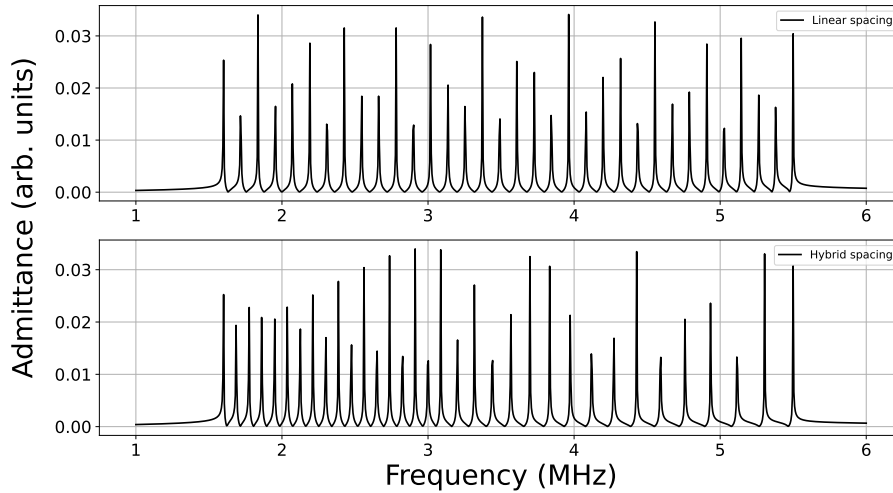


Figure 4.6: Simulated admittance for the frequency sequences with linear and hybrid spacing.

with a line width and spacing of $4 \mu\text{m}$ between adjacent coil lines. One end of each inductor is connected to a capacitor, while the other is routed to the bonding pads. The capacitors are interdigitated, with finger widths and spacing of $4 \mu\text{m}$, as shown

	Res freq (MHz)		Res freq (MHz)
1	1.60	18	3.61
2	1.72	19	3.73
3	1.84	20	3.85
4	1.95	21	3.96
5	2.07	22	4.08
6	2.19	23	4.20
7	2.31	24	4.32
8	2.43	25	4.44
9	2.55	26	4.55
10	2.66	27	4.67
11	2.78	28	4.79
12	2.90	29	4.91
13	3.02	30	5.03
14	3.14	31	5.15
15	3.25	32	5.26
16	3.37	33	5.38
17	3.49	34	5.50

Table 4.2: Resonance frequency and capacitance values for the LC filters in the designed chip.

in detail in Fig. 4.10-(c) and (d). The bonding pads are $200\ \mu\text{m}$ wide, and the interconnecting micro-strips has a thickness of $10\ \mu\text{m}$. Figure 4.10 also shows the final design for the complete LC chip with linear spacing.

The level of the electrical crosstalk is calculated for the three contributors as described in Chapter 3. The electrical crosstalk level is calculated by taking into account all phase components and using the magnitude of the detector currents. The results are presented in Figure 4.7 for the linear spacing, in Figure 4.8 for hybrid spacing. In both cases the dominant component is the common impedance crosstalk, with maximum values for the higher frequency filters around 1.5×10^{-2} . Some elements of the mutual inductance crosstalk matrix also exhibit relatively high levels of crosstalk, due to the spatial proximity of detectors close in the multiplexing frequency domain, and peaks around 2×10^{-3} . A difference can also be observed by comparing the average crosstalk values of the two components: it is approximately 10^{-5} for mutual inductance crosstalk and 10^{-3} for common impedance crosstalk, indicating a general higher level for the latter. As expected, carrier leakage crosstalk

	Res freq (MHz)		Res freq (MHz)
1	1.6	18	3.09
2	1.69	19	3.2
3	1.78	20	3.32
4	1.86	21	3.44
5	1.95	22	3.57
6	2.04	23	3.7
7	2.12	24	3.83
8	2.21	25	3.97
9	2.3	26	4.12
10	2.39	27	4.27
11	2.48	28	4.43
12	2.56	29	4.59
13	2.65	30	4.76
14	2.74	31	4.94
15	2.82	32	5.12
16	2.91	33	5.31
17	3.0	34	5.5

Table 4.3: Design resonance frequencies for an hybrid configuration: first half linearly spaced, second half logarithmically spaced.

is the least significant component.

The three plots in Figure 4.9 show the ratio of the crosstalk of the hybrid configuration to that in the linear configurations. Mutual inductance crosstalk (reported in logarithmic scale for clarity) exhibits significant variation due to changes in the shapes of the LC filters and their relative position, which depend on both geometry and multiplexing frequency. However, this variation does not show a clear dependence on frequency. In contrast, when shifting from the linear to the hybrid configuration, both carrier leakage and common impedance crosstalk increase at lower frequencies and decrease at higher ones. This mainly depends on the fact that the hybrid sequence, while increasing the spacing at higher frequencies, exhibits a reduction at lower ones. This case will be of interest for future measurements with this chip to determine whether the experimental results reproduce this pattern.

The complete design of the wafer is shown in Figure 4.11. Each wafer contains two

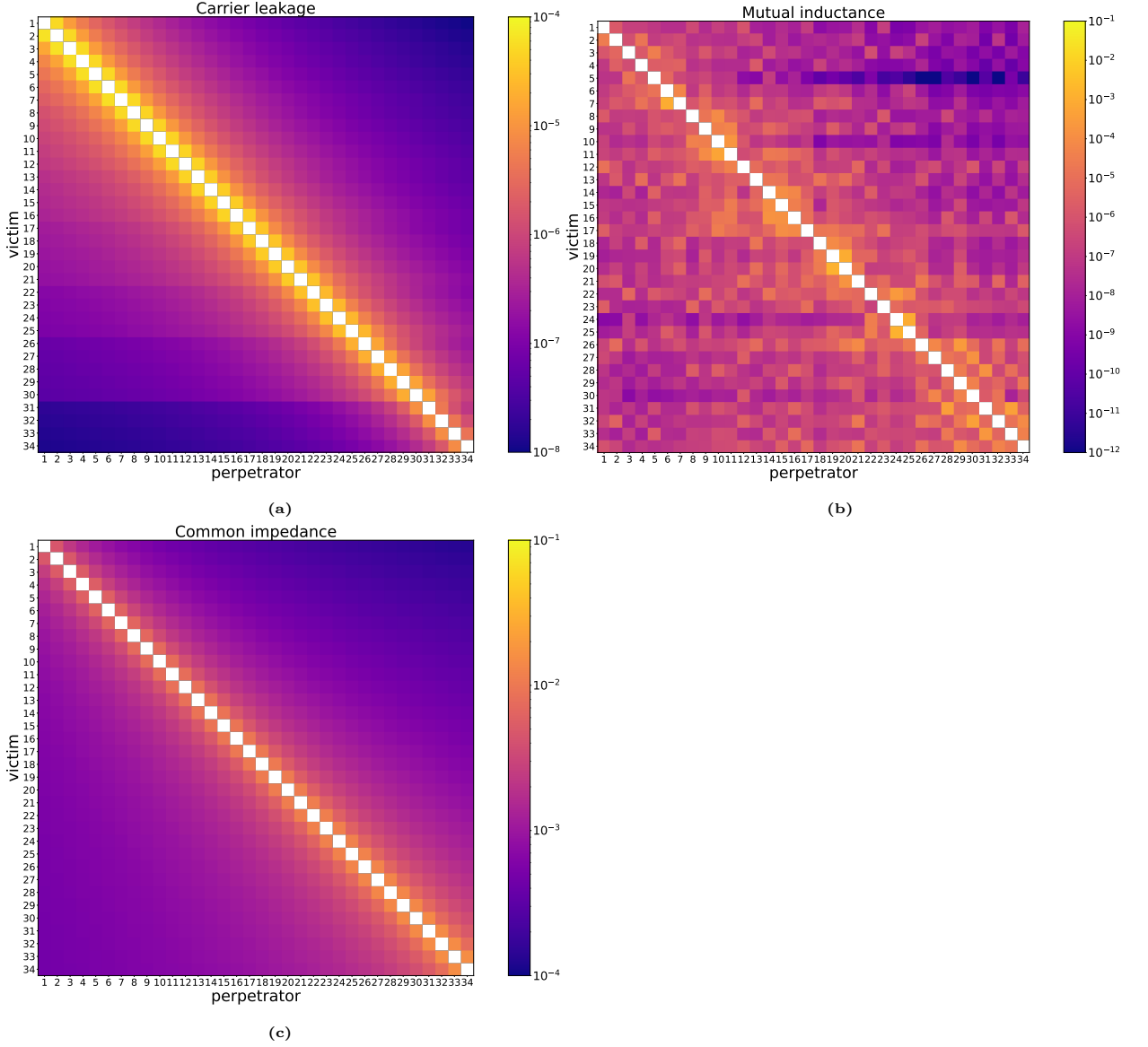


Figure 4.7: Crosstalk matrices for the linear frequency sequence. (a) Carrier leakage crosstalk, (b) mutual inductance crosstalk, (c) common impedance crosstalk.

linear LC chips and two *hybrid* LC chips, which appear as the four square patterns at the center of the wafer. Additional structures are placed along the wafer edges; these are specifically designed and tuned to investigate how crosstalk varies with the distance and orientation of the LC filters. Three wafers were fabricated based on this design. Two large view pictures of one of the delivered wafers are shown in Figure 4.12

The wafers were manufactured at the Lawrence Berkeley National Laboratory (LBNL) based on the design presented in this chapter. The entire design was patterned on a 6-inch silicon (Si) wafer using direct-write photolithography. The LC filters and

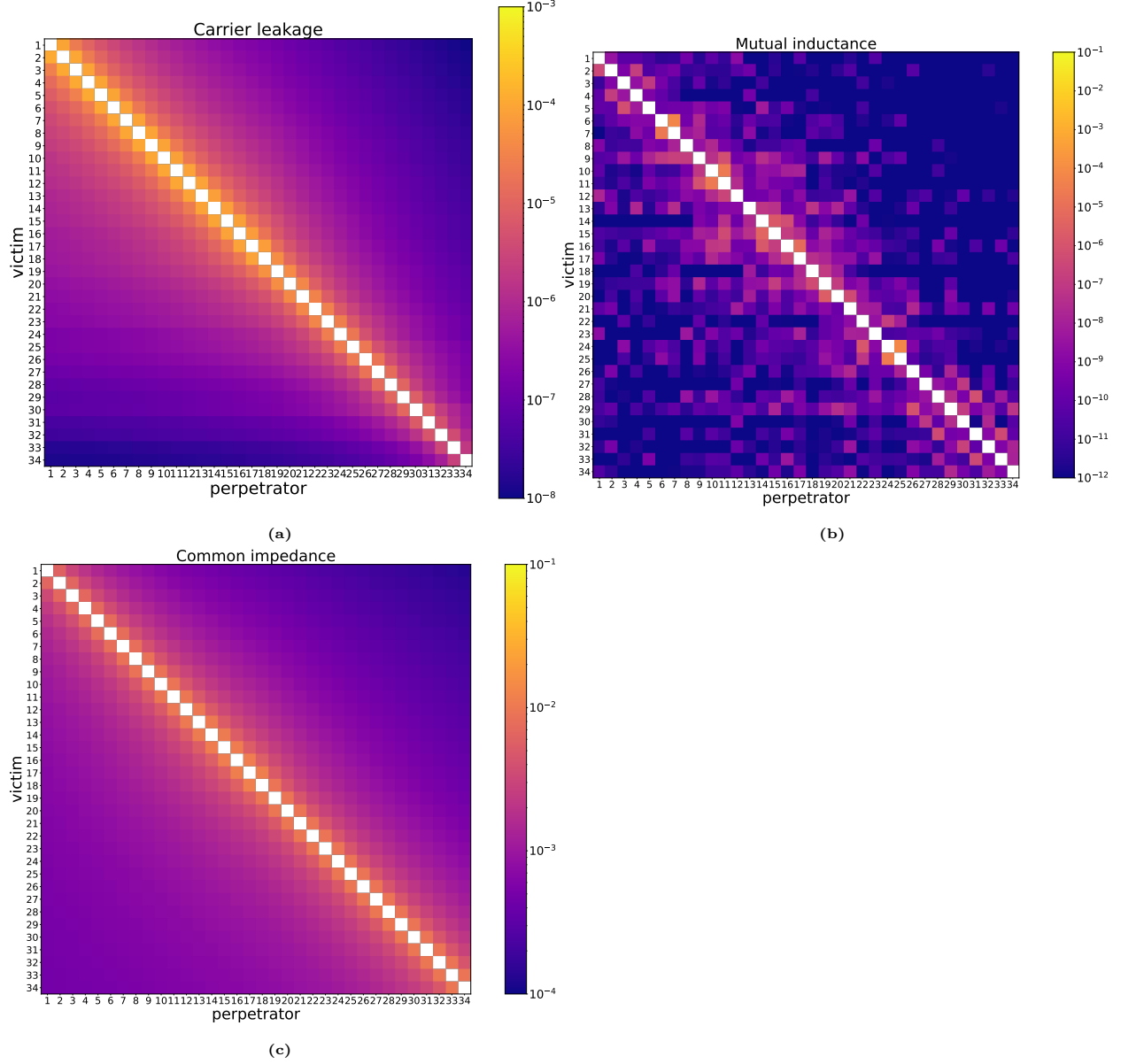


Figure 4.8: Crosstalk matrices for the linear frequency sequence. (a) Carrier leakage crosstalk, (b) mutual inductance crosstalk, (c) common impedance crosstalk. Hybrid spacing.

microstrips are made of niobium.

4.3 Conclusions

This chapter has outlined the motivation behind the multiplexing scheduling and the rules for minimizing electrical crosstalk. The analytical calculations of the power spectra were presented, with emphasis on the crosstalk contribution. The algorithm developed to define the multiplexing frequency sequence and the LCs–TESs assign-

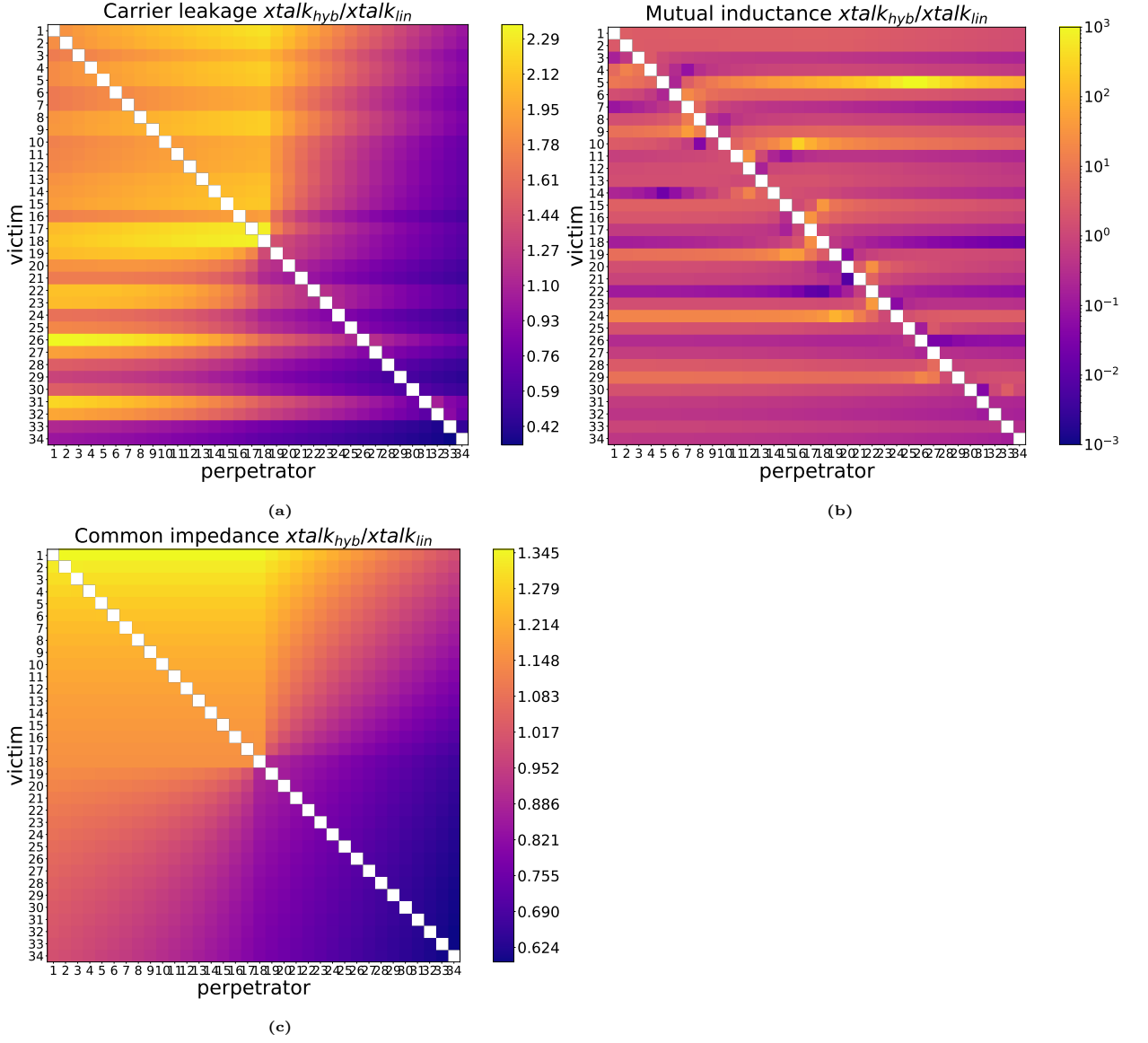


Figure 4.9: Ratio of hybrid crosstalk to linear crosstalk. A linear scale is used for (a) and (c). A logarithmic scale is used for (b).

ment was described, and two alternative LC chip designs were described and presented. The following chapters will focus on measurements and the characterization of LC chip properties.

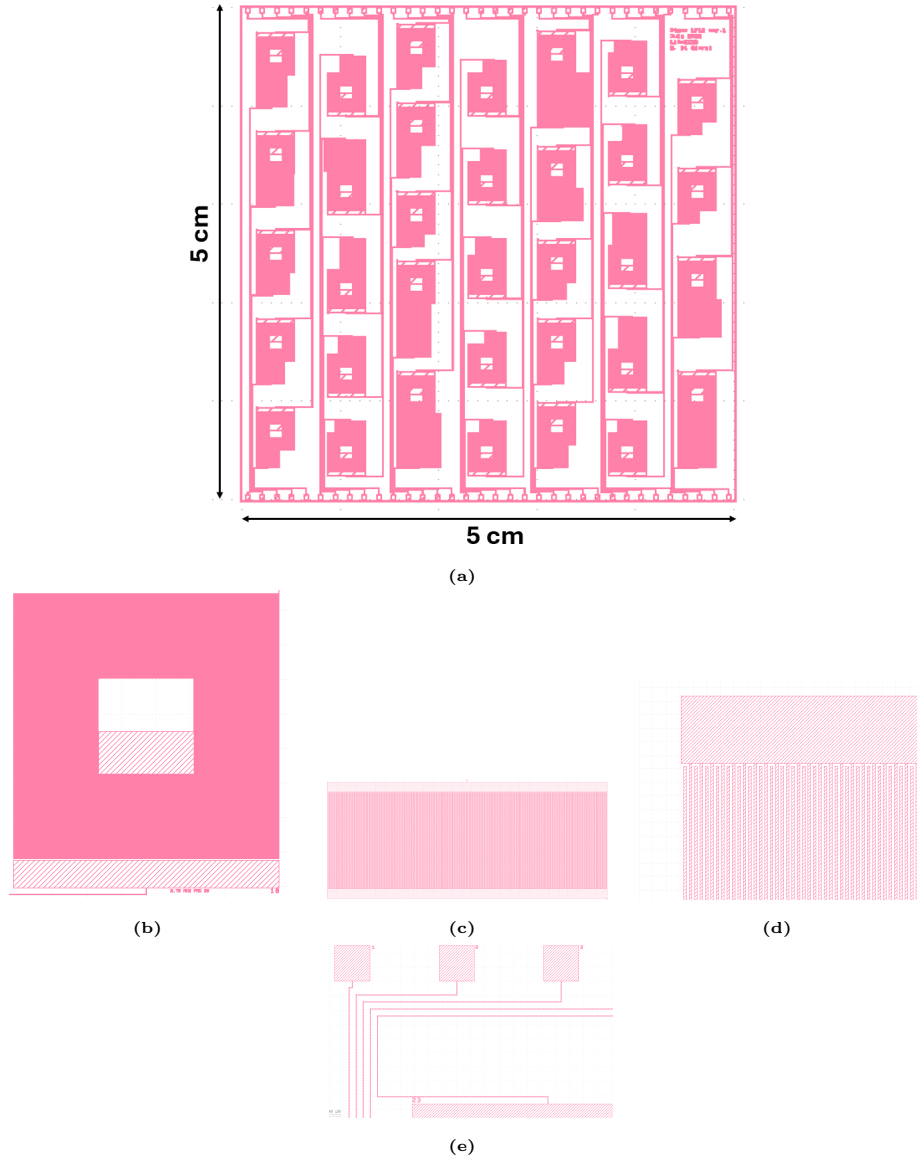


Figure 4.10: Complete final design for the LF12 LC chip (a). Multiplexing factor is equal to 34. Details in the following figures: (b) inductor, (c) capacitor, (d) detail of the capacitor fingers, (e) wiring and pads.

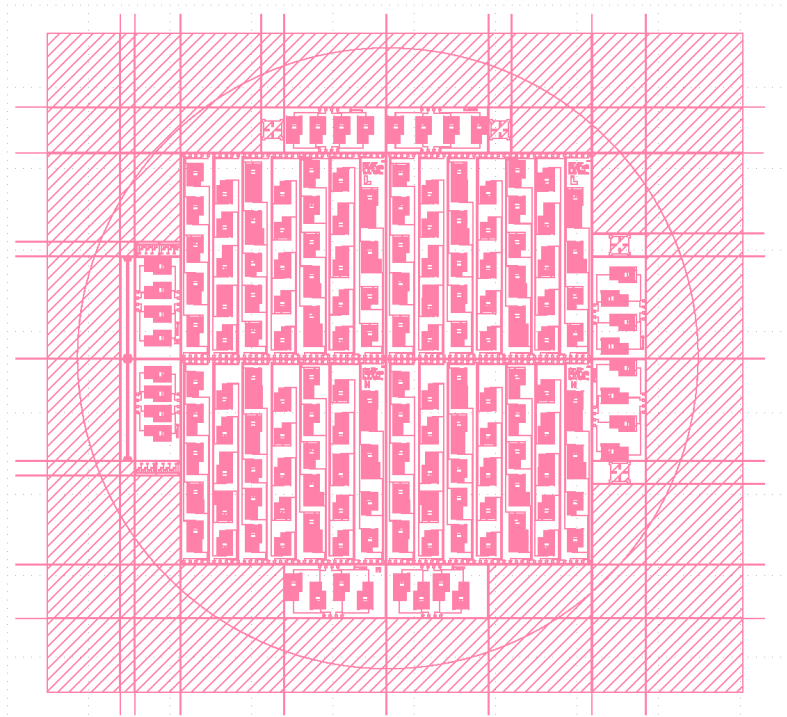
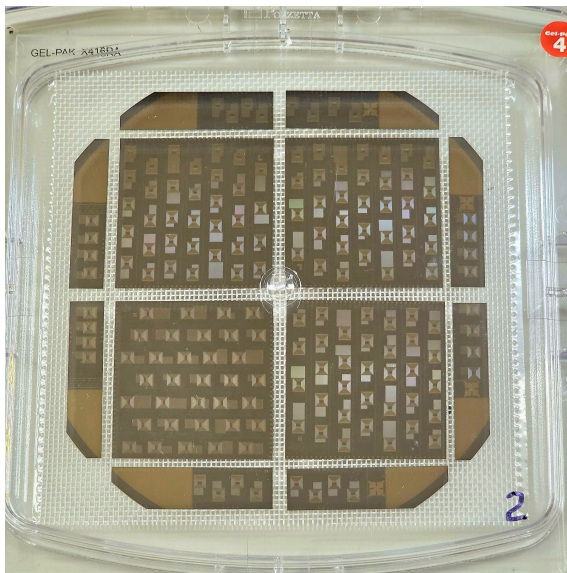


Figure 4.11: Complete wafer design. At the center, two copies of the 34 LC filters with hybrid and linear configurations are placed. Along the edges, groups of four LC filters with different orientations and spacings are arranged. Resistance-testing crosses and test microstrips are also included.



(a)



(b)

Figure 4.12: Pictures of the entire 6 inches wafer taken after manufacturing.

Chapter 5

Characterization and crosstalk measurements for an LC filters chip with parallel capacitors

In this chapter, the results of the measurements conducted at SRON (Space Research Organization Netherlands) on a LC chip entirely fabricated by lithography in the same facility are presented. The work includes device testing, characterization, and, finally, the evaluation of electrical crosstalk.

5.1 A lithographed LC chip for the LSPE/SWIPE experiment

These measurements were performed in the context of the LSPE-SWIPE experiment (Large Scale Polarization Explorer/Short-Wavelength Instrument for the Polarization Explorer) [45], a balloon-borne experiment expected to be operated in the stratosphere. The main scientific goal is to observe and measure the polarization B-modes of the CMB at a level corresponding to a tensor-to-scalar ratio $r = 0.03$ at 99.7% CL. Alternatively, the objective is to set an upper limit $r < 0.015$ at 95% CL combining observations from ground and data from *Planck* and QUIJOTE [46]. The detection and readout chain of SWIPE will be very similar to the one described for LiteBIRD. Superconducting TESs connected to SQUIDs will be used. In particular the two 90 degrees focal planes shown in Figure 5.1 will be equipped with 326 detectors read out by 24 SQUIDs with Frequency Division Multiplexing. Despite having fewer detectors than LiteBIRD, SWIPE achieves comparable sensitivity by

employing *multi-mode* detectors, which allow about 8800 radiation modes to be collected by a limited number of bolometers [45].

The FDM is performed with an hybrid LC filters configuration, composed of ce-

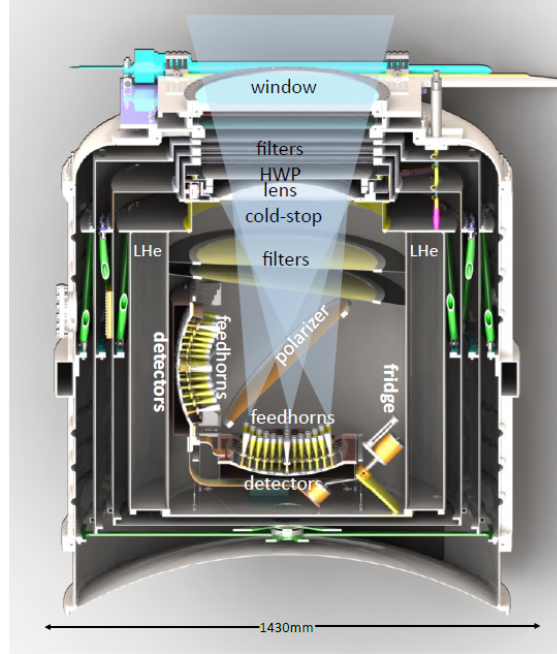


Figure 5.1: The payload of LSPE/SWIPE. The two 90 degrees focal plane are shown, together with the polarizer selecting the linear polarization and the complete optical chain. This includes an Half Wave Plate to rotate the polarization and shift the frequencies above the problematic $1/f$ noise level.

ramic capacitors and lithographed inductors. In order to reduce parasitics, a new kind of filters, entirely fabricated by lithography, was commissioned by INFN Pisa and produced at SRON for comparison and is described in the following sections. This technology is well established at SRON and was originally proposed for Athena; the novelty lies in the frequency range, which is shifted towards lower values than usual, and the larger inductors, to match the different resistance of the TESs of SWIPE, which is expected to be $R \simeq 500 \text{ m}\Omega$.

5.2 The LC filters geometry

Unlike the LC filters we described in the previous sections, the SRON LC filters are produced coupling an inductor, composed of 4 spirals wound in opposite directions, to a parallel plane capacitor, as described in [41]. This design minimizes the inductive coupling between the coils. The inductors are designed for an inductance value of $L \simeq 15 \text{ }\mu\text{H}$. A total of 21 LC filters is present on the chip, with 18 of them

resonating in the range expected for SWIPE, between 0.3 and 2 MHz. Three additional LC filters are designed to resonate at higher frequencies, respectively around 3, 4 and 5 MHz. This procedure is performed to investigate systematic effects and frequency-dependent properties, and study its possible applications to LiteBIRD.

The most relevant difference with the LiteBIRD style design is the complexity of the layout. SRON LCs are based on a multi-layer process and need more than one iteration of lithography. The big advantage is that they can be packed into a much smaller space, since the capacitor shape requires less scaling to adjust the capacitance value, unlike interdigitated capacitors. For example, interdigitated capacitors coupled with $60 \mu\text{H}$ inductors, in order to achieve a resonance frequency around 1.6 MHz, typically have dimensions on the order of $5 \text{ mm} \times 5 \text{ mm}$ on Si, whereas parallel-plate capacitors with inductors of $15 \mu\text{H}$ can be reduced to about $420 \mu\text{m} \times 420 \mu\text{m}$ on SiC.

The capacitors are composed of NbTiN (niobium titanium nitride) electrodes and a SiC (silicon carbide) dielectric layer. The TESs and SQUIDs are voltage biased with the use of a SMD shunt resistor $R_{bias} \sim 750 \text{ m}\Omega$ connected to ground in parallel to the TESs. A second capacitor is designed on the LC chip to implement a capacitor divider and reduce the effective bias resistance down to $1.1 \text{ m}\Omega$. This additional capacitor is connected in parallel with the LC filter and grounded. As a result, the total impedance of each leg to ground is modified, leading to a smaller “effective” shunt resistance for the circuit.

A microscope picture of the lithography of two LC filters of similar design is shown in Figure 5.2 [41].



Figure 5.2: Two LC filters. On the left two pairs of spirals for two inductors are shown, on the right two capacitors (white) differently shaped to reach different capacitances. On the far right, bonding pads are visible.

5.3 The experimental setup

The instrumentation setup for this measurements run in the SRON laboratories is mainly composed of a dilution Leiden Cryogenics refrigerator, operated, in this specific case, at 70 mK. The chip is mounted on a copper holder, hosting also the SQUIDs, the transformers and the TESs. A detailed scheme of the components arrangement and a microscope image of the lithographed chips are presented in Figure 5.3.

The transformers are used to allow convenient tuning of the inductance value across

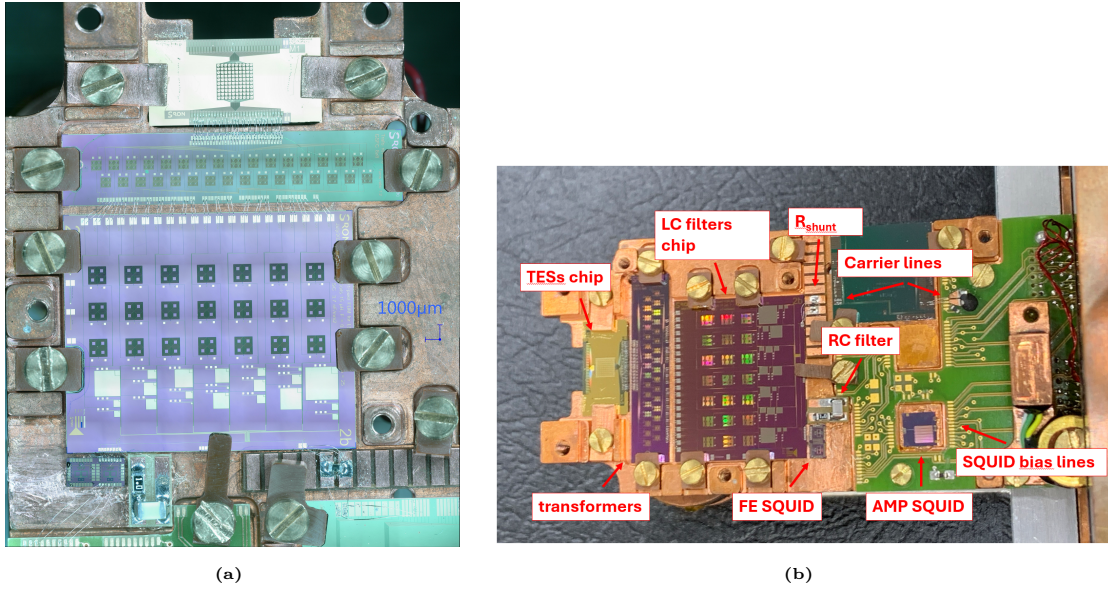


Figure 5.3: The holder arrangement. (a) A microscope picture of the lithographed components. From the bottom, the front-end SQUID (left) and the bias resistor (right). The LC chip with the inductors on top and the capacitors on the lowest part. Moving upward, the transformer chip and the TESs chip on top. Everything is kept in place with vespel clamps and thermalized on the copper surface. (b) An detailed scheme of the position of the elements on the holder.

different set-ups. The detectors used in this configuration are the Athena-like microcalorimeters, described in Chapter 2. They are employed here for convenience, as the purpose of this measurement is to characterize the cold readout chain rather than the detectors themselves. The SQUIDs are fabricated by *VTT Finland*, used in a dual stage configuration [47]. The *front-end SQUID* is directly glued on the copper interfacing the LC filters and is voltage biased. A RC filter, also known as *snubber*, is used to suppress the high frequency noise. The second one, called *amplification SQUID*, is current biased and provides the largest gain amplification. All the components are kept in place on the copper surface by *vespel* clips. Vespel is a polyimide-based plastic material with a good heat resistance and insulation. It

has the important role of securing the thermal contact of the chip and the holder. The components are enclosed in an aluminum box. The holder is suspended from the mixing chamber using Kevlar wires, which help in minimizing mechanical oscillations. Thermalization is ensured by means of copper strips.

The warm readout part employed for these measurements is a custom-made electronics [48]. At this stage, the input bias signals are generated, and an additional amplification of the output signal is eventually carried out.

The *base-band feedback* (BBFB) [49], [50] is employed. The BBFB is a technique used to increase the linearity and dynamic range of the SQUID by nulling the carrier signal. The TES signals are demodulated, re-modulated, and fed back through the SQUID feedback coil after compensating for phase shifts and correcting for delays. The main purpose of this procedure is to cancel the detectors' sinusoidal bias signal, by isolating the modulations caused by the sky signal and minimizing any residual phase errors. A complete block diagram is shown in Figure 5.4, from [51].

Based on this scheme, the biasing procedure of the SQUIDs can be illustrated. The

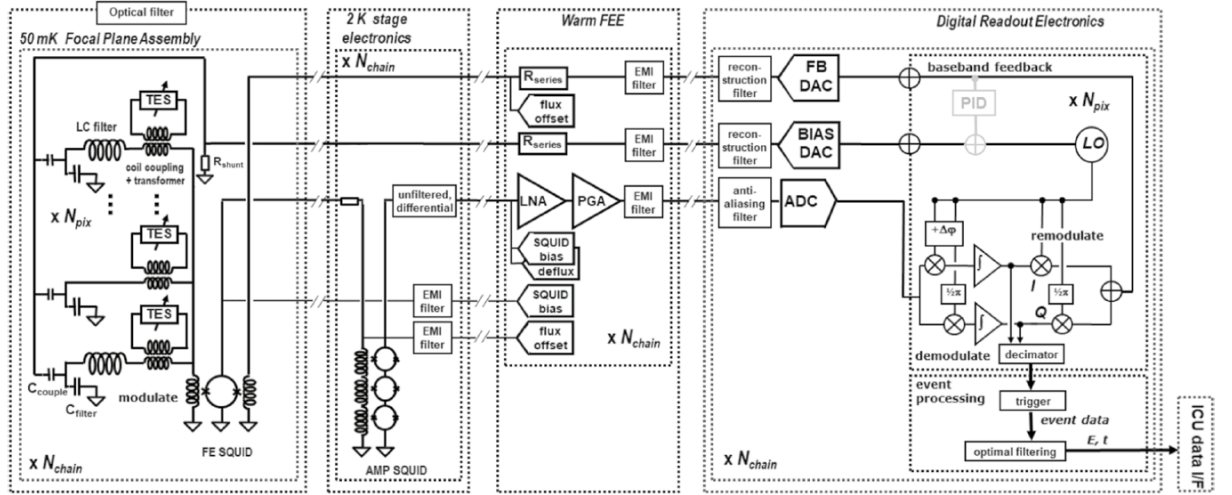


Figure 5.4: The circuitual scheme of the set-up. The cold part (TESs, multiplexing, SQUIDs) on the left, the warm part (FPGA, DACs and ADCs) on the right. [51],[49]. The front-end SQUID and the amplification SQUID are named respectively FE and AMP SQUID.

bias current $I_{\text{bias,AMP}}$ for the AMP SQUID, corresponding to the $V-\Phi$ curve with the largest peak-to-peak amplitude, is selected and the $I_{\text{flux,AMP}}$ is set to the value corresponding to the steepest slope of the $V-\Phi$ curve. Once the AMP SQUID is switched on, the FE SQUID bias current $I_{\text{bias,FE}}$ is determined and $I_{\text{flux,FE}}$ is adjusted according to the same procedure described above. The best values selected for the following measurements are presented in Table 5.1.

$I_{\text{bias,FE}}$	$I_{\text{flux,FE}}$	$I_{\text{bias,AMP}}$	$I_{\text{flux,AMP}}$
200 μA	8.8 μA	100 μA	14 μA

Table 5.1: SQUID settings.

5.4 Measurements

In the next sections, the measurements that were performed to test the LC chip properties and, as mentioned, the level of crosstalk will be presented. The measurements were performed using two configurations: without any connections between the LC filters and the TESs (the LC filters are connected to each other by wire-bondings, as shown in Figure 5.5), in order to investigate the intrinsic properties of the filters, and with the connection to the detectors and R_{shunt} restored, to enable comparison and the study of systematic effects. Moreover, the electrical crosstalk level was assessed in the presence of TESs.

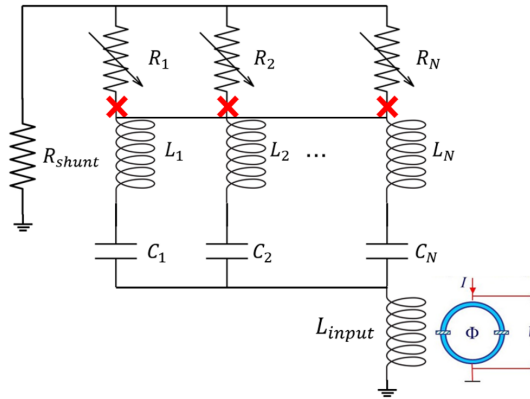


Figure 5.5: Configuration with TESs and R_{shunt} disconnected and LC filters lines shorted.

5.4.1 Resonance frequencies, quality factor and parasitic resistance

A highly effective way of illustrating the characteristics of a LC chip is to calculate the quality factor Q of the filters.

The Q -factor of a resonator gives an estimation of the efficacy in narrowing the bandwidth and indirectly of the electrical resistance of the circuit. It can be analytically calculated from eq. 5.1.

$$Q = \frac{\omega_{\text{res}}}{\Delta\omega} \quad (5.1)$$

where ω_{res} is the resonance angular frequency and $\Delta\omega$ is the bandwidth. Since the width of the resonance depends on the resistance as $\Delta\omega = R/L$, combining these two equations, we can retrieve the parasitic resistance level.

$$r_{par} = \omega_{res}L/Q \quad (5.2)$$

where L is the inductance of the leg.

This value specifically corresponds to the parasitic resistance; when the TESs are not part of the circuit, the response reflects only the intrinsic losses of the LC circuit. These values can be estimated through the measurements. In particular, the evaluation of the Power Spectral Density (PSD), which is a measurement of the noise level related to the SQUIDS outputs, is performed through the recording of the noise through the feedback coil. The results in the 0.2-6 MHz frequency range are reported in Figure 5.6. The plot displays more than the 21 expected resonance fre-

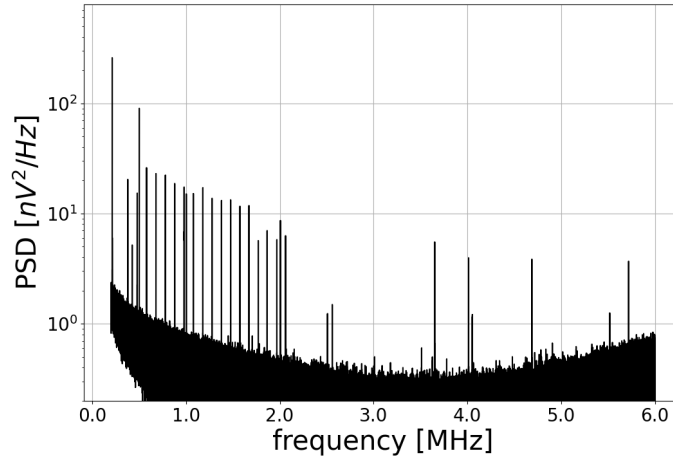


Figure 5.6: PSD for SQUID noise measurements in the resonance frequencies range. In this data the baseline was subtracted and the y axis data are represented in logarithmic scale.

quencies. The additional resonances are attributed to the warm electronics or to electromagnetic pick-up. The latter resonances exhibit narrower bandwidths, which is reflected in the corresponding Q -factors.

Figure 5.7 shows the Q -factor and the parasitic resistance values as a function of the resonance frequencies, calculated by performing a Lorentzian fit over the PSD data. The order of magnitude seems to be in line with previous works about FDM systems done in the same configuration [52].

As already mentioned, the spurious resonances (red dots in Figure 5.7) have higher Q -factor values and smaller parasitic resistance. The parasitic resistance r_{par} exhibits a frequency-dependent increase, consistent with the rise in eddy current losses.

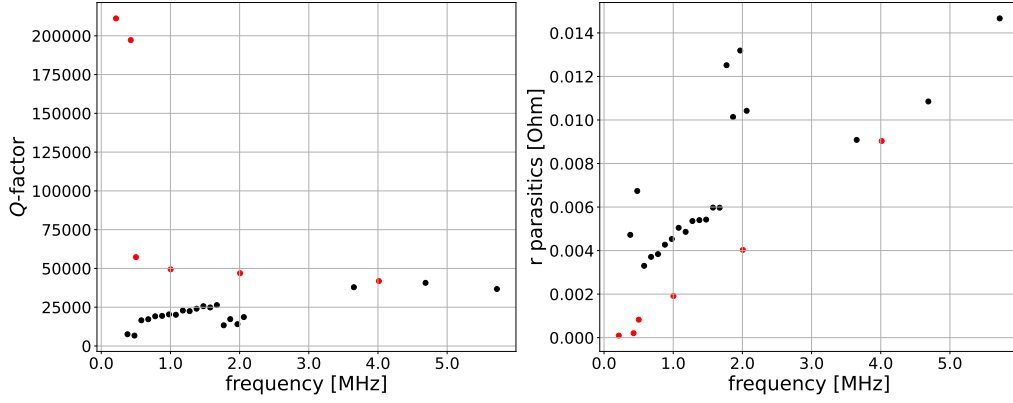


Figure 5.7: Left, The Q -factor calculated from the fit parameters. The black dots show the values related to the LC frequency resonances, the red dots refers to the warm electronics related resonances. Right, parasitic resistance in the LC filters circuit calculated from the fit parameters and Q factor values. Black dots are the values related to the LC resonance frequencies, red dots refers to the warm electronics related resonances.

Among the LC resonances values, four resonances near 2 MHz present lower Q and higher r_{par} , possibly due to the LC chip fabrication, and deserve further investigation.

5.4.2 Temperature scan

This measurement was repeated at different bath temperatures and with two SQUID bias settings (see Table 5.2), in the ranges 70–120 mK and 90–135 mK, respectively, to study the dependence of the LC filters properties from temperature.

We show exclusively the results related lowest resonance frequency (at $379.3394 \pm$

$T_b = 90 - 135$ mK			
$I_{bias,FE}$	$I_{flux,FE}$	$I_{bias,AMP}$	$I_{flux,AMP}$
350 μA	24 μA	100 μA	8 μA
$T_b = 70 - 120$ mK			
$I_{bias,FE}$	$I_{flux,FE}$	$I_{bias,AMP}$	$I_{flux,AMP}$
200 μA	0.002 μA	100 μA	11 μA

Table 5.2: SQUID settings and bath temperature range.

0.0002 kHz), since it is representative for the general behaviour.

A frequency shift is observed as the bath temperature T_b varies, accompanied by a change in the Q -factor. This behavior is attributed to a reduction in the filters capacitance, resulting from the temperature dependence of the dielectric permittivity ϵ_r [53]. In fact, the capacitance value is calculated as $C = \epsilon_r A/d$, where A is the

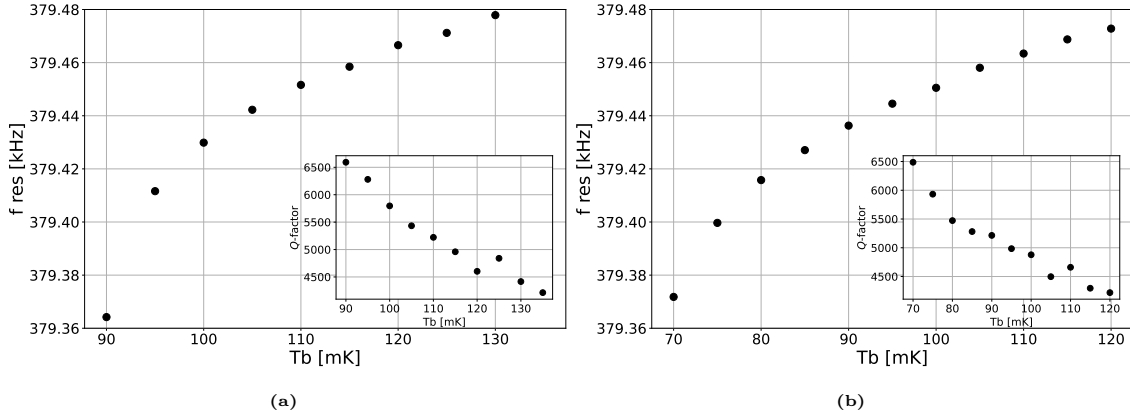


Figure 5.8: Resonance frequency and corresponding Q factor values shown as a function of the bath temperature. The plot on the left and right show respectively results obtained with SQUID high and low bias.

electrodes area and d the distance between them. As the temperature increases, the capacitance decreases, which in turn shifts the resonance frequency upward. In both the SQUID bias conditions, the resonance frequencies change by approximately 100 Hz.

These results have the same pattern of those reported by [52] for a similar LC chip at SRON, although in different frequency and temperature ranges.

5.4.3 Frequency scan and common inductance evaluation

The transmission coefficient in a two-port network S_{12} was characterized through a frequency sweep using a network analyzer, by injecting current signals at the resonance frequencies of the LC filters. The measurement was performed separately via the feedback line and the TES “carrier” line, through the R_{shunt} , covering the 0.1–6 MHz range, as illustrated in Figure 5.9.

The scan resolution was set to 2 Hz. Figure 5.9(a) and (b) present coarse scans through the R_{shunt} and TESs lines and the feedback lines. A fine sweep around the resonance frequencies was also performed.

The scan through the feedback coil shows that each resonance appears as a peak and a valley, as illustrated in Figure 5.10. The difference in the bandwidth of the resonances shown in the two plots arises from the TES side of the circuit being shorted in the left plot, which results in a narrower resonance, while the resistance of the TES increases its width. The shape showing a peak and a valley in the resonance originates from the presence and the specific configuration of the SQUID input coil and the wiring inductance, which gives rise to an additional contribution to the resonance.

From this analysis the quantity called *common inductance* can be retrieved [54]. [55], using Eq. 5.3.

$$L_{stray} = L \left[\left(\frac{f_0}{f_0 - \Delta f} \right)^2 - 1 \right] \quad (5.3)$$

where f_0 is the resonance frequency of each filter and Δf is the separation between the peak and the valley. f_0 is retrieved from the position of the peak. L is the LC filter inductance. Common inductance is the same quantity responsible for the *common impedance crosstalk* contributor, described in Eq. 3.24 from Chapter 3. The analysis of these measurements was carried out by applying a moving average to

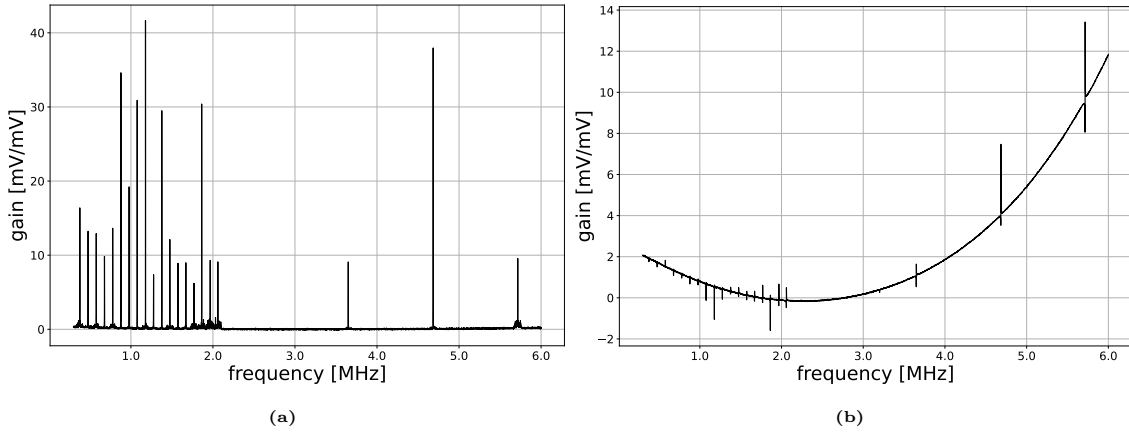


Figure 5.9: Network Analyzer scan through the TESSs line (a) through the feedback line (b).

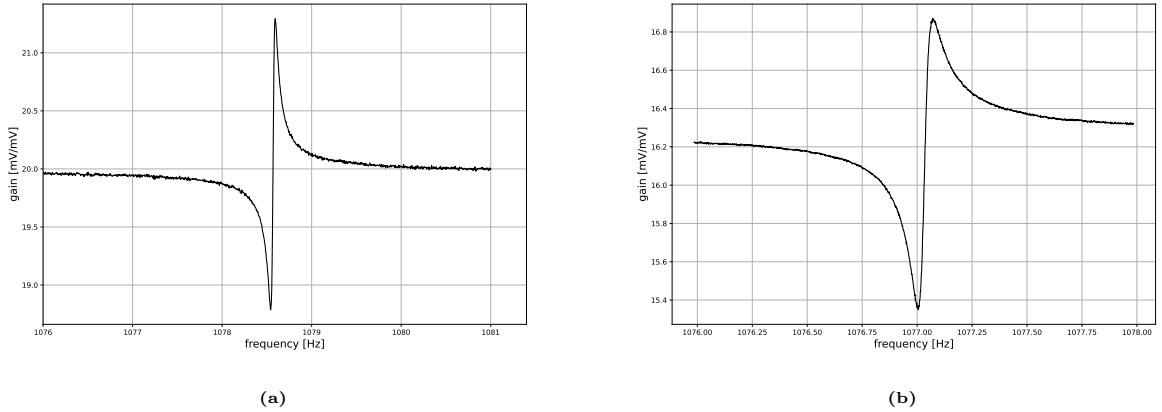


Figure 5.10: Detailed gain and phase for the NWA scan through the feedback line in case of AC bias disconnected (a) and connected (b).

the signal and estimating the peak and valley frequencies, as shown in Figure 5.11. From these measurements and from eq. 5.3 we can calculate the common inductance values corresponding to each resonance frequency. The result is shown in Figure

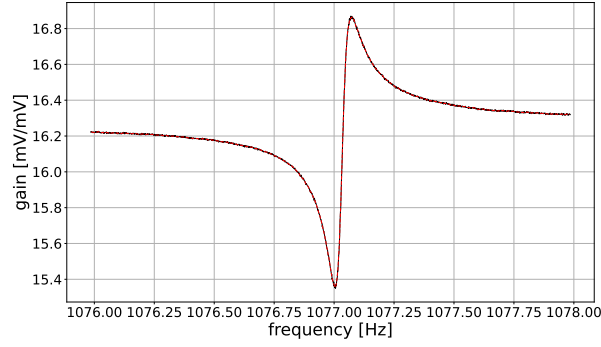


Figure 5.11: Moving average (red dashed line) over single resonance plot.

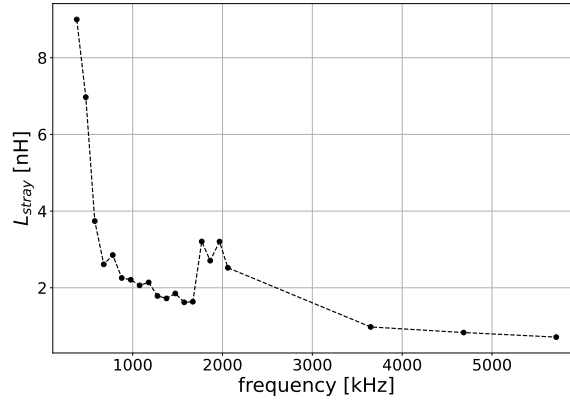


Figure 5.12: Common inductance as calculated using 5.3 from the NWA scan performed through the feedback line in the AC bias disconnected case.

5.12. The level of common inductance is of the order of the nH and it seems to be consistent with other measurements previously performed using the same circuitual design [55]. It shows an increase toward lower frequencies. This requires further investigation, since the design of the LC chip and SQUID is optimized for frequency ranges above 1 MHz. The 4 values around 2 MHz which slightly deviate from the other points, are the same that presented a smaller Q factor (Figure 5.7).

From the scan performed by injecting the signal into the carrier lines, the resonance frequencies and bandwidths were extracted from the resonance Lorentzian fit. These results will be used as a reference. In Figure 5.13 and in Table 5.3 the fit parameters are displayed.

Pixel	f_{res} [kHz]	Δf (kHz)
1	377.24 ± 0.04	0.120 ± 0.012
2	478.18 ± 0.03	0.133 ± 0.012
3	577.82 ± 0.03	0.144 ± 0.009
4	678.10 ± 0.02	0.154 ± 0.011
5	778.25 ± 0.02	0.133 ± 0.008
6	878.15 ± 0.02	0.075 ± 0.003
7	978.28 ± 0.02	0.103 ± 0.004
8	1077.01 ± 0.02	0.079 ± 0.003
9	1177.95 ± 0.02	0.07 ± 0.002
10	1276.129 ± 0.009	0.17 ± 0.008
11	1376.582 ± 0.014	0.083 ± 0.002
12	1474.51 ± 0.02	0.136 ± 0.006
13	1573.462 ± 0.011	0.153 ± 0.007
14	1669.589 ± 0.015	0.163 ± 0.008
15	1768.78 ± 0.02	0.211 ± 0.019
16	1862.89 ± 0.04	0.087 ± 0.005
17	1966.86 ± 0.04	0.196 ± 0.019
18	2059.75 ± 0.04	0.194 ± 0.016
19	3649.374 ± 0.006	0.156 ± 0.004
20	4684.700 ± 0.013	0.079 ± 0.002
21	5715.41 ± 0.02	0.21 ± 0.02

Table 5.3: Resonance frequency and resonance width values from the fit. The uncertainties are estimated from the standard deviation.

5.5 Electrical crosstalk measurements

Once the main characteristics of the LC chip were established, the measurement of the electrical crosstalk was carried out. At the end of the chapter, the experimental results will be compared with the analytical calculations that were described in Chapter 3.

For this tests, all TESs were biased at 30% of their normal resistance ($R_n \approx 25 \text{ m}\Omega$), to be operated in a region which provides a large dynamic range. The detector biasing is achieved by injecting the input signal onto the carrier lines at the tuned frequency of each pixel's LC filter. The presence of R_{shunt} , which is smaller than the impedance of each leg, ensures a stable voltage bias. The SQUID output is

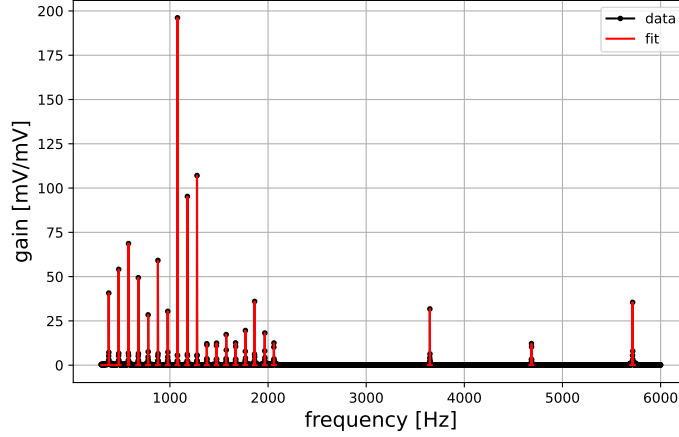


Figure 5.13: AC bias scan data (black dots) with fitting curves (red solid line).

read out, demodulated at the pixels' frequencies, then re-modulated and re-injected after correcting for phase shifts and delays to null the carrier signal.

In order to assess the level of electrical crosstalk, measurements were performed by modulating the carrier signal. Different amplitude-modulation shapes, such as pulses and sine waves, were used. While pulse-shaped tones are expected to be easily detectable and suitable for measuring crosstalk, sine-wave signals may lead to a more complex interpretation. On the other hand, the assumptions underlying the analytical calculations in this case refer to small-amplitude and frequency modulation of sine-wave signals. A complete comparison can therefore be obtained by analyzing both pulse-shaped and sine-wave modulations. As shown in the results presented in the following sections, the former provides more robust and reliable outcomes, whereas the latter exhibits the expected limitations.

5.5.1 Pulse shape modulation

The pulse modulation was injected into a single channel and shaped with a rise time of $100 \mu\text{s}$, a fall time of 10 ms , and a period of 500 ms . The response time of these TES microrcalorimeters is expected to be $\tau \simeq 10 \mu\text{s}$, while the thermal conductance is $G \approx 0.85 \text{ pJ/K}$ [51]. The output signals from all channels were then recorded. The modulated channel is referred to as the *perpetrator pixel*, while the remaining ones are denoted as *victim pixels*.

At first only five pixels were considered, in particular pixels 7, 8, 9, 10, 11, whose corresponding resonance frequencies are reported in Table 5.4. The electrical crosstalk was quantified by averaging the output recorded at the ADC level over the total

pixel 7	pixel 8	pixel 9	pixel 10	pixel 11
978.282 ± 0.017 kHz	1077.006 ± 0.018 kHz	1177.953 ± 0.02 kHz	1276.129 ± 0.009 kHz	1376.582 ± 0.014 kHz

Table 5.4: Resonance frequencies for the involved pixels, as measured with a Network Analysis scan through the carrier line (see Table 5.3).

number of acquired events and normalizing it to the peak height of the perpetrator channel response. Measurements were carried out with 50000 events and pixel 9 was selected to be the perpetrator pixel. The results are shown in Figure 5.14. In this

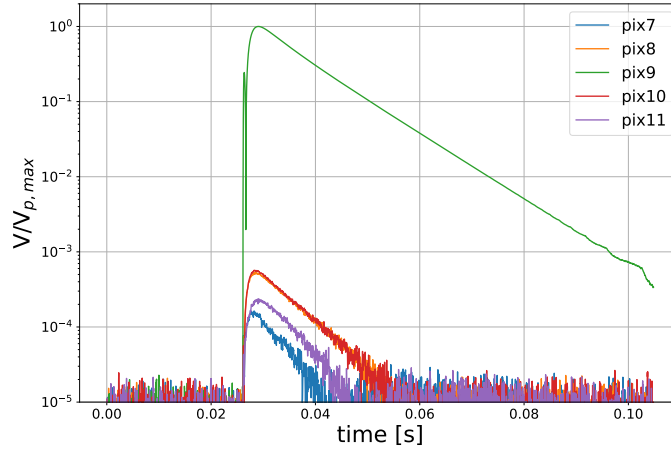


Figure 5.14: Response from the perpetrator (green line) and victim pixels averaging over 50000 events recorded.

case, the electrical crosstalk was estimated as the ratio between the signal maxima of pixels 7, 8, 10, and 11 and to that of pixel 9 (the peak has been evaluated with a moving average algorithm).

The error bars on the crosstalk ratio were calculated from the RMS value of the noise signal, which corresponds to the 1σ width of the distribution. To obtain a more robust estimate of the uncertainty, the error was expressed in terms of the FWHM, using the relation $\text{FWHM} = \sigma \times 2\sqrt{2 \ln 2}$.

In Figure 5.15 we report the results. As expected, crosstalk ratio for the closest neighbors is more significant: the closer the pixels are in the frequency space, the more the crosstalk will impact the measurements.

This measurements were repeated including all the 21 pixels in the matrix, modulating one pixel at a time and registering the response from all the other channels. The results are arranged in a matrix with a $N_{pixel} \times N_{pixel}$ size, where N_{pixel} is the number of channels in the LC chip. Each spot of the matrix describes the level of crosstalk related to a detector pair. The pixels are ordered by multiplexing frequency. This means that pixel 1 is associated to the lowest multiplexing frequency and pixel 21 to the highest frequency.

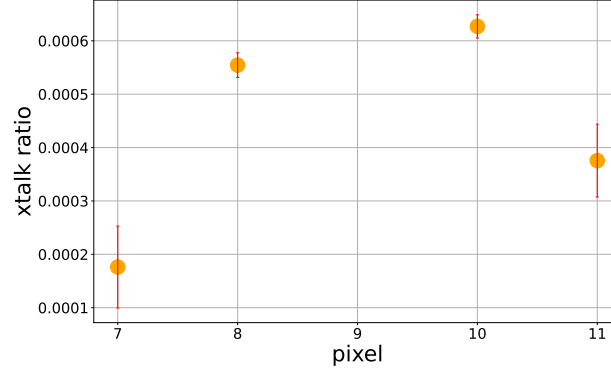


Figure 5.15: Crosstalk ratio for measurements.

The diagonal elements of the matrix are equal to 1, since they represent the auto-

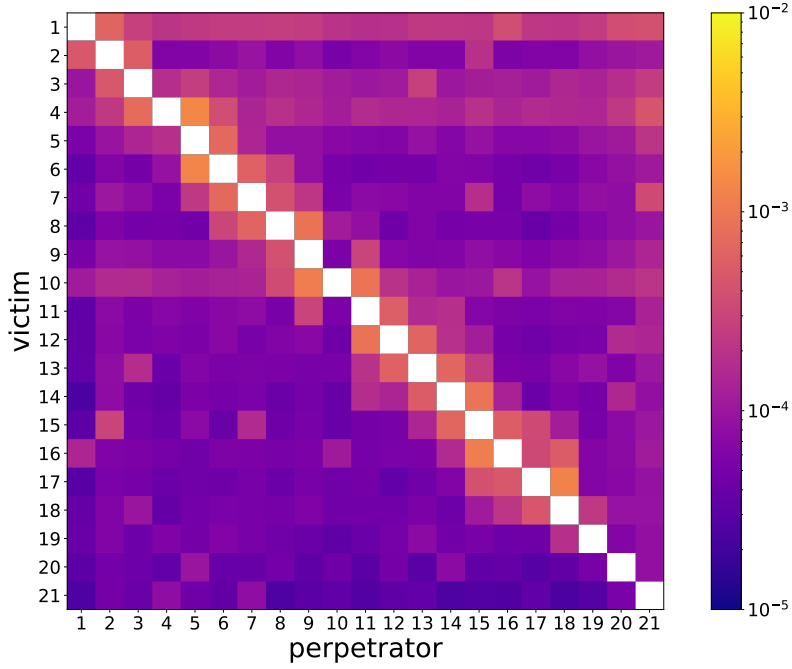


Figure 5.16: Crosstalk matrix built from experimental measurements. Each element of the matrix represents the level of crosstalk of a detector pair.

correlation of each pixel, and they are masked in this plot to emphasize the more relevant range of values.

5.5.2 Analytical model and measurements: a comparison

Using the methods described in Section 3.2.3 and Eq. 3.24, we analytically calculated the carrier leakage and common impedance crosstalk contributions for the crosstalk matrices presented in the previous sections, considering both the 5-pixel

and 21-pixel cases. The magnitude of the signals was taken into consideration for calculating crosstalk in this analysis. As can be seen in Figure 5.17, which shows the analytical crosstalk matrices for the 21-pixel measurement, the common impedance crosstalk is approximately three orders of magnitude larger than the carrier leakage crosstalk, indicating it is the dominant contribution. As already mentioned, this result is not entirely unexpected, since carrier leakage crosstalk primarily depends on the bandwidth and the victim pixel impedance, which are specifically tuned to minimize coupling between neighboring channels. Therefore, in the remainder of this section, the electrical crosstalk denomination will refer exclusively to the common impedance crosstalk.

The comparison between measured and modeled values for the 5-pixel case is reported in Table 5.5.

It is worth noticing that the results are satisfactory, with model and measurement

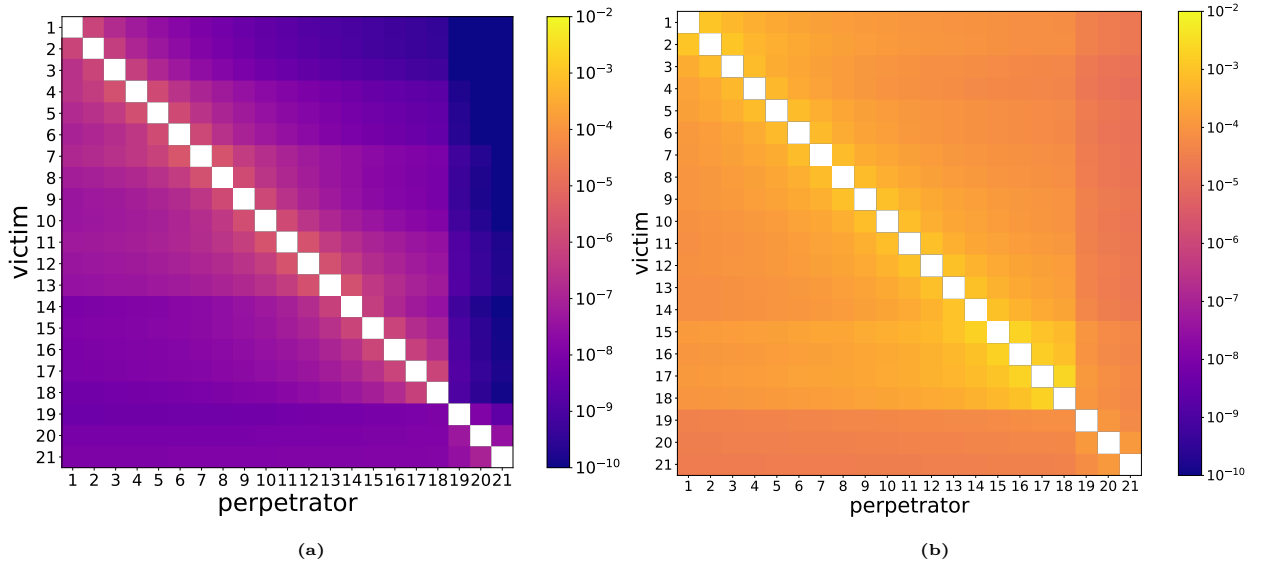


Figure 5.17: Carrier leakage crosstalk (a) and common impedance crosstalk (b) matrices, built from the analytical calculations. The same scale is used for both plots.

	pixel 7	pixel 8	pixel 10	pixel 11
Analytical results	3.8×10^{-4}	7.7×10^{-4}	7.6×10^{-4}	3.8×10^{-4}
Measurements	$(1.7 \pm 0.2) \times 10^{-4}$	$(5.3 \pm 0.2) \times 10^{-4}$	$(5.7 \pm 0.2) \times 10^{-4}$	$(2.4 \pm 0.2) \times 10^{-4}$

Table 5.5: Comparison between analytical model and measurements.

values agreeing within the same order of magnitude.

The analytical matrix in Figure 5.17 shows a dependence from the frequency, that

comes from the shape of the analytical formula:

$$xtalk_{COM} = \left| \frac{1}{2} \frac{\omega_i}{\Delta\omega} \frac{L_{stray}}{L} \right| \quad (5.4)$$

This pattern is not visible in the measurements (Figure 5.16), the crosstalk level for first neighbors does not increase with the multiplexing frequency.

It can be assumed that this inconsistency is related to contribution to crosstalk we are not taking into consideration. Additionally, the noise affecting the measurements can mask the expected behaviour. As mentioned, the assumptions underlying the model may not fully correspond to the conditions associated with this type of amplitude modulation.

Figure 5.18 shows a comparison between the model and the measurements for the 21-pixel array. The figure presents the ratio of the analytical results (Figure 5.17(b)) to the measured results (Figure 5.16). Overall, the results show good agreement,

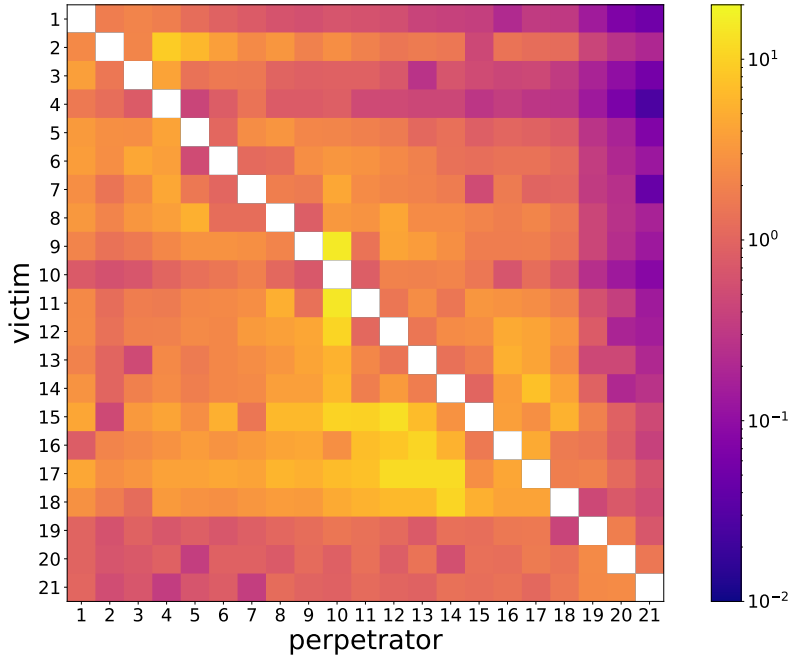


Figure 5.18: Ratio of the analytical results over the measurements results. The detectors are ordered by multiplexing frequencies sequence. On the diagonal the crosstalk level related to the pixels with themselves, it would be equal to 1 but the values is masked in order to show the other values.

with the ratio being close to unity for the first off-diagonal components and ranging between 2 and 10 along the edges. Some specific cases, such as the crosstalk between pixel 10 (perpetrator) and pixels 9 and 11 (victims), or between pixel 21 (perpetrator) and all other pixels (victims), differ by one order of magnitude.

5.6 Sine-wave modulation

The TESs used as bolometers, as mentioned before, will detect small variations in the temperature when the optical power is dissipated over its absorber. The TESs will be operated on the steep part of their transition, to obtain a broad dynamic range. When photons hit the detector and the temperature varies, the resistance of the TES changes and the signal will be modulated.

A similar behavior can be simulated in the laboratory, by electrically modulating the perpetrator carrier signal with a sine-wave and reading the output on the neighbors. The characteristics of the sine-wave modulation comprehend an amplitude equivalent to the 0.2% of the bias signal and a frequency of 120 Hz.

The latter was chosen because it is sufficiently far from both the resonance frequencies and the 50 Hz power line and its harmonics. An example of the perpetrator output when modulated at 120 Hz is shown in Figure 5.19.

Measurements were carried out on three pixels, 7, 8, and 9, with pixel 8 used as

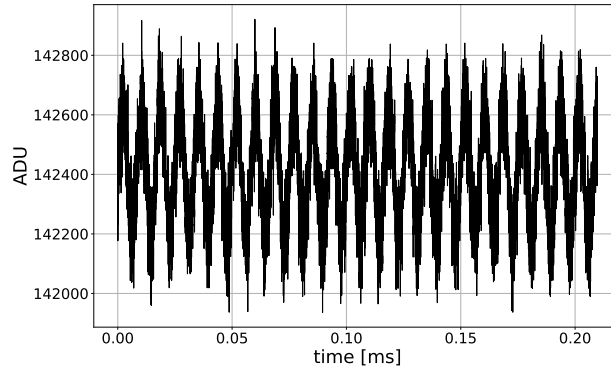


Figure 5.19: Raw data from the modulated output of the perpetrator pixel number 8.

the perpetrator. The demodulated output of each pixel was recorded over 100000 individual events. A Fast Fourier Transform (FFT) was performed on the data from each event, and the resulting spectra were then normalized to the maximum peak of the perpetrator curve and averaged over the entire set of events. The FFT is carried out to isolate the frequency contributions and in particular the modulation at 120 Hz on the perpetrator and victim pixels output. The average reduces the noise contribution to the results. The result are shown in Figure 5.20.

Examining the resulting plot, the peak at 120 Hz is clearly visible for the perpetrator, indicated by the orange curve. The baseline level is around $1 \cdot 10^{-2}$ and no response on the victim pixels curves is observed.

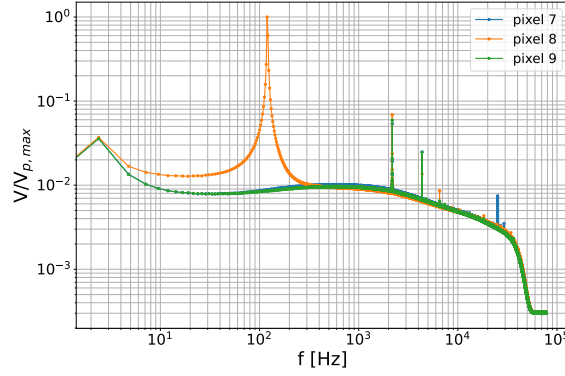


Figure 5.20: Average FFT from 100000 events. A modulation at 120 Hz is injected to pixel 8 (orange), the output from the first neighbors is measured.

The amplitude of the modulation was varied, increased to 0.5%, 1%, and 5% of the voltage bias. In these cases, applying the modulation resulted in a rise of the perpetrator's baseline, likely due to the TES being driven continuously away from its operating point.

Figure 5.21 presents the results of a measurement conducted with a sinusoidal mod-

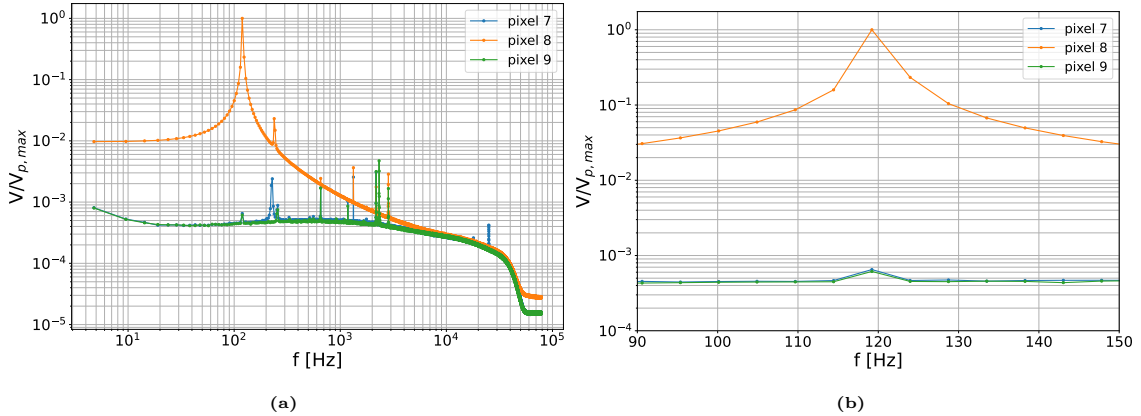


Figure 5.21: Crosstalk from a 5% sine-wave modulation (a), with a zoomed and detailed image (b). The effect of electrical crosstalk is visible in the neighbor curves, in particular the green (pixel 7) and the purple (pixel 9), which are the closest neighbors.

ulation of amplitude corresponding to 5% of the bias voltage. As evident from the plot, the baseline of the perpetrator's FFT curve, averaged over all events, is higher relative to the other pixels. In the zoomed view on the right, a non-zero crosstalk level can also be observed.

The combination of these results provides insight into the behavior of the system. The electrical crosstalk level is below 10^{-2} , although this represents a rather loose upper limit. By comparison, the level measured in the previous section was of the order of 10^{-4} , consistent with the discussion in Chapter 3. For readout chains of

this kind, with stray inductances L_{stray} of a few nH and LC filter inductances in the μH range, a typical lower limit is around 10^{-3} .

Increasing the modulation amplitude to 5% of the bias voltage seems to disrupt the TES electrothermal feedback, preventing it from remaining at its operating point. Consequently, the crosstalk level observed in Figure 5.21-(b) is not representative, as it is normalized with respect to an incorrect point on the perpetrator curve.

Previous studies, such as [54], report results obtained using the same method. The main difference lies in the resonance frequency spacing: in the system studied here, the spacing is approximately 100 kHz on average, whereas in [54] it was 35–40 kHz. A modulation amplitude of 1% of the bias voltage was employed in that case. This is not feasible in our system without displacing the TES from its operating point.

These results confirm both the expected magnitude of the crosstalk level in an FDM readout chain designed for CMB experiments and the anticipated limitations of the measurement setup. The next steps will build upon the measurement procedure developed here and will focus on possible optimizations aimed at achieving a consistent and comprehensive characterization of the electrical crosstalk.

Moreover, the level of electrical crosstalk obtained here is consistent with the requirements for CMB experiments, which set the acceptable crosstalk level at approximately 10^{-3} . This value is comparable to that of the optical crosstalk. Although no dedicated studies are currently available for SWIPE, similar requirements can be found for LiteBIRD [16].

5.7 Conclusions

In this chapter, we presented the measurements on a fully lithographic LC chip, consisting of coil inductors and parallel-plate capacitors. The results included the characterization of the main filter properties and of the multiplexed circuit scheme, as well as the evaluation of the crosstalk level. The latter was estimated through measurements performed by injecting a modulated electrical signal in two different configurations: using a pulse-shaped signal and a sine-wave modulated signal.

In the next chapter, we will focus on measurements related to a different type of LC chip, based on coil inductors and interdigitated capacitors. We will present the characterization of its main properties and preliminary results on electrical crosstalk measurements.

Chapter 6

Characterization and crosstalk measurements for an end-to-end readout chain of LiteBIRD

In this chapter, the measurements obtained using a complete end-to-end readout chain of LiteBIRD, including the cold cryogenic section and the room temperature electronics, composed of the SQUID controller assembly (SCA), the SQUID processing assembly (SPA), the digitizer assembly (DA) and the Mission Digital-Processing unit (MDPU), will be presented. The results will provide information on the status of the warm electronics development, as well as on the interfaces between the cold and warm sections, and between the detectors and the cold readout.

6.1 The experimental set-up

The measurements set-up includes the LC filters and one SQUID, tested at 4 K and 30 mK. These components are mounted on a PCB and a copper holder. The wiring on the PCB is routed to a Micro-D 25-pin (MDM-25) connector. Figure 6.1 shows the cryogenic section of the chain. The LC chip is entirely fabricated by lithography on a 150 mm silicon (Si) wafer and includes 40 niobium LC filters, designed to have resonance frequencies between 1.6 and 4.2 MHz [56]. A 100 Å layer of Titanium, followed by a 1400 Å layer of Aluminum were deposited, in order to minimize the eddy current losses in the mounting board in contact with the chip. This has an impact on the inductance value, which is decreased by 25%. Each LC filter is composed by a **coil inductor** and a **interdigitated capacitor**. The inductance is equal for all

resonators and it is $L = 60 \mu\text{H}$. The capacitors are shaped to have values between ~ 18 and 150 pF . The SQUID is a SA13 Series Array SQUID produced at *NIST* (National Institute of Standard and Technology). The voltage bias on the FDM scheme is performed with a shunt resistor $R_{bias} = 0.025 \Omega$ in parallel to the LC branches. The measurements were performed using two configurations: one with a row of resistors, having resistance values in the $1\text{--}2 \Omega$ range, connected in series with the LC filters to mimic the TESs, and another including six bolometer TESs to study the detectors' behaviour within a complete readout chain and to assess the level of electrical crosstalk in the system. The LC filters resonance bandwidth varies with the resistance of the channel, $\Delta\omega = R/L$, and it reflects the transition of the TES from the normal to the superconducting stage. This property will be presented in the next sections.

The PCB is attached to a copper holder for thermalization, enclosed in a Nb case

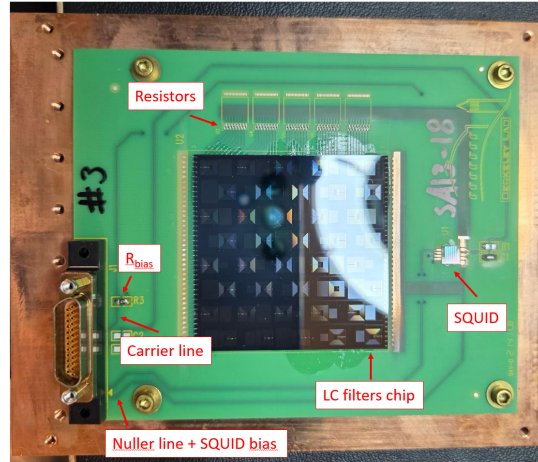


Figure 6.1: A picture of the cryogenic elements on the PCB: the LC filters chip in the middle, the SQUID on the right and the bias resistor on the left. On top the row of resistor can be observed.

for magnetic fields shielding, and mounted on the Mixing Chamber (MXC) of a Bluefors dilution cryostat.

The signal is routed outside the cryostat through the 4 K stage using a shielded superconducting cable that terminates on a Fischer 24-pin connector flange. The connection is then extended to the electronics by a room temperature cable terminated on the other side with a D-subminiature (DB) 25-pin connector.

The SCA (SQUID Controller Assembly) is designed to read out four channels, i.e. four SQUIDs; in this case exclusively *channel A* is used. The bias signals are sent to the SQUIDs through *channel 3* on the SPA (Signal Processing Assembly). Figure 6.2 shows the measurement set-up at cryogenic and room temperature.

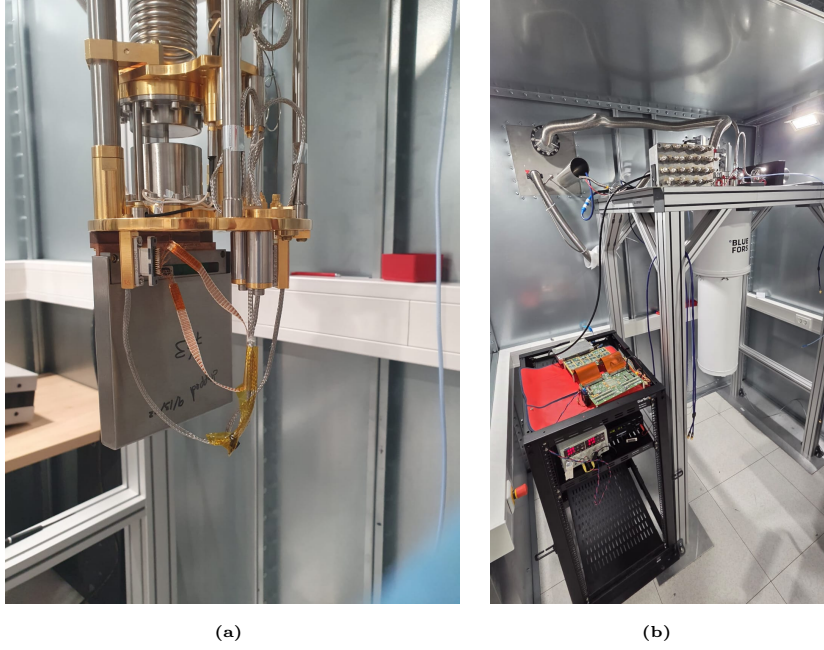


Figure 6.2: (a) The Nb case attached on the mixing chamber inside the cryostat. (b) View of the set-up at room temperature. On the left the warm electronics elements are shown.

The power for the SPA and DA is provided by a laboratory supplier and a MDPU simulator. The entire equipment is arranged in a EMI (electro-magnetic interference) shielded chamber, to isolate the instrumentation. The analysis is conducted on a laboratory laptop through a custom software¹ that interfaces with the FPGA.

6.2 Measurements and tests

A full readout chain, comparable to the one described above, was previously tested at the laboratories of McGill University [57]. Subsequently, the warm electronics were optimized, and a different cold readout setup was employed for the measurements presented in this work.

6.2.1 SQUID properties

The SQUID properties were evaluated through measurements of $V - \Phi$ curves at different bias currents, as detailed in Chapter 2. This measurement was performed at the bath temperature $T \simeq 4$ and 0.03 K.

A bias current range of $38.15 - 50.24 \mu\text{A}$ and a bias flux range of $0 - 33 \mu\text{A}$

¹The software and the firmware for the FPGA were developed by Graeme Smecher, Jean-François Cliche, Joshua Montgomery, Matt Dobbs, and the team at McGill University.

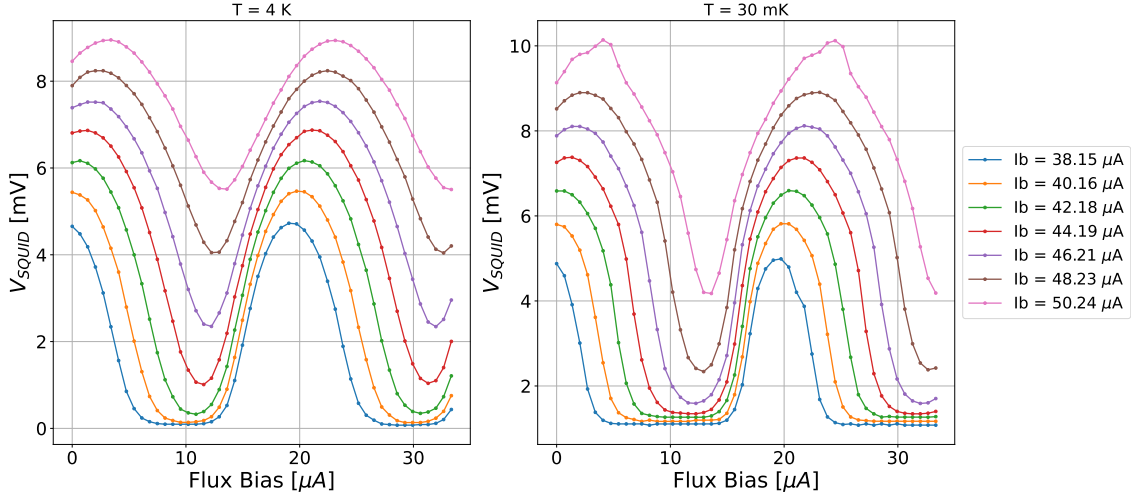


Figure 6.3: SQUID $V - \Phi$ curves at different bias current I_b at $T=4$ K and $T=30$ mK.

were explored and the results are shown in Figure 6.3. At lower bias currents the SQUID remains inactive or show a small amplitude. The maximum peak-to-peak response is achieved at $I_b \simeq 44.19 \mu\text{A}$ at $T=4$ K and $I_b \simeq 48.23 \mu\text{A}$ at $T=30$ mK. A larger peak-to-peak extension corresponds to a wider linear region and to a higher transimpedance, defined as $Z_{trans} = V_\phi \equiv dV(\Phi)/dI_{flux}$, which quantifies how the SQUID voltage varies with flux. An increase in Z_{trans} implies stronger amplification of small current variations. This quantity was evaluated for the measurement runs carried out at the two temperatures.

The transimpedance values for the curves with maximum amplitude on the negative and positive slopes are respectively $Z_{trans,n} = 1319 \Omega$ and $Z_{trans,p} = 1120 \Omega$ at $T = 4$ K, and $Z_{trans,n} = 2210 \Omega$ and $Z_{trans,p} = 1904 \Omega$ at $T \simeq 30$ mK. Figure 6.4 show the largest transimpedance positions in the $V-\Phi$ curves.

Another relevant parameter when operating the SQUIDs is the dynamic resistance, defined as $Z_{dyn} = dV(\Phi)/dI_{bias}$, which describes how the voltage across the SQUID varies with changes in the bias current when I_{flux} is fixed. We calculated Z_{dyn} using the $V-\Phi$ curve at the bias current corresponding to the maximum peak-to-peak amplitude, together with the two adjacent curves at neighboring bias currents. We set I_{flux} as the value corresponding to the maximum $Z_{trans,n}$ and $Z_{trans,p}$. As shown in Figure 6.5, we denote $Z_{dyn,1}$ as the dynamic resistance calculated on the maximum amplitude $V-\Phi$ curve and the one at the lower I_{bias} , and $Z_{dyn,2}$ as the dynamic resistance calculated relatively to the curve at the higher I_{bias} , at $T = 4$ K and $T = 30$ mK. The results are summarized in Table 6.1. As shown in Figure 6.3, higher Z_{dyn} values are expected on the negative slopes due to the larger separation

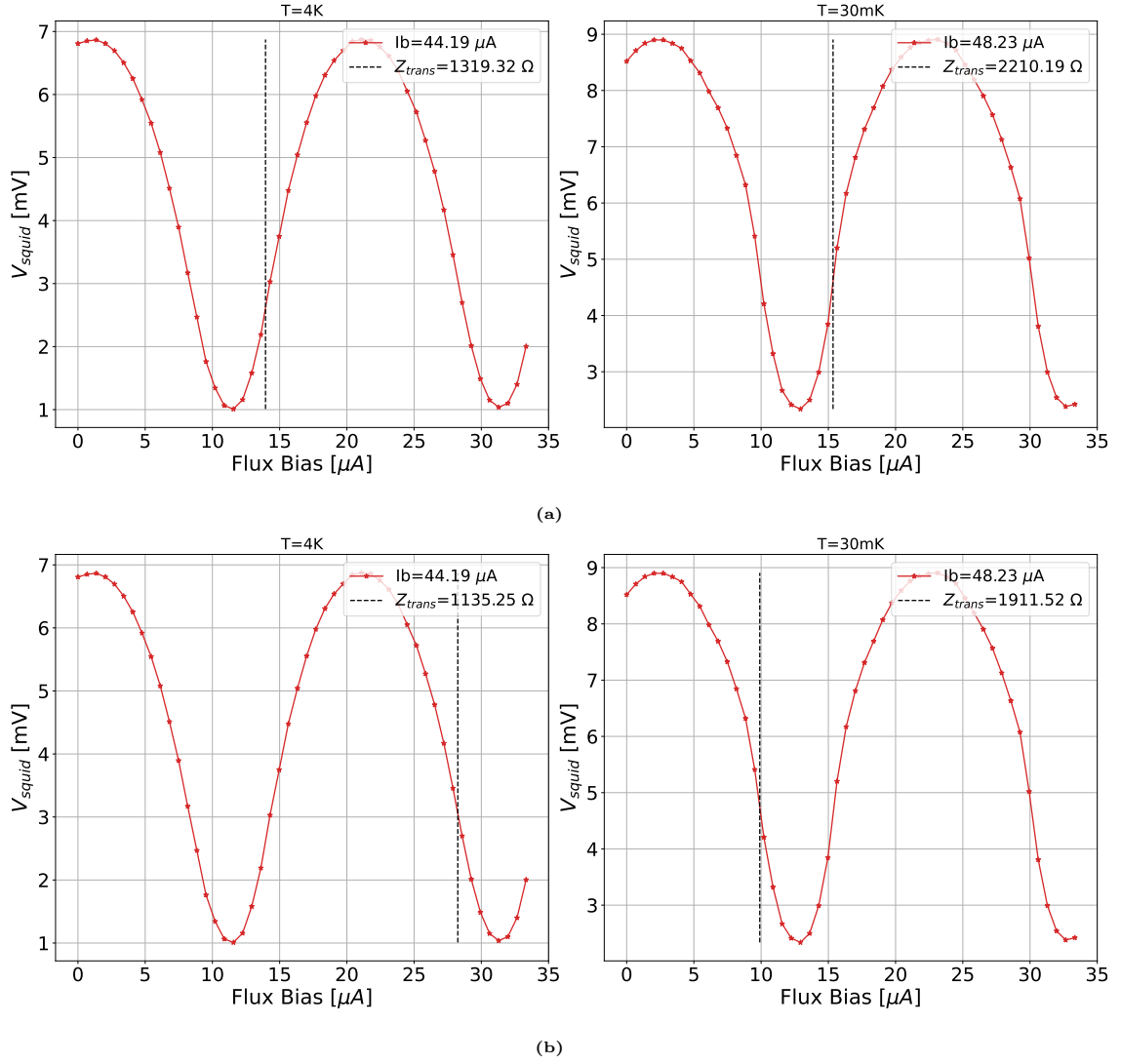


Figure 6.4: SQUID V - Φ curves at I_{bias} corresponding to the maximum peak-to-peak amplitude at $T = 4$ K (left) and $T = 30$ mK (right). The black dashed line marks the point of maximum $dV/d\Phi$ on the positive (top) and negative slope (bottom).

between these curves.

$Z_{dyn,1}$			$Z_{dyn,2}$		
T (K)	Pos. slope (Ω)	Neg. slope (Ω)	T (K)	Pos. slope (Ω)	Neg. slope (Ω)
4	384.48	1004.51	4	415.63	928.0
0.03	621.24	1192.15	0.03	676.98	1346.57

Table 6.1: Dynamic resistance for the V - Φ curves adjacent to the maximum peak to peak one.

The output noise signal was recorded on the working points to assess the overall noise level of the system. In this setup, the noise may arise from external electro-

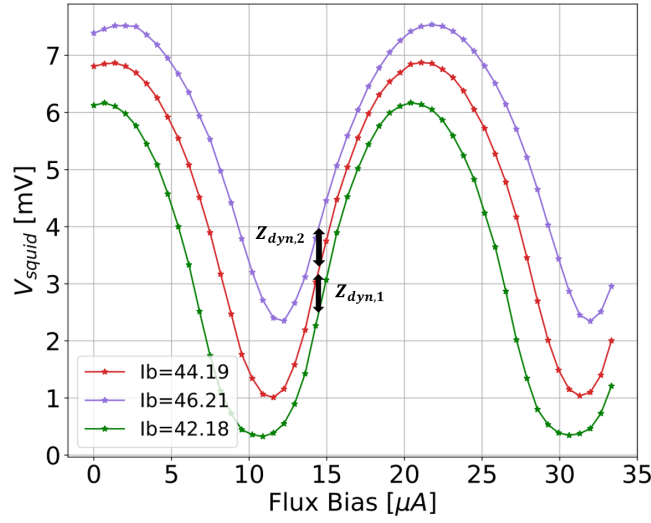


Figure 6.5: The definition of $Z_{dyn,1}$ and $Z_{dyn,2}$. The maximum amplitude (red) curves and the first neighbor are shown.

magnetic pick-up, poor grounding, or thermal noise originating from unconnected elements in the cryostat, such as thermometers or heaters.

Given the analysis on the measurements conducted in several configurations, the RMS noise level was reduced by optimizing the ground connections related to the warm electronics and by disconnecting the cryostat thermometric system. Peaks in the noise FFT curves were found to be recurrent and could not be eliminated at this stage. The noise minimization process is an important step in characterizing the readout chain, and will be investigated in greater detail in future works.

Figure 6.6 shows the lowest noise level achieved, along with the FFT of the corresponding data stream. In the latter, the primary noise peak is observed at approximately 140 kHz.

6.2.2 Network analysis

The Network analysis was employed to investigate the main characteristics of the LC filters. Following this approach, a sinusoidal signal of known amplitude is generated over the frequency range of the resonant filters and injected either through the *carrier* lines, i.e. the LC filter inputs, as a voltage signal, or through the “nuller” line, at the SQUID input, as a current signal. The *nulling signal* is a key parameter, implemented through a technique similar to the base-band feedback filter described in the previous chapter.

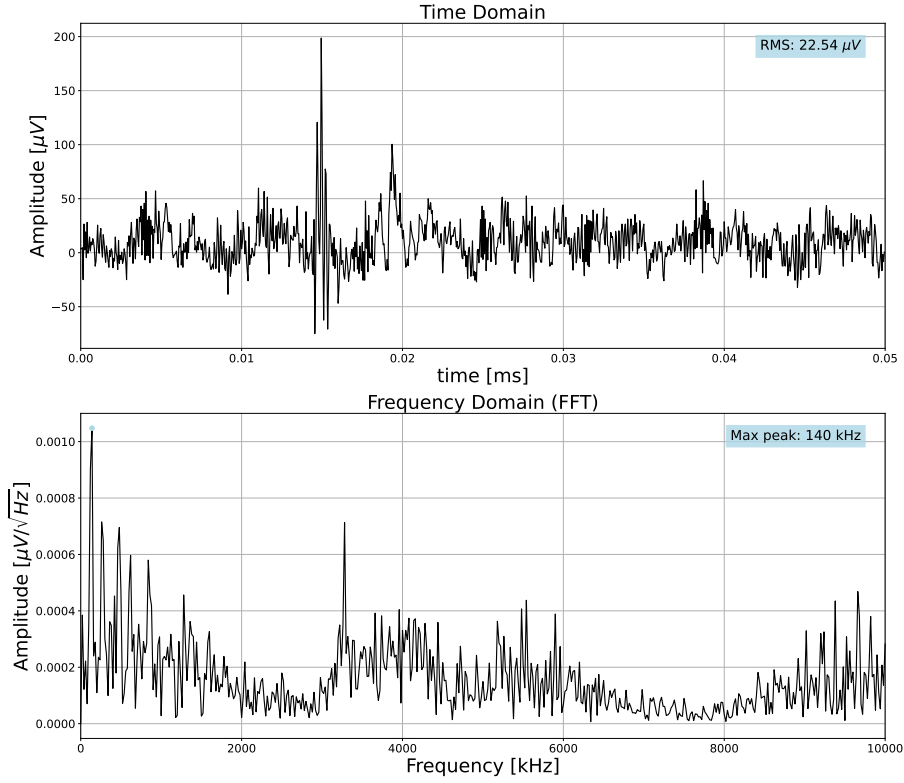


Figure 6.6: Data time stream for noise estimation (above), FFT of noise recurrent contributors (below).

The goal is the same: to “null” the bias signal at the resonance frequency to isolate the sky signal and linearize the response of the SQUID. The method used in Lite-BIRD is called **Digital Active Nulling** or **DAN** [58], [16]. While the basic *static nulling* technique exclusively cancels the signal at the bias frequency selected by the LC filter tuning, the DAN is extended to a wider bandwidth. The main difference with respect to the base-band feedback lies in the implementation of the nulling: while the former employs a SQUID feedback coil, the DAN method injects the signal directly into the circuit, immediately before the SQUID input coil, through the *nuller lines*, as shown in Figure 6.7. Figure 6.1 show the nuller line on the board. Effective cancellation requires the DAN signal and the SQUID output to be summed with a 180° phase difference, with additional phase shifts carefully characterized. These shifts are measured by injecting amplitude-modulated signals through the nuller lines and determining the phase offset relative to the SQUID output. The DAN is implemented in the FPGA, which continuously adjusts the phase and amplitude of the nuller current at a rate much faster than the detector time constant. The SQUID output, containing the carrier and nuller tones together with the sky-signal modulation, is demodulated and used to define the signal to be re-injected through the nuller line.

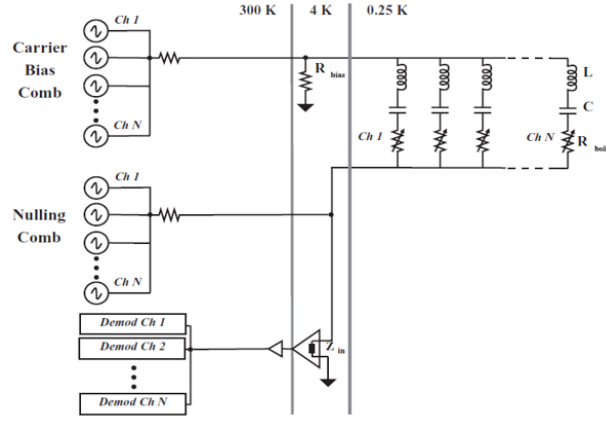


Figure 6.7: A scheme of a multiplexing system, with the carrier and nuller lines, from [43].

When the DAN is enabled, the network analysis requires additional considerations. In particular, the resonance frequency may be shifted with respect to the case without DAN. At the level of the SQUID input coil, the carrier and nuller currents can be defined as I_{car} and I_{nul} , respectively, as illustrated in Eq. 6.1 and 6.2 and Figure 6.8.

Equations 6.1 and 6.2 are derived from the SQUID outputs when a voltage V_1 is

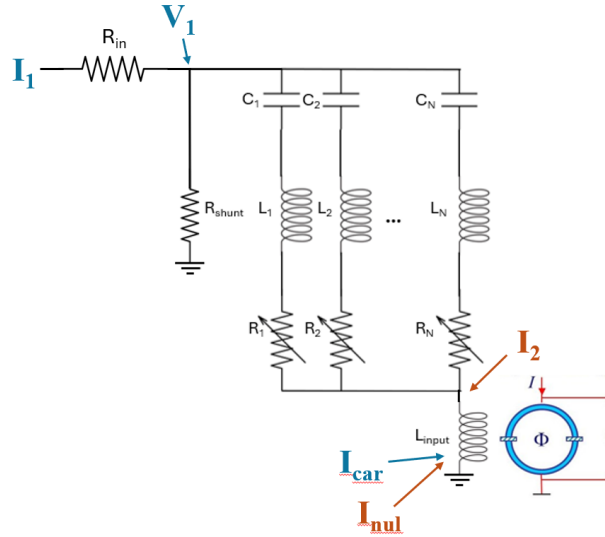


Figure 6.8: Schematic circuit representation. I_1 denotes the carrier input, with I_{car} the corresponding output on the carrier line. I_2 represents the nuller input, while I_{nul} indicates the output of the nuller current measured at the SQUID level.

applied through the carrier line and a current I_2 through the nuller line.

$$I_{car} = V_1 \cdot \left(\frac{1}{R_{bias} + Z_{RLC} + Z_{stray}} \right) \quad (6.1)$$

$$I_{nul} = I_2 \cdot \left(\frac{Z_{RLC} + Z_{stray} + R_{bias}}{R_{bias} + Z_{RLC} + Z_{stray}} \right) \quad (6.2)$$

The value of admittance $Y_{DAN} = 1/(Z_{stray} + R_{bias} + Z_{RLC})$ for network analysis is obtained from the ratio between the carrier and nuller currents I_{car}/I_{nul} .

$$Y_{DAN} = \frac{I_2 I_{car}}{V_1 I_{nul}} = \frac{1}{Z_{stray} + R_{bias} + Z_{RLC}} \quad (6.3)$$

The network analysis was performed after identifying the $V-\Phi$ curve with the maximum peak-to-peak amplitude and selecting, on this curve, the flux point corresponding to the maximum transimpedance. The resulting response is shown in Figure 6.9. The peaks in the plots correspond to the frequencies where the admittance with the

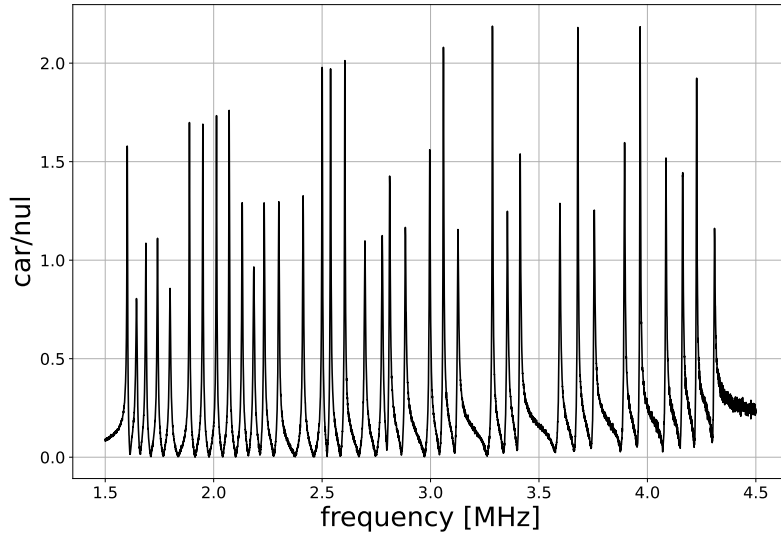


Figure 6.9: The Network Analysis performed at T=4 K.

DAN enabled (Y_{DAN}) is maximized. The network analysis data are analyzed using an analytical fitting function to extract the resonance frequencies and the corresponding bandwidths for each resonance. The results as reported in Table 6.2 and the fitting function over the data is shown in Figure 6.10. As mentioned, the LC filters chip should host 40 LC filters, whereas the measurements revealed only 36 resonances. Although no physical defects were observed on the chip, the missing resonances may be attributed to possible damages affecting either the inductors or the capacitors. Additionally, a shift in the resonance frequencies with respect to the

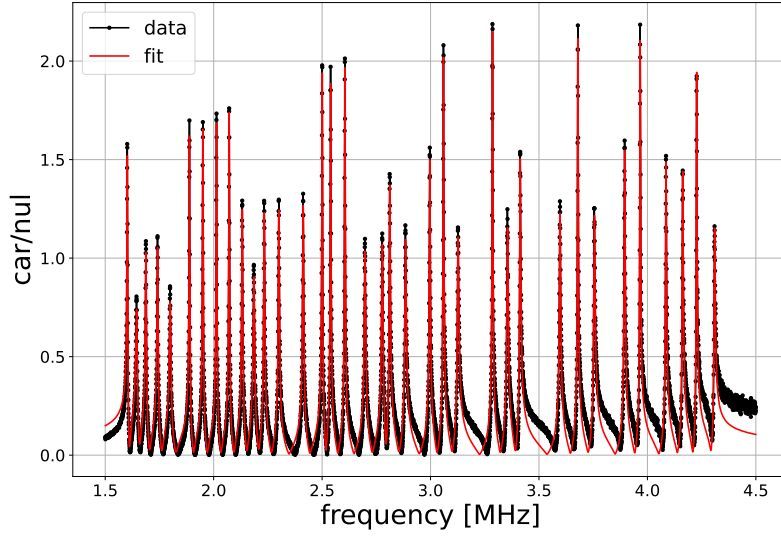


Figure 6.10: The analytical fitting function overlapped to the data obtained performing the network analysis.

#	Res frequencies (kHz)	Bandwidth (kHz)	#	Res frequencies (kHz)	Bandwidth (kHz)
1	1601.907 $\pm$ 0.069	18.607 $\pm$ 0.898	19	2778.126 $\pm$ 0.138	36.156 $\pm$ 1.794
2	1645.287 $\pm$ 0.216	42.445 $\pm$ 2.921	20	2811.971 $\pm$ 0.096	28.601 $\pm$ 1.238
3	1687.857 $\pm$ 0.13	29.31 $\pm$ 1.681	21	2884.584 $\pm$ 0.138	38.32 $\pm$ 1.766
4	1741.514 $\pm$ 0.126	28.998 $\pm$ 1.617	22	2996.726 $\pm$ 0.084	26.5 $\pm$ 1.063
5	1799.551 $\pm$ 0.206	42.983 $\pm$ 2.659	23	3059.638 $\pm$ 0.053	19.526 $\pm$ 0.672
6	1888.506 $\pm$ 0.064	18.451 $\pm$ 0.818	24	3127.008 $\pm$ 0.138	39.377 $\pm$ 1.764
7	1950.186 $\pm$ 0.064	18.666 $\pm$ 0.809	25	3285.765 $\pm$ 0.049	18.458 $\pm$ 0.616
8	2013.211 $\pm$ 0.062	18.335 $\pm$ 0.788	26	3355.028 $\pm$ 0.128	36.428 $\pm$ 1.626
9	2071.534 $\pm$ 0.06	18.219 $\pm$ 0.762	27	3413.427 $\pm$ 0.086	28.067 $\pm$ 1.097
10	2131.954 $\pm$ 0.1	26.251 $\pm$ 1.283	28	3597.194 $\pm$ 0.117	34.817 $\pm$ 1.483
11	2185.643 $\pm$ 0.17	39.133 $\pm$ 2.192	29	3679.28 $\pm$ 0.05	18.68 $\pm$ 0.627
12	2233.261 $\pm$ 0.105	27.877 $\pm$ 1.34	30	3755.65 $\pm$ 0.12	35.594 $\pm$ 1.522
13	2300.456 $\pm$ 0.104	27.904 $\pm$ 1.317	31	3895.081 $\pm$ 0.082	26.769 $\pm$ 1.031
14	2412.842 $\pm$ 0.102	28.276 $\pm$ 1.296	32	3966.528 $\pm$ 0.049	18.493 $\pm$ 0.626
15	2500.595 $\pm$ 0.053	18.237 $\pm$ 0.678	33	4085.899 $\pm$ 0.082	25.157 $\pm$ 1.037
16	2539.275 $\pm$ 0.056	18.664 $\pm$ 0.704	34	4162.888 $\pm$ 0.086	25.634 $\pm$ 1.089
17	2604.913 $\pm$ 0.052	17.817 $\pm$ 0.655	35	4227.016 $\pm$ 0.053	17.697 $\pm$ 0.668
18	2698.043 $\pm$ 0.144	37.181 $\pm$ 1.836	36	4309.811 $\pm$ 0.116	30.318 $\pm$ 1.493

Table 6.2: The resonance frequencies and the corresponding bandwidth values obtained by fitting the data. The uncertainties are calculated as the standard deviation on the fitting parameters.

values reported in [59] is observed. Because of this displacement, it is not possible at this stage to identify the missing resonators. This limitation is partially addressed by replacing the resistors with TESs, which will provide reference points for the resonance positions.

As anticipated, six TESs were mounted on the LC chip board in place of six resistors, that were unsoldered. The wiring bondings were installed to connect the detectors to the LC filters, as schematically shown in Figure 6.11. The TESs are bilayer AlMn detectors designed and manufactured at *NIST* [26], and they have a critical temperature $T_c \simeq 210$ mK and normal resistance $R_n = 1 \Omega$. The TES chip mounted in the setup hosts 12 TES devices, with six bonding pads on each side. For the measurements, we employed the six TESs located on one side of the chip. The assignment between TESs and LC filter resonances is reported in Table 6.3,

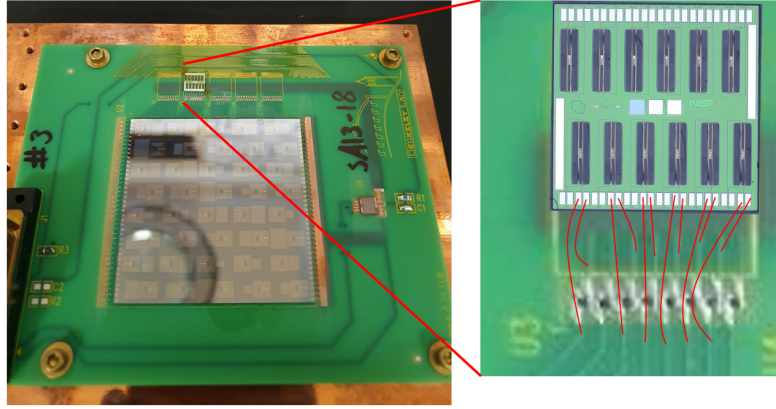


Figure 6.11: A picture (left) of the CIMM with the TESs chip glued on the board. A schematic representation of the wire bonding positions (right).

where the sequence position of each TES is matched to the corresponding LC filter. The LC filter number and the associated resonance frequency are the expected values and they are identified from the design of the LC filters chip.

The network analysis results for scans at $T=4$ K and $T=30$ mK are shown in Figure

# TES	# LC _{exp}	freq _{exp} [kHz]
5	6	1951.492
6	12	2315.489
2	16	2583.653
1	21	2952.006
3	23	3111.085
4	33	4016.317

Table 6.3: Expected values for the position in the sequence of the TESs, position in the sequence of the LC filters and corresponding frequencies.

6.12. The TESs shift from the normal ($R_n=1.2 \Omega$) to the superconducting regime. Accordingly, the quantity of current flowing across each leg grows.

This effect is observed in 5 channels (out of 6 we were expecting), with the cor-

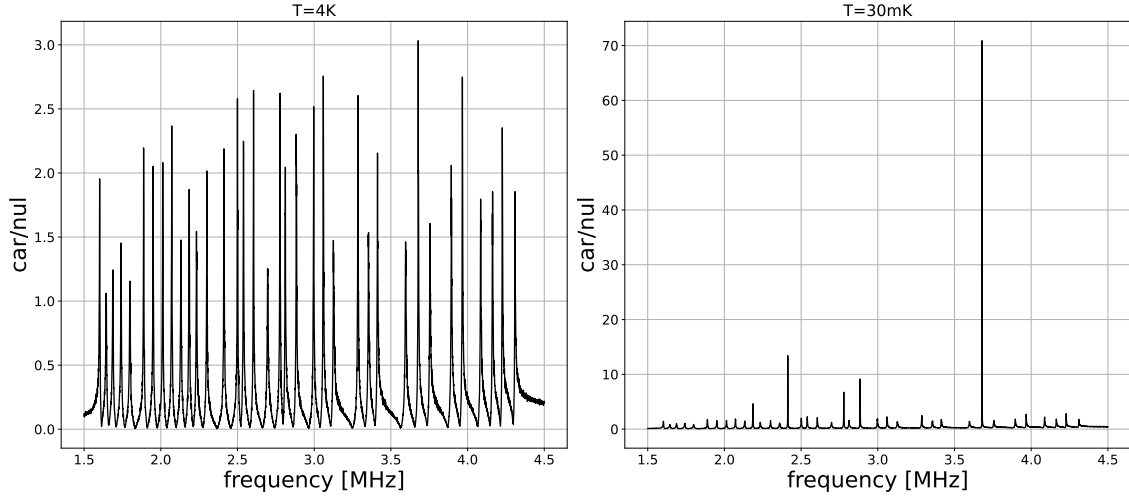


Figure 6.12: The network analysis with TES mounted at $T=4$ K and $T=30$ mK.

responding resonance frequencies deviating from the expected values. The results are displayed in Table 6.4 and Figure 6.13, where the position of the TESs in the resonance frequencies is highlighted by red dashed lines.

It can be observed that the TES connected to LC filter number 6 was not detected.

# TES	#LC _{exp}	#LC _{meas}	freq _{exp} [kHz]	freq _{meas} [kHz]
5	6	-	1951.492	-
6	12	11	2315.489	2185.519±0.017
2	16	15	2583.653	2413.044±0.012
1	21	20	2952.006	2778.093±0.014
3	23	22	3111.085	2884.806±0.012
4	33	30	4016.317	3680.079±0.008

Table 6.4: Expected and measured values for the position in the sequence of the TESs, position in the sequence of the LC filters and corresponding frequencies.

By analyzing the frequency sequence and the measurement positions of the TESs, we conclude that the LC filter number 6 is missing from the sequence. Consequently, the frequencies of the LC filters up to number 22 appear to be shifted by one. Moreover, the LC filter expected to be number 33 in the sequence is measured as number 30, indicating that, in addition to the 6th filter, two other LC filters are missing in the 22–33 range. The fourth and final missing LC resonance is expected to be among the higher-frequency filters.

As discussed in Chapter 2, the main tool to study the properties of a TES is the

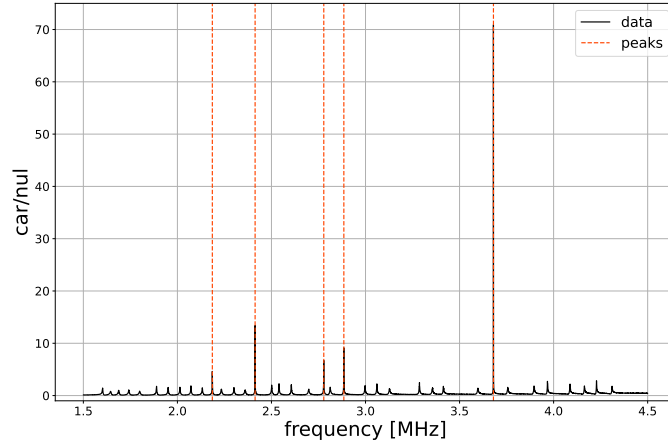


Figure 6.13: Network analysis with TES at $T=30$ mK. The red dashed lined highlight the position of the TESs in the resonances sequence.

I-V curve, obtained by measuring the output current I when the bias voltage V is varied. Figure 6.14 shows the I-V and the ratio of the resistance to the normal resistance for the five channels connected to a TES.

The measurements presented here are not calibrated, as the transfer function of the

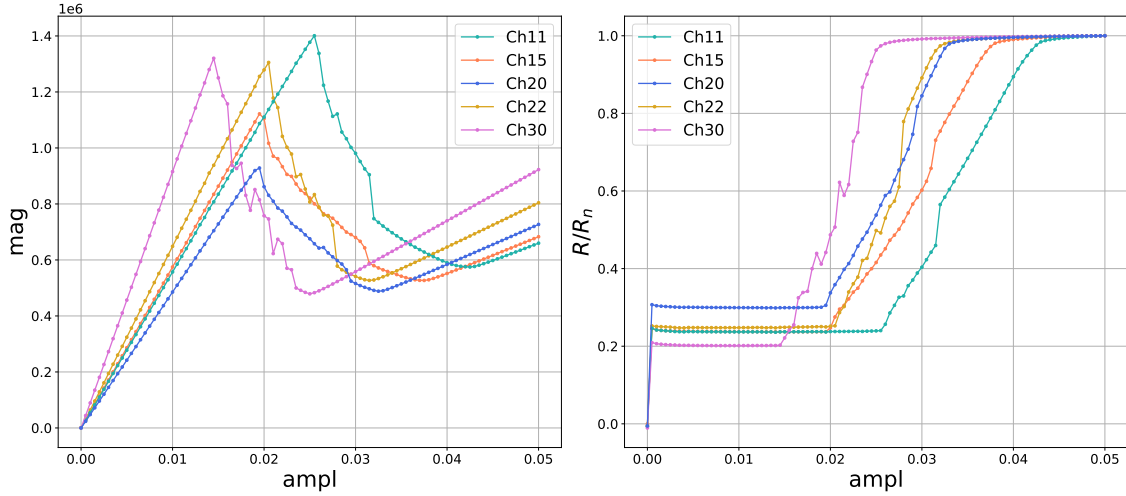


Figure 6.14: I-V curves for the five TESs mounted in series to the LC filters.

system has not yet been determined. Deriving it would require a detailed characterization of the entire electrical scheme, including the cables and the connections among the warm electronics components, which has not yet been performed. Consequently, the amplitude and magnitude values reported on the plot axes should not be interpreted as calibrated physical quantities, such as voltage (V) or current (A). Nonetheless, this level of precision is adequate for the present study, whose main objective is to analyze the overall shape of the I-V curves. In particular, from the

results, multiple transition “jumps” can be observed, together with a series resistance that reflects in the shape of the I-V curve at low amplitude, which should drop rapidly straight to zero. Additionally, the R/R_n ratio exhibits a residual value different from zero. These characteristics are likely linked to the status of the TESs chip and the residual noise captured in these measurements.

Nevertheless, the TESs exhibit predictable behavior, confirming their correct mounting and operation in series with an LC filter and biased through the custom LiteBIRD warm electronics.

6.2.3 Preliminary assessment of Electrical Crosstalk

The level of electrical crosstalk in an end-to-end readout chain of LiteBIRD can be quantitatively assessed through dedicated measurements performed on the system. In Chapters 2 and 5, the crosstalk analytical model and the methods to measure this quantity were described. Unfortunately, the FPGA hosted on the current version of the SPA (Signal Processing Unit) is not equipped with an algorithm to perform frequency or a shaped amplitude modulation of the carrier signals. We made a preliminary estimation of crosstalk by decreasing the amplitude of the bias input of a channel and recording the response of the adjacent pixels at the corresponding frequencies. The measurements were performed by setting the initial bias amplitude such that the TESs resistance was 80% of the normal value. For each input signal value, 1000 events were recorded and averaged to maximize the signal-to-noise ratio. We did this in three configurations, considering that there are 5 TESs and 31 resistors in the system:

1. **Perpetrator:** TES, **victims:** resistors
2. **Perpetrator:** resistor, **victims:** TESs
3. **Perpetrator:** resistor, **victims:** resistors

The last case is the simplest, as no effect is expected on the victims’ signal outputs. The circuit impedance remains unchanged, preventing any variation in current at frequencies other than that of the perpetrator (here called ω_i). Figure 6.15 shows consistency with this assumption. While for the perpetrator (pixel 34, blue dots) the current output changes with the amplitude, accordingly to $I(\omega_i) = V(\omega_i)/R_i$, the victims output at $\omega_{i\pm 1}$ (pixel 33, orange markers, and pixel 35, red markers) only varies for noise fluctuations.

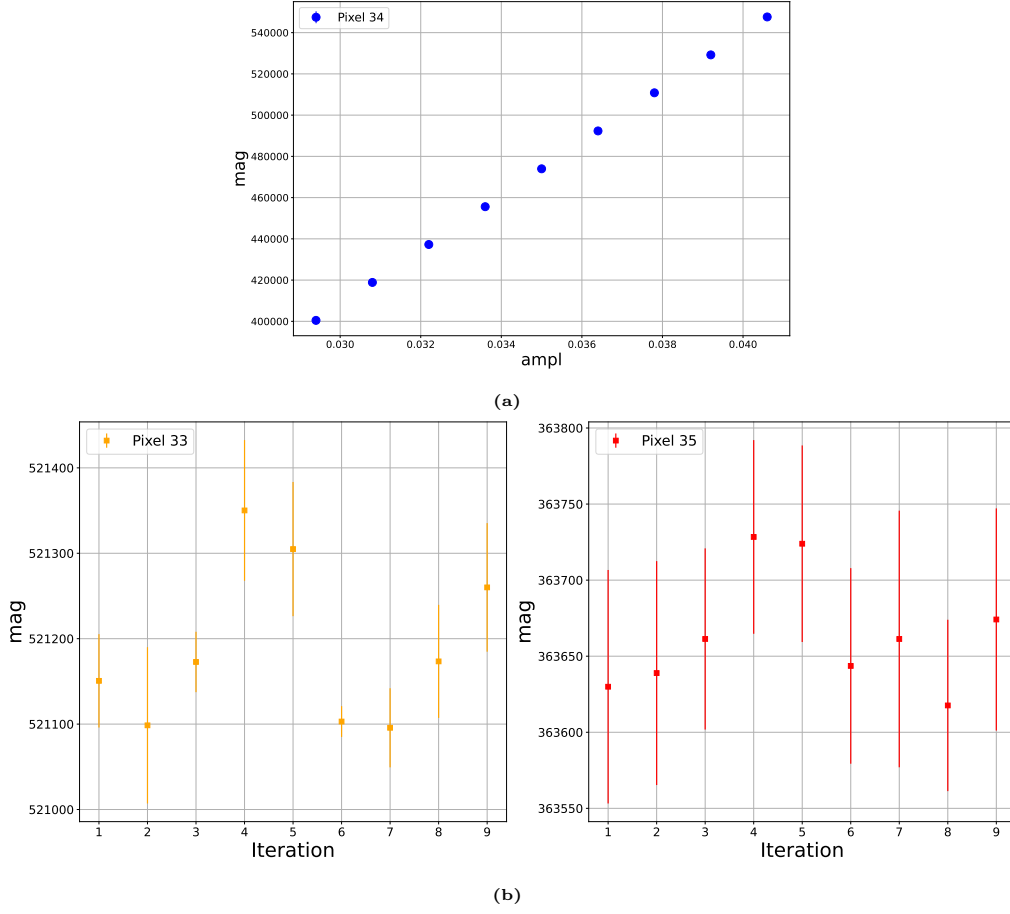


Figure 6.15: Crosstalk measurements for three resistors channels. (a) the perpetrator, following the Ohm's law when the input amplitude varies. (b) the victims output is not amplitude modulated. The error-bars were calculated as the standard deviation over the measured variation due to noise. These are not significant on the perpetrator.

In the second case (with the resistor acting as the perpetrator on channel i and the TESs as victims on channels $i \pm 1$), the output at frequencies $\omega_{i\pm 1}$ is expected to change. Due to electrical crosstalk, part of the current at ω_i flows into the adjacent channels $i \pm 1$. As a result, the amplitude modulation of the bias voltage at ω_i modifies the power dissipated in these TESs. This variation alters their resistance and, consequently, the effective impedance of the circuit at $\omega_{i\pm 1}$. For example, if the bias voltage $V(\omega_i)$ is decreased, the current $I(\omega_i)$ through the circuit will also decrease. The power dissipated in $R_{TES,i\pm 1}$ will be reduced as well. This leads to the shrinking of the resistance of these TESs and, accordingly, to a decrease of the impedance $Z(\omega_{i\pm 1})$. As a consequence, the current $I(\omega_{i\pm 1})$ increases. To summarize, a reduction of the input amplitude $V(\omega_i)$ results in a increase of the current measured at $\omega_{i\pm 1}$. Schematically:

$$V(\omega_i) \Downarrow \quad I_i(\omega_i) \Downarrow \quad I_{i\pm 1}(\omega_i) \Downarrow \quad P_{i\pm 1}(\omega_i) \Downarrow \quad R_{TES,i\pm 1} \Downarrow \quad Z(\omega_{i\pm 1}) \Downarrow \quad I(\omega_{i\pm 1}) \Uparrow \quad (6.4)$$

The measurements were carried out by decreasing the bias amplitude at the input of pixel 21 while recording the output of pixels 20, 21, and 22, as shown in Figure 6.16. The plots display a response that is consistent with the expected behavior. The victims plots in Figure 6.16(b) shows the steps corresponding to the reduction of the perpetrator amplitude on the x -axis: 1 corresponds to the largest amplitude, 9 to the smallest one.

Electrical crosstalk can be quantified by performing a linear fit of the victim signal

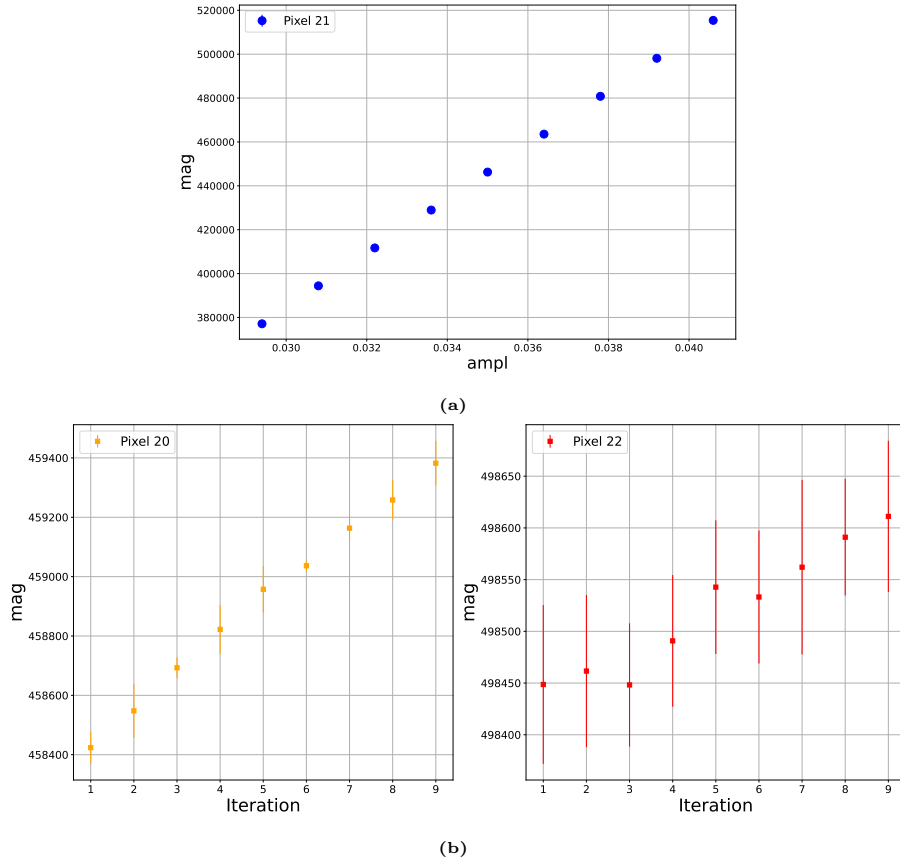


Figure 6.16: Crosstalk measurements for one resistor and two TESs channels. (a) the perpetrator, following the Ohm's law when the input amplitude varies. (b) the victims outputs modulated accordingly. The error-bars were calculated as the standard deviation over the measured variation due to noise. These are not significant on the perpetrator.

magnitude as a function of the perpetrator signal magnitude, with the crosstalk level determined from the slope of the resulting fit: the quantity is $\Delta I_{vict}(\omega_{vict})/\Delta I_{perp}(\omega_{perp})$ considering an offset. The results are presented in Table 6.5. Each row corresponds to the crosstalk between pixel 21 and either pixel 20 or pixel 22. The frequency spacing df between the perpetrator and the victim is listed, along with the measured electrical crosstalk.

The last case to be examined is the one in which the input of the TES is amplitude

#Ch	df [kHz]	Crosstalk level
20	34.84 ± 0.12	$(6.37 \pm 0.24) \times 10^{-3}$
22	71.87 ± 0.12	$(0.96 \pm 0.16) \times 10^{-3}$

Table 6.5: Level of crosstalk between pixels 21-20 and pixels 21-22. The frequency spacing, the level of measured crosstalk are presented.

modulated, and the outputs of the neighboring resistors are read out. These measurements were carried out with channels 10 and 12 as victims, and channel 11 as the perpetrator.

When the bias voltage on the TES is changed, the resistance of the detector $R_{TES,i}$ varies accordingly. This also modifies the impedance of the circuit $Z(\omega_{i\pm 1})$, causing a modulation of the output at the frequencies $\omega_{i\pm 1}$. In practice, assume that the TES is biased to operate along the transition slope, which, as shown in Figure 6.14, is defined as the region where a decrease in voltage amplitude corresponds to an increase in current magnitude. When the input voltage $V(\omega_i)$ is decreased, $R_{TES,i}$ decreases and the current $I(\omega_i)$ increases. Due to the variation in the TES resistance, the impedance of the circuit decreases at all frequencies, including $\omega_{i\pm 1}$. As a consequence, the current at $\omega_{i\pm 1}$ is expected to increase as well. Schematically:

$$V(\omega_i) \downarrow \quad R_{TES,i} \downarrow \quad I_i(\omega_i) \uparrow \quad Z_i(\omega_{i\pm 1}) \downarrow \quad I(\omega_{i\pm 1}) \uparrow \quad (6.5)$$

From the results presented in Figure 6.17, an opposite trend can be observed. When, in transition, the input bias voltage on the TES decreases, the output magnitude of the adjacent channels is reduced. The x -axis of the victim pixels plots reports the iterations corresponding to each successive decrease in the TES input amplitude.

Numerical simulations were performed to verify this dependence. For instance, by solving the circuit equations, one can determine how the impedance and current of the circuit at ω_{10} vary as $R_{TES,11}$ changes.

$$Z_{par}(\omega_{10}) = \left(\frac{1}{R_{10} + j\omega_{10}L_{10} + \frac{1}{j\omega_{10}C_{10}}} + \frac{1}{R_{TES,11} + j\omega_{10}L_{11} + \frac{1}{j\omega_{10}C_{11}}} \right)^{-1} \quad (6.6)$$

$$I(\omega_{10}) = I_0(\omega_{10}) \frac{1}{1 + \frac{L_{stray} + Z_{par}(\omega_{10})}{R_{bias}}} \quad (6.7)$$

where L_k and C_k , with $k = 10, 11$, are the inductance and capacitance value of the LC filters in leg 10 and 11. Figure 6.18 shows results obtained if $R_{TES,11}$ is decreased from 0.8 to 0.4 Ω , considering $f_{10} = 2132.8$ kHz and $f_{11} = 2185.52$ kHz, which are

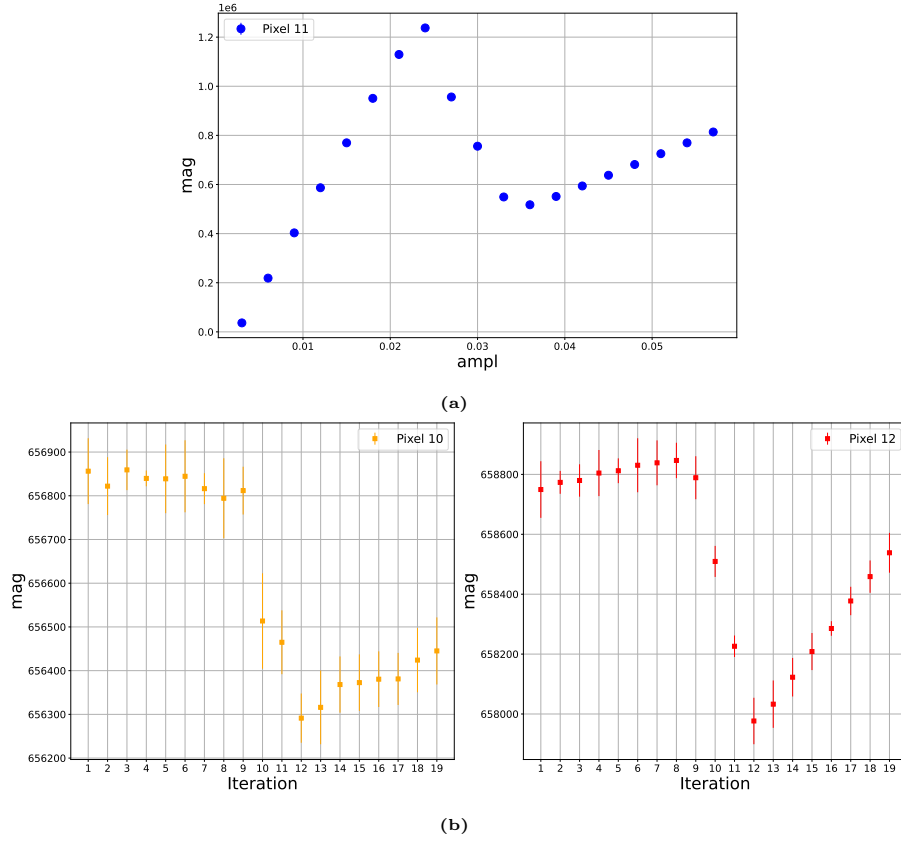


Figure 6.17: Crosstalk measurements for one TES and two resistors channels. (a) the perpetrator, following the I-V curve when the input amplitude varies. (b) the victims output. The error-bars were calculated as the standard deviation over the measured variation due to noise. These are not significant on the perpetrator.

the realistic values for the LC filters resonance frequencies considered in the measurements. The numerical results show consistencies with the measurements. The

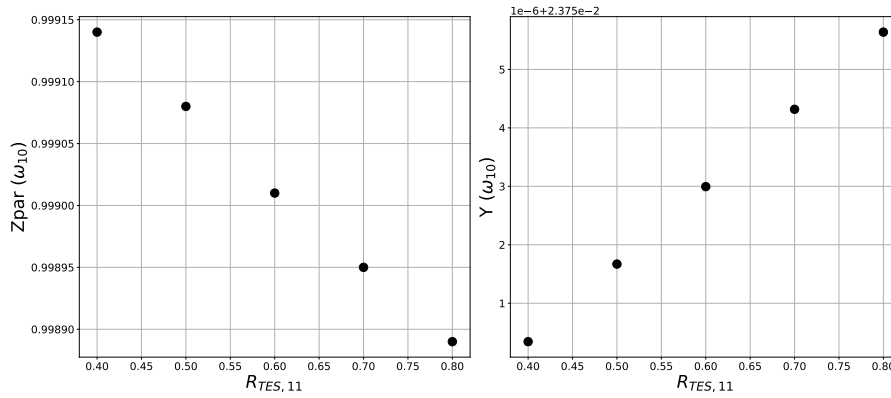


Figure 6.18: Variation of the impedance $Z_{par}(\omega_{10})$ (left) and admittance $Y(\omega_{10})$ (right) when the TES resistance $R_{TES,11}$ varies.

difference between the expectation and the results lies in the fact that the channels cannot be treated as independent branches. The proximity of the two resonances

causes a correlated variation in the impedance. When $R_{TES,11}$ is increased (simulating a grow in $V(\omega_{11})$), the sum of the two parallel legs at ω_{10} impedance decreases, causing the admittance to increase.

The level of crosstalk is calculated for pixel 11 coupling with either pixel 10 and 12, the results are shown in Table 6.6. The frequency spacing between the pixels is reported. It can be noticed also in this case that, the more the pixels are close in the frequency domain, the more the level of crosstalk is increased. The order of magnitude is however consistent with the results related to pixel 20, 21 and 22, presented in Table 6.5, and is in both cases around 10^{-3} .

#Ch	df [kHz]	Crosstalk level
10	52.73 ± 0.11	$(0.69 \pm 0.16) \times 10^{-3}$
12	48.93 ± 0.15	$(1.19 \pm 0.09) \times 10^{-3}$

Table 6.6: Level of crosstalk between pixels 11-10 and pixels 11-12. The frequency spacing, the level of measured crosstalk are presented.

6.2.4 Magnitude and in phase signals for crosstalk analysis

In this evaluation, we considered the total magnitude of the output signals to calculate electrical crosstalk, including all phase components. An alternative approach would be to consider only the victim outputs projected in phase with the perpetrator, since, after demodulation, only the in phase component is physically relevant. At this stage, however, the readout chain is not yet fully characterized or complete, and therefore it is not certain that the measured response strictly corresponds to the in phase component.

For this reason, we chose to use the total magnitude, which accounts for all possible signal contributions.

6.3 Conclusions

The measurements carried out on an end-to-end readout chain, derived from the LiteBIRD design, have been presented in this chapter. The results related to the preliminary noise characterization and the SQUID properties were displayed, showing the limits and possible improvements. The measurements on the electrical crosstalk level were presented and the preliminary results obtained were shown. The next

steps will focus on further investigating electrical crosstalk and create a solid procedure for the crosstalk level estimate.

Chapter 7

Propagation of crosstalk on the CMB maps and spectra

In order to evaluate how electrical crosstalk propagates into the final observational results, we need to assess its impact on the CMB temperature and polarization power spectra.

In Chapters 1 and 4 the theory describing the CMB power spectra was illustrated, with an insight on how the effect of crosstalk can be propagated in these. In this chapter the methods that the LiteBIRD collaboration provides to simulate these effects and the pipeline used to perform these calculations will be presented. An approach to calibrate and eliminate crosstalk from the power spectra is proposed.

7.1 Pipeline to simulate the propagation of crosstalk on the maps

The LiteBIRD simulation framework (LBS) ([60], [61]) is developed in Python and coupled to AstroPy and Healpix to track time, convert coordinates and handle sky maps.

As mentioned in the previous sections, the simulations are built from a time-order data (TOD) stream for each detector. The classes *Observation* and *Simulation* produce a complete simulation starting from the given data: the frequency channel, the polarization orientation and the position, together with all the other characteristics of the detectors and focal planes. The user can choose to include the $1/f$ noise or the foregrounds components such as the synchrotron radiation and the thermal dust emission. The resolution (typically $N_{side} = 512$), the length of the observation (1

year in our specific case) and the sampling rate (19 Hz for LiteBIRD) are set. As reported in Chapter 4, the electrical crosstalk affects the TODs in the following way:

$$\tilde{d}^j = \sum_i X_{ij} d^i \quad (7.1)$$

where X_{ij} is the element of the crosstalk matrix, d^i the TOD of the i -th detector, $\tilde{d}^j$ is the TOD of the j -th detector with crosstalk. Based on these “new” TODs, the sky maps and the power spectra can be built.

First of all, the choice of the detectors is crucial, and will be done with precise denominations which contain all the information.

From the TODs, we can build the sky coverage map for each detector, as well as the binned maps, either including or excluding the electrical crosstalk. One map for each Stokes parameter (denoted as I , Q and U) is then calculated using the *binned* function from the *Healpix* library. This function assigns the time-ordered data to the corresponding pixels on the sky, averaging all the observations data that fall within the same pixel, and producing a map that summarizes the signal measured by each detector.

From these maps, the power spectra, defined by the coefficients C_ℓ^{TT} , C_ℓ^{EE} , and C_ℓ^{BB} as functions of the multipole ℓ , are computed using the *anafast* function from *AstroPy*, which quantifies how the variance of the signal is distributed across different angular scales. In other words, the *anafast* function measures the amount of power contained at each multipole moment ℓ , providing a statistical description of the temperature and polarization anisotropies observed in the sky.

7.2 Simulations and results

The simulations were performed with two detectors, adjacent in the multiplexing frequency domain. We take into consideration the design and the multiplexing scheduling described in Chapter 4, section 4.1.1. Two electrical crosstalk matrices were considered in these simulations: 1) the carrier leakage crosstalk, 2) the common impedance crosstalk. They are shown in Figure 7.1. We focused on these two contributors because they represent the lowest and highest levels of crosstalk for the design considered, and their different impacts on the power spectra can be studied. In these matrices the detectors are ordered by multiplexing frequency, the number 1 corresponds to the LC filter resonating at 1.6 MHz, the number 34 to the one resonating at 5.5 MHz. As mentioned in Chapter 4, the multiplexing resonance

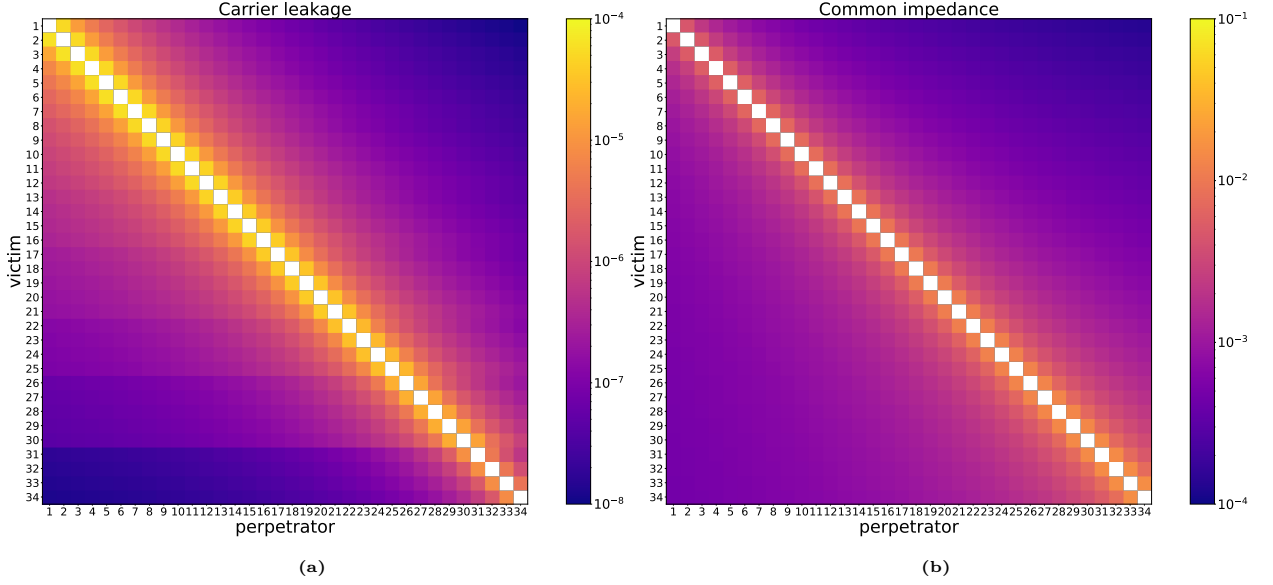


Figure 7.1: Carrier leakage and common impedance crosstalk matrices for the 34 LC configuration calculated in Chapter 4

frequencies is defined starting from the detectors sequence. The LiteBIRD sub-assembly LF12 is composed of optical and dark TESs with 2 calibration resistors in a specific sequence¹.

As extensively explained in Chapter 2, the majority of the LiteBIRD detectors are connected to antennas oriented to detect the sky signal with a specific polarization angle and a bandpass filter to select the frequency channel. The sequence of the detectors routing follows the position on the chip; an example of TESs chip for LF12 is shown in Figures 2.14 and 4.3. In this context, the term “pixels” refers to clusters of two or three detectors connected to the same antenna.

This implies that detectors within the same pixel and observing the sky in the same frequency channel, are oriented with a relative polarization angle of 90° . Further details are provided in Chapters 2 and 4 [25]. Detectors belonging to different pixels may instead be oriented at a relative angle of 45° .

7.2.1 Detectors on the same pixel with 90° polarization separation and identical frequency channel

Given the list of detectors, the element 3 and 4 in the multiplexing frequency sequence, whose crosstalk levels are represented by the X_{34} and X_{43} elements in the

¹The LF12 detectors sequence was obtained via private correspondence with Tommaso Ghigna.

matrices of Figure 7.1, correspond to the 000_000_006_UA_040_B and 000_000_006_UA_040_T detectors, respectively. As can be seen in Table 7.1, these denominations contain all the useful information needed for the simulations. From these, the simulator extracts the position of the detectors on the focal plane and the angle of the direction of observation in respect to the boresight of the satellite. Moreover, the number of the pixel and the frequency channel can be retrieved.

The sky coverage in galactic coordinates for the detectors on pixel 6 observing at

Pixel	6
Orientation	U
Handedness	A
Frequency channel	40 GHz
Polarization angle	B (0°) or T (90°)

Table 7.1: Informations contained in the names 000_000_006_UA_040_B and 000_000_006_UA_040_T of the two detectors.

40 GHz, is shown by a hit map in Figure 7.3. The time-order data (TODs) for both detectors corresponding to one year of simulated observation are presented in Figure 7.2. The spikes in the plot represent the signal from the Galaxy, which exhibits a higher intensity.

As already discussed in this work, the detection of CMB B-modes is particularly

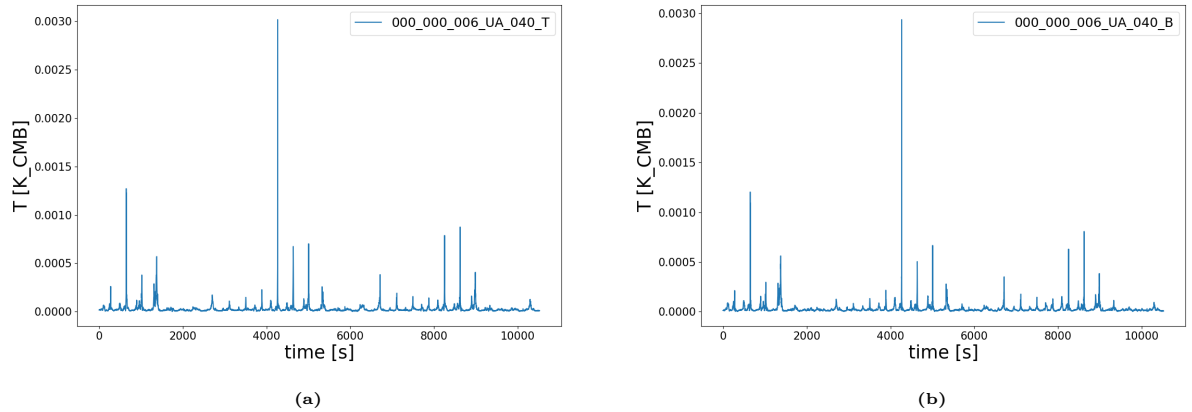


Figure 7.2: TODs of two detectors on pixel 6 observing the sky in the 40 GHz channel.

challenging: not only the signal is extremely faint, but it is also obscured by polarized galactic radiation emitted in the same frequency bands. This is visible in Figure 7.4. The plot illustrates the synchrotron radiation dominating and the thermal dust emission being negligible below ~ 80 GHz. The input maps for the three

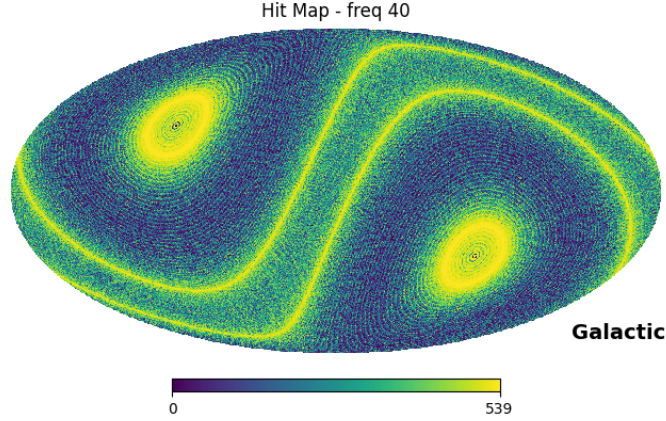


Figure 7.3: Sky coverage (galactic coordinates) for detectors belonging to pixel 6 observing the sky at 40 GHz.

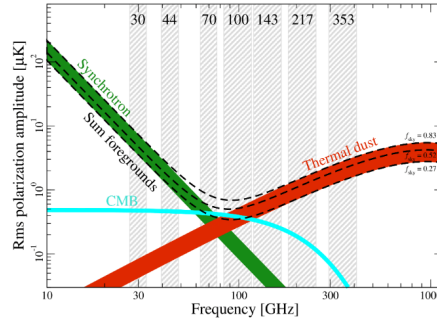


Figure 7.4: The frequency spectrum of the polarization amplitude for the CMB B-modes (light blue), the synchrotron (green) and Thermal dust radiation (red).

Stokes parameters, containing both the CMB and synchrotron radiation signals and derived from the TODs, are presented in Figure 7.5.

Crosstalk is injected into the TODs, and residual maps are obtained as the difference

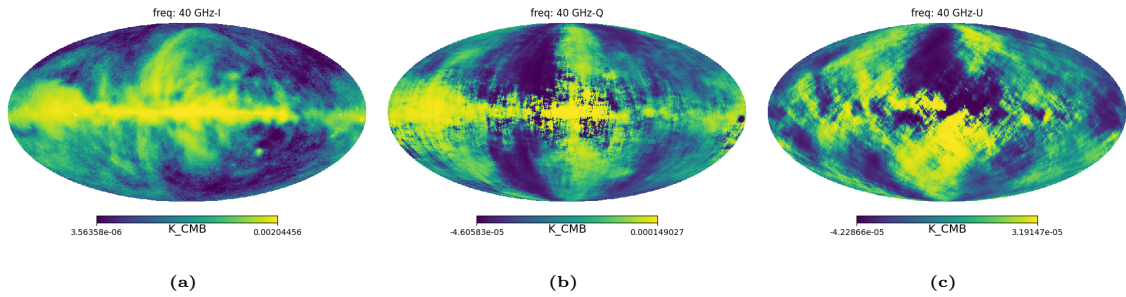


Figure 7.5: The input maps built from the detectors TODs for each Stokes parameter at 40 GHz: I (a), Q (b) and U (c).

between maps produced with and without crosstalk. This calculation is carried out for both carrier leakage and common impedance crosstalk, in order to study their

impact on the results. Since the differences between detectors observing the sky in the same frequency channel are negligible, we show the power spectra for a single detector only. The results are presented in Figure 7.6.

The plots show C_ℓ^{TT} , C_ℓ^{EE} , and C_ℓ^{BB} computed from the input maps, alongside the

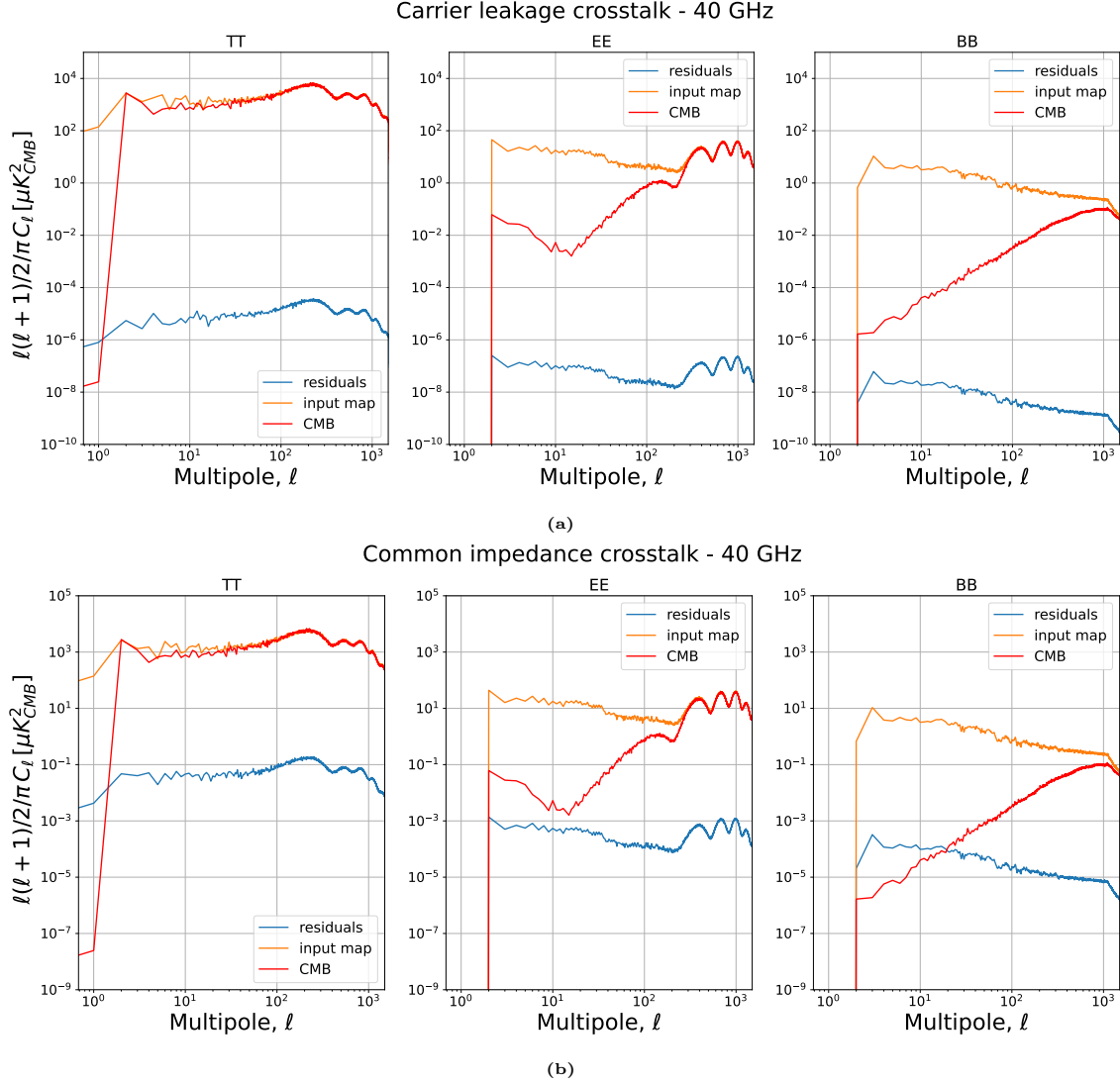


Figure 7.6: Power spectra for the TT, EE and BB correlation components. The residuals for crosstalk are shown with blue line, the synchrotron signal and the CMB radiation (in the BB case coming from gravitational lensing of the E-modes) at 40 GHz. The input detectors are 000_000_006_UA_040_B and 000_000_006_UA_040_T.

crosstalk residuals derived from the residual maps, which quantify the impact of the electrical crosstalk. For reference, the CMB signal alone is also shown. The CMB curve in the BB spectrum corresponds to the B -modes generated by gravitational lensing of polarized E-modes.

To estimate the impact of electrical crosstalk on the BB power spectrum, the integration of the residual curves is carried out. In Table 7.2 the results of the calculations

are shown for both the crosstalk contributors. Since the carrier leakage crosstalk is smaller than the common impedance contribution, its propagation onto the power spectra results in curves with reduced amplitude, as expected.

Crosstalk type	Crosstalk level [X_{34}]	Res. Amplitude [μK_{CMB}^2]
Car. leak.	6.3×10^{-5}	1.51×10^{-6}
Com. imp.	5.2×10^{-3}	1.4×10^{-2}

Table 7.2: Amplitude values obtained by integrating the residual curves on the BB spectrum for two detectors on pixel 6.

7.2.2 Detectors on different pixels with 45° polarization separation and identical frequency channel

The simulations were repeated to study the impact of crosstalk to the spectra for two detectors hosted in two different pixels, for instance pixels 6 and 3, but observing the sky in the same frequency channel. These pixels are spatially adjacent, and their antennas are oriented at a relative angle of 45 degrees. The names of the two detectors are 000_000_006_UA_040_T and 000_000_003_QA_040 and the TODs are shown in Figure 7.7. The power spectra corresponding to these detectors are presented in Figure 7.8.

The evaluation of the level of crosstalk on C_ℓ^{BB} is performed with the same integra-

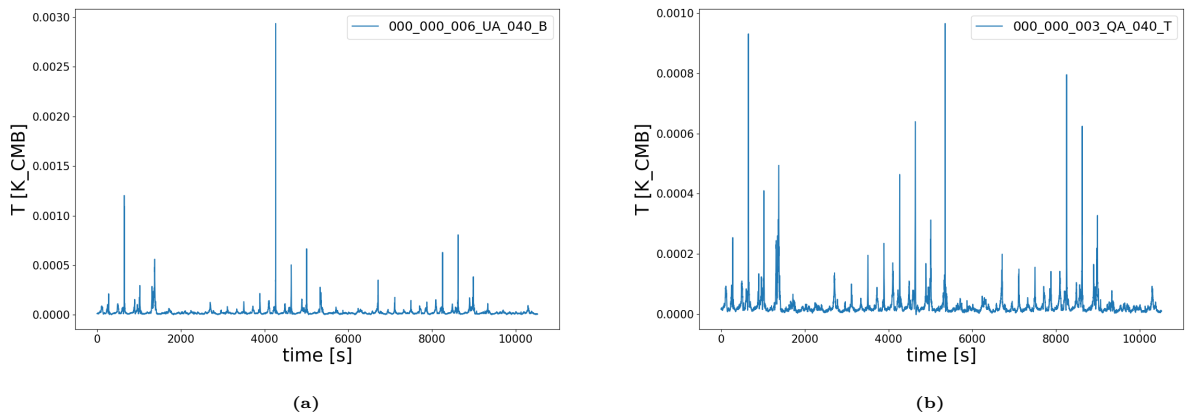


Figure 7.7: TODs of detectors on pixels 6 and 3 observing the sky in the 40 GHz channel.

tion method. Results are presented in Table 7.3. Both components, corresponding

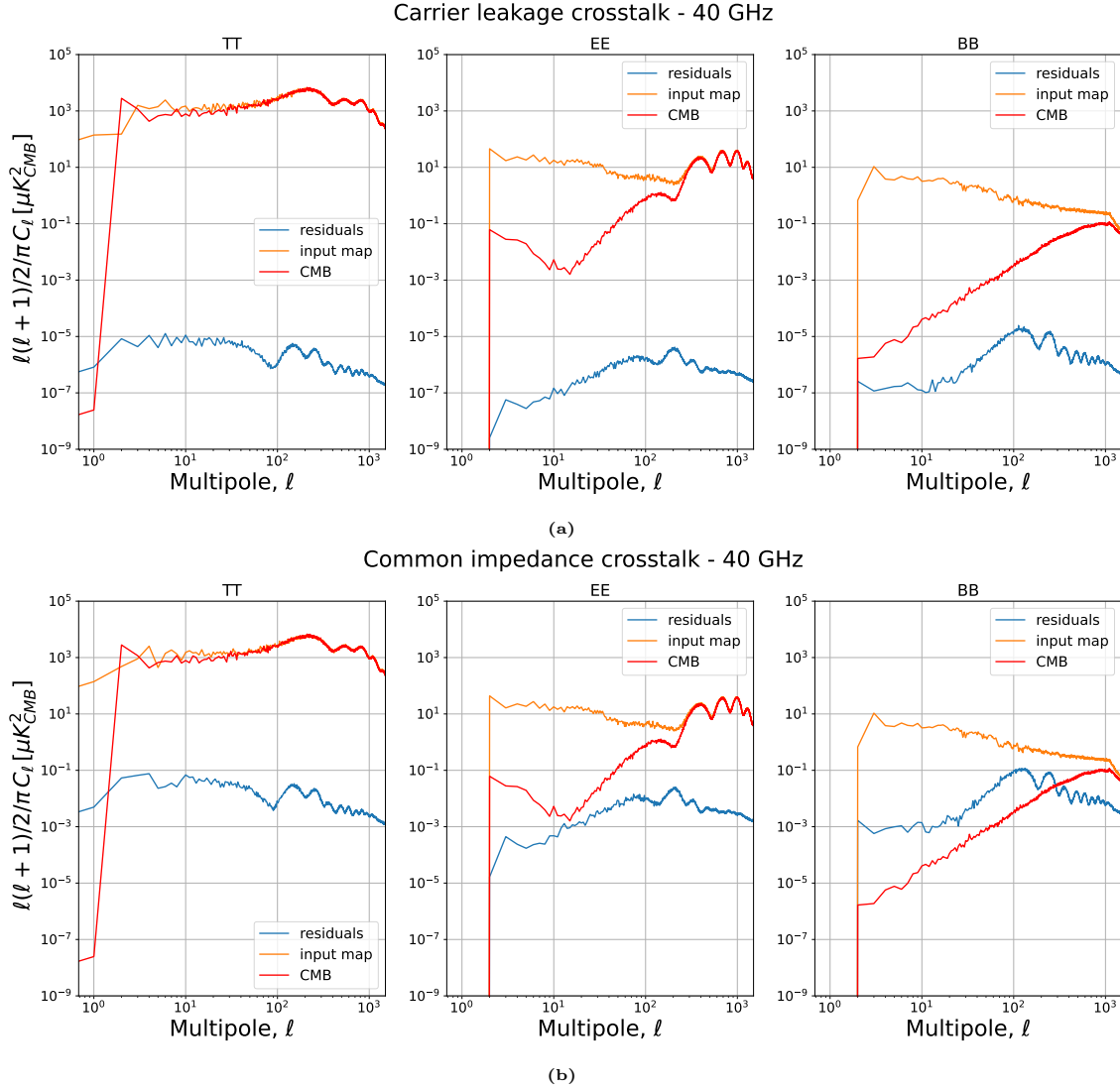


Figure 7.8: Power spectra for the TT, EE and BB correlation components. The residuals for crosstalk are shown with blue line, the synchrotron signal and the CMB radiation (in the BB case coming from gravitational lensing of the E-modes) at 40 GHz. The two detectors are 000.000.006.UA.040.T and 000.000.003.QA.040.T.

to carrier leakage and common impedance crosstalk, increase compared to the previous case shown in Table 7.2. Electrical crosstalk has more impact on detectors belonging to separate pixels. Regarding the rules enumerated in Chapter 4, the first one is violated in this case: the detectors are adjacent in the multiplexing frequency sequence but belong to different pixels.

Since the antennas are spatially separated, an additional effect is introduced: the propagation of their relative positions on the focal plane into differences in the observation directions. Considering realistic parameters for the LiteBIRD telescopes, we can estimate the multipole moment ℓ at which this effect becomes relevant. For a stop size of 300 mm, an f-ratio of 2.2, a focal length of 660 mm and a spatial

separation on the focal plane of 32 mm, the corresponding angular spacing on the sky is approximately $\theta \sim 2.75^\circ$, which correspond to $\ell = 65$, being $\ell \simeq \pi/\theta$. A small bump appears in the spectra around this multipole value and could be attributed to this effect.

Crosstalk type	Crosstalk level [X_{45}]	Res. Amplitude [μK_{CMB}^2]
Car. leak.	6.6×10^{-5}	1.09×10^{-4}
Com. imp.	5.5×10^{-3}	6.57×10^{-1}

Table 7.3: Amplitude values obtained by integrating the residual curves on the BB spectrum for a pairs of detectors belonging respectively to pixel 6 and 3.

7.2.3 Detectors on different pixels with 90° polarization separation and different frequency channels

The simulations were carried out considering two detectors on two different pixels (7 and 6) with a relative polarization orientation of 90 degrees and observing the sky in different frequency channels, specifically at 40 and 60 GHz. The input maps at 60 GHz, including CMB and synchrotron radiation, are presented in Figure 7.10 and the sky coverage at 60 GHz is shown in Figure 7.9. These figures complement those provided for the 40 GHz frequency channel. The two detectors are denomi-

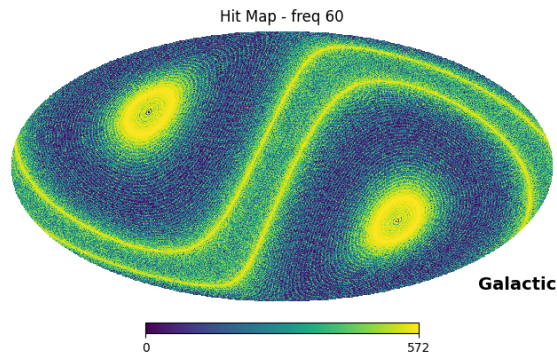


Figure 7.9: Sky coverage (galactic coordinates) for detectors belonging to pixel 7 observing the sky at 60 GHz.

nated 000_000_007_UB_060_B and 000_000_006_UA_040_T and the TODs plots are presented in Figure 7.11. The results related to the power spectra for the two frequency channels are presented in Figures 7.12 and 7.13. In Table 7.4 the results of the integration of the residuals amplitude on the BB power spectrum are shown.

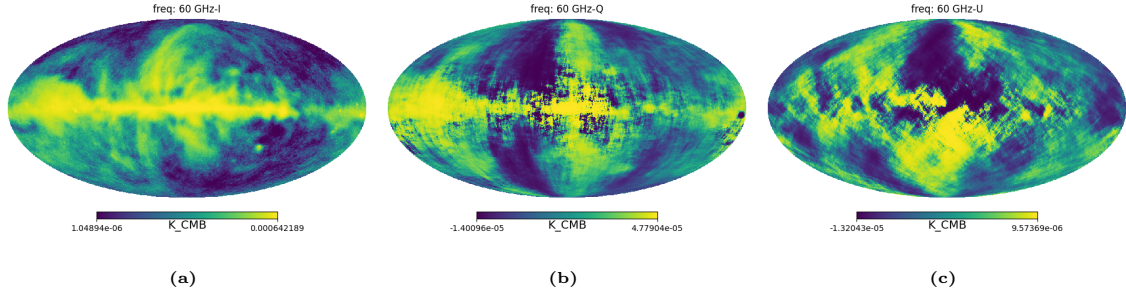


Figure 7.10: The input maps built from the detectors TODs for each Stokes parameter at 60 GHz: I (a), Q (b) and U (c).

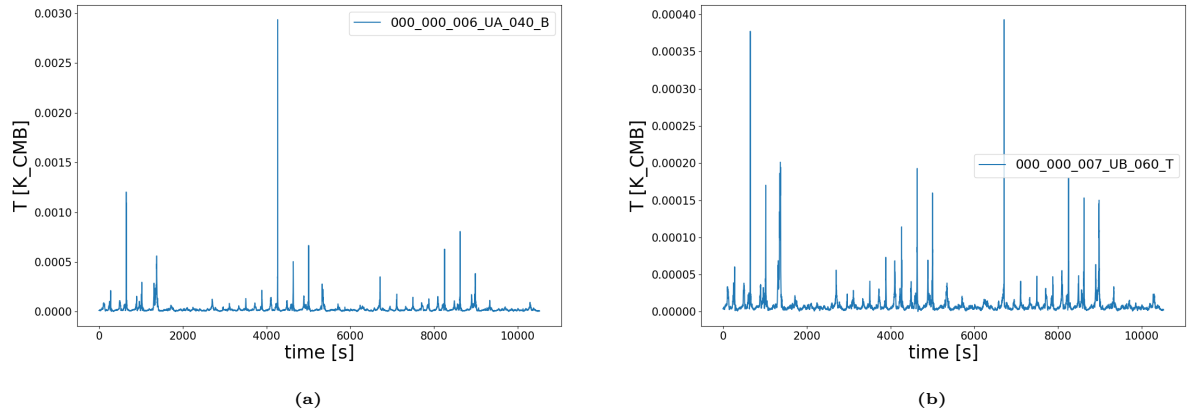


Figure 7.11: TODs of detectors on pixels 6 and 7 observing the sky in the 40 GHz channel.

The difference between the two frequency channels is related to variations in the spectral amplitude of the input signal, as well as to the level of crosstalk. The latter varies because, for each element of the crosstalk matrix, one detector is considered the perpetrator and the other the victim, and vice versa. The matrices do not show complete symmetry, as each element depends on the impedance and multiplexing frequency of the corresponding channels.

Crosstalk type	Crosstalk level		Res. Amplitude [μK_{CMB}^2]	
	40 GHz [X_{32}]	60 GHz [X_{23}]	40 GHz	60 GHz
Car. leak.	6.2×10^{-5}	7.2×10^{-5}	7.12×10^{-6}	9.14×10^{-5}
Com. imp.	4.8×10^{-3}	5.2×10^{-3}	4.29×10^{-2}	4.7×10^{-1}

Table 7.4: Amplitude values obtained by integrating the residual curves on the BB spectrum for a pairs of detectors belonging respectively to pixel 6 and 7, observing the sky in the 40 GHz and 60 GHz frequency channels.

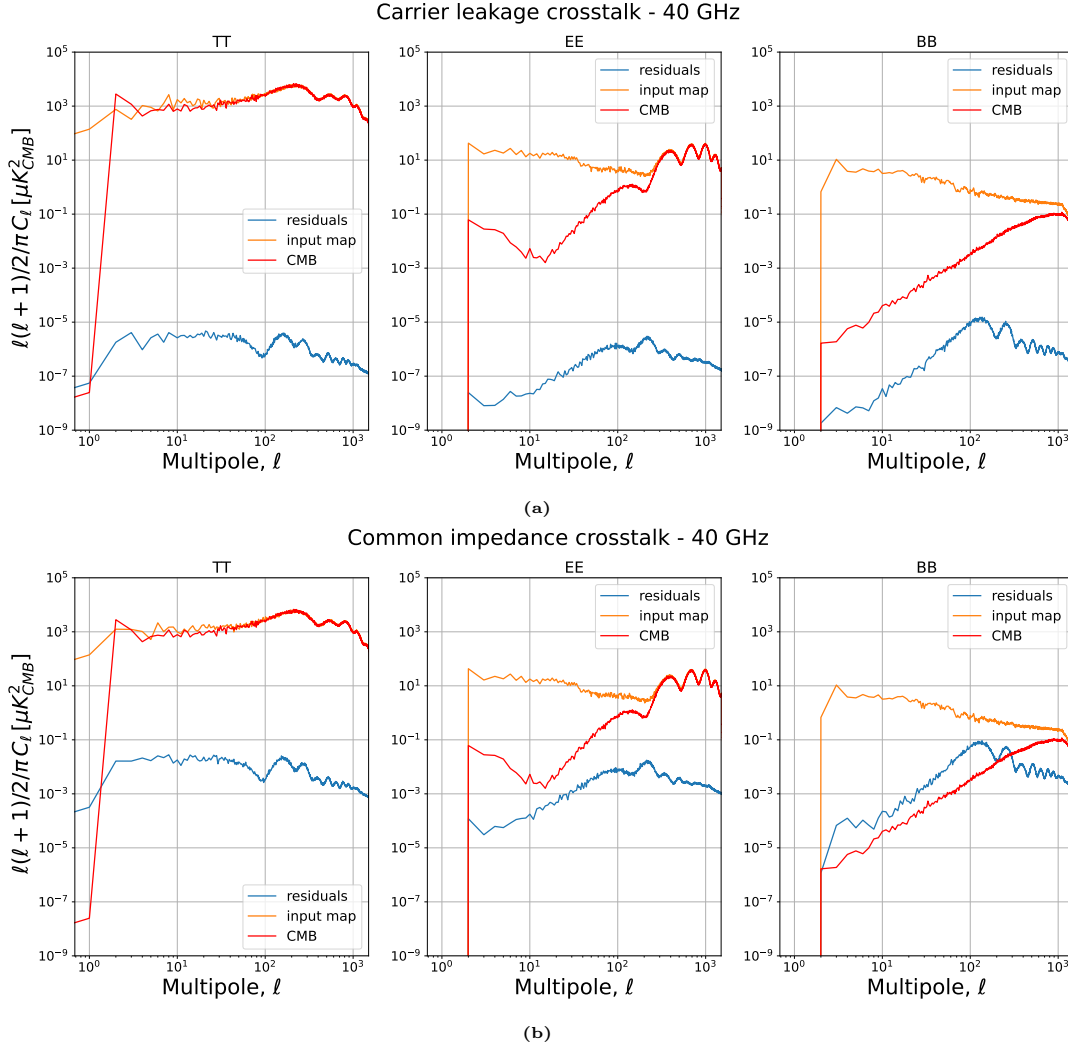


Figure 7.12: Power spectra for the TT, EE and BB correlation components. The residuals for crosstalk are shown with blue line, the synchrotron signal and the CMB radiation (in the BB case coming from gravitational lensing of the E-modes) at **40 GHz**. The two detectors are 000.000.006-UA.040-B and 000.000.007-UA.040-T.

7.2.4 Detectors on different pixels non adjacent in the multiplexing frequency sequence

The dependence of the power spectrum amplitude on the crosstalk level was investigated using simulations of detectors that are not immediate neighbors in the multiplexing frequency sequence, in contrast to the configuration used in the previous sections. Maps and corresponding power spectra were generated for detectors associated with crosstalk elements indexed as 4 and 6 (second neighbors) and 4 and 7 (third neighbors) in the matrices presented in Figure 7.1. The detector considered were 000.000.006-UA.040-B and 000.000.003-QA.040-T as second neighbor, and 000.000.006-UA.040-B and 000.000.003-QA.040-T as third neighbor. The result-

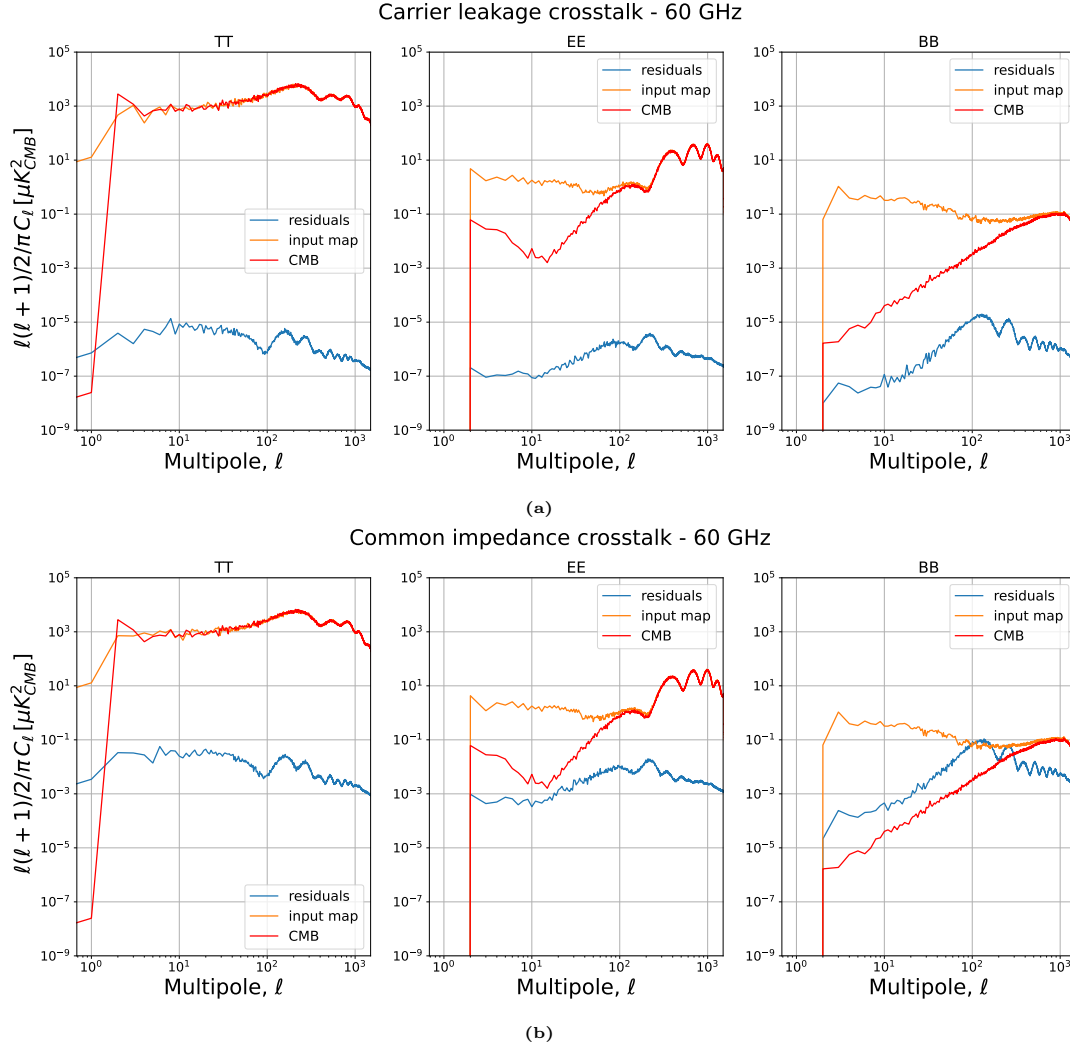


Figure 7.13: Power spectra for the TT, EE and BB correlation components. The residuals for crosstalk are shown with blue line, the synchrotron signal and the CMB radiation (in the BB case coming from gravitational lensing of the E-modes) at **60 GHz**. The two detectors are 000_000_006_UA_040_B and 000_000_007_UA_040_T.

ing integration from the residuals curve, for carrier leakage and common impedance crosstalk, is presented in Table 7.5. The amplitude of the residuals decreases as the

Index	Crosstalk		Res. Amplitude [μK_{CMB}^2]	
	Car. leak.	Com. imp.	Car. leak.	Com. imp.
$[X_{46}]$	1.48×10^{-5}	2.59×10^{-3}	7.66×10^{-6}	1.64×10^{-1}
$[X_{47}]$	6.2×10^{-6}	1.73×10^{-3}	1.65×10^{-6}	8.01×10^{-2}

Table 7.5: Amplitude values obtained by integrating the residual curves on the BB spectrum for a pairs of detectors non adjacent in the multiplexing frequency sequence.

crosstalk level is reduced, as expected.

7.3 Mitigation of Electrical Crosstalk

The procedure for isolating and mitigating the electrical crosstalk effect on the power spectra is carried out at the TOD level. The crosstalk measurement methods described in the previous sections constitute the preliminary steps for subtracting this systematic effect from the power spectra. Once the measured crosstalk matrix is created, it can be inverted and multiplied by the TODs, derived from the observational campaign data, to “subtract” the crosstalk contribution.

$$d_r^j = \sum_i X_{ij,meas}^{-1} \tilde{d}^i \quad (7.2)$$

where $\tilde{d}^i$ denotes the TODs calculated with Eq. 7.1 and represents the TODs with crosstalk, $X_{ij,meas}$ is the measured crosstalk coefficient for each detector pair and d_r^j is the TOD after the removal of crosstalk.

In the simulations, this procedure is modeled by taking the theoretical analytical crosstalk matrix and creating a modified version whose elements are taken from a normal distribution centered on the first matrix values and with a standard deviation in the range from 1% to 5% percent of the mean value. Denoting the modified matrix as $X_{ij,smeared}$, the resulting TOD for the j -th detector is:

$$d_r^j = \sum_i X_{ij,smeared}^{-1} \tilde{d}^i \quad (7.3)$$

where d_r^j is the TOD of the j -th detector, where the crosstalk contribution has been “removed”. At this stage, this element and the original input TOD are expected to be approximately the same.

An example is illustrated in Figure 7.14, calculated using the detectors 000_000_006_UA_040_B and 000_000_006_UA_040_T and the common impedance crosstalk matrix. The input power spectra without crosstalk, the power spectra obtained after crosstalk removal, and the resulting module of the difference between the two curve amplitudes are presented in the plot. The first two curves coincide within the plotted scale, with the residuals illustrating the remaining differences.

As mentioned, this technique mimics a possible calibration method for the electrical crosstalk starting from a crosstalk matrix. The smearing of the matrix represents the possible not perfect correspondence between the measured matrix and the actual crosstalk level of the system. In general, the lower the residuals, the more effective the method is at removing the crosstalk contribution from the spectra. Residuals that are larger than expected could indicate incomplete correction or systematic

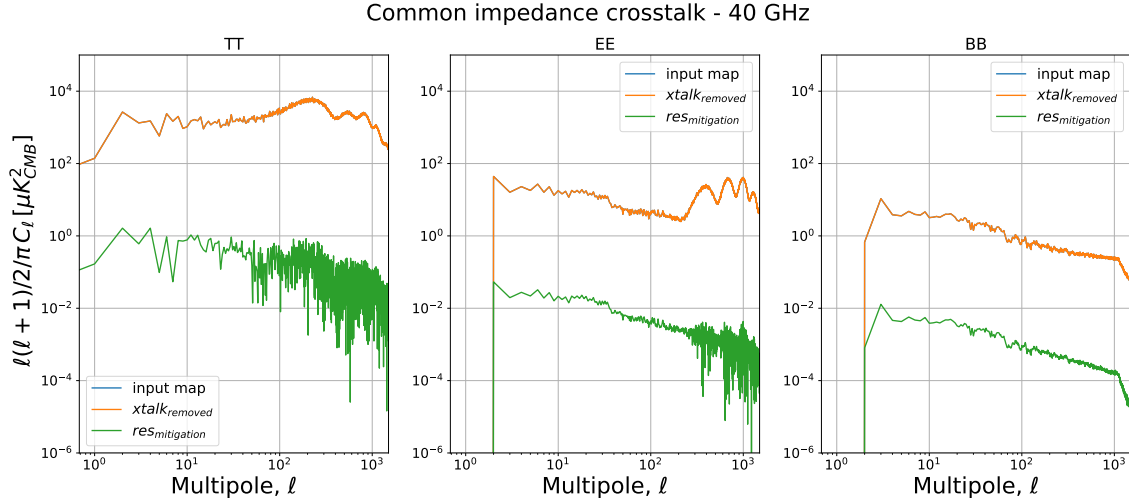


Figure 7.14: Power spectra for C_ℓ^{TT} , C_ℓ^{EE} and C_ℓ^{BB} . The orange and blue line overlap. The green line represents the residual between the two curves. The simulations were carried out using the 000_000_006_UA_040_B and 000_000_006_UA_040_T detectors, with common impedance crosstalk.

effects, and therefore require further investigation to ensure the crosstalk has been fully mitigated.

7.4 Conclusions

In this chapter the pipeline adopted in this work to inject the crosstalk in the TODs and to estimate its impact on the power spectra was presented. The preliminary results for different types of detector pairs were illustrated. We also described a procedure to calibrate and mitigate the crosstalk at the TODs level. The impact of the residuals must be investigated in more detail.

The next steps of this work will include further studies to evaluate the impact of the level of crosstalk on the mixing between frequency channels and polarization orientations. More simulations on the mitigation techniques will be run to assess the implementation of this technique.

Chapter 8

Conclusions

In this thesis we conducted an extensive investigation of the instrumental electrical crosstalk effect for a TES detector chain readout by Frequency Division Multiplexing, examining different aspects from multiple perspectives.

In order to provide context, the motivations and advantages to implement Frequency Division Multiplexing (FDM) were discussed, together with the techniques related to the design of the implemented band-pass filters. Two different configurations, using interdigitated and plane parallel capacitors, for LC filters chips entirely fabricated by lithography were described. The contributions to electrical crosstalk related to this kind of designs were enumerated, providing analytical model and calculations for each one. As discussed, the level of each component relative to the others was distinguished, and the common impedance crosstalk was recognized to be the dominant one.

We presented a method to optimize the multiplexing frequency sequence by minimizing the impact of electrical crosstalk on the system. The scheduling of the multiplexing frequencies, which corresponds to the LCs position on the chip, and the LC - TES assignment are crucial. We implemented an algorithm, on the model of the *simulating annealing*, to obtain the best LC filters configuration. The first results obtained on one of the LiteBIRD focal plane subassemblies were illustrated. The results look promising, since the combination found in a finite timescale mostly satisfy the rules. We also described the development and design of an LC filters chip, which consists of spiral inductors and interdigitated capacitors, and the corresponding chip was designed and fabricated.

Experimental measurements were performed on an end-to-end chain including TESs, an LC filters chip similar to the one described above, the SQUID and the LiteBIRD warm electronics. The results of these are discussed in this work. A preliminary

assessment of electrical crosstalk in this system was also presented. Although, as discussed, a more solid set-up for these measurements need to be implemented, the first results are consistent with the expected order of magnitude.

Similar measurements on another LC filters chip, consisting of spiral inductors and plane parallel capacitors, were carried out. The properties of the filters, such as the Q -factor, the intrinsic parasitic resistance and the *common inductance*, were quantified. An electrical crosstalk characterization was carried out. Different amplitude and frequency modulations were applied to the input line to study the leakage in the adjacent branches of the circuit. The results were generally consistent with the expected values, coming from previous works on similar configurations. The level of crosstalk retrieved is in the range expected from the analytical calculations.

Finally, the development of a pipeline to study the impact of crosstalk on the CMB power spectra was presented. Detectors with different characteristics were used to simulate the propagation of electrical crosstalk on the projected LiteBIRD observational results. A technique to calibrate and mitigate crosstalk was also illustrated. The resulting impact of crosstalk need to be studied in a broader context, by comparing its level to the one of other systematics.

The activities presented in this work can be further developed and constitute a foundation for several future studies. The investigation of electrical crosstalk can be extended in multiple directions, from improvement in the design, to systematic simulations for mitigation, going through more robust experimental measurements methods. Throughout the discussion of these aspects, we emphasized the strong interconnection between each step. Overall, this work is a comprehensive analysis that addresses different but well connected aspects and proposes possible future developments.

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