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Microwave SQUID Multiplexing for the Simons Observatory Cosmic Microwave
Background Telescopes

A dissertation submitted in partial satisfaction of the
requirements for the degree Doctor of Philosophy

in

Physics

by

Maximiliano Silva-Feaver

Committee in charge:

Professor Kam Arnold, Chair
Professor Raphael Flauger
Professor Ian Galton
Professor Karin Sandstrom
Professor Daniel Sievenpiper

2023

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2023

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ABSTRACT OF THE DISSERTATION

**Microwave SQUID Multiplexing for the Simons Observatory Cosmic
Microwave Background Telescopes**

by

Maximiliano Silva-Feaver

Doctor of Philosophy in Physics

University of California San Diego, 2023

Professor Kam Arnold, Chair

Measurements of the faint polarization signal from the cosmic microwave background (CMB) have become increasingly more sensitive largely due to the ability to package and read out large arrays of cryogenic detectors. The Simons Observatory, pushes this trend further by deploying a multiplexed readout technology called the

microwave squid multiplexer capable of reading out 1000 pixels on a single readout line; this has never previously been demonstrated. In this dissertation I overview the physics of cosmic microwave background, present the design and science reach of the Simons Observatory telescopes, present the design and integration of the multiplexer into the Simons Observatory telescopes, and present a predictive noise model for general multiplexer design. Lastly, I also outline the initial in field instrument commissioning data and pipeline work.

Chapter 1

Introduction to Cosmology and the Microwave Background

1.1 A Brief History of Cosmology

Cosmology is the study of our universe, including what its made of, how it was formed, how it evolves in time, and how it will end. Cosmological questions have captivated humankind since before recorded history. Historians have found that the earliest societies had formed some concept of our place in the wider universe. Early cosmological pictures were typically very earth, sun, or human-centric[3]. The study was largely religious or philosophical due to the limited observational evidence ob-

tainable with our eyes alone. In the early 20th century, this changed dramatically due to two major advances in physics that occurred around the same time: 1) the formulation by Albert Einstein of a General Theory of Relativity (GR)[4] in 1915, which shortly after its publication he turned toward the cosmological question (described further in section 1.2[5] and 2) the observation by Edwin Hubble 14 years later that our universe is filled with dense luminous “nebulae” which are all flying away from us with increasing rate as a function of distance[6].

These scientific advancements transformed Cosmology from a qualitative study reserved for religious and philosophical thinkers to a quantitative one that could be described by physical theories, formulated with mathematics, and capable of producing testable hypotheses. The GR picture allows our universe to be in one of three scenarios 1) expanding, 2) contracting, or 3) static, depending on the balance of the energy content of our universe. Hubble’s observations implied that our universe is expanding and that there must have been a time when all of the universe’s content was contained in a small volume. This was the first observational evidence for the picture of our universe’s origin that would become known as the big bang.

In the 1930-1940s, many theorists applied the recently developed quantum theory toward practical problems in nuclear physics[7], and experimentalists used early particle accelerators and radioactive sources to discover new isotopes and measure

nuclear decay rates and scattering cross-sections[8]. Separately, by 1937 using measurements of meteorites, earth samples, and the sun, chemists and physicists had decent constraints on the relative cosmic abundances of chemical elements[9]. In 1947-1948 Ralph Alpher, along with his colleague Robert Herman and former PhD advisor George Gamow used these new advances in theoretical and experimental nuclear physics to explain the cosmic chemical abundances through evolving the universe adiabatically from a very hot temperature ($\sim 10^8 - 10^{10} \text{ }^\circ\text{K}$) to modern day and balancing the rates of neutron scattering and capture given the mean temperature, the density of matter and radiation, and Hubble expansion rate. As a byproduct of Alpher, Herman, and Gamow's Big Bang Nucleosynthesis (BBN) theory, they predicted that the bath of radiation energy density left over from the big bang process is expected to be observable today as isotropic background radiation well described by a blackbody spectrum with a mean temperature of $\sim 5 \text{ }^\circ\text{K}$ [10, 11, 12]*. Seventeen years later, Arno Penzias and Robert Wilson published a measurement of isotropic excess antenna temperature of $3.5 \text{ }^\circ\text{K}$ [13] with a companion publication from Robert Dicke, James Peebles, Peter Roll, and David Wilkinson which connected this measurement to the cosmic microwave background (CMB) explanation of Alpher[14].

The initial measurement of the CMB in 1965 led to a series of successively more

*They were off by a bit in this initial estimate. We now know this temperature to be 2.725K , but given the experimental data, theoretical information, and computational methods they had at the time, this was shockingly close.

precise measurements of the CMB. So far, all measurements are consistent with a six-parameter description of our universe, the Λ -cold dark matter (Λ -CDM) model. In the following sections, I will quantitatively lay out these recent measurements, the Λ -CDM model, and the goals of contemporary experiments to search for extensions to Λ -CDM.

1.2 Homogeneous and Isotropic Universe

Einstein's theory of general relativity describes how space is curved due to the presence of a given energy-density distribution and, in turn, how matter will move in this curved space. The below generally follows [15]. Einstein's equation can be written as:

$$G_{\mu\nu} = 8\pi G T_{\mu\nu} \tag{1.1}$$

$G_{\mu\nu}$ is the Einstein tensor describing how space is curved, and $T_{\mu\nu}$ is the stress-energy tensor. There is no general solution to this equation, but for a few cases, we have closed-form solutions[15]. For the case of a homogeneous and isotropic universe, we can use the Freidman-Lemaitre-Robertson-Walker (FLRW) metric and solve Einstein's equation exactly. The FLRW metric (equation 1.2, diagonal in spherical polar coordinates) describes a space in which the distance between points grows in time

with a scale factor $a(t)$ [16]

$$g_{00} = -1 \quad g_{rr} = \frac{a(t)^2}{1 - Kr^2} \quad g_{\theta\theta} = a(t)^2 r^2 \quad g_{\phi\phi} = a(t)^2 r^2 \sin^2(\theta) \quad (1.2)$$

And the stress-energy tensor in an FLRW type is given by:

$$T^{00} = \rho(t) \quad T^{ij} = g^{ij}(\mathbf{x})a^{-2}(t)p(t) \quad T^{0i} = 0 \quad (1.3)$$

Which describes a perfect fluid with energy density, ρ , and pressure, p . The energy content can be broken down into three main constituents: 1) cold matter (dark matter and baryons), 2) hot matter (photons and relativistic particles), and 3) vacuum energy. Each of these can be described by an equation of state of the form $\rho = wp$ with $w = 0$ for cold matter, $-\frac{1}{3}$ for radiation, and -1 for vacuum. Given this stress-energy tensor, energy conservation implies a scaling of the energy density with the scale factor as $\rho \propto a^{-3-3w}$ or $\propto a^{-3}$ for cold matter, a^{-4} for radiation, and constant (often referred to as the cosmological constant, Λ) for vacuum energy.

The time evolution of the scale factor in a homogeneous universe is given by solving the 00 component of Einstein's equation:

$$\left(\frac{\dot{a}}{a}\right)^2 = \frac{8\pi G}{3}\rho - \frac{K}{a^2} \quad (1.4)$$

Equation 1.4 is typically written in terms of the Hubble constant, $H(t) = \dot{a}(t)/a(t)$. The previously undefined term K describes the curvature of our universe, $K = 1$ corresponds to a closed spherical universe, $K = -1$ is an open hyperspherical universe, and $K = 0$ is a flat universe. We expect K to take some value between -1 and 1, and $K = 0$ is a special case.

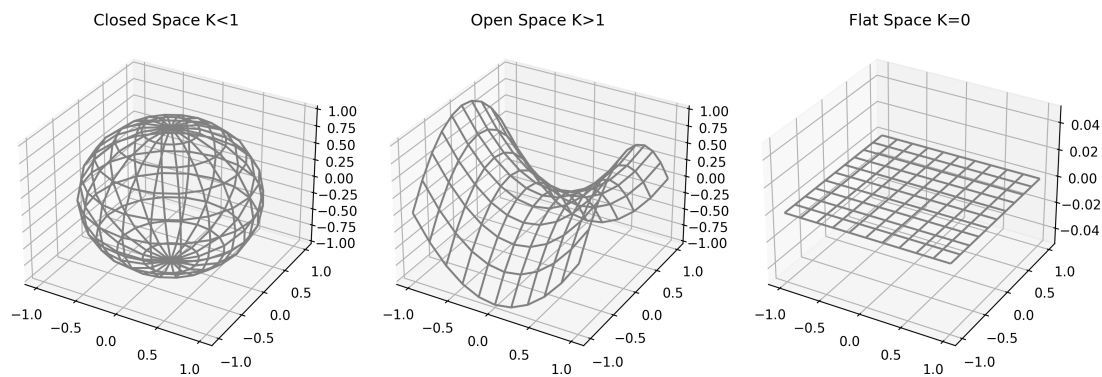


Figure 1.1: Schematic of a closed (left), open (center), and flat space corresponding to an FLRW metric with $K = 1, -1$, and 0 , respectively.

There is a critical density $\rho_{crit} = \frac{3H^2}{8\pi G_N}$ for which $K = 0$ and the universe is flat.

As cosmologists, we measure relative abundances of the constituent energy densities relative to this critical density $\Omega_i = \frac{\rho_i}{\rho_{crit}}$ where i takes the values m : Matter, γ : radiation, and Λ : vacuum. We observe that the sum of these three sources at the present day is extremely close to 1, implying we live in an exceptionally flat universe. Since in the context of GR this is such a unique case, we search for an explanation for that uniqueness – perhaps there is a physical mechanism that drives us to the flat

universe solution. The issue of why our universe is observed to be flat is referred to as the “flatness problem” in cosmology. A class of theories describing the dynamics in the first $\sim 10^{-30}$ seconds of our universe, inflation, described more in section 1.4.1, is one proposed mechanism to generate a flat universe.

Since the energy density of the various cosmological components dilute at different rates in an FLRW metric, and our universe is made up of all three components, our universe underwent three primary epochs during its expansion history for which we have a robust understanding: 1) radiation dominated (the first ~ 10 kyrs), 2) matter dominated (the next ~ 10 kyrs - 10 Gyrs), and 3) vacuum energy dominated (recently, in the last 1 Gyr) expansion.

1.3 Cosmic Microwave Background

After inflation and when the universe had cooled to the point that baryons were stable, and for the next 380,000 years, our universe consisted of a hot dense ionized plasma where all standard model particles (like photons, baryons, and neutrinos) were in thermal equilibrium by effectively exchanging energy through collisions and exchange interactions as described by the Boltzmann equation. As time evolves, the universe cools due to Hubble expansion, and populations of atomic nuclei and elements begin to form as the scattering rates for high energy photons to dissociate

them drop below the rates to bind. At $T \sim 1$ MeV, neutrons build up, and at $T \sim 0.1$ MeV, nuclei begin to form, leading to the formation of the light elements (H, He, Li, and their isotopes). However, these initial light elements are all ionized by the surrounding high-energy radiation leaving a large abundance of free electrons. At $T \sim 1$ eV, the free electron number rapidly drops as they become bound to the light nuclei. This event, called recombination[†], took place over a few 1000 years when the universe was around 380,000 years[17]. Due to the drop in free electrons during recombination, the plasma neutralized, and the photons were released and streamed largely unimpeded. We refer to this release of photons as the Cosmic Microwave Background (CMB) radiation. The CMB is the best-known blackbody radiation[‡], deviating from a Planck spectrum by <50 ppm[1]. When released, the CMB was ~ 3000 K and is measured today at ~ 2.725 K due to Hubble expansion. The peak of a 2.725K blackbody is at ~ 160 GHz or ~ 1.8 mm allowing us to use microwave/radio astronomy techniques to observe it. Figure 1.3a shows the blackbody spectrum measured by the FIRAS instrument on the COBE satellite. Since the big bang “banged” everywhere at once, or the early plasma was uniform and in thermal equilibrium over the entire observable universe, it is not localized. Unlike galaxies or stars, present-day cosmologist observes a constant 2.725K blackbody uniform in all directions on

[†]A bit of a misnomer as this was the first time that electrons bound to form stable atoms.

[‡]Because they were in thermal equilibrium with the plasma before recombination.

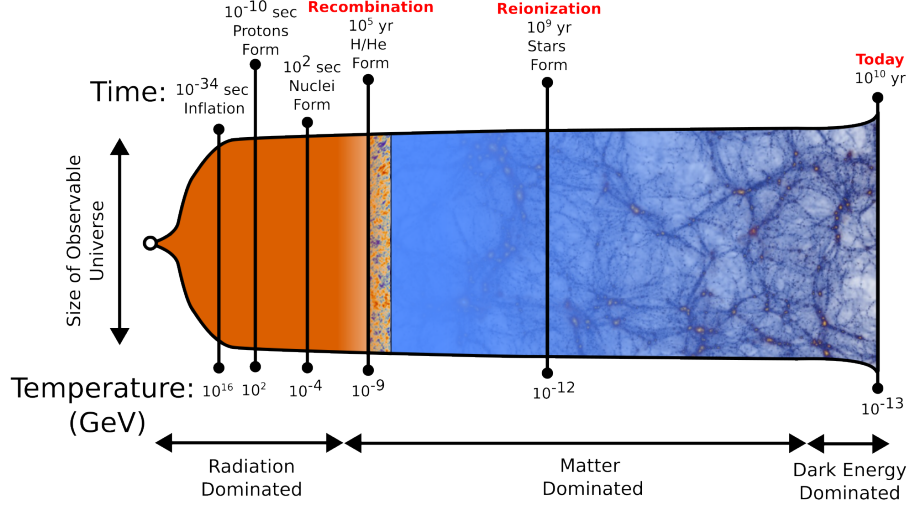


Figure 1.2: Schematic of the evolution of our universe. Time (temperature) is increasing (decreasing) from left to right, and the scale of the universe is increasing, shown (not to scale) as the y-axis. The radiation, matter, and dark energy-dominated epochs are roughly shown with solid double arrow bars at the bottom of the figure. Some key events are called out along the top, including from left to right: the end of inflation, the formation of protons, the formation of nuclei, the formation of the light elements (recombination), the formation of the first stars (reionization), and the present day. The background CMB image is a portion of the publicly releases all-sky Planck temperature anisotropy map and the background cosmic web image is a slice from the Millenium XXL simulation.

the sky and appear to be in the center of their observable portion of the CMB.

1.3.1 Primary Anisotropies

Despite the extreme uniformity, we predict small spatial anisotropies due to energy density, pressure, and velocity fluctuations in the early plasma. These anisotropies break our FLRW assumptions of isotropies, but since they are small, we can treat

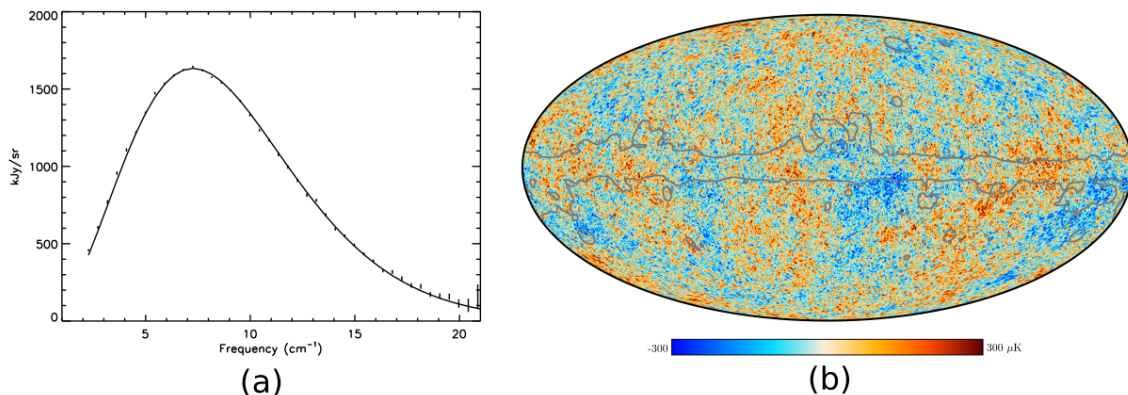


Figure 1.3: (a) The 2.725 Kelvin CMB power blackbody spectrum as measured by the FIRAS instrument on the COBE satellite[1] with error bars consistent with deviations from a perfect black body of <50 ppm. (b) The full-sky temperature anisotropy map (2018 SMICA temperature map) after subtraction of the monopole and dipole interpolated across the galactic plane and bright galactic and extragalactic sources (shown in grey outlines) as measured by the Planck satellite[18] showing deviations of $<300 \mu\text{K}$ from the 2.725K blackbody.

them as perturbative expansions to the Einstein and Boltzmann equations. These perturbations produce slight over (and under) densities in the matter distribution and slightly hotter (and colder) regions in the microwave background. We expect the source of these perturbations to be primordial in nature. A leading description of primordial perturbations is a class of theories that source perturbations from quantum fluctuations of an underlying field called inflation[19]. However, some alternative purely classical models have also been proposed to describe a primordial perturbation spectrum that fits our observations[20]

Independent of the underlying physical theory governing the primordial pertur-

bations, we expect (and observe) a distribution of perturbations on all angular scales. With the CMB, we can observe these perturbations by looking for differences in microwave power between different patches of the sky separated by different angular separations $\Delta T_{|d|} = \langle T(\vec{d}_2) - T(\vec{d}_1) \rangle_{|d|}$ where $\langle \rangle_{|d|}$ indicates an average over all points on the sky separated by distance, $|d| = |\vec{d}_2 - \vec{d}_1|$.

The dominant perturbations observed in the CMB temperature are acoustic oscillations. The acoustic oscillations manifest as slight baryon over- and under-densities in the plasma causing the photon wavelength to expand (or contract) due to the local contraction (or expansion) of space around the over (under)-dense regions. When the primordial perturbations are larger than the horizon size, they are frozen (do not evolve in time); once they enter the horizon, they begin to oscillate, and their value at recombination is what we observe in the CMB. The largest scale modes are the most pristine representations of the primordial perturbations because they had the least time (or no time) to evolve between entering the horizon and recombination. Due to this picture, we expect the CMB anisotropies to have maxima and minima at different angular separations ($|d|$) on the sky corresponding to acoustic modes at their oscillation maximum or minimum at the time of recombination.

In the plasma, photons have a finite scattering rate described by a mean free path, $\lambda_{\text{MFP}} = (n_e \sigma_T)^{-1}$, that they travel between each scattering event where n_e is the elec-

tron density, and σ_T is the Thomson scattering cross-section. The photon trajectory in the plasma is well described by a random walk of scattering events. Within a given Hubble time, H^{-1} , the typical distance a photon travels then is $\lambda_{\text{MFP}}\sqrt{(n_e\sigma_T c)/H}$. This creates a loss in power of the smaller-scale acoustic oscillations imprinted on the CMB temperature. So overall, we expect the acoustic oscillations to imprint a series of peaks and troughs in the CMB temperature as a function of angular separation on the sky, with the amplitude dropping off at small angular separations. A map of these temperature anisotropies as measured by the Planck satellite is shown in figure 1.3b.

1.3.2 CMB Polarization

Due to the perturbations described in section 1.3.1, CMB photons are also expected to be slightly polarized due to Compton scattering off free electrons during recombination. In Compton scattering (figure 1.4a), in the electron rest-frame, unpolarized radiation incoming along the y-direction scattering with a free electron will produce output radiation in the z-direction purely polarized in the x-direction. This is a simple situation when you have a single unpolarized photon scattering off a single free-electron; however, in an isotropic background of radiation, equal amounts of radiation will scatter incoming from the x- and y-directions producing unpolar-

ized radiation out along the z-direction. However, in the presence of a quadrupolar anisotropy (figure 1.4b) where the incoming radiation from the x direction is a lower temperature than the y direction, then the outgoing radiation along the z-direction will be slightly polarized with more polarization along the x-direction than the y-direction.

In the epoch before recombination, when electrons and photons were tightly coupled, we expect some quadrupolar anisotropies. The quadrupole acoustic oscillations are proportional to the dipole oscillations, which trace the plasma baryon velocity and are expected to be out of phase with the monopole. Additionally, the quadrupole is suppressed by a factor of wavenumber-over-optical depth ($k/n_e\sigma_T a$) at the time of decoupling. So we expect to see the same peaks and troughs as a function of angular scale in CMB polarization as we see in the temperature but with peaks (troughs) at the locations of temperature troughs (peaks) and with reduced amplitude.

Linear polarization can be decomposed into stokes parameters T, Q, U, and V representing the intensity, two linear polarization states, and circular polarization,

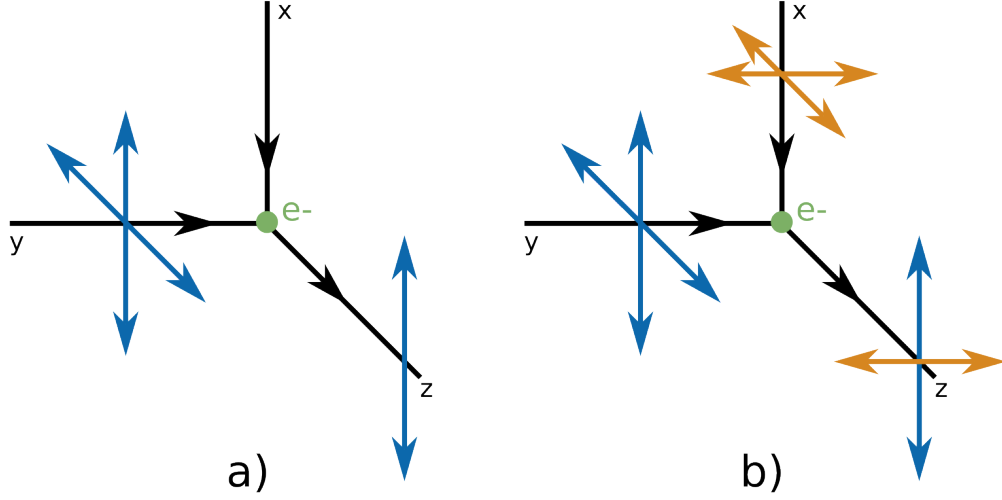


Figure 1.4: Diagram of inverse Thomson scattering of photons off of a free electron in the rest frame of the electron. a) Unpolarized light (depicted by equal blue vectors along the x and z directions) incoming along the y-direction (propagation direction depicted by the black arrows along the coordinate axes) scatters off of the free electron perpendicular outgoing along the z-direction linearly polarized along the x-direction. b) A quadrupolar anisotropy, depicted by unpolarized light of unequal amplitude incoming along the x- and y-axes, produces outgoing light in the z-axis with slightly more linear polarization along the x-direction than the y-direction resulting in some net outgoing polarization. If the incoming light were isotropic (equal in amplitude from both the x- and y-axes), the outgoing light would be unpolarized.

respectively.

$$\begin{aligned}
 T &= E_x^2 + E_y^2 \\
 Q &= E_x^2 - E_y^2 \\
 U &= 2\text{Re}(E_x E_y^*) \\
 V &= 2\text{Im}(E_x E_y^*)
 \end{aligned}
 \tag{1.5}$$

We expect no circular polarization from the CMB, so we ignore V for now and

focus on Q and U polarization. Note that, unlike the electric field direction, Stokes parameters describe power (average of the electric field square $\langle E \rangle^2$). Additionally, positive Q denotes polarization along E_x , negative along E_y , and the magnitude denotes the degree of polarization. U is the same as Q but with the $x - y$ axes of the polarization rotated by 45° as depicted in figure 1.5.

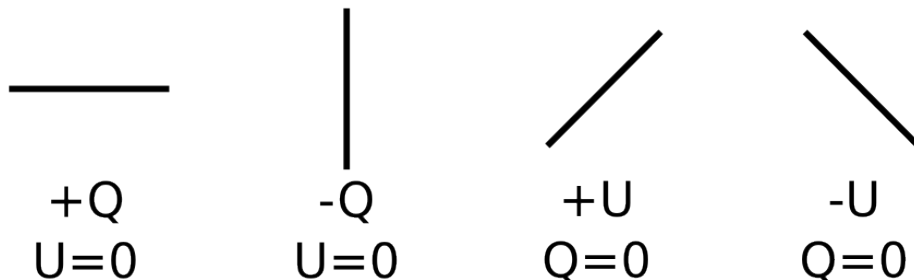


Figure 1.5: Stokes Q and U vectors are shown in the 2D x-y plane. +Q is purely along the E_x direction with the, -Q is along E_y , and +/-U are rotated 45° relative to Q.

We can measure and calculate CMB polarization maps in stokes Q/U; however, it is helpful to recast the polarization into a different basis denoted as E/B because they obey the properties $\nabla \times E = 0$ and $\nabla \cdot B = 0$ in analogy to the electric and magnetic fields in electromagnetism. Due to these symmetry properties, scalar perturbations—like the acoustic oscillations discussed thus far—will produce E modes but not B modes. Tensor perturbations—perturbations to the gravitational metric itself and manifest as gravitational waves—will generate both E and B modes. So far, E-modes consistent with the acoustic modes have been detected at high significance,

but gravitational wave-generated B-modes have yet to be detected. Inflation models discussed more in section 1.4.1 generically predict tensor perturbations. One aim of modern CMB experiments is to achieve greater polarization sensitivity to detect the B-mode signature and confirm the inflationary theory.

1.3.3 CMB Power Spectra

As discussed in section 1.3.1, the signature of plasma acoustic oscillations in both temperature and polarization is an average excess in signal between positions in the sky separated by particular angular separations. A convenient way to characterize this average is through a Fourier transform. Since the CMB is a 2-dimensional map on the surface of a sphere, the equivalent transform is a spherical harmonic decomposition:

$$T(\theta, \phi) = \sum_{\ell=0}^{\infty} \sum_{m=-\ell}^{\ell} a_{\ell m} Y_{\ell m}(\theta, \phi) \quad (1.6)$$

$$C_{\ell} \equiv \langle |a_{\ell m}|^2 \rangle_m$$

Where $Y_{\ell m}$ are the spherical harmonics[§], $a_{\ell m}$ are the spherical harmonic coefficients.

We further average over the azimuthal index (m) of the $a_{\ell m}$'s to produce a 1-D power spectrum just in the ℓ -coordinate represented by C_{ℓ} . Like in a Fourier transform,

[§]These are orthogonal and periodic in ϕ and θ in spherical coordinates in analogy to sin and cos in flat space. For polarization, these are replaced with spin-2 spherical harmonics because polarization is a tensor field.

small ℓ corresponds to the largest angular scale separation in the 2-D spherical map, and large ℓ corresponds to small angular separations. The polarization anisotropies are decomposed similarly but using spin-2 spherical harmonics since the polarization map is a tensor field.

Shown in figure 1.6 is the current state of the field measurements of the power spectra for temperature and E-mode polarization (left)[21] and B-mode polarization (right)[22]. The measurements are shown in points, and the best fit Λ -CDM expectation is plotted in solid lines. The data shows extreme agreement with the theory. The remaining uncertainty in the fitted cosmological parameters comes from the relatively poor constraints in: 1) high- ℓ TT and EE, 2) the lensing power spectrum, and 3) the low- ℓ BB. Modern CMB experiments (discussed more in chapter 2) fall into two categories: those targeting high- ℓ temperature and polarization and those targeting low- ℓ polarization signal.

Through these precise measurements of the CMB anisotropies, we can constrain the major properties of the universe we live in. In particular, we know now that we live in a very flat universe with the equivalent energy density produced by curvature, Ω_K , making up less than 0.26% of the energy density in the universe. Matter, Ω_m , accounts for 31.5% of the energy density, with only 4.9% coming from baryonic matter. The remaining 26.5% comes from dark matter[¶], and we are currently dominated

[¶]An unknown form of matter which interacts gravitationally but does not interact (or only

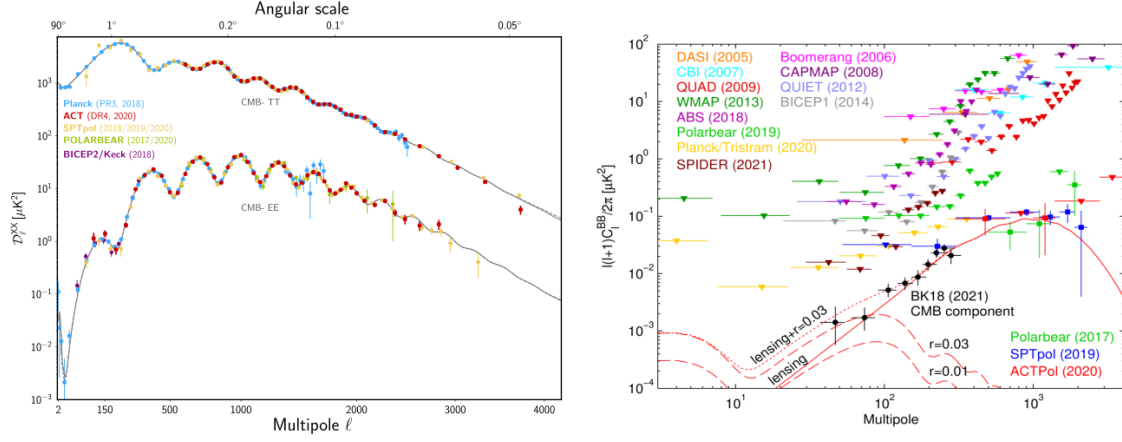


Figure 1.6: Current state of the field CMB power spectrum measurements (left) TT and EE power spectra from 6 experiments (colored point) with the tightest constraints at most scales coming from the Planck satellite (blue points) and showing strong agreement with Λ -CDM (grey lines) from Choi *et al.* 2020[21], (right) BB power spectra measurements from 12 (colored points) current and previous experiments. Detections of the high- ℓ lensing B-mode power spectrum have been made by several measurements, but the inflationary gravitational wave low- ℓ B-mode power spectrum has not been detected, and the tightest constraints come from the BICEP/Keck collaboration (black points) with an upper bound of $r < 0.036$ from BK18 2021[22].

by the vacuum energy with $\Omega_\Lambda = 68.5\%$. We date the age of the universe to be 13.787 Gyrs old, and the present-day value of the Hubble constant is $67.5 \text{ km s}^{-1} \text{ Mpc}^{-1}$. In short, we have learned an immense amount of information by measuring the CMB radiation and fitting it with the Λ -CDM model. However, there is still more to be learned from the CMB, and I will describe a few of the future goals of modern CMB experiments in section 1.4.

extremely weekly) through the other standard model forces.

1.4 CMB Science Probes

We now use the CMB and its interaction with large-scale structure to probe physics beyond-the-standard model including dark matter, neutrinos, and inflation at the highest energy scales inaccessible in terrestrial collider experiments. Current generation CMB experiments measure the large angular scale B-modes[22, 23] and bispectrum[24] to discover the appropriate model driving inflation (section 1.4.1)—a period of rapid expansion that took vacuum quantum fluctuations to horizon scales—by aiming to detect B-modes generated from primordial gravitational waves or primordial non-gaussianity. We also aim to better constrain the matter clustering scale, the growth of cosmic structure and feedback processes that effect galaxy evolution and star formation through the thermal and kinetic Sunyaev-Zeldovich effects[25]. At high angular the CMB damping tail and CMB lensing probe neutrino properties including the number of relativistic species, N_{eff} , and the sum of the neutrino masses, Σm_ν (section 1.4.2). Future CMB observatories aim to measure the sum of the neutrino mass states, Σm_ν , down to the level needed to distinguish between the normal and inverted mass hierarchy. We explain further in Sections 1.4.1 and 1.4.2 the inflationary and neutrino mass science cases.

1.4.1 Inflation

Inflation is a class of early[¶] universe models that describe a period of expansion of spacetime that proceeded at a rate far faster than the speed of light. It initially addressed some of the observational oddities observed in the CMB, including 1) the flatness problem, 2) the horizon problem, 3) gaussian primordial density perturbations, and 4) nearly scale-invariant density perturbations. As described in section 1.2, we live in an extremely flat universe; inflation accounts for this by rapidly expanding space so that any curvature becomes small within the Hubble radius (our observable universe), analogous to an ant on an expanded balloon or a human on the surface of the earth. The horizon problem refers to the extreme uniformity in the CMB temperature. Points on the sky that would appear to be causally disconnected^{**} without invoking some mechanism to connect them, show the same temperature, implying that they had to be in thermal equilibrium at some time in the past. Inflation provides a mechanism to resolve the horizon problem by expanding space faster than the speed of light for a brief period in the Universe's infancy.

Additionally, the well-established density perturbations described in section 1.3.1, which generated the CMB anisotropies and matter distribution, required a source,

[¶]First $\sim 10^{-34}$ seconds.

^{**}Causally connected objects are separated by distances shorter than the distance that light can travel over the age of the universe

and inflation provides this source. Generically, because inflation is a perturbation of the gravitational metric itself, it produces a spectrum of tensor perturbations (gravitational waves) along with the scalar perturbations (density perturbations discussed in section 1.3.1). These tensor perturbations produce some B-mode polarization in the CMB with an amplitude and power spectrum dependent on the specific inflationary model and the duration over which inflation dominated the universe’s dynamics. The amplitude of the tensor perturbations is represented by the parameter r , the ratio of the primordial tensor to scalar power spectrum amplitude at a particular angular scale of choice^{††}. So far, no detection of a primordial B-mode signal has been made, so only upper bounds on r have been placed. The BICEP/Keck collaboration has measured the tightest upper bound on $r \leq 3.6 \times 10^{-2}$. If the primordial B-mode signal is detected at a lower amplitude, then the spectral tilt of the tensor power spectrum, n_t , can also be measured to distinguish which inflationary model is correct. Inflation is a favored description of the early universe because few alternatives describe all of our current observations; however, some cosmologists have developed inflation alternatives that are classical in nature and do not generate gravitational waves[20]. A non-detection of B-modes by future experiments with higher sensitivities will imply either that these alternative models are a better description of our universe or that the particular inflationary model produces a tensor power spectrum

^{††}This choice varies in the literature, 0.002 and 0.05 Mpc⁻¹ are common choices.

with r too small for us to detect^{††}.

1.4.2 Neutrino Mass

The gravitational potential from large-scale structure deflects CMB photon's on their path from recombination to our telescopes. This phenomenon, called weak lensing, traces the large-scale structure^{§§} along the line of sight from the CMB at $z \sim 1100$ to $z = 0$. The deflection of the CMB from weak lensing converts E-modes into B-modes. These B-modes peak at the E-mode peak around $\ell \simeq 1000$, unlike the primordial B-modes, which peak at $\ell \simeq 100$. Using external tracers of the matter density^{¶¶}, or the CMB T, E, and B modes themselves, the underlying deflection field $\vec{d}$ and the gravitational lensing potential can be reconstructed. Since Neutrinos are relativistic at early times, they do not contribute to the cold dark matter that dominates the gravitational lensing potential, the lensing power spectrum is sensitive to the neutrino mass. From terrestrial neutrino experiments, we know neutrinos come in 3 mass eigenstates. We know the mixing angles between those mass states and that the lower bound on the sum of neutrino masses is $\Sigma m_\nu > 58\text{meV}$ [30]. However, we do not know the mass ordering of the states. Critically, the SO could provide

^{††}This will be fundamentally limited by cosmic variance, the statistical uncertainty due to the finite number of observable modes at a given scale in a finite-sized universe.

^{§§}Dark matter halos and connecting filaments.

^{¶¶}Such as galaxy distributions from optical large sky area surveys such as BOSS, DES, DESI, and LSST[26, 27, 28, 29].

a measurement neutrino mass $\sigma(\Sigma m_\nu) = 33$ meV, and with better constraints on the degenerate reionization parameter—from other experiments such as CLASS, Litebird, and BFORE[31, 32], $\sigma(\Sigma m_\nu) = 22$ meV which could distinguish between the normal and inverted mass hierarchy of the neutrino mass states[33].

1.5 Conclusions

Here we provided a qualitative historical perspective on the development of Cosmology (section 1.1) and a quantitative description of the gravitational properties of our universe on large scales (section 1.2). These statistical processes describe the big bang (section 1.3), the perturbations to this model that produce the CMB and large-scale structure we observe today (section 1.3.1), and precise measurements we’ve already made about our universe with current and past experiments (section 1.3.3). We described two additional tests of high energy physics (section 1.4), which we will constrain with future experiments. In the following chapters, I will introduce you to the Simons Observatory (chapter 2), a next-generation CMB experiment coming online in 2024, and provide an in-depth overview of the detector and readout system (chapter 3), readout noise modeling (chapter 4), time ordered data processing pipeline preparation (chapter 5), and instrument performance validation tests (chapter 6) required to develop and integrate an instrument, and observe with it to

constrain cosmological parameters with the CMB further.

Chapter 2

The Simons Observatory

2.1 Overview

The Simons Observatory (SO) is a set of four CMB telescopes at the Chajnantor Plateau in the Atacama Desert in Chile. The site was chosen for the observatory due to 1) the low precipitable water vapor (PWV), 2) access to a large fraction of the sky, and 3) the large overlap with other CMB, optical, and infrared surveys. The atmosphere's water and other elements (mostly O₂) absorb and emit in the microwave band, where we observe the CMB. Since these are molecular absorption lines, some frequencies have less absorption than others. We design our instrument to be maximally sensitive within these “windows” of high atmospheric transparency and

minimally sensitive outside them. Figure 2.1 shows an example of atmospheric transmission when observing through the atmosphere at an elevation above the horizon of 50° at variable PWV between 0 and 8mm at the SO site. The SO instrument band-passes are overlaid in red for low frequency (LF), green for mid frequency (MF), and blue for ultra high frequency (UHF), utilizing the atmospheric windows efficiently. The atmospheric emission is inversely proportional to the transmission, and the variation in emission due to atmospheric turbulence is much larger than the variation in power coming from the CMB anisotropies. The Chajnantor plateau has most days with $\text{PWV} < 2.0\text{mm}$ and a median of 1 mm minimizing atmospheric contamination relative to lower elevation and higher humidity sites. We additionally take advantage of the fact that the atmospheric variation is highly correlated between the detectors in our cameras and varies on long timescales to design our telescopes and data processing to mitigate further atmospheric contamination as discussed in section 2.2.2.

The SO survey sky coverage and overlap with other CMB and optical/infrared galaxy surveys are shown in Figure 2.2. The large sky area and survey overlap is enabled by the site’s proximity to the equator and the proliferation of South American astronomical observatories. Figure 2.2 also shows the survey strategy for the two types of telescopes built for the SO the Small Aperture Telescopes (SATs)

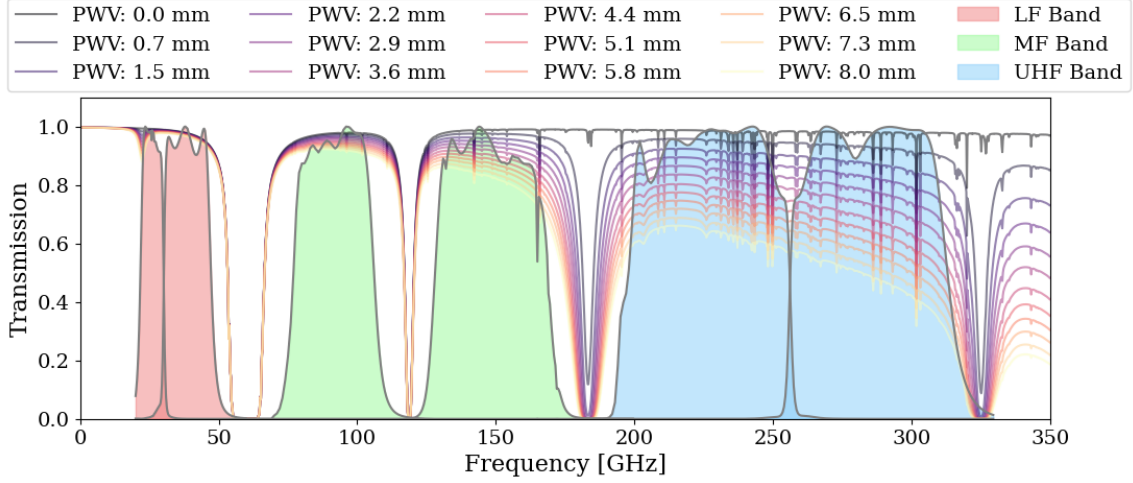


Figure 2.1: Colored lines show the atmospheric transmission in the Atacama desert at 50° observing elevation over a PWV range from 0 to 8mm. Shaded bands represent the six observing bands for the SO placed in the available atmospheric windows. Atmospheric data from the am atmospheric modeling code[34].

shown on the left and the Large Aperture Telescope (LAT) shown on the right. The SATs are three low-resolution ($30'$ at 90 GHz) telescopes that scan two patches that cover $\sim 10\%$ of the sky optimized for detection of the inflationary B-mode signal peaked at degree angular scales and are discussed further in Section 2.2. The LAT is a single high-resolution ($2.2'$ at 90 GHz) telescope that surveys one large patch that covers $\sim 40\%$ of the sky optimized for high- ℓ CMB observables, including measurement of the thermal and kinetic Sunyaev-Zeldovich effect in galaxy clusters, neutrino mass constraints through weak-lensing signal, radio transients, and galactic science. We will not discuss the detailed LAT design in this dissertation; for further detail, refer to Orłowski-Scherer 2022[35], Zhu 2022[36], Zhu 2021[37], Xu 2021[38],

and Gudmundsson 2021[39].

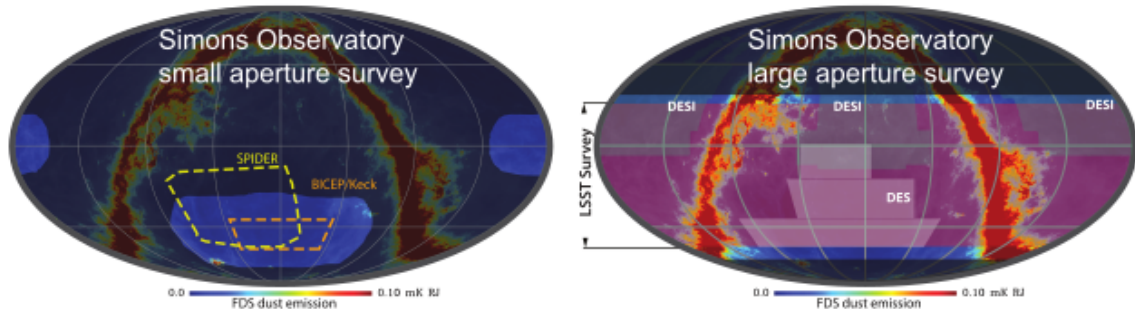


Figure 2.2: SAT survey area includes two patches (unshaded regions) which avoid areas of high dust contamination near the galactic plane (red strip), covers $\sim 10\%$ of the sky, and has a large overlap with the SPIDER and BICEP/Keck CMB surveys (left). The LAT survey area is a large patch that covers $\sim 40\%$ of the sky but minimizes the time spent in the galactic plane. It fully overlaps with the DES, DESI, and LSST optical/infrared galaxy survey for cross-correlation analyses especially beneficial for extracting the weak lensing potential (right). Figure reproduced from Simons Observatory Collaboration 2019[33].

Polarization anisotropies due to the dust and electron synchrotron emission within our galaxy is everywhere brighter than the polarization anisotropies from the CMB we are trying to measure. Since this emission obscures our ability to observe the CMB, they are called galactic foregrounds. While there are many sources of unpolarized foregrounds, there are only two sources of polarized foreground emission 1) spinning dust grains and 2) synchrotron emission from charged particles accelerated around galactic magnetic fields. Dust emission follows a modified blackbody that peaks in the THz ($T \simeq 20$ K), synchrotron emission follows a falling exponential while the CMB is flat up to ~ 100 GHz and then falls off at a higher frequency in

Rayleigh Jeans brightness temperature as shown in Figure 2.3 at a degree angular scale. The amplitude of the foreground anisotropies varies across the sky, with the largest foreground contamination in the galactic plane. The minimum of the two polarized foreground components is around 100 GHz; however, even in the patches we’ve chosen to minimize foreground contamination, there is no frequency at which the CMB dominates over the polarized foreground anisotropies. To recover the CMB anisotropies from the foregrounds, the SO observes in six observing bands; due to the different spectral shape between the foregrounds and the CMB, we can use combinations of observations in the six bands to clean the foregrounds from the CMB. The six spectral bands, shown in Figures 2.1 and 2.3, are chosen to fit within the available atmospheric windows and provide high sensitivity to the frequencies dominated by dust (UHF band) and synchrotron (LF band) for optimal foreground subtraction.

2.2 SAT Instrument Design

Now that we have outlined the science and observational constraints that drive the SO design, we will briefly overview the SAT design.

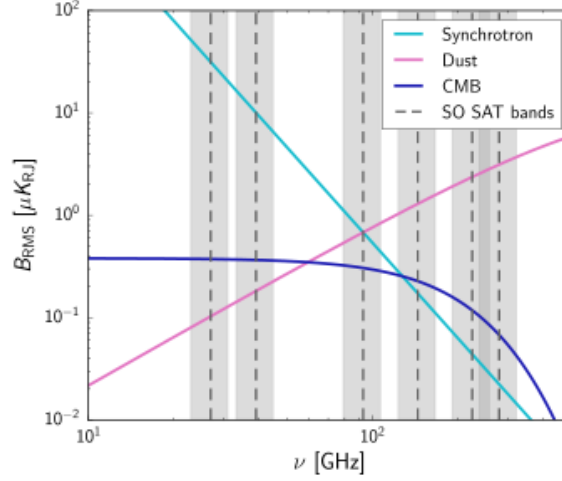


Figure 2.3: The polarized galactic foregrounds at a degree angular scale: dust (pink) and synchrotron (cyan) along with the CMB (dark blue) SED in Raleigh-Jeans brightness units for the SO SAT sky patch at $\ell=80$ with the SO observing bands shown in light grey. Figure reproduced from Simons Observatory Collaboration 2019[33].

2.2.1 Optics design

Each SAT is a refracting telescope whose optical design achieves $f/\# = 1.4$ and a 35° diffraction-limited field of view (for the 90 GHz-center band) using a 42 cm aperture, 36 cm focal plane, and three cryogenic (1 K) silicon lenses[40]. The lenses are anti-reflection coated using a diced metamaterial structure on both of their vacuum-to-dielectric surfaces[41]. In the reverse-time sense, the Gaussian beams from each detector feed are truncated by a cryogenic (1K) stop at ~ -4.5 dB and -8.0 dB for the 90 and 150 GHz central pixels. Millimeter-wave absorbers surround the optics, on the cryogenic stop, and on a series of cryogenic baffles to minimize diffraction-induced

far-sidelobe features[42].

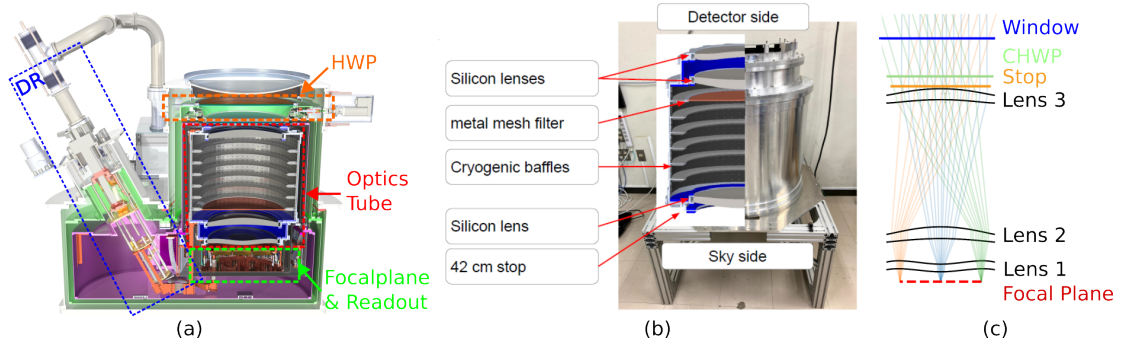


Figure 2.4: (a) Computer-Aided Drafting (CAD) cross-section of the SAT cryogenic receiver with the vacuum shell shown in grey, 40K radiation shield shown in green, 4K radiation shield shown in pink, the dilution unit outlined in dashed blue, optics tube outlined in dashed red, HWP outlined in dashed green, and the focal plane and cryogenic readout outlined in dashed orange. (b) The fully constructed optics tube with a CAD cross-section of the internal optical elements overlaid on the left half of the optics tube. The silicon lenses, IR blocking filters, baffling, and stop are called out with red arrows. The orientation is flipped 180 degrees relative to figures (a) and (c), with the sky side on the bottom of the image and the focal plane on the top. (c) A ray trace from a central pixel and two edge pixels on the focal plane through the three silicon lenses with each of the lenses, the cryogenic stop, HWP, and primary aperture (window) labeled to the right of the image. (b) reproduced from Kuichi *et al.* 2020[40], (c) was generated in Python using publicly available code `sosat_optics`[43]

The SO SAT scan strategy consists of scanning two patches (Figure 2.2—left) in repetitive scans left-and-right in azimuth at a fixed elevation. Each ~ 10 -40 minute observation covers the full declination range and a sub-range in right-ascension of the full patch set by the amount of sky rotation over the observation. We continue this pattern stepping the center and span of the observation in azimuth until the full patch is observed or the remainder of the patch drops below the horizon or behind

the sun or moon. Since the emission from the ground is slightly polarized and produces an azimuth synchronous signal, it is an important systematic to control for recovering low- ℓ B-modes. To mitigate ground pickup, there is a series of three warm baffles: 1) an absorbing conical baffle that attaches to the aperture, 2) a reflective scoop-shaped baffle that extends in front of the interface baffle and is always oriented to shield from the ground (Figure 2.5b), and 3) a cylindrical ground screen that is fixed to the ground around the telescope platform (Figure 2.5a). This warm baffling design was selected to require any light from the ground to diffract twice off any baffle surface before entering the aperture at the SAT's nominal observing elevation of 50° .

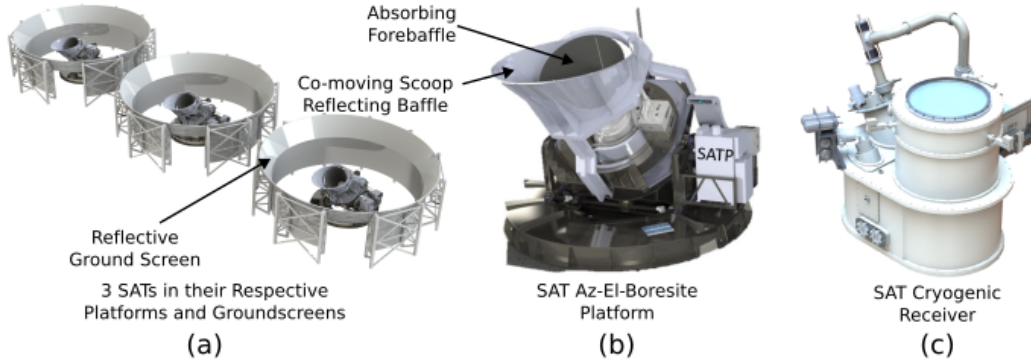


Figure 2.5: (a) All three SATs inside their reflective ground screens. (b) A single SAT with its Az-El-Boresite mount is shown along with the absorbing forebaffle and scoop baffles called out. (c) One SAT cryogenic receiver outside of its Az-El-Boresite mount. All figures are CAD renderings.

2.2.2 Half-waveplate

To minimize the spurious signals from the turbulent atmosphere, a cryogenic continuously rotating half-waveplate (HWP) is placed at the optical stop of the SATs (Figure 2.4c). The HWP is made from a birefringent material (sapphire) which produces a difference in wave speed between polarization aligned and anti-aligned with the crystal axis. In a typical HWP, the thickness of the material is chosen so that the delay between the two polarizations at a single frequency is a half wavelength. However, our observing bands in each SAT are ~ 100 GHz wide, so our HWP is constructed from three layers of sapphire with the crystal axis angle rotation and thickness of each sapphire plate chosen to maximize the modulation efficiency over the full observing bandwidth.

The HWP is placed at the aperture stop so that all detectors on the focal plane uniformly illuminate the HWP and aren't sensitive to variations in the optical properties in the sapphire stack or anti-reflection coatings. The HWP is also placed on the sky-side of all optical elements except for an infrared blocking filter at 40 K and 300K and the vacuum window. This design removes all instrumental polarization except from the few elements on the sky-side of the HWP and the HWP itself. To minimize the instrumental polarization, the HWP would be the first element in the optical path; however, the temperature of the HWP impacts the detector loading

and spurious signal induced from the HWP itself, so placing it at the 40K stage behind the minimum number of optical elements was a compromise between the two competing constraints. For a review of the expected spurious HWP signal leakage, see Salatino *et al.* 2018[44].

The sapphire stack is attached to a ring of permanent neodymium magnets, which is levitated by a Yttrium barium copper oxide (YBCO) ring on the stator. When cooled below its superconducting transition (92 K), the YBCO ring pins the magnetic field in place due to the Meisner effect, which acts as a very stiff and low dissipation bearing on which the HWP is rotated. A ring of solenoids on the stator outside the YBCO ring is driven in a sequence of positive and negative polarity to spin the rotor. The rotor can spin up to 3 Hz in both directions and typically operates around 2 Hz during CMB observations. The continuous rotation modulates incoming polarization at four times the rotation rate, much faster than the slowly varying signal from the atmosphere, effectively mitigating atmospheric leakage into the CMB polarization signal. The SO HWP is based on the design for the POLARBEAR-2B receiver; for more details refer to Hill *et al.* 2016[45].

2.2.3 Cryomechanical design

To minimize the emission from the optical elements and utilize the low noise transition edge sensor (TES) detectors discussed in Section 2.2.4, all optical elements are housed in a cryogenic receiver (Figure 2.5c). The receiver follows a typical nested design, starting with an outer 6082 Al* vacuum shell to minimize convective heating from room temperature to the colder cryogenic stages. Inset from the vacuum shell by a ring of G-10 tabs is a high-conductivity Al (1050 H24) shell held at ~ 40 K to reduce radiative heating to the colder cryogenic stages. The heat lift for the 40 K stage is provided through a connection with a flexible stack of 99.9999% pure Al sheets to the first stage of a pulse tube cryocooler[†]. Inset with another G-10 tab ring from the 40 K stage is a ~ 4 K shell constructed identically to the 40 K shell but connected to the 2nd stage of a pulse tube to reduce the radiative loading on sub-Kelvin stages. The G-10 tabs provide a minimally thermally conductive structural support for all hardware held at colder stages (optics, detectors, and readout electronics).

The 40 and 4 K radiation shields are wrapped in multi-layered insulation (MLI) constructed from layers of aluminized mylar separated by fiber mesh to reflect infrared power from the hotter thermal stages. Thirty layers of MLI are used at 40 K,

*Aluminum is chosen over steel to minimize weight, and the 6000 series is chosen for its balance of strength and weldability.

[†]Cryomech PT420: www.cryomech.com/products/pt420/

and ten layers are used at 4 K. The contribution from radiative loading is summarized in column 4 of Table 2.1 and dominates the loading at 40 K.

There are two sub-4K thermal stages, one at ~ 1 K and one at ~ 100 mK, cooled through gold-plated copper connections to the still and mixing chamber stages of a Bluefors SD400 Dilution Refrigerator. The low thermal conductivity mechanical supports between the 4–1 K and 1–0.1 K stages are made through circular trusses constructed from carbon fiber struts described in Crowley *et al.* 2022[46]. The 4–1 K truss holds a 1 K high permeability magnetic shield, a 1K superconducting magnetic shield, and the optics tube (shown in Figure 2.4b). The 1–0.1 K truss holds the focal plane and detector arrays. The second column of Table 2.1 summarizes the thermal loading through the G-10 and carbon fiber support elements for each inter-thermal-stage support. The G-10 supports constitute the largest component of the thermal budget at the 4 K stage.

An RF shield is established in the backend cavity behind the focal plane, where the detectors and readout are housed to minimize RF interference at the focal plane. A thin aluminized mylar foil is stretched between the mechanical support trusses to create this RF cavity. The aluminization is chosen to be as thin as possible to minimize excess thermal loading and the number of tears in the aluminum layer during installation that compromise the RF shield. A $6.3 \mu\text{m}$ layer was chosen to

balance these competing constraints, and the thermal loading from this shielding is summarized in column 9 of Table 2.1.

In addition to the re-imaging and polarization modulation optics described in section 2.2.1, a series of infrared-blocking filters are used at each thermal stage to minimize the amount of out-of-band infrared radiation from hitting the coldest radiation stages. The infrared blocking filters are low-pass filters that reflect[47] or absorb[48] power above the mm-wave observation band. The contribution to the total loading due to infrared loading is summarized in column 5 of Table 2.1 and is the dominant source of loading at 1 K and a significant fraction of the loading at 4 and 40 K.

Table 2.1: SAT design estimated loading between the five thermal stages in the cryogenic receiver. The last two columns show the total loading between each stage relative to the heat lift from the PT420 pulse tube cryocooler for the 40 and 4 K stages and the SD400 dilution refrigerator for the 1 and 0.1 K stages. The central columns break down the total loading between the subsystems described in this chapter that contribute to the total stage loading.

Thermal Stage	Mechanical Supports	Wiring	Radiative	Optical	Amplifiers	RF/DC Readout Power	HWP	RF Shield	Unit	Total	Cooling Power
300-40K	5.3	1.9	6.0	5.2	0.8	0.0	0.3	0.2	W	20.5	55
40-4K	112.6	74.2	0.5	94.6	56.0	1.3	N/A	86.1	mW	425.2	2000
4-1K	.37	.05	0.00	3.78	N/A	0.02	N/A	0.03	mW	4.22	20.00
1-0.1K	12.3	4.2	0.0	6.8	N/A	1.8	N/A	2.9	μ W	28.1	400

Cabling must be run between thermal stages to read out the detectors and the thermometers and heaters used to operate and monitor the refrigerators (housekeeping). The cabling materials are chosen to minimize thermal conduction within the electrical requirements; however, they still constitute a non-negligible fraction of the thermal budget as summarized in column 3 of table 2.1. The housekeeping cabling is made from 38 AWG Manganin. There are 4 x 50 wire twisted pair looms from 300–4 K and a series of 2- and 4-wire twisted looms that breakout connection to 1 K and 0.1 K from a set of breakout PCBs at 4 K. The detector readout wiring consists of DC twisted pair looms and coaxial cabling and is discussed in further detail in Chapter 3. The loading from the housekeeping and focal plane readout wiring combined is summarized in column 3 of table 2.1.

2.2.4 Detectors and Readout

The SO MF and UHF focal plane arrays comprise seven universal focal plane modules (UFMs)[49], each housing 432 dual-polarization dichroic pixels. Each pixel consists of a machined gold-plated aluminum profiled horn[50] coupled to an ortho-mode transducer (OMT) through a waveguide choke and a quarter-wave back short, which maximizes the coupling efficiency. Each of the two OMT orthogonal polarization antennas feeds a planar microstrip mm-wave feedline which splits power to two

feedlines, each with a series of unique stub filters that define the band edges. The power from each microstrip is deposited on the TES bolometer through a matched load on the bolometer island. The OMT coupled feedhorn design is based on the successfully fielded design for the Atacama Cosmology Telescope polarization (ATCpol) experiment[51]. In total, a SAT MF or UHF focal plane consists of 12,096 optically coupled bolometers. To minimize the number of readout wires running to the 100 mK focal plane and contributing to the conductive loading budget, we use the microwave multiplexed (μ mux) readout scheme[52]. The SO detector and readout technologies and their implementation is reviewed in detail in Chapter 3.

2.2.5 Control, Data Acquisition, and Analysis Software

The SO uses the Observatory Control System (OCS) software framework for data acquisition and control[53]. OCS is a flexible framework that allows communication with distributed devices in a network. Each device has an associated "agent" that is run within a docker image[54]. Each agent has a set of exposed tasks and processes. Tasks are a command that is run once, typically a commanded change in device state, and processes are commands that start a process running continuously until commanded to stop, typically acquiring data from the device. Upon startup a specific set of task and processes are run (typically configure acquisition and start

acquisition). A client can also be written which runs a set of agent tasks. The client can be an interactive notebook, script, or web interface. We use OCS to control and monitor all of the SO devices including the cryogenic receivers, telescope platforms, computing servers, power systems, and detector and readout systems. While OCS monitors health status associated with the readout such as die temperatures, fan speeds, and power consumption the main detector data rate is too high to share local network bandwidth with the rest of the observatory sensors. The readout software stack consists of `pysmurf` built on top of `pyrogue` built on top of `rogue` and provides a control and high bandwidth data acquisition for the readout system. This software is described more in Chapter 3. Specific high level `pysmurf` operations are exposed through a wrapper OCS agent as tasks. These tasks include detector tuning, detector time constant estimation, starting data acquisition and stopping data acquisition. These exposed tasks allow OCS to control the detector operation in concert with rest of the observatory devices. As data is streamed to disk data packaging is performed automatically combining multiple detector data streams from many wafers into a single data package and cosampling relevant secondary data including telescope point and HWP angle. This package data consists of an observation book and is references in an observation database which can be used to load relevant calibration information and additional housekeeping fields associated with the given observation. From

the observations we run a series of time domain filtering operations and statistical analyses on the data to produce data quality statistics, and daily maps that can be used as an early commissioning pipeline and for data quality checks for the lifetime of the SO. This set of operations is described further in Chapter 5.

Chapter 3

Detectors and Readout

Here we describe the detector and readout system used in the SO. We introduce bolometric transition edge sensor detectors in Section 3.1. In Section 3.5, we introduce multiplexed readout and the microwave squid multiplexer (μmux) used in the SO. In Section 3.6, we describe the SO readout and detector implementation, including the wiring that brings signals from 300K to the cryogenic focal plane and the universal focal plane modules (UFMs) that package the readout and detectors in a common unit. Section 3.7 introduces the custom digital signal processing electronics developed to read out the μmux circuit in the SO. Lastly, in Section 3.8, we describe the detector and readout tuning procedure used to bring the SO cameras into their observation-ready states.

3.1 Detector Overview

When optimizing any sensor for detecting any energy deposition (electromagnetic radiation, particles, and phonons, for instance), one must balance a few universal design parameters: 1) noise, 2) gain, 3) bandwidth, and 4) dynamic range. Sensor noise refers to fluctuations in the sensor (thermal carrier—electron or phonon—fluctuations, film impurities, magnetic flux vortices, to name a few) manifesting as a voltage or current noise on the output of the sensor independent of the signal detected by the sensor. To maximize the signal to noise of the detection of the signal of interest, the sensor noise must be much smaller than the intrinsic statistical noise of the signal. The sensor’s gain is the output current or voltage per input signal unit. The gain must be sufficiently high such that the amplified sensor plus signal noise is much larger than the noise added by the next amplification stage of the sensor readout system. The sensor’s bandwidth refers to the range of signal timescales over which the sensor’s gain is high enough to achieve sensor noise domination over the next readout amplification stage. Input signals changing faster than the sensor bandwidth are not detected. The dynamic range is the range of input signal amplitudes over which the sensor *linearly* (i.e., $\text{Signal}_{\text{Out}} = \text{Gain} \times \text{Signal}_{\text{In}}$) amplifies the input signal divided by the sensor noise floor. Input signals larger than the maximum sensor-signal amplitude will suffer from significant non-linearity, and signals smaller than

the noise floor will be dominated by sensor noise.

3.2 Coherent vs Bolometric Detection

For the detection of CMB radiation, two types of detectors have been used; coherent detectors[55] and bolometers[56]. Coherent detectors employ an RF amplifier, often a High Electron Mobility Transistor (HEMT) or Superconductor-Insulator-Superconductor (SIS) mixer, as the sensor element and can detect both the amplitude and phase of the incoming signal. Coherent detectors have a larger dynamic range than bolometers; however, their noise has not reached the quantum limit in the microwave frequencies $\gtrsim 90$ GHz. Coherent detectors are the detectors of choice for mm/sub-mm observatories that study bright astrophysical sources, such as the Atacama Large Millimeter Array, because, with bright astrophysical sources, the noise levels are sufficiently low relative to the source noise and the larger dynamic range allows for detection of a bright source in a field with a low emission background while maintaining sensor linearity over the whole field. The phase that is accessible with coherent detection also allows for multiple telescope signals to be combined through interferometry to achieve a higher spatial resolution than reasonably achievable with any single pointing telescope. Because the CMB is extremely faint and does not vary much over any single observation field, all modern CMB observatories observe using

bolometers to maximize instrument sensitivity.

A bolometer consists of a thermally absorbing element well thermalized to a material with resistance that varies with temperature, which is in turn attached to a thermal bath via a weak thermal link. The heat capacity of the absorber, C , and the thermal conductivity of the thermal link, G , set the intrinsic thermal time constant of the bolometer $\tau_{\text{thermal}} = C/G$. An electrical bias is applied to the resistive element to provide joule heating. The balance of received optical signal and joule power creates an electrothermal feedback (ETF) loop. A bolometer operated with a high ETF loopgain effectively balances changes in power incoming from the absorber with an inverse change in the joule heating[57, 58]. If the sensor’s resistance increases with increasing power on the absorber, a voltage bias is used so that the joule power $P_{\text{Joule}} = V^2/R$ decreases as the sensor’s resistance increases. If the sensor’s resistance decreases with increasing power on the absorber, a current bias is used so that the joule power $P_{\text{Joule}} = I^2 R$ increases as the sensor’s resistance decreases with increasing power on the absorber.

3.3 TES Bolometers

The first bolometers used for CMB measurements were semiconductor-based Neutron Transmutation Detectors (NTDs), which have decreasing resistance with in-

creasing temperature and require a current bias[59]. These sensors, however, suffered from large low-frequency noise, limited sensor bandwidth, and a shallow resistance vs temperature curve (gain). Additionally, the sensors were high impedance which required readout by a high input impedance transistor amplifier which performed poorly at the coldest cryogenic temperatures and suffered from high power dissipation and high input noise. In contrast, the superconducting Transition Edge Sensor (TES) bolometers used by all modern CMB experiments operate on the superconducting-to-normal transition of a thin superconducting film with a resistance that increases steeply with temperature and can be tuned to a wide range of operating temperatures. They can be designed to have much higher bandwidths than NTD bolometers and are low-impedance sensors, so low-noise SQUID-based current amplifiers can be used to read them out, which also operate well at mK temperatures and easily achieve noise subdominant to TES sensor noise.

The ETF can be represented by the power balance equation, which states that the combination of the optical power (P_{opt} , from the CMB radiation through the weak thermal link) and the electrical power (P_{bias} , from joule heating) sums to the saturation power, P_{sat} [57].

$$P_{sat} = P_{opt} + P_{bias} \quad (3.1)$$

If P_{opt} exceeds P_{sat} , the TES film will be completely in the normal conducting regime

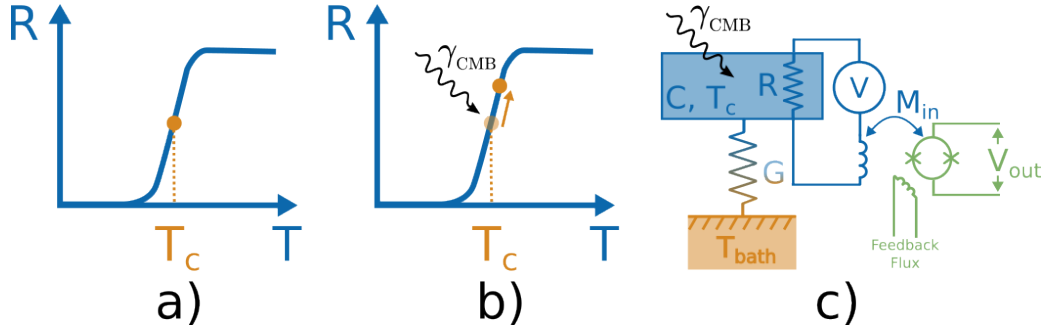


Figure 3.1: Transition edge sensor bolometer schematic operation. a) A superconducting film is characterized by a steep change in resistance from zero resistivity to a normal metal resistivity at a characteristic temperature T_c . b) As some CMB radiation is absorbed by the TES the TES island heats up causing a large change in resistance when biased on its superconducting transition. c) A bolometer consists of a superconductor well thermalized to an absorber (blue) with a heat capacity C that is attached to a thermal bath T_{bath} (orange) through a weak thermal link G . The power absorbed in the island from the CMB heats up the film changing its resistance. Voltage biasing the superconducting film produces Joule heating that acts in the opposite direction from the absorbed optical power creating a negative electrothermal feedback loop. The change in current through the TES is measured by transformer coupling to a SQUID amplifier (green).

and have no sensitivity to the CMB. The bath temperature is chosen to achieve low enough sensor noise. The saturation power is tuned by adjusting the transition temperature of the film, T_c , and the G of the weak link to a value that is larger than the largest expected mean optical power at the Chilean observing site but as small as possible to minimize the thermal carrier noise which increases with increasing P_{sat} as $NEP_G \propto \sqrt{P_{sat}}$, where NEP is the noise equivalent power which is the root of the spectral density of the power fluctuations in units of $W/\sqrt{Hz}$. To balance dynamic range and noise, $P_{sat} \approx 2P_{max}$ for the SO where P_{max} is the expected optical loading on the detectors from the atmosphere and instrumental emission when observing at 35 deg elevation with 1mm PWV at the Atacama site.

3.4 TES Readout with a SQUID Amplifier

Because TES bolometers are voltage biased to provide negative electrothermal feedback, the first stage amplifier must be an ammeter. An ideal ammeter has a much lower input impedance than the sensor. TES bolometer resistances are in the range of 4-400 m Ω when biased in the middle of their superconducting-to-normal transition requiring an ammeter with an input impedance of $\lesssim 100 \mu\Omega^*$. As mentioned in Section 3.1, the current sensor must also have input noise much less

*One order of magnitude smaller than the sensor impedance.

than the sensor noise and high gain. While there are newer alternatives[60, 61], the mature technology that achieves these requirements is a Superconducting Quantum Interference Device (SQUID) based amplifier.

A SQUID is a superconducting loop interrupted by at least one normal or insulating section forming a Josephson junction (JJ). The phase (ϕ) of the tunneling supercurrent (I_{sc}) across a JJ obeys the Josephson relations[62, 63]:

$$I_{sc} = I_c \sin \phi \quad (3.2)$$

$$\frac{d\phi}{dt} = \frac{2eV}{\hbar} \quad (3.3)$$

Where I_c is the material- and geometry-dependent critical current above which superconductivity breaks down, e is the electron charge, V is the voltage across the junction, and $\hbar$ is Planck's constant.

A DC-SQUID has two junctions, one on each branch of the superconducting loop. The supercurrents through the two junctions are bias-current and flux dependent. The interference of the two supercurrents makes the SQUID an effective flux to voltage transducer with a gain dependent on the bias current and mean flux operating point. The output voltage of the SQUID can then be amplified with a conventional low-noise OpAmp circuit to monitor changes in input flux. An RF-SQUID has one junction in its superconducting loop and is coupled to an LC resonator circuit.

The impedance of the RF-SQUID varies as a function of external flux modifying the circuit's resonant frequency or quality factor. The transmission or reflection from the tank circuit is monitored with an RF probe tone tuned to the resonance frequency. The transmitted or reflected tone's resultant amplitude or phase variation is amplified with conventional low-noise RF amplifiers and used to reconstruct the input SQUID flux.

Transformer coupling the TES circuit to the SQUID allows for current changes in the TES circuit to be detected as flux variation in the SQUID. The SQUID then acts as an ideal ammeter because its input impedance is just the impedance of the superconducting input coil $|Z_{in}| = \omega_{TES}L_{in}$. Typically the TES bolometer bandwidth is <1 kHz, and the input inductance of the SQUID is <100 nH, so the input impedance of the SQUID is $<10 \mu\Omega$. For CMB bolometers, typical maximum sensor signals are $\lesssim 10$ Hz such that the input impedance of the SQUID can be ignored entirely. Figure 3.1c shows a typical SQUID coupled TES circuit for the DC SQUID case. The TES bolometer is voltage biased. A superconducting transformer coupled to a SQUID is placed in series with the TES resistance. As the sky signal γ_{CMB} varies, the bolometer resistance varies, which in turn varies the current through the transformer. As current varies through the transformer, the SQUID transduces the coupled flux to a voltage output which can then be amplified and read out with

a conventional OpAmp circuit and Analog-to-Digital Converter (ADC). A second transformer couples additional flux into the SQUID to keep the average flux at the maximum gain point of the SQUID.

The SO employs RF SQUIDs due to their superior multiplexing ability. Multiplexing refers to the ability to read out multiple sensor signals on a single readout circuit. Multiplexing minimizes the conductive loading through the readout circuit wiring from room temperature to the cryogenic focal plane and reduces the space and complexity required for readout wiring and electrical components. Section 3.5 provides a detailed overview of the RF-SQUID readout implemented for the SO.

3.5 The Microwave SQUID Multiplexer

The Microwave Squid Multiplexer (μmux) is a form of frequency division multiplexing that employs RF-SQUIDs operated at microwave frequencies, $\mathcal{O}(\text{GHz})$. Sensors are each coupled to an RF-SQUID with a unique resonance frequency. All RF-SQUIDs are coupled to a common RF transmission line, and the TES signals are encoded as a phase modulation on a frequency comb of $\mathcal{O}(1000)$ interrogation tones. Here we build up the circuit design and operating principles starting from the resonator circuit (Section 3.5.1), then adding the RF SQUID (Section 3.5.4), and lastly linearization using flux ramp modulation (Section 3.5.6). Much of the deriva-

tion here is a condensed version of Mates 2011[64] reproduced here as a reference for our discussion of the implementation details for the SO. I provide additional example plots to build intuition for the foundational aspects of the technology, which are built on in the following sections and chapter 4.

3.5.1 Capacitively Coupled $\lambda/4$ Resonator

In place of a lumped element LC-tank circuit, SO μ mux uses distributed coplanar waveguide (CPW) quarter-wave ($\lambda/4$) resonators capacitively coupled to a CPW transmission line. A quarter-wave transmission line transforms a load impedance[65] as:

$$Z_{in} = Z_1 \frac{Z_L + iZ_1 \tan\left(\frac{\omega l}{v_p}\right)}{Z_1 + iZ_L \tan\left(\frac{\omega l}{v_p}\right)} \quad (3.4)$$

Where Z_1 is the characteristic impedance of the CPW, Z_L is the load impedance, v_p is the phase velocity in the CPW, l is the length of the transmission line, and ω is the angular frequency ($\omega = 2\pi f$). When terminated with a short ($Z_L = 0$), $Z_{in} = iZ_1 \tan\left(\frac{\omega l}{v_p}\right)$. The total impedance presented to the transmission line when including the series capacitance is then:

$$Z_{tot} = \frac{1}{i\omega C_c} + iZ_1 \tan\left(\frac{\omega l}{v_p}\right) \quad (3.5)$$

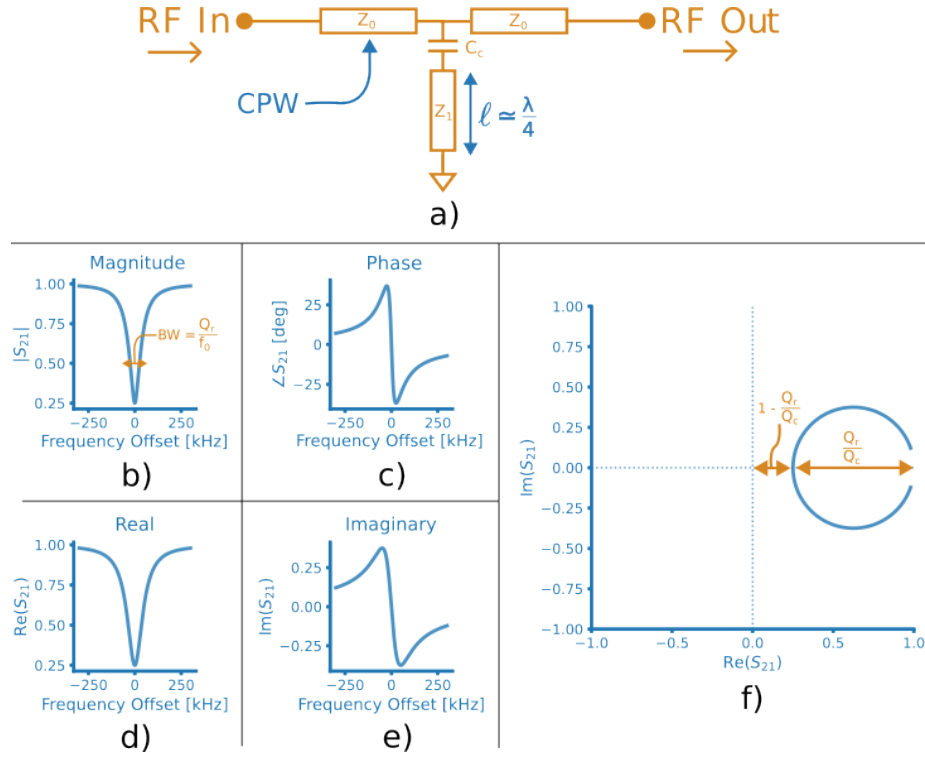


Figure 3.2: The quarter wave resonator a) circuit diagram of a coplanar waveguide (CPW) quarter wave resonator capacitively coupled to a CPW feedline. The magnitude (b), phase (c), real (d), and imaginary (e) part of the forward scattering parameter S_{21} (RF Out/RF in) vs frequency. The bandwidth of resonance is equal to the ratio of the resonator quality factor (Q_r) to it's characteristic frequency (f_0). In the complex plane the S_{21} vs frequency traces out a circle (f) The circle's diameter is equal to the ratio of the total quality factor (all losses in the resonator) to the coupled quality factor (loss through the coupling port).

Resonance occurs when the capacitively coupled transmission line presents a short to the primary transmission line ($Z_{tot} = 0$). This sets the resonance condition:

$$\omega_0 C_c Z_1 = \cot \left(\frac{\omega_0 l}{v_p} \right) \quad (3.6)$$

where C_c is the coupling capacitor, and ω_0 is the resonance frequency. This is a transcendental equation but under the assumption of small coupling capacitance ($\omega_0 C_c Z_1 \ll 1$), the cotangent can be expanded about $\pi/2$ in orders of the difference between the uncoupled ($\omega_{\lambda/4} = \frac{\pi v_p}{2l}$) and coupled resonance frequency (ω_0):

$$\omega_0 C_c Z_1 = \cot \left(\frac{\pi}{2} + \frac{\pi}{2\omega_{\lambda/4}} (\omega_0 - \omega_{\lambda/4}) \right) = \tan \left(\frac{\pi}{2\omega_{\lambda/4}} (\omega_{\lambda/4} - \omega_0) \right) \simeq \frac{\pi}{2\omega_{\lambda/4}} (\omega_{\lambda/4} - \omega_0) \quad (3.7)$$

This gives a modified resonance frequency:

$$\omega_0 \simeq \frac{\omega_{\lambda/4}}{1 + \frac{2\omega_{\lambda/4} C_c Z_1}{\pi}} \quad (3.8)$$

3.5.2 Transmission Coefficient

The forward transmission coefficient along the feedline, S_{21} , of the capacitively coupled quarter wave resonator can be written as:

$$S_{21}(x) = 1 - \frac{Q_r}{Q_c} \frac{1}{1 + 2iQ_r x} \quad (3.9)$$

Where $x = \frac{\omega - \omega_0}{\omega_0}$ is the reduced frequency offset, Q_r is the total resonator quality factor, and Q_c is the coupling quality factor. Q_r sets the resonator bandwidth (BW) as $BW = \omega_0 / Q_r$. Q_r is a combination of the intrinsic resonator quality factor Q_i

and the coupling quality factor Q_c .

$$\frac{1}{Q_r} = \frac{1}{Q_i} + \frac{1}{Q_c} \quad (3.10)$$

3.5.3 Quality Factor

The quality factor is a measure of the loss in the resonator. A higher-quality factor resonator has less loss and will take longer for its resonant oscillations to dissipate once excited. Q_c parameterizes losses from the resonator out the coupling capacitor port, and Q_i is all intrinsic loss mechanisms in the resonator itself, including two-level system loss, radiative losses, magnetic vortex losses, and quasiparticle losses[66, 67, 68, 69]. Each internal loss mechanism can be associated with its own quality factor summing to a total resonator Q_i as given in Equation 3.11.

$$\frac{1}{Q_i} = \frac{1}{Q_{i,\text{TLS}}} + \frac{1}{Q_{i,\text{Radiative}}} + \frac{1}{Q_{i,\text{Magnetic}}} + \frac{1}{Q_{i,\text{qp}}} + \dots \quad (3.11)$$

Design and fabrication parameters affect the resonator Q_i , including resonator geometry, dielectric and superconducting material choice, oxide growth, and microwave package design. In the SO μ mux resonator design, densely packaging many resonators on the same silicon chip and a reliable SQUID fabrication coupled to sensor

bond pads constrains Q_i to roughly $2\text{-}4 \times 10^5$. Once Q_i of a particular resonator fabrication is known, the coupling capacitance can be varied to set the resonator bandwidth. For the SO μ mux implementation, Q_c is varied as a function of resonance frequency to maintain a constant bandwidth of ~ 100 kHz. This is a considerable readout channel bandwidth relative to the $\mathcal{O}(1 \text{ kHz})$ sensor bandwidth but provides noise advantages when considering flux ramping as discussed further in Sections 3.5.6 and 4.3.3.

3.5.4 RF SQUID Loaded Resonator-Resonance Condition

To read out an RF SQUID with a capacitively coupled $\lambda/4$ resonator, the resonator short to ground is replaced by an inductive load (L) to ground, which is transformer-coupled to an RF SQUID loop. Since the quarter wave transformer is no longer loaded with a short but instead with the impedance of the inductive load to ground, the resonance condition is modified to:

$$\omega_0 \simeq \frac{\omega_{\lambda/4}}{1 + \frac{2\omega_{\lambda/4}C_c Z_1}{\pi} + \frac{2\omega_{\lambda/4}L}{\pi Z_1}} \quad (3.12)$$

Due to the changing impedance of the RF SQUID as a function of input flux, the inductive load seen by the resonator is flux dependent; thus, the resonance frequency is also flux dependent.

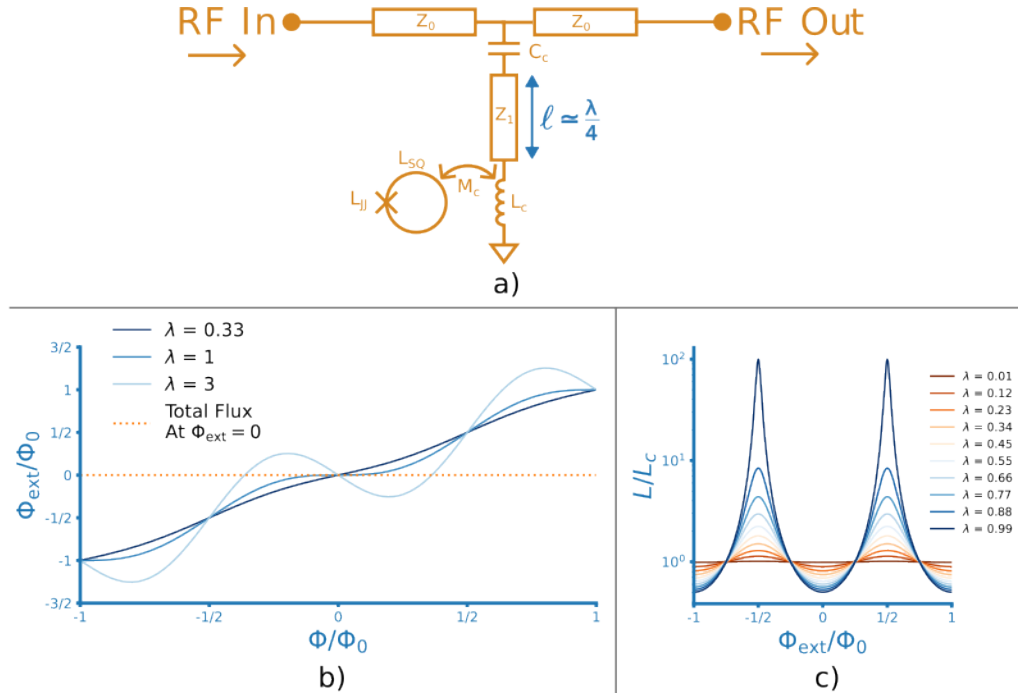


Figure 3.3: a) A quarter-wave resonator capacitively coupled to a transmission loaded with an inductively coupled RF-SQUID. b) Total flux in the SQUID (x-axis) is a combination of the external flux (y-axis) and the flux from the supercurrent through the josephson junction. Only when $\lambda = L_{SQ}/L_J < 1$ is this a single valued function. When the RF-SQUID is designed with $\lambda < 1$ it is non-hysteretic. c) The total effective inductance presented as a load to the quarter wave resonator as a function of the external flux (x-axis) and λ (color scale).

3.5.5 Flux Variable Inductance

In μmux , a dissipationless RF SQUID is used, which means no explicit shunt resistor is placed across the JJ, and only the parasitic capacitance and resistance of the junction itself are present. This minimizes the loss from the junction shunt resistance to maximize the resonator Q_i and does not produce any changes in resonator Q_i as a function of flux at first order. To determine the frequency shift as a function

of flux, we can use the Josephson equations (Equation 3.2) to solve for an effective flux variable inductance of the JJ (L_{JJ})

$$L_{JJ} = \frac{\Phi_0}{2\pi I_c} \sec\left(\frac{2\pi\Phi}{\Phi_0}\right) = L_J \sec\left(\frac{2\pi\Phi}{\Phi_0}\right) \quad (3.13)$$

Where L_J is the Josephson inductance and Φ is the total flux through the SQUID loop. The total flux includes a contribution from the externally applied flux through the loop (Φ_{ext}) and the flux induced by the tunneling supercurrent through the SQUID loop.

$$\Phi = \Phi_{\text{ext}} + L_{SQ} I_c \sin\left(\frac{2\pi\Phi}{\Phi_0}\right) \quad (3.14)$$

Where L_{SQ} is the self-inductance of the SQUID loop. Since equation 3.14 is a transcendental equation, in general, Φ is not a single-valued function of Φ_{ext} . When there are multiple values of Φ for a given Φ_{ext} , the inductance is hysteretic. L_{JJ} takes a different value when approached from increasing Φ_{ext} relative to approaching from decreasing Φ_{ext} . To avoid the hysteretic operation of the RF SQUID, Φ_{ext} must be monotonically increasing as a function of Φ .

$$\frac{d\Phi_{\text{ext}}}{d\Phi} = 1 - \frac{2\pi L_{SQ} I_c}{\Phi_0} \cos\left(\frac{2\pi\Phi}{\Phi_0}\right) = 1 - \lambda \cos\left(\frac{2\pi\Phi}{\Phi_0}\right) > 0 \quad (3.15)$$

Where we've introduced the hysteresis parameter $\lambda = L_{SQ}/L_J$. Equation 3.15 requires $\lambda < 1$ for non-hysteretic behavior. When the RF-SQUID is coupled to the resonator through a mutual inductance M_c , the effective load impedance presented to the quarter wave transformer is a flux variable inductance $L(\Phi)$:

$$L(\Phi) = L_c + \frac{M_c^2}{L_{SQ}} \left(\frac{\lambda \cos\left(\frac{2\pi\Phi}{\Phi_0}\right)}{1 + \lambda \cos\left(\frac{2\pi\Phi}{\Phi_0}\right)} \right) \quad (3.16)$$

Replacing L in the resonance condition from equation 3.12 with $L(\Phi)$ in equation 3.16, we see that the μ mux is a flux to resonance frequency shift transducer as shown in figure 3.4. By monitoring changes in the probe tone transmission (equation 3.9) amplitude and phase as a function of time, we can reconstruct the resonance frequency shifts and recover the sensor-coupled flux signal.

In the limit of λ , $M_c \rightarrow 0$, $L(\Phi) \rightarrow L_c$, and the resonance frequency becomes insensitive to SQUID flux, so we need both λ and M_c to be non-zero with larger values translating to a larger SQUID gain. To avoid hysteresis, however, we target $0.3 < \lambda < 0.6$, so the peak-to-peak resonance frequency shift, df_{pp} is tuned by adjusting the mutual inductance, M_c , between the resonator and SQUID. We target df_{pp} to match the resonator bandwidth (~ 100 kHz).

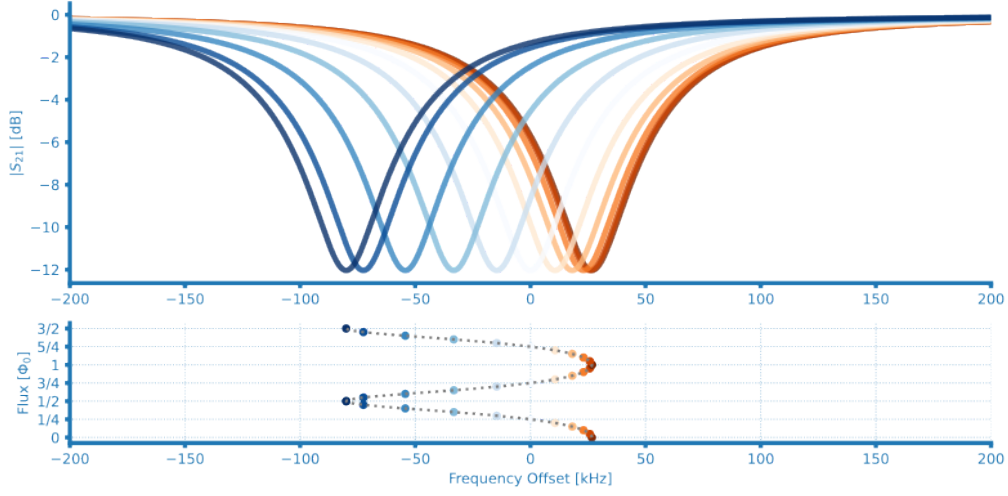


Figure 3.4: The magnitude of S_{21} of a SQUID coupled resonator as a function of flux through the SQUID (top) and the resonance frequency as a function of the flux through the SQUID (bottom).

3.5.6 Flux Ramp Modulation

Since the frequency shift is a periodic function of flux, there are points of maximum sensitivity ($df/d\Phi|_{\max}$) and zero sensitivity ($df/d\Phi|_{\min} = 0$) to input sensor flux. One would apply a feedback flux through a second transformer coil to maximize the SQUID sensitivity to keep the SQUID operating at a $df/d\Phi|_{\max}$. However, this would require an extra flux feedback circuit addressing each SQUID, defeating the thermal loading and wiring complexity reduction benefits of multiplexing. A compromise is to couple flux to all SQUIDs on a single RF feedline (~ 1000) through a common circuit. The current through all SQUID can then be ramped with a sawtooth signal with an amplitude of multiple flux quanta ($N_{\text{Ramp}} = \text{number of } \Phi_0$

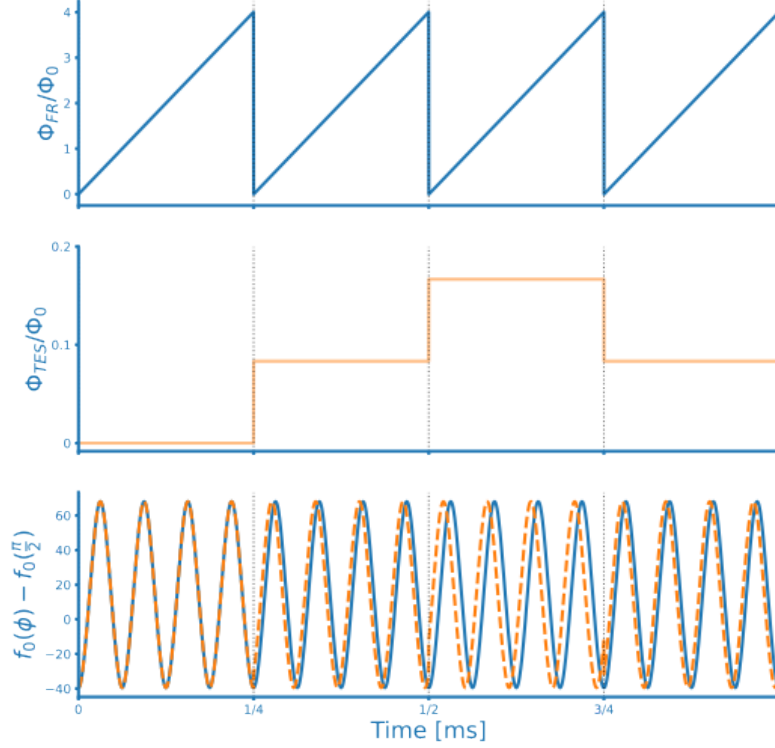


Figure 3.5: Flux ramp sawtooth input signal with an amplitude of $4 \Phi_0$ and reset rate of 0.25 ms (top). An example flux input from the TES input coil changing much slower than the flux ramp reset rate (middle). The resonance frequency shift vs time given the flux ramp signal in the top plot (bottom) if no TES signal is present (blue solid curve) and with the TES signal present in the middle plot (orange dashed curve). The addition of a TES signal produces a constant phase offset to the resonance frequency modulation in each ramp reset frame (vertical dotted grey lines) which can be demodulated to recover the TES input signal.

periods in the sawtooth amplitude) at a rate ($f_{\text{Ramp}} = 1/T_{\text{Ramp}}$ with T_{Ramp} = the reset period of the sawtooth) much faster than the bandwidth of the sensor. Since the TES signal changes at a rate much slower than the flux ramp rate, the added flux from the TES, Φ_{TES} , contributes a constant phase offset to the f - Φ modulation.

An example of flux ramp modulation is illustrated in figure 3.5.

The $f_0(t)$ response can be demodulated at the squid modulation rate $f_{SQ} = N_{\text{Ramp}} f_{\text{Ramp}}$ and the phase of the demodulated signal encodes the TES signal.

$$d_I + id_Q = \langle f_0(t) e^{i\omega_{SQ} t} \rangle_{T_{\text{Ramp}}} \quad (3.17)$$

$$\frac{\Phi_{\text{TES}}}{\Phi_0} = \frac{\arctan\left(\frac{d_Q}{d_I}\right)}{2\pi} \quad (3.18)$$

Equation 3.17, shows the demodulate real (d_I) and imaginary (d_Q) signal at the flux ramp rate with t the sample rate of the f - Φ curve and $\langle \rangle_{T_{\text{Ramps}}}$ indicating an average over the flux ramp reset period so that one sample of the TES signal is returned per ramp reset. The TES signal is just the phase of the demodulated data indicated in equation 3.18.

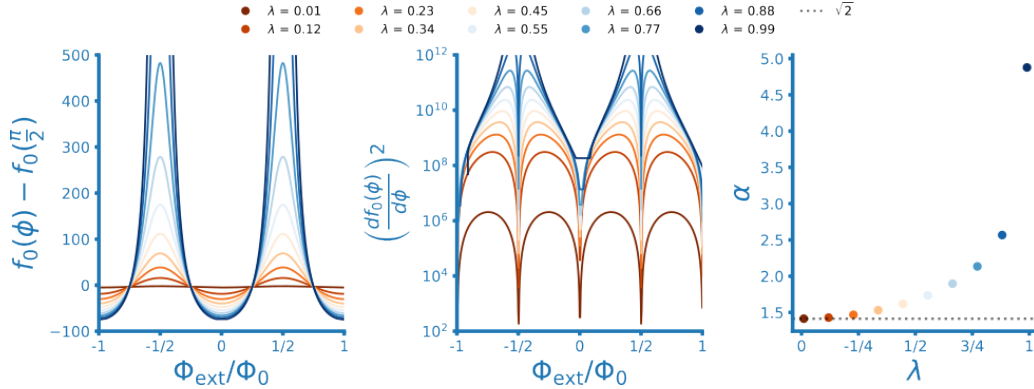


Figure 3.6: (left) The frequency shift vs external flux over two periods for $0.01 \leq \lambda \leq 0.99$ (color scale). (center) The square of the SQUID gain as a function of flux becomes more sharply peaked as λ increases. (right) The flux ramping penalty, α increases as λ increases and asymptotes to $\sqrt{2}$ in the limit $\lambda \rightarrow 0$.

The SQUID sensitivity when flux ramping, however, is reduced relative to biasing at the maximum f_0 - Φ derivative ($\frac{df_0}{d\Phi}|_{\max}$). When flux ramping, the sensitivity is the RMS derivative of the SQUID response as given by equation 3.19.

$$\langle \frac{df_0}{d\Phi} \rangle_{\Phi_0} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} \left(\frac{d}{d\phi} f_0(\phi) \right)^2 d\phi} \quad (3.19)$$

Where $\langle \rangle_{\Phi_0}$ indicates an average over one Φ_0 period. The penalty from flux ramping relative to applying feedback at the maximum sensitivity point can be characterized by the parameter α defined in equation 3.20.

$$\alpha \equiv \frac{\frac{df_0}{d\Phi}|_{\max}}{\langle \frac{df_0}{d\Phi} \rangle_{\Phi_0}} \quad (3.20)$$

For a perfectly sinusoidal response $f_0(\phi) = \sin(\phi)$, $\alpha = \sqrt{2}$. At small λ the f - Φ response is more sinusoidal and α asymptotes to $\sqrt{2}$ and as λ increases α increases. The design parameters for our μ mux chips are summarized in Dober *et al.* 2021[70].

3.6 Simons Observatory Readout Design

Here we discuss the specific implementation of the RF and DC components in the SO cryostats required for reading out $\mathcal{O}(10,000)$ detectors in each receiver's focal

plane.

3.6.1 Universal Focal Plane Modules

The SO focal planes are divided into modular units called universal focal plane modules (UFMs) due to the fact that their mechanical footprint is identical between the 3 dichroic detector array styles and is interchangeable between the different telescopes. They are hexagonal to maximize packing density by tiling in a tessellating pattern. The physical space constraints also required that all of the readout fits in a package with a height $\lesssim 5$ cm. The compact SO-UFM design contains the detectors, multiplexer chips, and the interconnecting RF and DC connections between the readout, detectors, and the outside world.

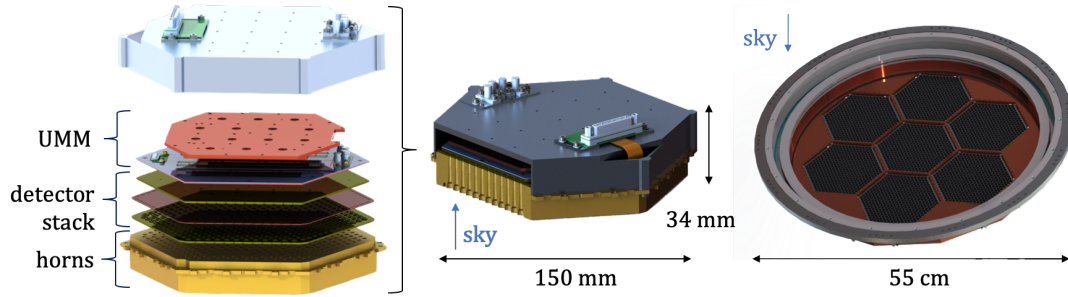


Figure 3.7: (Left) Exploded view of universal focal plane module (UFM) calling out the horn, detector stack and readout assembly (UMM) with the lid and connections to the outside world shown on the top. (Middle) Cross-section of the UFM with the profiled horns and internal silicon wafer stack seen from the side. (Right) Seven of the UFMs tiled into the focal plane of a SO-SAT. Figure reproduced from McCarrick *et al.* 2021[49].

The UFM's contain layered stacks of hexagonal silicon components as shown in Figure 3.7. Each silicon component is cut from 150mm diameter silicon wafers and is aligned via precision dowel pins press fit into the center and one edge of the assembly base. The dowel pins interface with a hole in the center and a slot on the edge of each silicon wafer to maintain alignment and alleviate stresses due to differential thermal contraction when the assembly is cooled from 300K to 100mK. The silicon stack is surrounded by a gold-plated aluminum housing to provide heat sinking and an RF-tight environment. On the sky side of the UFM's is a set of optical coupling components. For the MF and UHF UFM's, this consists of a gold-plated aluminum profiled horn block that interfaces with the on-chip planar orthomode transducers (OMT) and provides a mounting flange around its perimeter to attach the UFM's into the focal plane.

The OMT's are fabricated on the detector wafer which includes 432 pixels and sits on top of the horn block interfaced with a few deep reactive ion etched (DRIED) wafers that optimize the OMT to horn coupling. Each pixel contains an OMT with two orthogonal polarization probes. The horn and OMT structure impedance match free space radiation to an on-chip superconducting microstrip transmission line. The microstrips from each OMT probe are split with an on-chip power splitter and two separate microstrip lines carry the mm-wave signal to a matched impedance

termination on a TES bolometer. In between the power splitter and termination, a series of microstrip stub filters set the passband (for MF band centers are 90 and 150 GHz and 220 and 270 GHz for UHF) for the radiation into each bolometer. In addition to the 1728 optically coupled bolometer, 36 non-antenna-coupled bolometers are placed across the wafer to monitor the thermal drifts of the detector wafer. The detector wafer contains a pair of superconducting traces from each of the 1764 TES bolometers to a dense array of bond pads on the perimeter of the wafer.

On top of the detector wafer are two additional DRIEd silicon wafers that provide a back short structure to increase the coupling efficiency of the horn and OMT. On top of the backshort wafer is a self-contained copper enclosed assembly called the universal multiplexing module (UMM) that contains all of μ mux resonators and associated wiring required to read out the 1764 bolometers.

The UMM sits on a copper base that contains eight rows of 65 resonator channel multiplexer chips. Each half of the base contains 14 chips whose individual CPWs are tied together through wirebonds so that in total 910 resonators with resonances between 4-6 GHz share a common transmission line. The CPWs between the rows of multiplexer chips are connected via a silicon wafer with cutouts for the multiplexer chip rows called the routing wafer. The routing wafer connects through wire bonds to a low-loss printed circuit board that includes a CPW to subminiature push-on (SMP)

impedance-matched transition. The routing wafer also contains a dense array of bond pads on its perimeter, which are bonded down to the detector wafer perimeter bond pads and then routed to the TES-SQUID input bond pads on the multiplexer chips. A second flexible printed circuit board carries 14 twisted pair signals from a 37 pin micro d-subminiature (MDM) connector to the routing wafer. Two twisted pairs for the flux ramps on each half of the UMM, and twelve pairs provide a common voltage bias to ~ 150 TES detectors each. The shunt resistors in parallel with each TES is also fabricated on the routing wafer. A copper lid screws into the copper base of the UMM which provides an RF-enclosed cavity around each of the eight multiplexer chip rows and provides a common RF microwave ground path for the multiplexer chips.

Gold-plated aluminum walls are then mounted around all six edges of the UFM and a lid with Beryllium-Copper tripod springs is placed over top of the entire stack. The outer walls and lid serve to conceal the wirebonds, provide an improved thermal path across the silicon stack, and mechanically constrain the silicon stack. On the top side of the UFM four RF-SMP connectors and one 37-MDM connector interfaces the UFM to the rest of the cryostat.

3.6.2 RF Chain

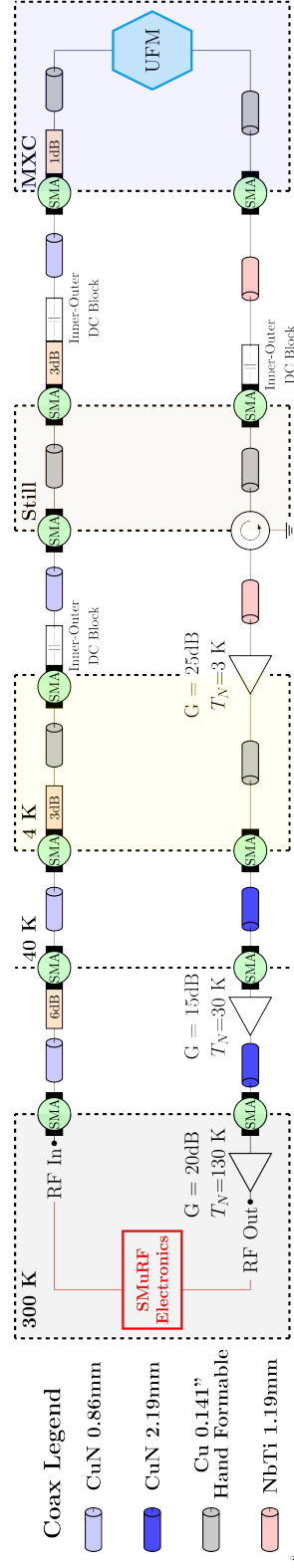


Figure 3.8: The SO RF chain. The RF chain includes all coaxial cables, attenuators, amplifiers, DC blocks, isolators, UFM's, and SMuRF electronics connected together to transmit tones to the UFM's and back to the SMuRF without degrading the readout performance or adding too much thermal loading to the cryogenic stages.

The RF chain refers to all transmission lines and microwave components "chained" in series to carry signals from the digital signal processing (DSP) electronics at 300K (section 3.7) to the UFM's and back from the UFM's to 300K. The final design of the RF chain is summarized in figure 3.8. The design principles that resulted in the chain shown in figure 3.8 are:

1. All components should have a nominal bandpass of ≥ 4 -8 GHz.
2. The optimum output power from the DSP electronics is 0 dBm. To utilize this full dynamic range with 2000 tones being generated, that sets an optimum output power per tone output = -33 dBm.
3. The power at the resonator, which minimizes the readout noise, ranges from -70 to -80 dBm.
4. Room temperature (noise generated by a 300K impedance matched termination) to be attenuated (including a 6 dB resonator dip) to 10% of the amplifier input noise temperature (including follow-on amps).
5. We want the power back into the smurf to be 0 dBm (summing all tones such that each tone is -33dBm and after the dc attenuators which have a 15 dB range).

6. The input noise temperature of the amplification chain should be as low as possible while still providing sufficient gain to achieve item 5 and linearity given item 3, 6 dB resonance depths, and 2000 tones within the 4-8 GHz bandwidth specified by item 1.
7. Assuming all other items are met, thermal conductivity of coaxial cabling and power dissipation of active components at cryogenic temperatures should be chosen to minimize power deposited by the RF chain on all cryogenic stages with a stricter requirement at the lowest cryogenic stages where the cooling power is the smallest.
8. Low Voltage Standing Wave Ratio (VSWR) < 1.2 components should be utilized along the chain to minimize standing wave ripples in the transmission. When higher VSWR components must be used to meet the other design criteria or VSWR at cryogenic temperature is unknown attenuators or isolators should be used in between these components to mitigate standing waves.

Sections 3.6.3, 3.6.4 and Rao and Silva-Feaver *et al.* 2020[71], walk through how we used the design requirements here to come to the RF chain design summarized in figure 3.8.

3.6.3 Input Transmission

Section 3.6.2 items 2 and 3 set the total input attenuation to be in the range 37-47 dB. It does not, however specify how much of that attenuation should be cold or at room temperature. A minimum cryogenic attenuation specification can be set using item 4, given that the best commercially available amplifier that meets the specifications of item 6 has an input noise temperature of ~ 3 K. At least $10\log_{10}(300K/0.3K) - 6$ (resonance depth in dB) = ~ 24 dB of cryogenic attenuation is needed to achieve item 4. To be able to tune across the 10 dB range specified in item 3[†] assuming we need at least 3 meters[‡] of room temperature PTFE coaxial cabling with a typical loss of 3 dB/m between the DSP electronics and the cryostat input this sets an upper bound of the cryogenic attenuation of 34 dB.

We can then further subdivide the cryogenic attenuation between the coaxial transmission lines and discrete attenuators in the cryostat. Given the geometry between the cryogenic stages and assuming use of typical low thermal conductivity coaxial cables constructed from CuNi between 300-4K and NbTi from 4K-0.1K the coaxes alone contribute 14-20 dB of the 24-34 dB budget. We chose then that the total cryogenic fixed attenuation should be 13 dB such that the total cryogenic

[†]Using variable attenuators at room temperature on the output of the DSP electronics

[‡]A conservative assumption based on the geometric constraints of the mounting location for the DSP electronics relative to hermetic RF vacuum feed-throughs on the cryogenic receivers.

attenuation should span 27-33 dB within our specified range.

The distribution of discrete attenuators between cryogenic stages is then chosen to satisfy item 7 and 8 while not requiring custom attenuators[§]. The chosen distribution is 6, 3, 3, 1 dB at the 40, 4, 1, 0.1 Kelvin temperature stages respectively.

Table 3.1 summarizes the total input attenuation and it's distribution between the different sub-assemblies, cabling loss, and chosen attenuators.

Table 3.1: Summary of input attenuation breakdown

Subsystem	Max [dB]	Typ [dB]	Min [dB]	Non-cable Attenuation [dB]
Input Chain Total	-37	-40	-47	-13 to -28
Input Chain Warm	0	-3	-3	Variable 0 to -15
Input Chain Cryogenic	-24	-30	-34	-13
300-4K	-15	-16.25	-17.5	-9
4-0.1K	-10	-12.5	-15	-4
Isothermal 4-4K	0	-0.5	-1	-

3.6.4 Output Transmission

Item 3 of section 3.6.2 combined with an assumption of 6 dB dip depths sets the power out of the mux chip and into the output RF chain to be -76 to -86 dBm. The desired per-tone input power into the DSP electronics is -33 dBm so the output chain must provide 43 to 53 dB of gain total. Linearity (discussed in section 3.6.5) and previous noise demonstrations with the mux drove us to select the first stage

[§]standard values are 0.5, 1, 3, 6, 10, 20 dB

amplifier to be LNF-LNA4_8LG which has 20 - 27 dB gain over its 4-8 GHz bandpass with a ~ 2.1 K input noise temperature when operated at 4 K. The input reflections for this component are relatively high so a cryogenic isolator is placed at 1K between the UFM's and 4K amplifier input. Arizona State University designed a follow-on amplifier to match the gain and linearity of this amplifier and operate at the 40K stage. It has a gain of 14-15 dB with an input noise temperature of 26-30 K. We set the allowed total 4 K amplifier input referred noise to be < 3 K which requires that the added noise from all follow-on amplifiers < 0.9 K. We allocate 0.6K of the 0.9K allocation to the 40K amp and 0.3K for any additional room temperature amplifiers. Assuming the minimum gain of the 4 K amplifier, a 0.6 K noise temperature refers to 60 K on the output of the 4 K amplifier, which is 3 dB above the maximum input noise temperature of the 2nd stage amplifier, so the maximum allowed loss between the two amplifiers is 3 dB. The total gain at this point (including the maximum loss) is 31-42 dB so an additional 12-16 dB of gain is needed to achieve the desired gain of 43-53 dB. We use the ZX60-83LN12+ amplifier from MiniCircuits at 300 K, which has a gain of 21-22 dB and an input noise temperature of 126 K. In order for the 300 K amplifier to contribute < 0.3 K at the 4K LNA input the loss between 40-300K must be less than 4.8 dB. To keep the loss between amplifiers below these allowed ranges no discrete attenuation is used and the cryogenic isothermal coaxial cables are

chosen to keep the loss within these ranges while minimizing thermal conductivity given the physical distances between the cryogenic stages within the receivers. For typical values, we take the range of expected components, not any upper bounds set by linearity or noise temperature described here. All of these output numbers are summarized in Table 3.2.

Table 3.2: Summary of output transmission breakdown

Subsystem	Max [dB]	Typ [dB]	Min [dB]	Non-cable Attenuation [dB]
Output Chain Total	43	-	53	Variable 0 to -15
Output Chain Warm	18	18.5	22	Variable 0 to -15
Output Chain Cryogenic	25	34	46	-
300-4 K	7	11	13.5	-
4-0.1 K	19.5	23	36	-
4-4 K	0	-0.5	-1	-

3.6.5 Linearity

The RF linearity is only impacted by active components in the system: ADCs, DACs, and amplifiers[¶]. For the cryogenic system this is just the amplifiers and should be limited by the lowest temperature amplifier because it has the tightest gain and noise constraints which are hard to optimize simultaneously with linearity.

The high end of the range in section 3.6.2 item 3 sets the ideal input third-order

[¶]The TES itself has important linearity limitations not discussed here, see, for instance, Takakura *et al.* 2017[72]

intercept point (IIP3) for the 1st stage amp. $-70 \text{ dBm} - 6 \text{ dB (dip depth)} + 33 \text{ dB (2000 tones)} = -43 \text{ dBm}$ is the maximum expected power at the amp. Then we want the input 1 dB compression point (IP1dB) to be greater than the expected input power. Typically IIP3 is $\sim 10 \text{ dB}$ higher than the IP1dB, so that sets the IIP3 for the amp to be $> -33 \text{ dBm}$ (which is achieved by the 4K amplifier). All follow-on amps are designed for this to be the limiting IP3 of the system.

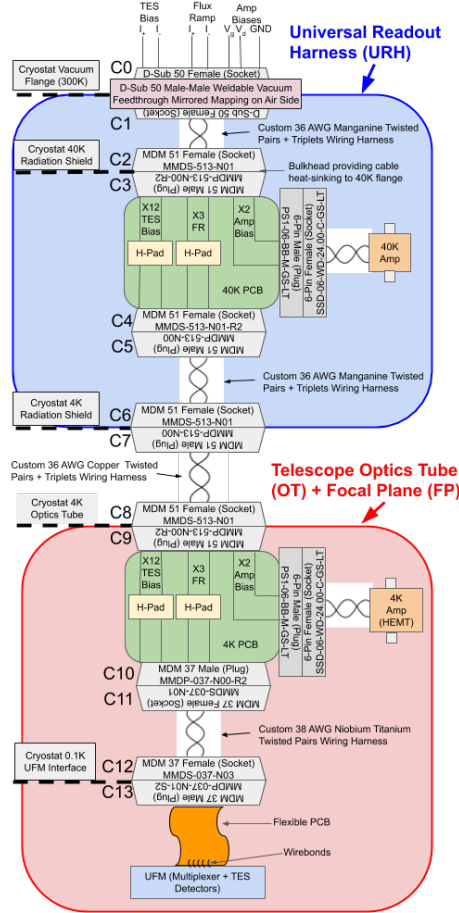


Figure 3.9: Schematic summary of readout DC wiring connections from the vacuum flange to the UFM. The type of connector is called out for each of the connections C0-C13 and the breakout PCBs (green) and connections to the 40 and 4 K LNAs are shown (orange on right).

3.6.6 DC Wiring

The readout DC wiring refers to the cryogenic wiring running from the 300K digital signal processing electronics to the UFM's. It must provide amplifier bias triplets (V_d , V_g , ground) for both the 40K and 4K amplifiers, voltage biases for the TES bolometers, and flux ramp input lines. The signal generation and filtering from the digital signal processing electronics to the cryostat vacuum hermetic feed-throughs is covered in section 3.7, so here we discuss the wiring from the vacuum feed-throughs to the UFM's as shown in Figure 3.9. The output from the DSP electronics and input to the UFM's assume that all of the signals for one UFM are connected through a single connector which requires the cable loom to contain four amplifier bias triplets, twelve TES bias pairs, and two flux ramp pairs (a minimum of 40 contacts). At 300K, 50-pin Dsub connectors were chosen because they are more common than MDM for hermetic weldable connectors, and inline pi filters are commercially available. At the cryogenic stages, 51-pin MDMs were chosen to achieve higher packing density. At 40 K and 4 K, inline PCBs are used to break out 6-pin connectors (2 x 3 pin for V_g , V_d , and Gnd) to bias the cryogenic amplifiers as shown in Figure 3.10. After the 4K PCB, the remaining TES bias and flux ramp signals are carried on 37-pin MDMs.

We use 36 AWG manganin wiring for 300-40K and 40-4K wiring because of its

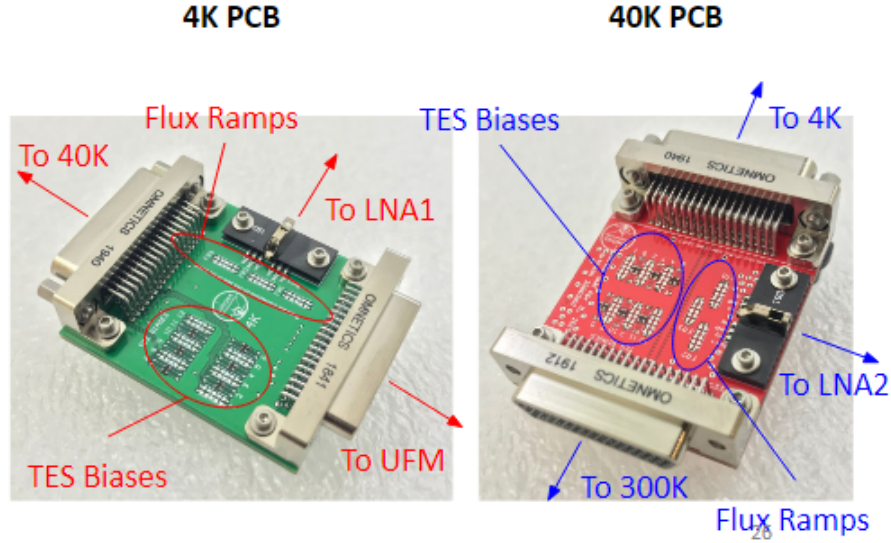


Figure 3.10: Photo of the compact 4 (left) and 40 (right) K PCBs used to pass the TES biases and flux ramp signals and break out bias signals for the cryogenic LNAs.

good ratio of thermal-to-electrical conductivity and its demonstrated performance in wiring harnesses in many cryogenic experiments. For all isothermal sections where we do not need low thermal conductivity material, we use 36 AWG copper wire. For all inter-temperature stage connections below 4 K, we use 36 AWG NbTi because it provides the best possible thermal-to-electrical conductivity ratio at these temperatures. All TES bias and flux ramp pairs and amplifier triplets are twisted to maximize the common mode rejection ratio, and all pairs/triplets are held together in their loom using a woven fiber weave. The wiring harness-to-connector interface is potted with a Stycast epoxy for strain relief.

3.7 Digital Signal Processing Electronics

Digital signal processing electronics capable of synthesizing and digitizing 2000 RF tones between 4-8 GHz, DC biases for the TES bolometers and RF amplifiers, and a sawtooth signal for the flux ramp with sufficiently low noise required to read out the μ mux circuit for the SO were not available at the start of the project. To enable μ mux readout for the SO, we developed with the engineering team in the technology innovation directorate at SLAC national accelerator laboratory the SLAC Microresonator RF (SMuRF) readout electronics between 2015-2020. In section 3.7.1 we overview the SMuRF overall system design, section 3.7.2 describes the design of the RF-analog front end, section 3.7.3 describes the low-frequency signal generation, and 3.7.4 describes the firmware and software design[73].

3.7.1 SMuRF Architecture

The SMuRF electronics are designed to operate in Advanced Telecommunication Computing Architecture (ATCA) crates. ATCA crates are a commercially available telecommunications architecture commonly used in accelerator applications that consist of a backplane that provides power, timing, networking, and sensor distribution to a series of "slots"^{||}. In addition, ATCA crates contain power supplies, inlet and

^{||}ranging from 1-14, the SO uses 7-slot ATCA crates

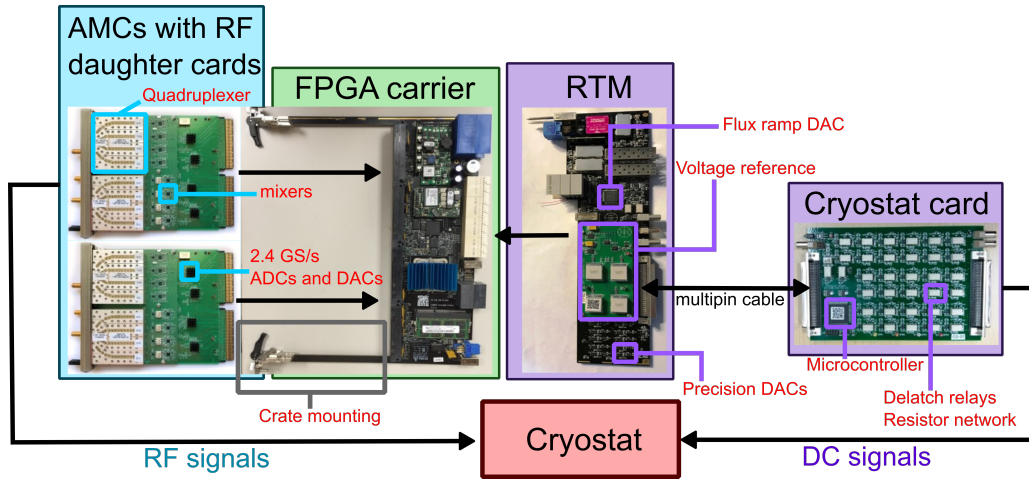


Figure 3.11: A picture of the hardware for one SMuRF system accommodating up to 3328 resonator channels across one or two RF chains, excluding peripherals such as the server, data streaming, and timing. The pictured contents excepting the cryostat card occupy one slot in a multi-slot ATCA crate. The Advanced Mezzanine Cards (AMCs) with RF daughter cards mount to the front of the carrier and present SMA connectors for RF connection to the cryostat, while the RTM mounts to the rear of the carrier and presents a multipin cable to a cryostat card, which mounts directly to the cryostat. The RTM and cryostat card combination supplies the low-frequency signal generation and conditioning.

outlet fan trays, and an onboard computer running a bare-bones Linux operating system called a shelf manager that manages sensor alarms, powering slots, and the fan cooling program. Each slot can power a carrier that inserts into the front of the crate and can support two advanced mezzanine cards (AMCs). Each slot can also power a rear transition module (RTM), which inserts into the rear of the crate.

The SMuRF carriers and AMCs support the RF frontend (section 3.7.2) and an FPGA that runs custom firmware (section 3.7.4). The SMuRF RTMs provide low-frequency signal generation (section 3.7.3). Between the RTMs and the 50 DSub

connectors is a small board mounted right on top of the vacuum feedthrough into the cryostat called the Cryostat Card (Cryocard). The Cryocard provides adjustable filtering of the TES bias and flux signals out of the RTM, and control of the cryogenic amplifier biases. Figure 3.11 shows all of the boards that make up one SMuRF ATCA slot.

The first slot of each crate is reserved for a commercial network switch that provides a 10 Gbit/s fiber connection to each of the slots for data streaming. A server computer is connected via a PCIe card running SLAC-designed custom firmware that provides up to a 70 Gbit/s connection through multi-fiber MPO connectors. A standard 1 Gbit/s LAN connection is provided from the server to the network switch for loading new firmware onto the carriers, and another LAN connection communicates between the server and the crate shelf manager.

A custom SLAC carrier and RTM generate a timing signal that can be carrier over fiber. Another custom SLAC carrier in the SO-office crate provides timing fiber signal fanout from one input to multiple outputs. From the office fanout, a timing fiber is run to each of the telescope platforms. Another fanout carrier in each of the platforms distributes the timing signal from one crate to the RTM in slot 2 in all of the others on the platform. The timing within each ATCA crate is then distributed to all other slots through the crate backplane.

3.7.2 RF Frontend

The AMCs contain the ADCs and DACs for the main RF signal path, clock generation system, and LO generation for the RF daughter cards. Each AMC consists of a base board interfaced with an RF daughter card supporting a total of 2 GHz of bandwidth between 4-6 GHz. The RF daughter card provides the up- and down-mixing between the ADC and DAC band centered at 750 MHz and the 4-6 GHz RF frequencies of the microwave resonators.

AMC Base Board

The AMC base board contains the hardware for interfacing the digital logic on the carrier (section 3.7.4) with the analog electronics. A block diagram is shown in figure 3.12.

Each AMC baseboard features four 16-bit 2.4 GS/s ADCs (TI ADC32RF45) and four 16-bit 2.4 GS/s DACs (TI DAC38J84). The SMuRF implements I/Q data separation digitally, using *Digital Down Conversion* (DDC) rather than using integrated analog I/Q combiners and separators. In this scheme, rather than implementing separate I and Q hardware paths, a single high-speed ADC samples the full input signal, which is then separated into orthogonal components via multiplication with a directly synthesized intermediate frequency (IF), filtered, and downsampled to obtain

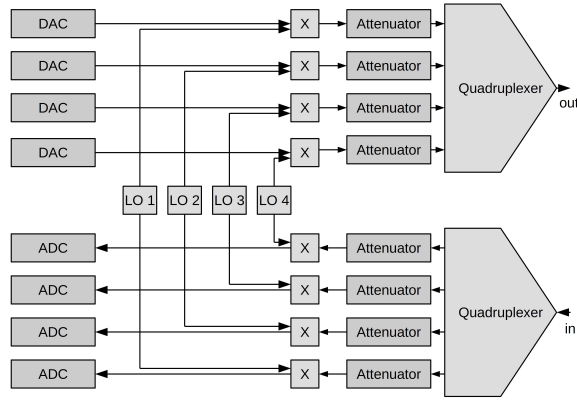


Figure 3.12: Block diagram of a SMuRF Advanced Mezzanine Card (AMC), which is responsible for the RF signal input/output. RF signals enter the system between 4-6 GHz and are first split into 500MHz bands via a quadruplexer (bottom right). Each band is then processed in parallel through digitally controllable attenuators and mixers (denoted with X) each with an LO tuned to place the IF band in the 500-1000MHz range. The ADCs (bottom left) then sample the IF of each band at 2.4 GS/s. The RF output (up conversion) chain starts from the DACs (top left) and performs the same operations in reverse with the mixers fed by the same LOs as the down conversion. The separate bands are combined with the quadruplexer, resulting in a comb of output between 4-8 GHz at the RF output (upper right).

the complex baseband signal. This relaxes the requirement on precisely matching the gain and maintaining the phase offset between the I and Q components. However, DDC requires higher speed DACs and ADCs to do the I/Q combination and separation in firmware.

The DACs convert input complex data from the FPGA carrier card at 614.4 MS/s to a 500 MHz-wide filtered signal between 500 MHz and 1 GHz. The ADCs convert the unaliased input signal in the 500 MHz-1 GHz frequency range up to a complex data stream at 614.4 MS/s to pass to the carrier. The ADCs and DACs are run with 4x digital decimation and interpolation filters, respectively, to define the Nyquist band and simplify analog filtering requirements. Data bandwidth limits prevent the ADCs from transmitting wholly unprocessed ADC data and the DACs from processing arbitrary waveforms at the full data rate. The communication between the carrier and the ADCs and DACs is therefore routed through a 12 Gb/s custom, open-source JESD protocol.

Each DAC/ADC pair is associated with a unique local signal generator to provide the LO that the RF daughter card uses to upmix or downmix between the 500 MHz-1 GHz band and the desired GHz-frequency band. Each 500 MHz band has its own LO, which is offset by the ADC/DAC numerically-controlled oscillator (NCO) frequency of 2.4 MHz from the centers of 3.5 GHz, 4.0 GHz, 6.0 GHz, and 6.5 GHz.

We use the lower sidebands for the first two LOs and the upper sidebands for the second two such that the LOs fall outside the resonator bandwidth. These LOs are generated via a fractional-N locked phase-locked loop IC (Analog Devices ADF5355) with integrated low-noise voltage-controlled oscillators (VCOs).

Finally, the AMC card contains a clock generation system (TI LMK04828) for the digitizer and associated JESD204b links. This clock, as well as all other SMuRF system clocks, are ultimately derived from a common timing reference (Crystek CVHD-950).

AMC RF Daughter Cards

The AMC RF daughter cards mounted to the AMC base boards provide the RF frequency tone generation and readback. The RF mixers (Analog Devices LTC5548) that provide the up- and downmixing between the DAC/ADC baseband from the AMC base board and the LOs are contained on the RF daughter cards. Since the RF mixer typically sets the linearity of the system, the choice of mixer here is critical, as is the attenuation of DAC output power prior to passing to the mixer.

Quadruplexers (Lark Engineering SMCQ819-110) interface from the mixers on the four ADC/DAC pairs to a single input/output coaxial pair that connect with the cryostat or other test system via SMA connectors. Cavity filters within the

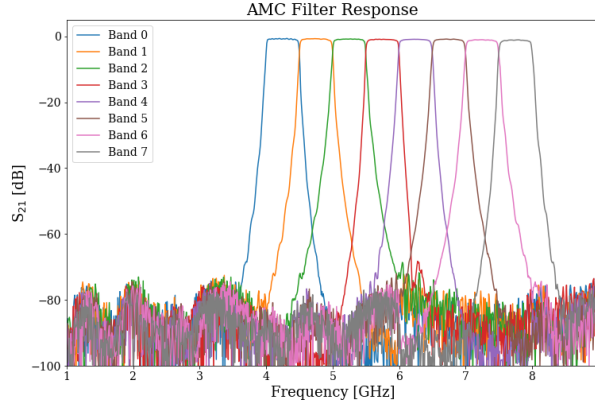


Figure 3.13: The response of the AMC band-defining filters for each of the 500 MHz wide bands across the full 4-6 GHz SMuRF band. Band 0 corresponds to the 4-4.5 GHz band, and the bands count upwards in 500 MHz increments from there.

quadruplexers define overlapping bandpass filters that split the RF frequency band into each 500 MHz band that is synthesized or channelized in firmware. Since the ADCs and DACs are sampled at 2.4 GS/s, the Nyquist frequency for the alias band is at 1.2 GHz.

The AMC S_{21} response for each 500 MHz band is shown in Figure 3.13. We note in particular, that the maximal passband of the filters is not quite 500 MHz wide to guarantee sufficient rejection out of band, set at 60 dB isolation between bands. The portion of the passband that is flat within -2 dB is about 460 MHz wide, while the edges of each band are about 3 dB down and isolation is > 65 dB between adjacent bands. The RF daughter cards additionally include programmable attenuators on both the RF input and output for each 500 MHz band to allow for flattening the RF

signal levels over the full SMuRF RF bandwidth.

Since the ADCs and DACs have optimal noise performance at a fixed output/input power level, respectively, adjusting power levels at the resonator is most easily accomplished via some combination of external attenuation/amplification and these programmable attenuators. With existing default settings, the optimal output/input power levels are about -33 dBm per tone. Individual tone powers are adjustable in firmware in coarse (3 dB) steps, while the programmable attenuators act on each 500 MHz band defined by one ADC/DAC pair and can be tuned in finer (0.5 dB) steps. The size of firmware steps can be modified easily in firmware, but the present combination of coarse and fine steps has been sufficient for user needs.

3.7.3 Low-Frequency Subsystem

Rear Transition Module

The rear transition module (RTM) generates the following low-frequency signals for the SMuRF system: (1) Detector biases, (2) RF amplifier biases, and (3) flux ramp. It additionally provides control and clocking to the cryostat card.

The RTM acquires its sequencing and interfaces via a complex programmable logic device (CPLD), which runs a simple, compact firmware programmed in VHDL. The CPLD firmware cannot be updated remotely.

The DC TES biases and cryogenic amplifier biases are provided by 33 low-noise 18-bit DACs (Analog Devices AD5780). To reduce low-frequency noise, the TES bias DACs are operated differentially.

The RTM allocates eight DACs to bias up to four total RF amplifiers, two amplifiers placed at 4K and 50K for two separate RF chains. Each amplifier is allocated two DACs for the gate and drain voltages. These DACs, referred to as “slow”, can generate arbitrary waveforms up to about 1 kHz for calibration and testing purposes.

An additional 50 MS/s 16-bit differential current DAC (Analog Devices LTC1668) drives a differential pair that serves as the flux ramp. Since the flux ramp is required to be much faster than the detector and amplifier biases and operates at lower full-scale voltages, it is higher speed and, therefore, higher noise than the DC bias DACs. The flux ramp is implemented as a sawtooth function. It is generated by the RTM CPLD with control and triggering from the carrier card FPGA. The RTM finally provides DC power and a low-speed SPI serial link for communication to the cryostat card.

A daughter card on the RTM consists of a voltage reference providing buffered $\pm 10\text{V}$ reference signals to the precision DACs. The card uses two commercial voltage references (Analog Devices LTC6655) in parallel to reduce noise, and low-pass filters with low acoustic sensitivity capacitors (Rubycon 35MU106MD35750) to mitigate

vibrational pickup. A daughter card implementation for the voltage reference provides the flexibility to easily and cheaply upgrade the performance of the voltage reference if required for a particular application.

The RTM presents a 100-pin SCSI connector containing the low-impedance flux ramp, TES bias, amplifier bias, DC power supplies, and digital communication to the cryostat card. The SPI communication link between the RTM and the cryostat card is default quiet to reduce noise during operation, with no active polling. An SFP port receives timing data that may be used for synchronizing multiple SMuRF systems together.

Cryostat Card

The cryostat card interfaces signals from the RTM directly with the cryostat. It conditions the TES bias and flux ramp signals from the low-impedance voltage sources at the output of the RTM to filtered, high-impedance current sources. Since the bias lines are particularly susceptible to noise pickup after converting to current sources, the cryostat card was designed to be placed as close to the cryostat as possible, with a separate card and enclosure from the main RTM to allow for shielding and thermal regulation. The connector on the cryostat side is a 50-pin dsub to matching SO receiver 300 K vacuum plate feedthrough.

The cryostat card additionally provides regulated power supplies and conditions gate bias voltages from the RTM DACs to bias the cryogenic RF amplifiers. The card is controlled and monitored via an on-board microcontroller (Microchip PIC32MX795 32-bit microcontroller) and interfaces to the rest of the SMuRF system via an SPI serial link to the RTM CPLD.

The detector bias is a differential output voltage using two opposing $\pm 10\text{V}$ RTM DACs. For each detector bias line, mechanically latching relays on the cryostat card (Axion 5-1462037-4) controlled by the microcontroller allow for switching between a low-pass filtered, low-current bias circuit (“low current mode”) and a higher current, minimally filtered ($\sim 100\text{ kHz}$ bandwidth) circuit (“high current mode”). The relays are mechanically latching to reduce power dissipation that could couple to temperature variation in the board, meaning the state cannot be queried during operation. The default state is in low current mode.

In low current mode, added parallel capacitance low pass filters the detector bias outputs with a frequency cutoff $f_{3\text{dB}} \sim 8\text{Hz}$. The output current levels in both low and high current modes are set with inline series resistors, typically in the 1-50 $\text{k}\Omega$ range depending on the maximum current desired, providing a high impedance source. The maximum current is thus determined by the resistor selection, which is set by individual detector operating parameters, up to op-amp limits of $\sim 10\text{ mA}$.

This high current mode is used to drive TESs from their superconducting to normal operating states, or for high bandwidth measurements.

We set stringent requirements on the low-frequency performance of signals provided by the cryostat card. It is expected that the temperature variation of electronic components drives a large fraction of the variation in the low-frequency noise, particularly the temperature variation in the RTM voltage reference and cryostat card resistors. This sets a requirement of $\lesssim 20 \text{ ppm}/^\circ\text{C}$ on the temperature coupling coefficient of these resistors. The cryostat card resistors were chosen to be guaranteed to $\pm 10 \text{ ppm}/^\circ\text{C}$ (TE Connectivity RQ73 series). The capacitors on the cryostat card were similarly selected for low frequency noise performance and low acoustic sensitivity (Rubycon).

The 4K and 50K amplifier gate bias voltages are each driven single-ended via an RTM DAC with $20 \mu\text{F}$ bypass capacitors on the cryostat card for filtering. A resistor divider on the cryostat card allows for limiting the output gate voltages to less than the maximum amplifier ratings if required. The drain currents may be read back to allow for bias adjustment using an inline resistor on the input to the regulator, which is amplified by an instrumentation amp and read back by ADCs on the microcontroller.

All output voltages, including the RF amplifier drains, are disabled by default on

power-on and must be explicitly enabled. This is a requirement for many cryogenic RF amplifiers, which can draw large currents and generate substantial heat loads if the drain voltage is enabled before the gate voltage is set. For the RF amplifier drain voltages, which are generated on the cryostat card using adjustable linear voltage regulators, this is accomplished by toggling the enable pin of the linear regulators using the cryostat card PIC.

The flux ramp is a differential output current source drive, converted from the RTM low-impedance voltage source via a passive resistor H-pad attenuator. A mechanically latching relay allows for switching between AC and DC-coupled modes via large capacitors. In AC mode, large acoustic-sensitive capacitors (Rubycon) are used to mitigate against vibrational coupling. Operating in AC mode high-pass filters the differential flux ramp drive with an $f_{3\text{dB}} \sim 80$ Hz, attenuating 60 Hz power line noise and low-frequency noise pickup in the cables between the RTM and cryostat. Operating in the DC-coupled option enables low-frequency and fixed-current flux ramp biasing as required, such as when taking resonator frequency versus flux ramp bias curves or other device characterization data.

3.7.4 Firmware and Software Architecture

The SMuRF firmware was a significant development effort led by SLAC and is explained in detail in [73]. Here I briefly summarize the key aspects of the firmware relevant to this dissertation. The firmware is responsible for generating and processing the transmitted and received tones. Both tone synthesis and analysis is performed via a polyphase filter bank (PFB)[74]. Each analysis polyphase filter bank channel produces a complex datastream sampled at 2.4 MHz which is multiplied by a complex calibration parameter, η . η rotates and scales the complex data stream. The η -rotation is chosen such that changes in the transmission through the resonator due to the resonance frequency shifts are detected maximally in the imaginary component of the datastream and minimally (ideally zero) in the real component. The η -scaling rescales the complex data stream units into equivalent resonance frequency shift, Δf .

To estimate η we use a measurement of the complex analysis PFB signal as a function of synthesis PFB frequency around each of the μ mux resonators. From this measurement η is the vector that connects the points at $f_0 + df$ and $f_0 - df$ where f_0 is the resonance frequency and df is a tunable value typically chosen to be 10 kHz. Figure 3.14 shows the η estimation and rotation process. The original measured data is shown in blue with the points used to calculate η called out with yellow stars. The

angle of the vector is used to rotate the blue points to the red points so that the tangent to the circle is parallel to the Q-axis. The amplitude calibration factor— $2df/|\eta|$ —is used to scale the analysis PFB complex data stream to frequency-offsets Δf .

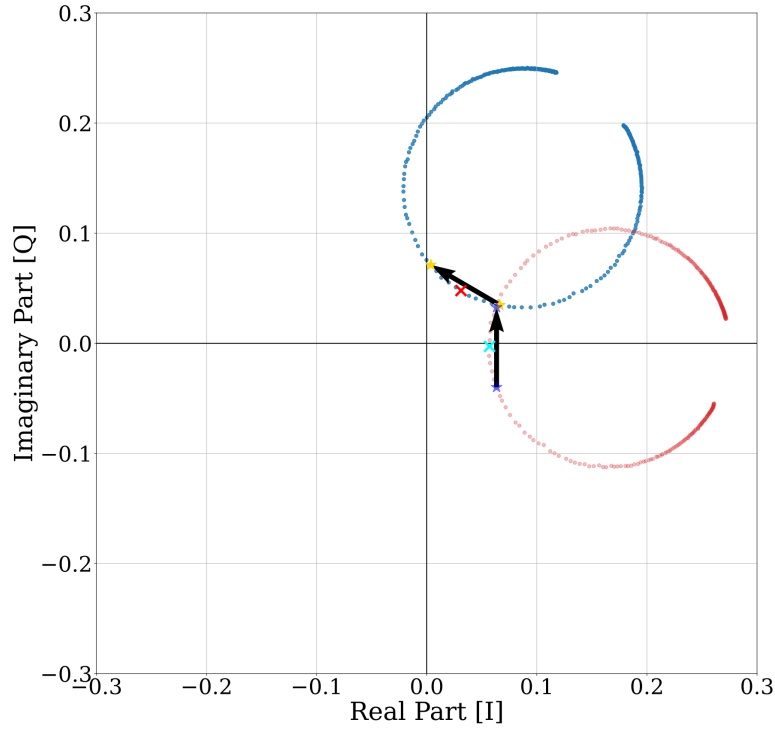


Figure 3.14: Example of the analysis polyphase filter bank calibration parameter— η —estimation and rotation process. The original measured data is shown in blue with the points used to calculate η called out with yellow stars. The angle of the vector is used to rotate the blue points to the red points so that the tangent to the circle is parallel to the Q-axis. The amplitude calibration factor— $2df/|\eta|$ —is used to scale the analysis PFB complex data stream to frequency-offsets Δf .

When the flux ramp is enabled a stochastic gradient descent (SGD)[75] is used to fit out the first three fourier coefficients of the flux ramp modulation frequency from the frequency-shift data stream. The arctangent of the first harmonic sine amplitude divided by the cosine amplitude, $\theta = \arctan\left(\frac{s_1}{c_1}\right)$, is the measured phase shift of the flux ramp modulation which encodes the detector signal. The phase measurement is then averaged over all points in a flux ramp reset time ($\sim 4kHz$). The phase data from all PFB channels can be directly written to disk at this rate or downsampled. We typically downsample the data from 4 kHz to 200 Hz for normal data streaming.

The sum of the fourier series from the SGD is the fitted value of the frequency shift, $\hat{\Delta f}$. The $\hat{\Delta f}$ estimated at each 2.4 MHz sample of the PFB frequency-shift data stream and added to the current tone frequency programmed for the synthesis, f_n , and then the updated estimate of the current resonance frequency, $f_{n+1} = f_n + \hat{\Delta f}_n$, is used to update the synthesis PFB tone frequency. This closed loop process ensures that the tone frequencies stay centered on the resonance frequency minima thereby maximizing the resonator phase sensitivity $\frac{d\text{Im}S_{21}}{df}$. Keeping the synthesized tones centered on the resonance frequencies also minimizes the power transmitted to the LNA which improves system linearity.

3.8 Readout Tuning and Setup

Here we discuss the steps required to configure the SMuRF electronics to prepare the readout and detectors for CMB observations.

3.8.1 Power Level Check

The first step in setting up the readout is to enable all of the warm and cryogenic amplifiers and check the transmission through the cryogenic receiver. This can be done by either setting on a number of fixed frequency tones using the filter banks or by playing noise from a lookup table on the DAC directly. The transmitted signal should not saturate the ADC and should show resonance line shapes. If the ADC is saturated then additional inline attenuation must be added and if no resonance line widths are found when below ADC saturation the amplifier bias and all RF connections should be checked.

3.8.2 Phase Delay Correction

As described in section 3.7.4, the SMuRF can correct for time delays between the commanded output and measured input by buffering data coarsely in the baseband processor with a resolution ($1/2.4 \text{ MHz} = 0.416 \text{ } \mu\text{S}$) and with a finer resolution ($1/307.2 \text{ MHz} = 3.2 \text{ nS}$) in the ADC data block. The total delay through the system

is a combination of cable delay, processing delay, and polyphase filter delay. In total we have $\sim 8 \mu\text{s}$ mostly from the polyphase filter bank. As described in section 3.8.3, we identify resonances by looking for large phase changes with frequency, but a large round trip time delay dominates over the resonance phase response, so we must first estimate the time delays and set the course and fine delay settings to compensate for them.

A time delay produces a linear phase as a function of frequency, as given by equation 3.21

$$\theta_n = 2\pi f_n \tau \quad (3.21)$$

where τ is the time delay, θ_n is the phase delay at frequency f_n . The SMuRF phase delay setup procedure includes 1) measuring the phase vs frequency using a noise file played from the DAC and read back by the ADC cutting out the firmware filter banks and 2) measuring the phase of a series of frequency-increasing tones generated by the synthesis filter bank and read back with the analysis filter bank. A line is fit to both to get two time delays. The first measurement gives us a measurement of the DAC and ADC latency as well as the cable delay, and the second measurement gives us the same as the first but also with the latency of the filter bank. The first measurement can vary from cryostat-to-cryostat, but the second value should be the same across different systems that are running the same version of the firmware.

3.8.3 Resonator Identification

The next step involves identifying the location of all of the resonances on each RF chain so that the initial tone frequencies for the output comb can be generated. This involves coarsely sweeping (in steps of ~ 40 kHz) a tone through each of the synthesis filter bank channels and reading the amplitude and phase back with the analysis filter bank. We then look for locations with a local minimum in S_{21} by checking for $\frac{d|S_{21}|}{df} = 0$ and $\frac{d^2|S_{21}|}{df^2} > 0$ and a maximum phase derivative $\frac{d^2\angle S_{21}}{df^2} = 0$ and $\frac{d^3\angle S_{21}}{df^3} < 0$. This produces a list of candidate resonance frequencies. Around each of the candidate frequencies, a fine resolution sweep of 600 kHz width in steps of 2 kHz is performed to refine resonance frequency. Any resonances with a shallow depth (low-Q) or a collision with a neighboring resonance are disabled here. The depth and collision distance for cuts is adjustable, but we typically use < 1 dB depth and < 200 kHz separation for SO.

3.8.4 η Calibration

From the fine sweep data the η parameters described in section 3.7.4 are calculated using points Δf above and below the each of identified resonances. SO uses a typical Δf of 20 kHz to calculate η . After this step the IQ streams from each channel are converted to f (frequency) and df (frequency errors) by applying the η rotation and

scaling and the slow (< 1 kHz) feedback is enabled to keep all tones on resonance.

3.8.5 Flux Ramp Setup

Next the flux ramp amplitude and sawtooth reset rate are set and the ramp signal is applied to all resonances. The truncated fourier series expansion at the tracking rate (ramp amplitude times ramp reset rate) is enabled and the output demodulated flux ramp phase signal can now be streamed to disk at the flux ramp reset rate. Any channels that 1) do not respond to the flux ramp 2) have too high or too low peak-to-peak frequency shift or 3) have too large of a frequency error are disabled. Channels that fall into category 1 are calibration resonators that are not inductively coupled to the flux ramp. Channels that fall into category 2 and 3 have anomalous SQUID properties (too high lambda, outlier values of SQUID to resonator coupling, or are too close to their resonance neighbors) which result in higher noise if left on. Only a small number of channels are typically cut at this step $\lesssim 5\%$.

3.8.6 Noise Estimation and Power Optimization

Next the demodulated phase from all channels are streamed at ~ 200 Hz for 10 seconds. The mutual coupling between the TES input and the SQUID is used to

calculate $\frac{dI_{\text{TES}}}{d\theta_{\text{Fluxramp}}}$

$$\frac{dI_{\text{TES}}}{d\theta_{\text{Fluxramp}}} = \frac{M_{\text{input}}}{2\pi\Phi_0} \quad (3.22)$$

where θ_{Fluxramp} is the phase of the demodulated flux ramp signal. $\frac{dI_{\text{TES}}}{d\theta_{\text{Fluxramp}}}$ converts the output demodulated phase to equivalent current through the TES and is roughly 1.4×10^6 pA/radian for the SO μmux design. The power spectrum of the timestream is estimated using the welch method and mean amplitude spectral density (root of the power spectral density) between 10-30 Hz is used as an estimate of the white noise component. As discussed further in chapter 4 we expect the readout contribution to the white noise to be $\sim 35 - 45$ pA/ $\sqrt{\text{Hz}}$. The white noise contribution measured at this step in the setup is a combination of the Johnson noise of the superconducting shunt resistors and the readout noise. At the focal plane base temperature of 100 mK we expect the shunt resistors ($\simeq 400 \mu\Omega$) to contribute $\text{NEI}_{\text{Johnsons}} = \sqrt{\frac{4k_B T}{R}} = 117.5$ pA/ $\sqrt{\text{Hz}}$ since this noise is much larger than the expected readout noise, uncertainty in the bath temperature and shunt resistance have a large effect on the estimated readout noise when we subtract (in quadrature) the Johnson noise from the total noise. As a cross-check we apply a large current to all of the TES bias lines or raise the bath temperature to drive the detectors normal and take another white noise measurement. In this state the total noise is the quadrature sum of the readout noise and the Johnson noise from the normal detectors (~ 8 m Ω) at a temperature of

$\gtrsim T_c \simeq 160$ mK which is closer to $35 \text{ pA}/\sqrt{\text{Hz}}$ and we have much higher sensitivity to the readout noise. If the superconducting and overbiased noise give roughly consistent numbers for the readout noise and they are much higher than the expected we perform a power optimization.

Since the RF power level at the resonator is one of the most important factors that determines the readout noise, as discussed in chapter 4, the first noise optimization step is a sweep over RF power levels. This optimization consists of sweeping the programmable attenuators on the input to the cryostat over their 15 dB tuning range to find a power level that minimizes the noise when the detectors are overbiased. The power optimization is done in each 500 MHz band because the input attenuation has a typical -2 dB/GHz input attenuation slope which requires more attenuation in the lowest band and less attenuation in the highest band. A $\sim 3\text{-}6$ dB fixed attenuator is typically placed on the output of smurf in order to have the noise minimum be roughly in the middle of the 15 dB tunable range of the smurf attenuators. After this step the readout is set up optimally and should be able to track the resonances indefinitely. The next operations are all related to configuring the TES bolometers.

3.8.7 Biasgroup Mapping

Next the detector bias lines are all set to 0 volts bringing the detectors into their superconducting state. In the superconducting state the bolometer resistance is much less than the shunt resistance and all current commanded down the detector bias lines goes through the TES input coil to the SQUIDs. A series of small bias voltage steps are commanded down one bias line at a time while data from all of the resonances is recorded to disk. The recorded timestreams are fit to identify which readout channels are connected to TESes and for the TES coupled readout channels which bias lines they are each attached to. In theory this is knowable from how things are wired up between the multiplexer and the detectors but in practice we have to get this from probing the system because the resonance frequency ordering can make identifying the design multiplexer channels challenging.

3.8.8 IV Curves

Next current-voltage (IV) curves are taken on each of the detector bias lines to determine the bias voltage to use for each bias line that will simultaneously maximize sensitivity** and operable†† detector count. During the IV procedure the detectors are

**Sensitive here refers to maximizing $s_I = \frac{dI_{\text{TES}}}{dP_{\text{TES}}}$ where P_{TES} is the optical power on the detector.

††Operable means biased in the superconducting transition, not in their normal or superconducting states

first brought to their normal state by switching to high-current mode and applying a current larger than the critical current of the TES film ($\sim 5\text{-}10$ mA). The current is held at this high level for a short time ~ 1 s and then dropped down to a smaller value to minimize the time left at the full joule heating when overbiasing ($P_{\text{Joule}} = N_{\text{TES per bias line}} I_{\text{overbias}} R_{\text{shunt}} \simeq 6 \mu\text{W}$). After waiting for the focal plane to cool the bias voltage is stepped down to 0V while data is streaming and the current through the TES is recorded for all channels.

In analysis, the normal, in-transition, and superconducting portions of the IV curve is identified. A linear fit is performed on the normal portion of the IV curve. The normal resistance is recorded from the slope of the fit and the offset is used to convert TES current to absolute units. With the absolute TES current and commanded current then the resistance, voltage, and power on the TES can be calculated at each voltage bias step. A range of fractional resistance ($R(V_{\text{bias}})/R_{\text{normal}}$) is used to identify a bias voltage that maximizes the number of channels biased in that range for each of the 12 bias lines. The current is then increased on all bias lines to drive all TESes normal and then dropped down to the identified bias points. At this point the detectors are ready to observe the CMB.

3.8.9 Rebiasing

As observing conditions (observing elevation and PWV) change the mean loading on the detectors change and the bias points must be updated to keep the maximum number of detectors operating at maximal sensitivity. A new set of IV curves can be taken at the new loading levels to establish the new bias points following the procedure in section 3.8.8 but the time it takes for the focal plane to thermalize after driving all detectors normal can make the whole process take $O(10-30)$ min. A faster procedure involves using a set of bias steps to choose the new bias point. Refer to Wang *et al.* 2023 for more details on rebiasing using bias steps[76].

Chapter 3.7, in part, is a reprint of the material as it appears in Cyndia Yu, Zeeshan Ahmed, Josef C. Frisch, Shawn W. Henderson, Max Silva-Feaver, Kam Arnold, David Brown, Jake Connors, Ari J. Cukierman, J. Mitch D'Ewart, Bradley J. Dober, John E. Dusatko, Gunther Haller, Ryan Herbst, Gene C. Hilton, Johannes Hubmayr, Kent D. Irwin, Chao-Lin Kuo, John A. B. Mates, Larry Ruckman, Joel Ullom, Leila Vale, Daniel D. Van Winkle, Jesus Vasquez, Edward Young, "SLAC microresonator RF (SMuRF) electronics: A tone-tracking readout system for superconducting microwave resonator arrays," Rev Sci Instrum 1 January 2023; 94 (1): 014712. <https://doi.org/10.1063/5.0125084>. The dissertation author was a first-tier author of this paper.

Chapter 4

Microwave SQUID Noise Modeling

As the first at-scale (1000x multiplexing factor) application of μ mux and the SMuRF readout electronics, the SO needed to develop a predictive model of the noise contributed by the readout system. The goal of this modeling effort was to be able to enumerate sources of noise in different aspects of the warm and cold readout system and refer them to equivalent noise at the TES bolometers through design and operating parameters of the readout. Such a model can then be used to: 1) ensure that the readout noise is kept below the fundamental sensor noise, 2) identify optimal operating parameters, and 3) propose modifications to the existing cold and warm circuit designs which would yield lower noise operation for future projects with a tighter noise budget.

The noise sources can be coarsely divided into white and low-frequency noise and further divided into single-tone and multi-tone noise contributions. Single-tone noise contributions are noise sources present when operating only a single readout channel and multi-tone noise contributions are noise sources that arise from the non-linear mixing between multiple interrogation tones. In this chapter we discuss the single-tone white noise contributions. In a multitone system the intermodulation products between the interrogation tones produce an effective "psuedo-noise" that scales with the number of tone and individual tone power cubed. For SO we have designed the system with sufficient linearity to avoid operating in this regime and will not discuss this form of multitone noise further.

Figure 4.1 is a schematic representation of the key elements of the warm and cold readout system used in constructing our model. Highlighted in yellow are the noise sources themselves which we refer between each other and TES equivalent noise using the transfer functions highlighted in blue. All sources are ultimately referred to equivalent sensor noise highlighted in green. The main design components considered in the model which are needed to estimate both the noise sources and transfer functions are highlighted in grey, they include the SMuRF electronics, the input RF-chain, the superconducting resonator, the RF SQUID, the TES coupling, and the output RF-chain.

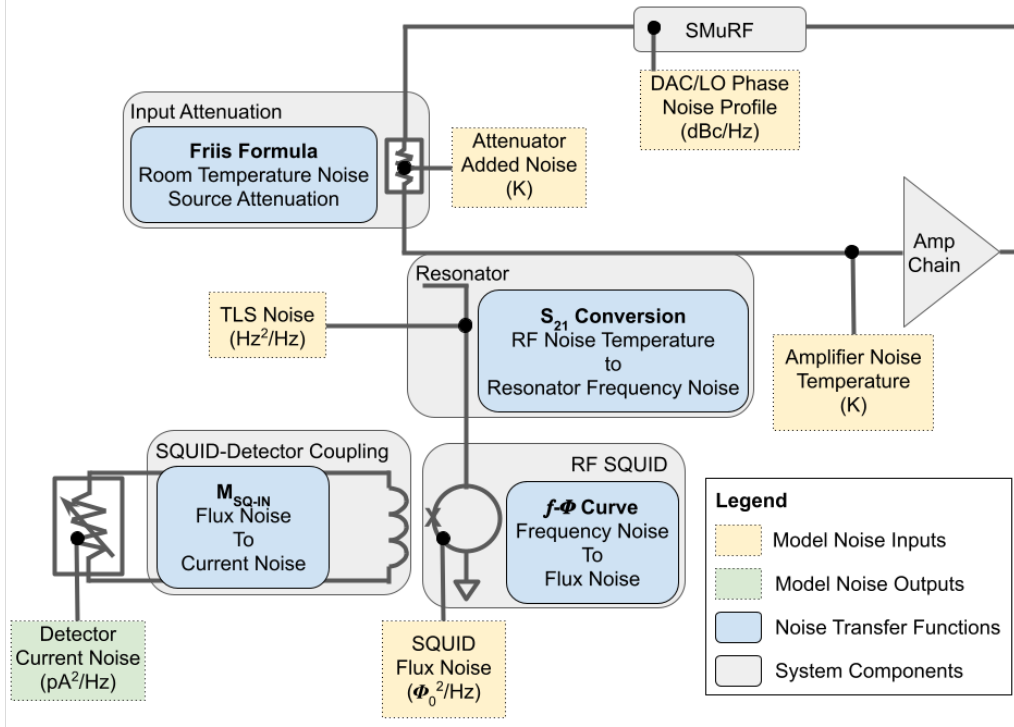


Figure 4.1: Schematic overview of the key of the construction of our noise model. The model includes noise sources (yellow - section 4.3), and transfer functions (blue - 4.2) that are used to refer noise between locations in the circuit. The system components that need to be specified in order to define the noise sources and transfer functions are shown in grey. All noise sources are referred to equivalent current noise in the detector (green).

In section 4.2, we derive the transfer functions required to refer noise through the circuit. These transfer functions include: 1) referring noise power through the RF-input and -output chains, 2) referring noise power at the multiplexer capacitive coupler to frequency noise in the resonator through the resonator transfer function (section 4.2.2), 3) referring frequency noise in the resonator to flux noise in the SQUID through the SQUID $f - \Phi$ curve (section 4.2.3) and 4) referring flux noise

in the SQUID to current noise in the sensor through the mutual inductance between the sensor and SQUID. In section 4.3 we describe how to calculate the three dominant noise sources which include: 1) Two-Level System (TLS) noise in the resonator dielectric, 2) RF tone-generation noise, and 3) RF amplifier chain noise. In section 4.4, we present measurements demonstrating the noise model transfer function accuracy (section 4.4.1) and show agreement between our modeled and measured noise and discuss sources of uncertainty in the model. Lastly, in section 4.5, we provide suggestions for improving the noise from the 3 dominant contributors and conclude.

4.1 Reference Device

All measurements presented in this chapter are performed on 62 resonators on an SO-style NIST μ mux 100kv3.2 chip with frequencies spanning 4250-4400 MHz. The chip is installed in a copper package on the mixing chamber of a dilution refrigerator kept at 100mK with a control loop. The DR has an RF-chain similar to figure 3.8 installed with a 0-30 dB digitally controlled variable attenuator.

Where helpful to demonstrate the general agreement over a population or quantify statistical fluctuations in the data or model I show results or statistics for all resonators however for most measurements it is sufficient and clearer to present our results on a single resonator. I have chosen one resonator to use for all of these results and will

refer to it hereafter as the reference resonator. The properties of the reference device are summarized in table 4.1 along with some typical operating parameters that are assumed for all data sets unless otherwise stated.

4.2 Noise Referral Model

4.2.1 RF Chain Model

Noise temperature is referred through a chain of attenuators or amplifiers using the Friis Formula[65].

$$T_{tot} = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \dots + \frac{T_n}{\prod_{i=1}^{n-1} G_i} \quad (4.1)$$

Where G_i is the linear gain of each component and T_i is the input noise temperature of each component. Attenuators are assumed to have $G_i < 1$ and the effective input noise temperature is calculated from their physical temperature (T_{Physical}) as

$$T_{\text{N,attenuators}} = T_{\text{Physical}} \times (1 - G) \quad (4.2)$$

This formula transfers noise through a chain of amplifiers and attenuators to its output and we use it to refer noise injected at room temperature or on the input side

Table 4.1: Properties of the reference resonator used in all result plots in this chapter. All of the fitted or derived properties for this resonator and it's SQUID are included along with operational parameters that are shared by all resonators. The distribution of these parameters among the 62 resonators are reported as the first and third quartiles except for the resonance frequencies which are the min and max of the population.

Device Parameter	Unit	Reference Resonator Value	Population Range
Resonator Parameters			
f_r	MHz	4288.51	4296.55, 4360.88
Q	$\times 10^3$	41.6	35.5, 42.0
$\text{Re}(Q_e)$	$\times 10^3$	56.4	45.3, 58.9
$\text{Im}(Q_e)$	$\times 10^3$	-3.2	-6.3, 4.0
δ	nS	10.5	7.2, 10.8
ϕ	Radians	1.8	0.6, 1.8
A_{mag}	ADC units	0.95	0.76, 0.9
A_{slope}	ADC units/MHz	-0.012	-0.010, 0.004
SQUID Parameters			
I_{Φ_0}	Fraction of Flux Ramp DAC Range	9.2e-2	9.19e-2, 9.0e-2
I_0	Fraction of Flux Ramp DAC Range	9.06e-2	0.34e-2, 5.02e-2
a_1	MHz	0.66e-1	0.65e-1, 0.72e-1
a_2	MHz	-1.39e-2	-2.14e-2, -1.38e-2
a_3	MHz	1.60e-3	1.78e-3, 5.36e-3
a_4	MHz	3.36e-4	-12.77e-4, 2.46e-4
a_5	MHz	-2.60e-4	-1.92e-4, 1.00e-4
df_{p-p}	kHz	66.8	66.2, 77.2
Rampheight M_{SQ-TES}	pH	227	-
$\langle df/dI \rangle_{\Phi_0}$	mHz/pA	14.7	14.5, 17.0
HHP	%	4.53	4.62, 9.35
Operating Parameters			
$P_{\mu W}$	dBm	-79.6	-80.7, -79.3
f_{FR}	kHz	20	-
T	mK	100	-

of the resonator down to the resonator. We invert this formula to refer noise from the output amplifier chain back to the input of the first amplifier.

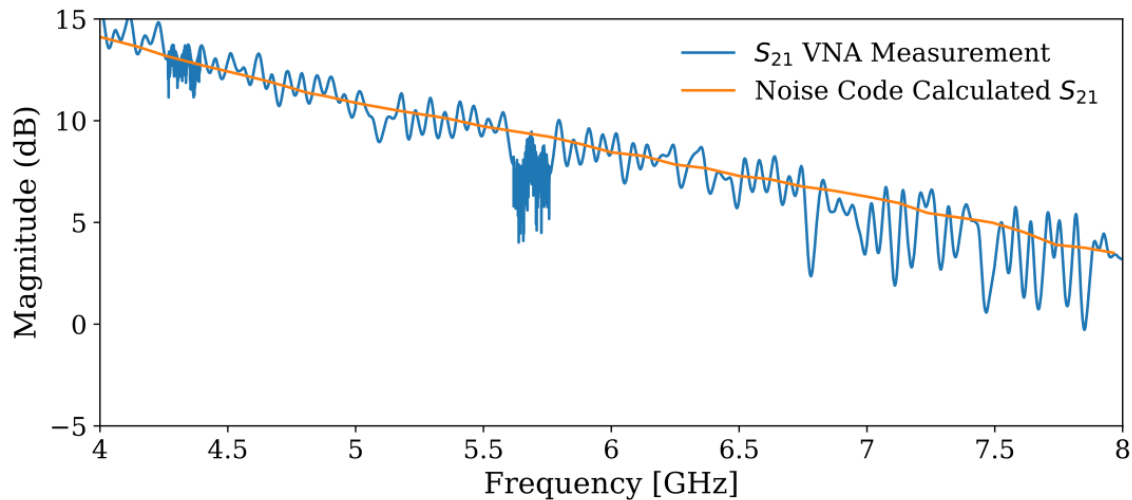


Figure 4.2: The accuracy of our component loss model is important for referring tone power, and noise through the system. The sum of our all of our modeled components in our input and output chains (orange) matches a calibrated VNA measurement of the round trip transmission (blue) to within 1 dB of the mean trend which confirms that we are correctly modeling our RF chain loss.

Accurately quantifying the insertion loss (gain) of all components in the RF-input (output) chains is an important input to our model as both the noise sources and transfer functions are dependent on the operating power. Ideally we would measured the S-parameters of each component over our operating frequencies using a vector network analyzer (VNA) with its reference plane calibrated to the end of its test leads. However, most of the components are operated at cryogenic temperatures and have temperature dependent loss (gain) which requires a cryogenic calibration

method suitable at all operating temperatures. The typical short-open-load-through (SOLT) VNA calibration scheme requires four measurements with the termination switched in between each measurement which is infeasible given the number of components and cryogenic cycling time of our test chambers. Instead we rely on interpolating the sparsely sampled (in temperature and frequency) manufacturer provided insertion loss (gain). To check the accuracy of this method for estimating insertion loss (gain) we measure transmission through the input RF-chain looped back to the output RF-chain through a short length of coaxial cable in place of the multiplexer module. We then compare this measured round trip transmission to the sum of the estimated component loss (gain) of all components. Figure 4.2 shows our estimated transmission is within < 1 dB agreement with the mean of the measured transmission. The deviations from our estimate come from: 1) standing waves due to reflections at component interfaces and 2) resonator transmission (around 4.3 and 5.6 GHz) both of which are not included in our loss estimates.

4.2.2 Resonator Model

The resonator transmission described in section 3.5.2 is used to compute the transfer function between a measured voltage in the Q-quadrature and a frequency shift of the resonance. In the SMuRF system this parameter is called η and is

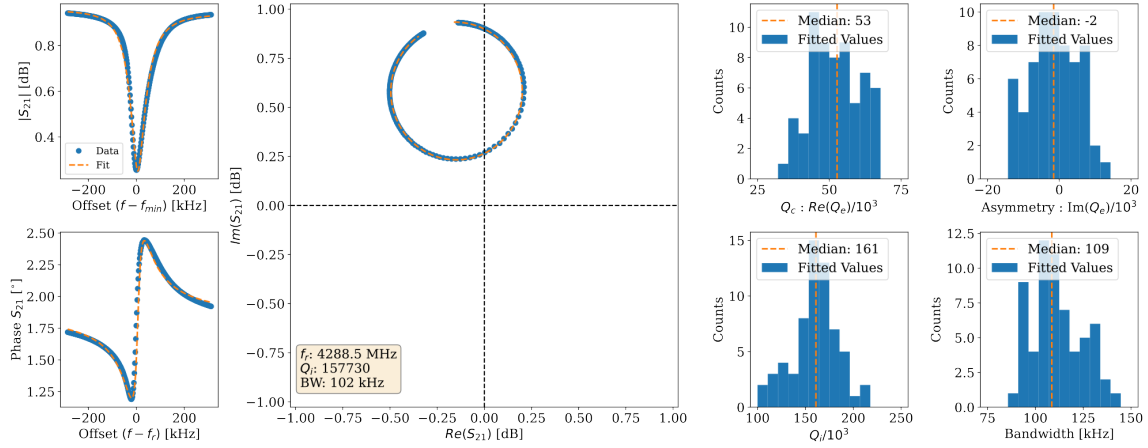


Figure 4.3: Resonator transfer function $T_{21, ADC}$ for the reference resonator as measured by the SMuRF (blue points) plotted as amplitude (upper left), and phase (lower left) vs frequency, and real vs imaginary (right). Our fit to equation 4.5 is overplotted on the data (orange dashed) with the fitted f_r , Q_i , and BW displayed. This resonator is used for all of the reported measurements that are from a single resonator for the remainder of this work. The distribution of a few of the fitted parameters for the 62 channels measured show our scatter in resonator parameters (right 4 panels). We aim to reduce discrepancies between our noise measurements and modeled estimates by using the fitted parameters for every resonator instead of a population average or design estimate. For our devices the coupling quality factors, Q_c (top left) is much smaller than the internal quality factor Q_i (bottom left) and therefor sets the resonator bandwidth (lower right). These measurements were taken with the power at the resonator, $P_{\mu W} = -80$ dBm where the resonator non-linearity is starting to be visible which induces some resonator asymmetry manifesting as a non-zero imaginary Q_e (top right). The population is still consistent with $Q_e = 0$ on average but at higher powers the asymmetry parameters takes much larger negative values for the entire population.

computed in the calibration steps described in section 3.8.4. Here we derive this transfer function from fits to the measured resonator transmission.

In equation 3.9 we present the transmission coefficient for an input and output impedance matched resonator which has a transfer function that is symmetric about

it's resonance frequency. In reality, the RF-chain provides attenuation and a phase shift and the input and output ports are not perfectly matched to the resonators characteristic impedance. At higher powers we also see evidence of a reactive non-linearity similar to that observed near bifurcation in kinetic inductance detectors[77] which distorts the measured resonance lineshape. This creates resonator asymmetry which we parameterize using the work of Kahlil *et al.*[78] by equation 4.3

$$S_{21,\text{Kahlil}} = \left(1 - \frac{Q\hat{Q}_e^{-1}}{1 + 2iQ\frac{\omega - \omega_0}{\omega_0}} \right) \quad (4.3)$$

Here Q_c is replaced with $\hat{Q}_e$ which is a complex number (denoted by $\hat{}$) with the real part equal to Q_c and the imaginary part parameterizing the resonator asymmetry. We find that for our μmux resonators we have to use equation 4.3 and not equation 3.9 to get low fit residuals.

To refer Q-quadrature voltage noise to frequency noise we also have to rotate the transmission into the real and imaginary axes of the SMuRF ADC and scale the transmission into the units of the ADC. This requires multiplying equation 4.3 by an empirically derived formula:

$$F_{\text{system}} = A_m (1 + A_s(f - f_{\text{min}})) e^{i(2\pi(f - f_{\text{min}})\delta + \phi)} \quad (4.4)$$

The exponential term accounts for the phase delay between the ADC and DAC which includes the cabling, and digital processing delays. Where $f_{\min}$ is the minimum frequency in the frequency window used for fitting (± 300 kHz around the resonance frequency), δ is the round trip time delay (in seconds), and ϕ is a phase offset. The term in parentheses accounts for the filters in the SMuRF AMCs, the cable loss slope, and the amplifier gain slope with the overall slope parameterized by A_s . A_m is an overall scaling which accounts for both the average loss through the RF chains and the conversion from power to ADC voltage units. So the product

$$T_{21,\text{ADC}} = F_{\text{system}} \times S_{21,\text{Kahlil}}^* \quad (4.5)$$

is the received signal at the ADC as a function of frequency for a specific output power setting of the DACs. The conversion from voltage noise as measured at the ADC to frequency noise in the resonator is the derivative of $T_{21,\text{ADC}}$ with respect to frequency evaluated at the operating frequency. We evaluate $\frac{dT_{21,\text{ADC}}}{df}$ numerically.

As an example, assume we have a noise source with power spectral density N_x (in units of W/Hz) and the probe tone power is P_c (in units of W) then the equivalent

*We used T here to denote that this is a unitful quantity while the scattering parameters, S are unitless.

frequency noise [†], N_f is given by:

$$N_f = \frac{\left(\frac{N_x}{P_c}\right) P_{\text{tone, ADC}}}{\left|\frac{dT_{21,\text{ADC}}}{df}\right|_{f=f_0}^2} \quad (4.6)$$

The ratio $\frac{N_x}{P_c}$ is the phase noise of the carrier at $f=f_c$ and $P_{\text{tone, ADC}}$ is the squared sum average amplitude in the two quadrature as measured by the ADC of a tone centered at $f = f_0$.

$$P_{\text{tone, ADC}} = \langle V_{\text{I, ADC}}^2 + V_{\text{Q, ADC}}^2 \rangle_t \quad (4.7)$$

Where $\langle \rangle_t$ denotes a time average and the subscript ADC indicates a quantity in ADC units. Multiplying the phase noise by $P_{\text{tone, ADC}}$ scales it to the units of T_{21} .

Estimating $T_{21,\text{ADC}}$ and $P_{\text{tone, ADC}}$

To apply the noise power to frequency noise transfer function given by equation 4.6 we need a way to estimate $T_{21,\text{ADC}}$ and $P_{\text{tone, ADC}}$. We outline two ways to derive this transfer function, one that requires the cryogenic multiplexer cooled down in a cryostat with a nominal RF-chain, and read out using the SMuRF electronics. The advantage of this method is that it requires only one measurement setup and will produce the most accurate values for $T_{21,\text{ADC}}$ and P_{tone} since it does not rely on any modeled or estimated parameters. Since this is sufficient to compare our modeled

[†]This is the power spectral density of the resonance frequency shift as a function of time $df(t)$.

noise to measured noise and allows us to disentangle multiple sources of noise I use this method for the remainder of this work. The disadvantage however is that it requires producing a working multiplexer and cooling it down in a cryostat. When designing a new readout it is useful to start from a set of target resonator parameters and estimate what the noise will be when read out with the SMuRF system without having to fabricate the devices. The design parameters can then be adjusted before fabrication to yield a device with sufficiently low noise performance with minimal costly fabrication cycles. With this in mind we provide a second method to estimate $T_{21,ADC}$ and $P_{\text{tone}, ADC}$ which relies on measurements with the SMuRF and a VNA but does not require any measurements with cryogenic resonators.

Measuring $T_{21,ADC}$ with SMuRF and cryogenic resonators:

In this method we connect the output of SMuRF to our cryostat in which an RF-chain and cryogenic amplifier are installed. To measure $T_{21,ADC}$ we output a tone from the DAC and sweep it over the range of each cryogenic resonance and measure the received complex amplitude at the ADC as described in section 3.8.3. The average complex amplitude measured at the ADC is $T_{21, ADC}$. We then fit equation 4.5 as shown in figure 4.3 from which we can extract the resonator parameters of our devices, a selection of which are shown in the right four panels. We then set a tone at each resonance frequency and stream a few thousand samples of it's complex I/Q

voltage data out of the analysis filter bank and to estimate P_{tone} . Since F_{system} and $S_{21,Kahlil}$ are linear operations and $S_{21,Kahlil}$ encompasses the parameters related to the cryogenic resonator design the transfer function (and modeled noise sources) can be scaled from this measurement by adjusting the modeled parameters in equation 4.3 relative to the fitted values to understand room for improvement/optimization in the cryogenic resonator design.

Measuring F_{system} with SMuRF, and VNA:

In this method we start by looping back the SMuRF output to it's input using a short length of coaxial cable and measuring the ADC signal as a function of swept DAC output frequency as we did in the previous method. Then we measure the true S_{21} (unitless) of the roundtrip cryostat transmission with a VNA calibrated to have it's reference planes at the location of the SMuRF input (output). The product of these two measurements gives us F_{system} . We can separately use our designed resonator parameters to calculate $S_{21,Kahlil}$ and estimate $T_{21,ADC}$ from the product of our calculation and measurement. $P_{tone, ADC}$ is the value of $T_{21,ADC}$ at the resonance frequency and can be scaled to different SMuRF operating powers by scaling A_m based on the change in output power settings (i.e. change in on-board digitally controlled attenuators, or added discrete attenuators at the input to the cryostat) relative to the SMuRF loopback measurement.

4.2.3 Flux Response Model

We can use the f - Φ response function of the microwave SQUID to convert from frequency noise to flux noise in the SQUID, and the mutual inductance between the SQUID and the TES to convert from flux to current noise in the detector. The mutual inductance is a geometric design parameter which is fixed at $227 \text{ pH} = 9.1 \text{ pA}/\Phi_0$ for our devices. Section 3.5.6 describes the integration of the f - Φ curve performed numerically to estimate $\langle df/d\Phi \rangle_{\Phi_0}$.

We can use multiplexer design parameters to estimate $\langle df/d\Phi \rangle_{\Phi_0}$ from the equations presented in section 3.5.6. However, in this work, we instead measure the f - Φ curve, fit the response function, and evaluate the integral of the fit. To measure the f - Φ response we track the resonance frequency shift as the commanded current through the flux ramp input line (I_{FR}) is stepped over $\sim 4 \Phi_0$ and fit the data to a harmonic expansion described by equation 4.8,

$$f(I_{\text{FR}}) = a_0 + \sum_{n=1}^N a_n \cos\left(\frac{2\pi n(I_{\text{FR}} - I_0)}{I_{\Phi_0}}\right) \quad (4.8)$$

where a_0 is the mean frequency offset of the data, a_n are the amplitudes of each of the harmonics, N is the maximum harmonic used in the fit, I_0 is the phase offset in commanded current units, and I_{Φ_0} is the commanded current that corresponds to

one Φ_0 . The data[‡] from the reference resonator (points) and harmonic expansion fits including up to the 5th harmonic (dashed lines) are shown in the left panel of figure 4.4.

We use all coefficients above the first harmonic to quantify the degree to which the SQUID curves deviate from a pure sine function (parameterized by λ in section 3.5.6) by constructing the metric, higher-harmonic-power (HHP):

$$\text{HHP} = \frac{\sum_{n=2}^N a_n^2}{a_1^2} \quad (4.9)$$

The HHP as a function of the tone power at the multiplexer chip is plotted in the right panel of figure 4.4 with the data points corresponding to the median of the 62 resonators and the error bars representing one standard deviation. We interpret the decrease in HHP and peak-to-peak frequency shift as due to circulating RF power in the resonator. One model of this behavior is presented in [79]. Essentially, the f - Φ curve we measure is really the average of the frequency shift over a flux range equal to the amplitude of the oscillating RF-coupled flux (Φ_{RF}) about the DC flux value. At low powers[§], Φ_{RF} can be ignored, and the equations presented in section 3.5.6 hold. However, as the total flux in the SQUID approaches Φ_0 , the f - Φ_0 response

[‡]Displayed points are decimated by a factor of 10 to make the fits easier to see on top of the data.

[§]Here small is in terms of Φ_{RF} so it's a function not just of the RF power but also feedline-resonator coupling strength, and the SQUID-resonator coupling strength.

approaches zero, so when modeling the SQUID response one must use a model that includes the averaging over Φ_{RF} .

The SMuRF system uses 3 harmonics in its frequency tracking loop and only uses the coefficients of the first harmonic to calculate the demodulated phase signal. Going into this modeling effort we suspected there was some loss in sensitivity due to not using all of the harmonics in the phase estimation. HHP was a useful metric for probing this. By comparing the tracked to untracked data presented in section 4.4.1 vs HHP we found much better agreement between the tracked and untracked data for power levels with significant HHP when we referred the untracked data to TES current using only the first harmonic to calculate $\langle df/dI \rangle_{\Phi_0}$. So even when using a modeled f - Φ curve as input one must still fit equation 4.8 and use the first harmonic to calculate $\langle df/dI \rangle_{\Phi_0}$ to avoid errors in frequency shift to TES current referral for high- λ SQUIDs or high power operation. In the future we aim to implement a version of the SMuRF frequency tracking algorithm that uses all of the harmonics to estimate the phase. With this update we will look for: 1) lower tracked noise at power levels with higher HHP and 2) better agreement between the tracked and untracked noise when referred using $\langle df/dI \rangle_{\Phi_0}$ calculated from the all of the harmonics.

While a more sinusoidal SQUID curve leads to a higher $\langle df/dI \rangle_{\Phi_0}$ all else held

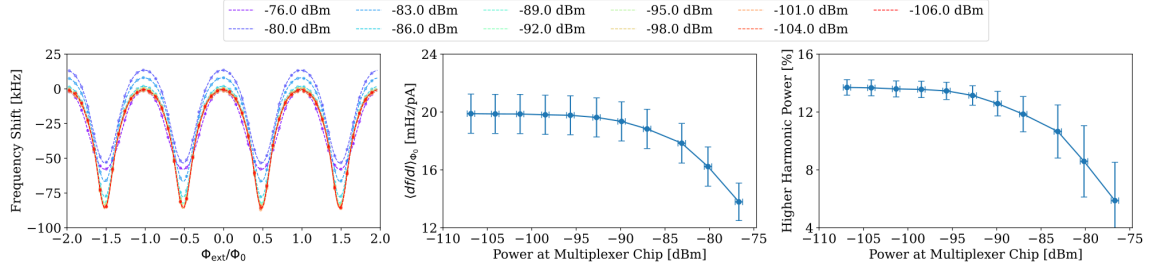


Figure 4.4: SQUID f - Φ response curves (left plot) measured by the SMuRF system over a 30 dB range of operating power. Measured data (points) with the fit to equation 4.8 using 5 harmonics overplotted (dashed lines). From the fits to all 62 resonators we estimate the degree of deviation from a sinusoidal response - HHP (right plot, and equation 4.9) and the average squid gain $\langle df/dI \rangle_{\Phi_0}$. The population median (points) and standard deviation (y-error bars) are plotted as a function of power clearly showing the power dependence expected from the extra flux loading from the RF microwave current coupled into the SQUID. The power dependence of $\langle df/dI \rangle_{\Phi_0}$ limits how high we can go in power to suppress our TLS noise before the tone-generation noise starts to contribute significantly.

equal, the accompanying peak-to-peak frequency shift means that the overall effect is to decrease the SQUID sensitivity as shown in the center panel of figure 4.4. This is an important effect because it drives us to want to operate at lower powers to maximize the SQUID sensitivity, but as discussed in section 4.3, two of our three noise sources decrease with increased power, and the SQUID sensitivity decreasing with increased power limits how much we can suppress these noise sources by adjusting our operating power.

4.2.4 Example converting dBc/Hz to pA/ $\sqrt{Hz}$

Combining the transfer functions described in the previous sections we can convert a general noise source in phase noise to current noise at the TES. We often present phase noise in decibel units which is defined as:

$$S_y^{\theta, dBc} \equiv 10 \log_{10} \left(\frac{S_y(f)}{\langle y^2 \rangle_t} \right) \quad (4.10)$$

where we use the superscript θ, dBc to indicate a phase noise in decibels relative to the carrier in units of dBc/Hz. This phase noise is of an arbitrary variable y evenly sampled in time (i.e. voltage on the feedline, voltage at the ADC, or frequency) and $S_y(f)$ is the power spectrum of variable y . Unlike the unitless phase noise, $S_y(f)$ without the superscript has units of y^2/Hz . $\langle y^2 \rangle_t$ is the RMS-average "power" in y . Given the phase noise the referral to NEI at the sensor is:

$$NEI = \frac{\sqrt{10^{(S_y^{\theta, dBc}/10)} P_{\text{tone, ADC}}}}{\left| \frac{dT_{21, \text{ADC}}}{df} \right| \left\langle \frac{df}{dI} \right\rangle} \quad (4.11)$$

4.3 Noise Model

Here we enumerate the noise sources in the system and estimate our typical sensor input referred noise level at our nominal operating parameters.

4.3.1 Amplifier Noise

On the output of the mux chip is a series of low noise amplifiers that take the microwave signals from the powers optimal for the resonator readout to a power level matched to the ADC dynamic range. Each amplifier adds some amount of noise to the system due to the DC bias circuit, and physical temperature of the transistor elements. The total noise contributed by a low noise amplifier is parameterized by a noise temperature defined as the physical temperature which a perfectly matched resistor would need to have at the input of the amplifier in order to generate the noise (assumed to be white) power measured at the output as described in equation 4.12:

$$T_N \equiv \frac{P_N}{k_b \Delta f_N} \quad (4.12)$$

The typical noise temperature of our first low noise amplifiers at its nominal operating temperature (4K) as reported in the data sheet from the manufacturer is shown in blue in the left plot in figure 4.5. As discussed in section 3.6.4 we have two cryogenic LNAs, one at 4K and one at 40K, as well as a room temperature LNA. In order to refer the noise temperature of each follow on amplifier to the input of the first stage amplifier the Friis formula described in section 4.2.1 is used. In our system the input noise temperature of the first stage amplifier is $\sim 2.1\text{K}$ but the total chain input referred noise is $\sim 3.4\text{K}$ so the follow on amplifiers contribute $\sim 38\%$ of the total

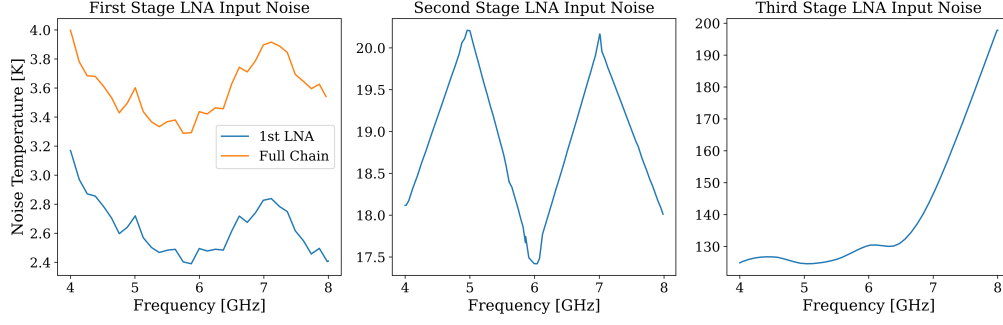


Figure 4.5: LNA noise temperature as a function of frequency for our 3 first amplifiers at 4K (left blue), 40K (center), and 300K (right) and the full chain noise referred back to the input of the first stage amplifier (left orange). The curves plotted here come from the manufacturer data sheets.

amplifier chain noise.

The added phase noise offset from the carrier at the flux ramp rate in dBc/Hz from the amplifier chain is:

$$S_{\theta, \text{LNA}} = 10 \log_{10} \left(\frac{k_B T_N}{P_{\mu W}} \right) \quad (4.13)$$

where $P_{\mu W}$ is the power into the mux chip and we use equation 4.11 to convert to TES noise equivalent current noise.

The noise temperature at the resonance frequency of our representative resonator is 3.7 K which corresponds to a phase noise on the feedline of -115 dBc/Hz and NEI at the detector of $\sim 8 \text{ pA}/\sqrt{\text{Hz}}$.

4.3.2 RF Tone Generation Noise

Tone generation noise refers mainly to jitter in the amplitude or phase of the tones generated in the SMuRF AMCs. The dominant source of noise comes from the onboard phase-locked loop (PLL) voltage-controlled oscillator (VCO) used to mix the baseband tones from the DACs up to the 4-8 GHz band. Instead of accounting for all of the on-AMC noise sources individually, we parameterize the noise by the phase noise as measured out of the SMuRF directly. This measurement is performed by connecting the output from the SMuRF to the input of a spectrum analyzer. We center the tone in the spectrum analyzer sweep window, set the resolution bandwidth to 1 Hz (so that each spectral bin is the average power in a 1 Hz bandwidth), average over 16 sweeps and save this data. We then load the data, find the peak and shift the spectrum left so that it's at 0 frequency. Then we measure the power in the bins from 100 Hz to 1 MHz which gives us the noise, $S_V(x)$ in units of W/Hz offset by x Hz from the tone center. We then measure the RMS averaged in a band ± 1 MHz[¶] from the tone center to get the power in the tone in W, P_{tot} . The phase noise at offset frequency x is then defined as $S_{\theta,SA}(x) = S_V(x)/P_{tot}$. The blue curve in figure 4.6 shows the $S_{\theta,SA}(x)$ spectrum analyzer measurement after logarithmically binning

[¶]this range is chosen to capture any power in the tone smeared into neighboring bins from the tone center, a smaller band size can also be used since bins far from the tone center contribute negligibly to the total band power

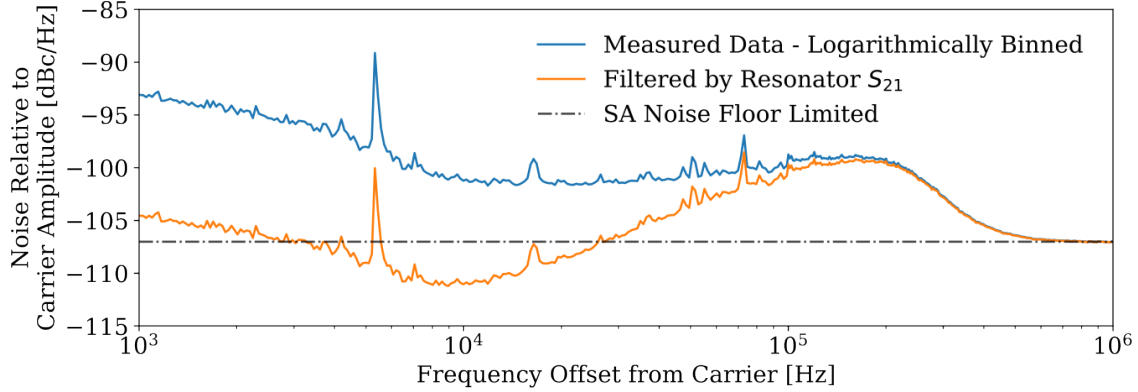


Figure 4.6: The phase noise profile out of smurf measured with a spectrum analyzer (blue). The data was logarithmically binned before plotting to make the effect of filtering the spectrum by the resonator transfer function $S_{21,Kahlil}$ clearer. The phase noise measured by the spectrum analyzer is dominated by the phase stability of the on-board VCOs used for up/down mixing. The gain of the PLL was adjusted to flatten the noise spectrum around 20 kHz where we typically flux ramp at the expense of increased noise around ~ 200 kHz. The rolloff at ~ 300 kHz is from the PLL as well and we expect the phase noise to continue decreasing above this but our measurement was limited by the noise floor of our spectrum analyzer. Since the frequencies where the SA noise floor dominates is outside the resonator bandwidth where we don't operate we didn't attempt to remove it and it appears in the total modeled noise.

the data.

The slight rise and fall of the tone profile come from the VCO PLL. We are operating the PLL toward its maximum bandwidth. Reducing the bandwidth (by decreasing the PLL gain) would move in the rolloff that starts around 300kHz and reduce the peak observed around 150 kHz. Increasing the bandwidth would have the opposite effect and risk instability. The flattening off of the noise profile at ~ 500 kHz comes from reaching the noise floor of the spectrum analyzer. The true phase noise

profile we expect to continue decreasing about 500 kHz until it hits the thermal noise floor of the system. We hit this noise floor well outside the resonator bandwidth (~ 100 kHz), however, so it's not important for our noise modeling as we won't operate our flux ramp at these frequencies.

To convert the S_θ as measured at the output of SMuRF to the phase noise profile as measured at the ADC, we must filter the phase noise profile by the resonance line shape

$$S_{\theta,ADC}(x) = S_{\theta,SA}(x) \frac{T_{21,ADC}(x)}{\max(T_{21,ADC}(x))} \quad (4.14)$$

where $T_{21,ADC}(x)$ is the resonator transfer function described in section 4.5 but indexed relative to the tone center frequency so that the left and right of resonance are folded over the center frequency and summed. Dividing by the maximum removes the unitful scaling A_m and just leaves the asymmetric resonator shape. Since dS_{21}/df is only non-zero in the imaginary direction the tone-generation noise is only present in the Q-noise direction after transmission through the resonator. The noise equivalent current referred to the TES is calculated using the transfer functions described in section 4.2:

$$\text{NEI}_{\text{Tone Generation}} = \frac{\sqrt{S_{\theta,ADC}(x) P_{\text{tone, ADC}}}}{\left| \frac{dT_{21}}{df} \right|_{f=f_0} \left\langle \frac{df}{dI} \right\rangle_{\Phi_0}} \quad (4.15)$$

The tone generation noise at $x = 20$ kHz^{||} for our reference resonator is -110

^{||}Our flux ramp modulation rate.

dBc/Hz in phase noise and $\sim 14 \text{ pA}/\sqrt{Hz}$ referred to NEI at the detector. Since this noise is dominated by the on-board VCOs phase noise at

4.3.3 Two-Level System Noise

The largest noise source in the system is two-level system (TLS) noise in the resonators. TLS's are excited states of the electron system in a dielectric. Tunneling between these states in the surface dielectric on top of the CPW resonators looks like an effective capacitance fluctuation which generates frequency noise in the resonator. A typical fabrication layer stack of a μmux CPW resonator is shown in Figure 3.1 of Mates 2011[64].

Since dS_{21}/df is only non-zero in the imaginary direction this frequency noise is present only in the Q-noise. The TLS frequency noise power spectral density $S_{\delta f_r}/f_r^2$ in units of 1/Hz is a function of operating power, operating temperature, dielectric and conductor material choice, and resonator geometry. When understading the TLS noise for our resonators we hold the temperature, geometry, and material choice fixed and consider only the power and frequency scaling which is modeled as:

$$S_{\delta f_r}/f_r^2 \propto \left(\frac{1}{f}\right)^\alpha \left(\frac{1}{P_{\mu W}}\right)^\beta \quad (4.16)$$

Where f here is the frequency offset from the resonance frequency f_r . From the

literature we expect $\frac{1}{2} > \alpha > 1$ and $\beta = \frac{1}{2}$ [66].

To estimate the TLS noise we measure in a regime where we know TLS is dominant and then scale it to more typical operating parameters. To measure the TLS frequency noise spectrum we output a fixed tone from SMuRF set to a resonance frequency and sample the voltage in both quadratures at the ADC. We take the PSD of this timestream and use $\frac{dT_{21,\text{ADC}}}{df}$ to rotate and scale the quadrature voltage spectral density into frequency-shift power spectral density. We choose 1 kHz away from the resonant frequency based on literature and examining our PSDs by eye as a point in our power spectra that are dominated by TLS for a wide range of probe tone power. We would like to compare to literature to confirm that we do not have anomalous TLS behavior. If we confirm our μmux resonators match literature for our resonator geometries and materials we can use literature values for other future designs using different materials and geometries. To compare to literature we must normalize this spectral density to a unitless phase noise by dividing by the resonance frequency

$$S_{\delta f_r}/f_r^2 = \frac{S_{V_Q}}{f_r^2 \left| \frac{dT_{21,\text{ADC}}}{df} \right|^2} \quad (4.17)$$

We choose a measure of power that is independent of the RF-chain or feedline coupling of a given resonator. Here we use the photon occupation number which is just

equal to the energy stored in the cavity[64] divided by a single photon energy quanta.

$$\frac{E}{hf_r} = \frac{2Q_c P_{\mu W}}{\pi h f_r^2} \quad (4.18)$$

Figure 4.7 shows our measurements of the amplitude of $S_{\delta f_r}$ at 1 kHz**. We measure all 62 resonators over a 30 dB range of power. Overplotted is a measurement of a 200 nm Nb thin film CPW quarter-wave resonator deposited on a silicon substrate as described in Gao *et al.*[80]. Most aspects of the resonator match our design well except for the resonator shape and the CPW center strip width, w , and ground plane gap, g . These effect the amount of the electric field concentrated in the dielectric where the TLSs are present. We have a larger w which decreases TLS when w/g is held constant. Our w/g however is larger which increases TLS. Additionally, fabricating SQUIDs on the same wafer as the resonator requires more fabrication steps that can degrade the dielectric quality. All these taken into account we expect our TLS noise to be the same order of magnitude but slightly larger than the resonator measured in Gao *et al.* which we confirm with the data in figure 4.7.

We also find our data fit well to $\beta = 1/2$ and $\alpha \simeq 1$ both consistent with the literature with α on the higher end of the range presented in the literature. By fitting

**This matches the measurement described in Gao *et al.* 2007[80].

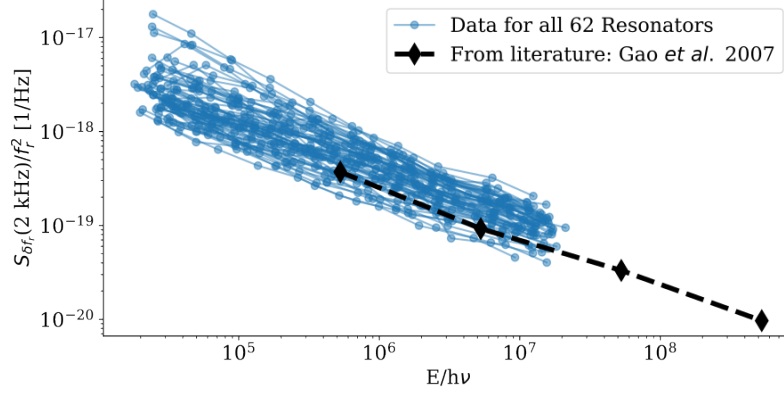


Figure 4.7: Frequency noise measurements from 62 resonators (blue dots). The frequency noise is measured from the Q-noise at 1 kHz offset from the carrier. Our data shows agreement in the power scaling and roughly with the amplitude of a measurement from a device presented in Gao *et al.*[80] made with the same dielectric and superconductor as our μ mux resonators (black diamonds). The slightly higher TLS levels for our resonators can be accounted for by the differences in geometry and poorer quality dielectric due to the larger number of fabrication steps required to fabricate a SQUID.

the data shown in figure 4.7 we then get a model for the TLS noise:

$$S_{\delta f_r}/f_r^2 = A_{1kHz} \left(\frac{1kHz}{f} \right) \left(\frac{1}{P_{\mu W}} \right)^{\frac{1}{2}} \left| \frac{1}{1 + \frac{i2Q_r f}{f_r}} \right|^2 \quad (4.19)$$

Where A_{1kHz} comes from the fits (or using values from literature) and the last term corresponds to the TLS fluctuations rolled off by the resonator bandwidth (which comes from our fits described in section 4.2.2). For our reference resonator the TLS noise contributes -104 dBc/Hz at the feedline which corresponds to $NEI = 28$ pA/ $\sqrt{Hz}$ at the detector.

4.4 Results

Here we validate the assumptions made about the noise referrals through the system and show that the model discussed in section 4.3 is predictive of the noise we measure over a range of operating parameters. We validate our model in two steps. First, given a measurement of the quadrature voltage spectrum we show that we can accurately recover the measured tracked TES current noise using the transfer functions given in section 4.2. Second, using the methods described in section 4.3 we show that we can accurately predict the quadrature voltage spectrum.

4.4.1 Tracked to Untracked Noise

Here we show that we can correctly calculate the flux ramp demodulated closed loop noise from the open loop quadrature noise spectrum. To do this we make four measurements for every resonator using the SMuRF system as described in sections 4.2 and 3.8:

1. Measure resonator T_{21} and fit to our model
2. Measure tracked closed-loop TES current noise spectrum
3. Measure the $f\text{-}\Phi$ SQUID response and fit to our model
4. Measure the open-loop quadrature noise spectrum

We repeat these four measurements at 10 different power levels. Measuring a whole chip of resonators gives us some statistics but also allows us to test the accuracy of our referrals with different parameters (i.e. quality factor, frequency separation, SQUID curve $f\text{-}\Phi$ response curve). Since both the SQUID and resonator transfer functions are power-dependent its also important for us to confirm though these measurements that we're capturing those dependencies correctly.

We summarize the results of these measurements in figure 4.8. The tracked noise is the average between $10 < f < 30$ Hz of the power spectrum of the demodulated flux ramp phase. Since the tracked data are modulated, we care about the referral from the Q-noise spectrum in a narrow ~ 1 kHz^{††} bandwidth around the flux ramp modulation. A vertical dashed line in the first plot shows the location of our flux ramp rate. The orange stars on the top two plots show the amplitude of the measured tracked noise for two operating powers of our reference resonator over top of the measured Q-noise spectrum. The left axes is the PSD in dBc/Hz and the right axes are referred to pA/rtHz using equation 4.11. We see that at both power extremes our referred Q-noise matches the measured tracked noise well. The lower left plot shows the distribution of tracked-noise over referred Q-noise ratio (tracked/untracked in figures) for the population of resonators at the highest power setting (where we typically operate). The vertical dashed orange line is the population median centered

^{††}The flux ramp sawtooth reset rate at which the phase data is sampled

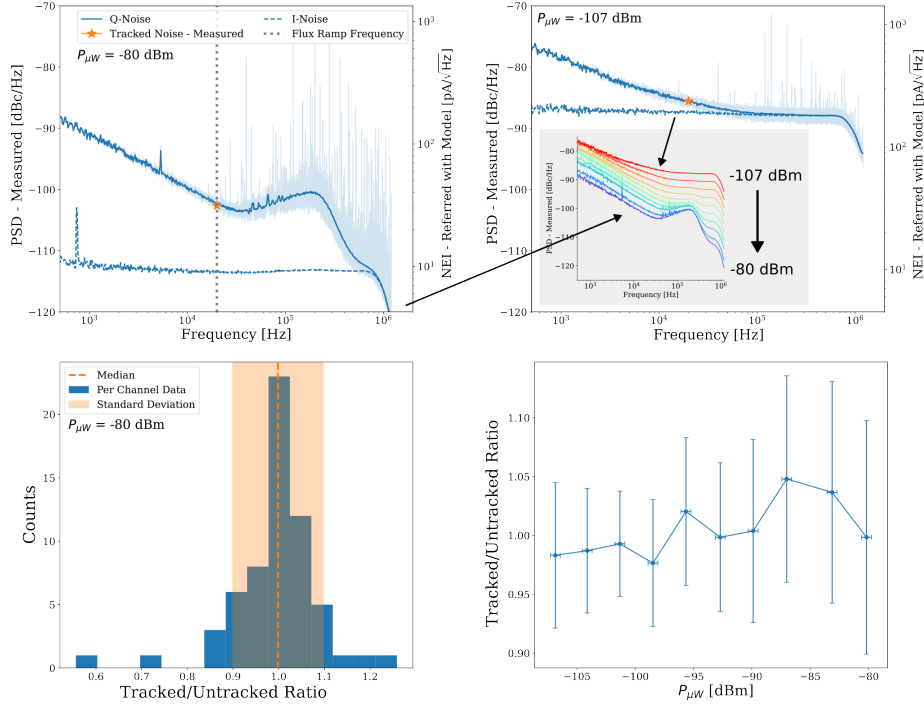


Figure 4.8: Comparison of noise measured with flux ramp demodulation to open-loop Q-noise referred to equivandalent sensor noise using the transfer functions described in section 4.2 for the reference resonator. (Top left) I and Q noise spectra for the highest measured power in phase noise on the left axis and equivalent sensor noise on the right axis. The tracked sensor noise is overplotted at the flux ramp frequency (orange star) and shows good agreement. (Top right) Same as the top left but at the lowest power used in the TLS-noise power sweep. The inset shows the Q-noise for all powers measured during the TLS-noise power sweep. The falling spectra of the TLS noise is present at all powers at low frequencies, the hump in the DAC noise becomes visible at high powers and the characteristic flat spectrum of the LNA is visible at the higher frequencies and powers. (Bottom left) histogram of the tracked/untracked noise ratio at $P_{\mu W} = -80$ dBm is clustered around 1 indicating our transfer functions are correctly referring to sensor noise. (Bottom right) the median (points) and standard deviation (y-error bars) of the tracked/untracked ratio as shown in the bottom left but displayed as a function of power. The data is clustered around 1 and doesn't show any strong trend with power indicating that our referral are properly accounting for the power dependence.

at 1 and the shaded orange region is the standard deviation. If our referrals capture the conversion from Q-noise to NEI perfectly then we expect the whole population sit in the bin at 1. The bottom right plot shows the median and standard deviation of the tracked/untracked ratio as a function of power. The population is centered near 1 for all powers with the standard deviation within $\pm 10\%$. There's no strong dependence on power which is an indication that we are capturing the power dependence in our referral.

4.4.2 Model Predicted Noise vs Tracked Noise

Having built confidence in our referral from Q-noise to NEI our next task is to show that we can predict the shape of the Q-noise spectrum using the estimates of the TLS, Tone Generation, and LNA noise discussed in section 4.3. We show that our modeled noise matches our measurements well over a 30 dB range in power. At higher powers the SQUID RF-flux coupling and resonator non-linearity produce a rapid drop in the resonator and SQUID transfer functions limiting our ability to push to higher powers to further improve our noise.

Figure 4.9 shows the modeled components estimated at our nominal operating power for our reference device. The LNA noise is flat as a function of frequency^{‡‡}

^{‡‡}While figure 4.5 shows a frequency dependence to the LNA noise temperature the magnitude of the variation vs frequency is small—O(10 ppm)—over the ~ 100 kHz bandwidth—typical SQUID modulation—around each resonance frequency so we take a single noise temperature for

up to the rolloff of the polyphase filter bank. We have multiplied all of the modeled noise components by the expected rolloff of a 4th order filter with a 1 MHz cutoff frequency representative of our PFB filter:

$$\text{LPF}_{\text{PFB}}(f) = \left| \frac{1}{1 + i \left(\frac{f}{f_c} \right)^4} \right| \quad (4.20)$$

In the top right plot in figure 4.8 and its inset we can see that the the I-noise is flat and equal in amplitude to the Q-noise at high frequencies. This is true at the powers and frequencies where the LNA noise is dominant. This is expected because as noted within the resonance the DAC and TLS noise are expected to contribute only to the Q-noise while the LNA contributes equally to both.

The peak in the tone generation noise around 200 kHz is clearly visible in the Q-noise especially at the higher powers where its hump becomes dominant at higher frequencies. Above ~ 400 kHz the DAC noise and total noise model diverges from the measured Q-noise due to the spectrum analyzer noise floor limiting our sensitivity to the DAC noise at those extreme offsets from the carrier. It is disadvantageous to flux ramp faster than the resonator bandwidth[81] so we do not worry about trying to better characterize the tone generation phase noise spectrum at those frequencies.

For higher bandwidth resonators where one may want to flux ramp at $O(\text{MHz})$ (i.e. each resonator.

calorimetry applications[82]) one can measure the phase noise profile out to higher offset frequencies using a lower noise spectrum analyzer or by adding additional amplification between the SMuRF output and the spectrum analyzer.

The dominant noise contributor is the TLS noise which has the expected falling $1/f$ spectrum filtered by the rolloff of the resonator bandwidth as described in section 4.3.3. The TLS noise accounts for 76% of the total noise for our reference device. When all three noise sources are summed we recover a model of the phase noise profile that matches our Q-noise measurement.

The LNA noise dominates at lower powers because it is a fixed noise power generated at the amplifier input so as the probe tone power decreases the phase noise increases linearly. The TLS noise on the other hand decreases as $P^{-1/2}$ and the tone generation noise is constant as a function of power because the noise and signal are attenuated equally. We clarify this picture in figure 4.10 where we plot the Q-noise at 20 kHz for each of the three modeled noise components, the sum of all of the modeled components, and our Q-noise measurement. In the left plot we can see the three described indices $\text{LNA} \propto 1/P$, $\text{TLS} \propto 1/P^{1/2}$, and tone generation $\propto 1/P^0$. Extrapolating the powers above our measurement we find that the TLS phase-noise dominates between $-103 < P_{\mu W} < -65$ dBm limited by the LNA below and tone-generation noise above.

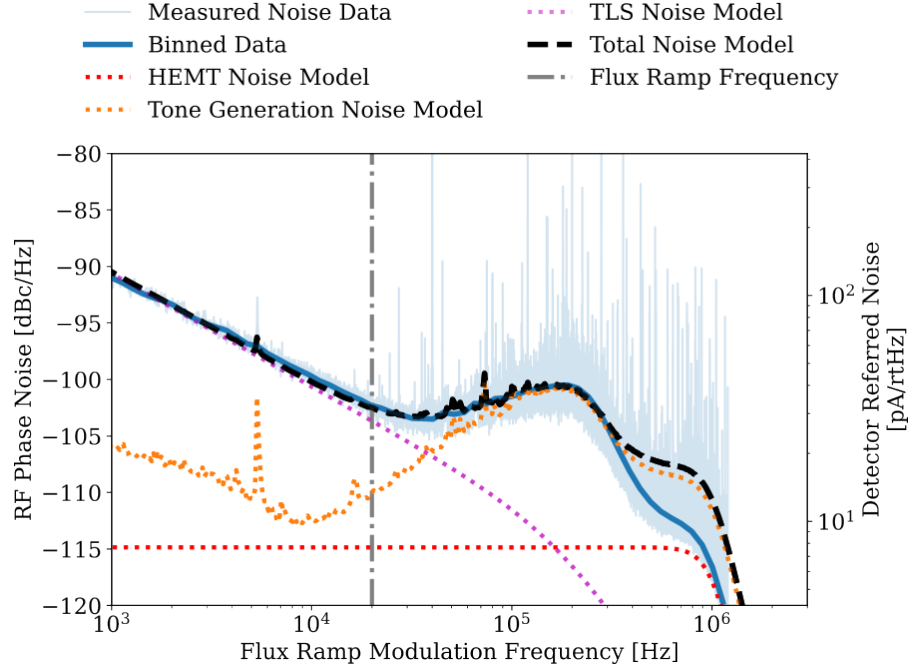


Figure 4.9: The measured Q-noise power spectrum for our reference resonator (light blue) and a binned version of the data (solid blue) in phase noise (left axis) and referred sensor noise (right axis). We have overlaid our estimates of the three noise sources we modeled: LNA (red dotted), tone generation (orange dotted), and TLS (magenta dotted) with the sum of the three sources (dashed black) matching our measured Q-spectrum well. The divergence of the model from the data at > 400 kHz comes from the tone generation noise where we derived the tone noise profile from a measurement on a spectrum analyzer that was limited by the analyzer noise floor at these frequencies. A vertical line (gray dot-dashed) indicates our flux ramp modulation rate where we evaluate the expected tracked noise and see that we can improve our noise by flux ramping closer to the Q-spectrum minimum of ~ 30 kHz.

The upper limit however can never be reached by our resonators because they become nonlinear and the SQUID response decreases at high powers. This is visible in the inset plot in the left panel of figure 4.10 which shows the deviation of the tone-generation phase noise from flat which comes from the variation in the resonance

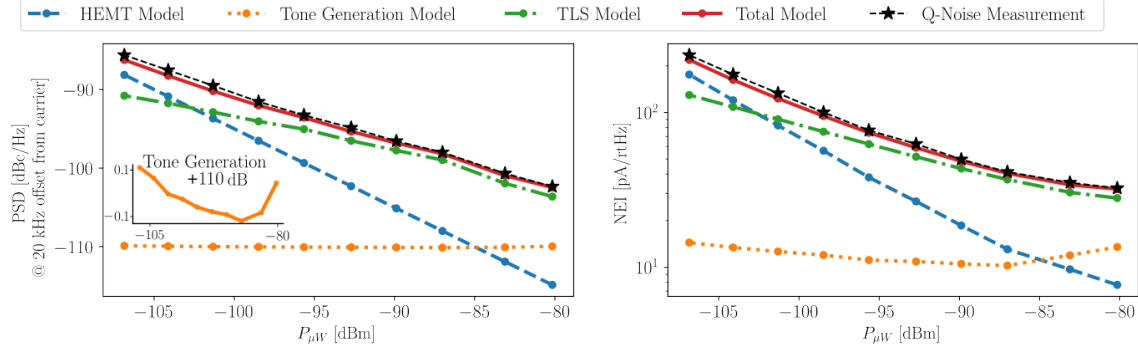


Figure 4.10: The power dependence of our 3 noise sources for our reference resonator. We plot the noise at the flux ramp rate (20 kHz) in phase-noise (left) and referred sensor noise (right). As seen in the inset of figure 4.8 the TLS noise dominates up to the two lowest powers where the LNA dominates. At the highest powers the TLS noise approaches the tone generation noise. The resonator non-linearity impact on the lineshape is seen in the zoom in of the tone generation phase noise plot. The resonance non-linearity and SQUID modulation degradation are especially apparent in the sensor referred tone generation noise. The sum of all modeled noise sources (red line with points) is compared to the measure Q-noise spectrum (black dashed with stars) and matches well over all powers demonstrating that we have the power scaling of both the noise referrals and noise models correct. While this plot seems to indicate that we can continue increasing power some to decrease the noise further the onset of resonator non-linearity shown in Figure 4.11 which decreases the resonator transfer function and a decrease the SQUID sensitivity shown in Figure 4.4 means that the sensor referred noise steeply increases at higher powers.

line shape. The rise in the NEI-referred noise plot most visible in the highest two powers also shows the impact of decreasing SQUID and resonator response. We further investigate the power dependence of the resonators using a VNA to take measurements at higher and lower powers than the SMuRF system. We took a fine sweep around each resonance and fit the data to extract an estimate of the frequency shift and resonance quality factor as a function of power. Figure 4.11 shows our data

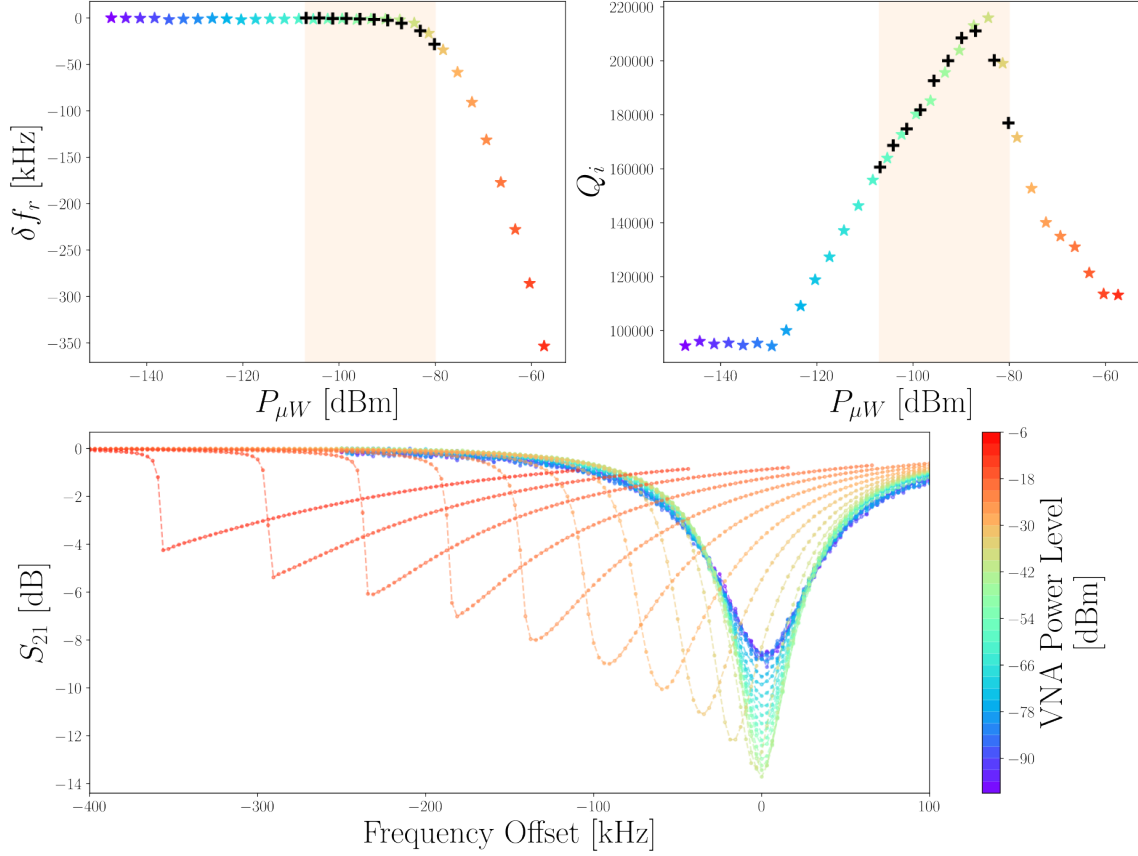


Figure 4.11: As the TLS noise depends strongly on power and it would be beneficial to extend to higher powers we explore the behavior of the resonators which prevent us from pushing to higher powers. We measure the resonator transmission (lower plot) with the SMuRF in a limited power range (black crosses) and a VNA over a wider power range (colored stars) and extract the resonance frequency and internal quality factor from fits to equation 4.5. The change in resonance frequency reflects a reactive nonlinearity dominating at powers above $\simeq -80$ dBm at the feedline similar to nonlinear kinetic inductance[77]. Quality factor changes dominate at powers below $\simeq -80$ dBm and indicate a dissipative non-linearity due to the saturation of TLS resonant transitions[83, 84]. At the lowest powers the nonlinearity is suppressed and Q_i is fixed to a minimum. The magnitude of the change in Q_i from -130 to -80 dBm indicates that the TLS loss is a large contribution to the total Q_i .

and fit parameters with the measurements using the SMuRF system overplotted. The distortions in the measured resonator transmission (bottom plot) are due to the power dependencies of their quality factor and resonance frequency. As the VNA or SMuRF probe tone is swept across the resonator, a different amount of power is coupled into the resonator which shifts the resonator Q_i and f_r such that at each point we are sampling a different underlying transmission profile. There are two nonlinear effects. The first is from the same TLS population responsible for most of our noise. The resonant transitions between TLS states contribute both a weak reactive (frequency shift) and strong dissipative (Q shift) non-linearity[83, 84]. At low temperatures the Q_i from TLS saturates to a low value and increases with increasing power, from the VNA data we see that Q_i saturates around -130 dBm to $\sim 80,000$ then increases by over a factor of two up until ~ -80 dB where another more dramatic non-linearity dominates. The second nonlinearity is a strongly reactive non-linearity shifting the resonance by many linewidths and displaying hysteretic behavior characteristic of kinetic inductance non-linearity[77]. These high power effects, reactive non-linearity in the resonator and flux modulation suppression in the SQUID, limit our ability to push to higher powers to further suppress the TLS noise to the level of the DAC noise.

4.5 Discussion and Conclusion

We have presented a model for the referral of noise through a μ mux resonator read out with the SMuRF electronics. We also presented methods for estimating the expected level of the three largest noise contributions in the case of a single tone readout. An additional noise due to nonlinear mixing can arise in a multi-tone system and is important at large numbers of tones and high readout power levels however we do not present a multi-tone model here as the SO- μ mux is designed to operate in a power limit well described by a single-tone noise model. We show that our noise referrals accurately predict the tracked noise from the open-loop noise and that the sum of our estimated noise sources match our measured open-loop noise spectrum. We have neglected some noise sources including the intrinsic SQUID noise and ADC noise but our agreement over a wide range in power between model and measurement where the DAC, LNA, and TLS all swap levels of importance at some portion of the frequency range demonstrates that any additional noise sources are sufficiently subdominant to exclude from our model.

From this model we have learned that the smurf tracking algorithm is only sensitive to the first harmonic effectively reducing $\langle df/dI \rangle_{\Phi_0}$ which we plan to address with an upgraded firmware that uses all harmonic coefficients to estimate the SQUID modulation phase. We also found that our nominal tracking frequency of 20 kHz is

not quite optimal and we could reduce the noise in our reference resonator from 31.9 to 29.0 pA/ $\sqrt{\text{Hz}}$ by increasing the flux ramp to the minimum in the Q-noise spectrum of 31 kHz. Our tone generation noise is also contributing non-negligibly at our optimal operating power and will be an even larger contributor at the higher tracking frequency where the TLS noise is smaller. The excess noise in the tone generation comes from using a VCO for the LO however we do not need to tune the LOs so we can swap the VCO for a lower phase noise fixed frequency low phase noise crystal oscillator which we expect could reduce the contribution from the LO completely reducing the noise at 20 kHz from 31.9 to or 29.1 pA/ $\sqrt{\text{Hz}}$ or from 29.0 to 23.3 pA/ $\sqrt{\text{Hz}}$ at the currently identified best tracking frequency of 31 kHz. With an LO that falls off fast enough to only consider the LNA and TLS above our 20 kHz tracking rate we could achieve ~ 13 pA/ $\sqrt{\text{Hz}}$ if we flux ramped near the resonator bandwidth. While our current noise levels are sufficient for the noise levels of SO detectors pushing to lower noise opens up the use of μmux for TES bolometers in lower photon loading environments when the readout noise requirements are more stringent.

We have also explored the high power behavior of our multiplexer and found that the power we operate at to minimize noise is at the boundary of the onset of significant resonator reactive non-linearity and SQUID curve modulation suppression.

This necessitates that we operate no higher than the reference power we presented here so we cannot push to higher powers to reduce the TLS noise. In multi tone readout we also must consider the third order intermodulation (TOI) products that fall within the readout band due to the nonlinear mixing in the amplifier chain and ADC. There are $\mathcal{O}(N_{\text{tones}}^3)$ intermodulation products, $\simeq 10^9$ for our 1000x multiplexer. The TOI products increase in power as P^3 rapidly overwhelming the tracked noise when they fall in the flux ramp demodulation bandwidth. The SO has shown that we can achieve readout noise levels similar to the single tone noise levels when operating all 1000x tones (see chapter 6) indicating that we are not limited by TOI induced noise near the reference powers presented here although the ability to decrease the noise while reducing or fixing the power levels is beneficial. The implementation of a better LO and faster flux ramp rates are both methods SO can employ to further mitigate TOI concerns.

Last the success in identifying the limiting noise sources and the ability to understand the tracked noise from the open loop Q-noise spectrum makes the methods presented here particularly useful for debugging excess noise in the integrated receivers. These excess noise sources could be due to broadband RF pickup or degraded resonator or SQUID performance for instance. Utilizing the tools presented here we can use the SQUID, resonator, and Q-noise spectrum properties to debug

and mitigate such integrated instrument issues.

Chapter 5

Small Aperture Telescope Data Processing

5.1 Introduction

Here we review the design and construction of the Simons Observatory (SO) Small Aperture Telescope (SAT) First Light Pipeline (FLP). The FLP is a software infrastructure implemented in a publicly available code called `sotodlib`[85], which we plan to use to make initial maps during instrument commissioning and daily maps for data quality checks during normal operations. A separate pipeline, the Data Release Pipeline (DRP), which includes transfer function estimation, a null

test framework, and observation matrix-based B-purification will be used for the full science analysis and is not discussed here. In particular, we discuss the time-ordered data (TOD) processing, which includes all data quality analyses and filtering applied to the raw data before projecting the TODs to maps.

5.1.1 Filter-and-Bin Map Making

The SO SATs will use a filter-and-bin map-making scheme. In CMB observations, we acquire data by scanning a telescope across predefined patches of the sky. In an ideal observation, the resulting timestream per pixel encodes the sky signal (CMB and foregrounds) and Gaussian noise. We then project the signal from each detector into a map based on where the detector is pointing at each time sample. Multiple detectors observe the same sky pixels over multiple observations, and the map contains the average signal in each sky pixel over all detector-samples. In the ideal case where the noise in each detector is uncorrelated between samples and the noise between detectors is uncorrelated, the variance in each map pixel decreases, and we get a higher significance detection of the sky signal. In reality, atmospheric emission[86], ground emission[87], RF interference[88], cosmic ray hits[89], and instrumental systematics[90, 91, 92] (i.e., gain drifts, timestamp jitter, HWPSS, HWP rotation instability, detector non-linearity) create correlated non-gaussian noise be-

tween samples and detectors. Timestream filtering aims to remove these correlations so that we can bin maps as described in the idealized case with Gaussian uncorrelated noise.

The filters remove some of the sky signals in addition to the noise. However, since the filters are linear operations, the ℓ -space transfer function can be estimated by applying the same filtering operations to many sky simulations and comparing the power spectrum of these simulations with and without the filtering applied. While the computational cost of running many simulations to estimate the transfer function is significant, the map-making operation is computationally much cheaper than maximum likelihood map-making (used for the SO LAT), which requires computation and inversion of a large noise covariance matrix that encodes all of the correlations removed by filtering in filter-and-bin map making[93].

We do not discuss the calibration of the TODs, but we review in detail the following steps, each of which I contributed significantly to the implementation of in the FLP, based on work by [94, 23]:

1. Glitch (and jump, trend, turnaround) flagging
2. Detector cuts
3. HWP synchronous signal (HWPSS) estimation + subtraction
4. Iterative glitch flagging + HWPSS estimation

5. Detrending
6. Apodization
7. Dephasing
8. Low-pass filtering
9. Decimation/downsampling
10. Scan synchronous signal (SSS) estimation + subtraction
11. Polynomial Az signal subtraction
12. Counter 1/f weighting

We also review the TOD preprocessing pipeline structure, which is a flexible way to string together calibrations, data quality checks, and filtering operations in a configurable order to be run automatically on data written to disk or to automatically apply pre-calculated calibrations and filter parameters to data when loaded from disk.

5.2 TOD Processing

Time-ordered data (TOD) here refers to data streamed out from the readout system in arrays shaped $N_{\text{detectors}} \times N_{\text{samples}}$ aligned in increasing time order from the time the observation began with a uniform spacing*.

*The sample rate is typically 200 Hz, but can be adjusted

5.2.1 Data Flagging and cutting

Flags are boolean arrays that identify a sample, or collection of samples, within a single detector or an entire detector with anomalous data that should be masked or cut from the final data product. We consider the following flags:

1. Glitches - one or a few samples with anomalously high or low values relative to the rest of the TOD of a given detector. Radiofrequency interference or cosmic rays are common sources of glitches.
2. Jumps - a place in the TOD of a detector where the mean value of the TOD shows a discontinuity. Phase jumps in the SQUIDs are one example.
3. Turnarounds - A collection of samples around where the telescope changes its pointing from left-going to right-going (or vice-versa) in azimuth.
4. Trends - A detector whose TOD continually ramps upward (or downward) in one direction over the whole TOD at a rate much faster than expected due to atmospheric drift or other low-frequency noise alone). These are associated with resonators that have unlocked frequency tracking loops.
5. Sources - A collection of points in a detector's TOD where it is scanning over a point source (planet).

6. High Noise - A detector whose TOD has anomalously high white or low-frequency noise relative to the main population of detectors.

This starting list of flags is not exhaustive, and we expect to implement additional flags similar to those implemented from previous experiments[94, 23], such as flags on the HWPSS and detector responsivity, as needed. We use detector-level flags to cut detectors out of the set of detectors used in the remainder of the processing. To avoid biasing the fits, we use sample-level flags to mask and exclude data points from fitting routines (the HWPSS, SSS, and Polynomial Az signal estimation). Additionally, when we apply any Fourier filter (i.e., low pass, high pass) to the TODs, they can produce ringing around discontinuities extending to samples outside the flags. We gap-fill sample-level flags before applying the filters to avoid this ringing. A polynomial interpolation or the first few modes of a singular value decomposition representing the correlated signals between all of the detectors (due to the atmospheric fluctuations) replaces the sample-level flags during gap-filling. Lastly, we exclude flagged samples from the map projection. Before HWPSS estimation, we only perform detector-level cuts.

Glitch Detection

The phenomenology of glitches (i.e., regularity, number of samples affected) varies depending on the source of the glitches and is a function of the environment and readout system. However, the detection scheme remains the same for many glitch phenomenologies. The data is first high-pass filtered and smoothed with a Gaussian filter, suppressing correlated low-frequency noise and reducing the variance of the TOD. The samples with glitches will remain significant outliers relative to the rest of the data points, so one can detect glitches by setting a signal-to-noise threshold for outliers and flagging samples within a buffered region around those outliers. The signal is the absolute value of the amplitude of the highpass-filtered TOD, and we estimate the variance from the interquartile range[†] of the highpass-filtered TOD. The `sotodlib` routine `get_glitch_flags` implements this glitch detection scheme and is illustrated in figure ??.

The top row of figure 5.4 shows glitch detection on a dataset with HWPSS. The left column shows the full TOD from one simulated observation (described in section 5.2.2) with multiple high-significance glitches. The center and right columns show a zoom-in around the first glitch. The orange data is the unprocessed TOD. The blue data is high-pass filtered with a cutoff of 0.5 Hz and Gaussian filtered with a width

[†]This is the difference between the 25th and 75th quartile of the data and is an estimate of sample variance, which is less sensitive to outliers.

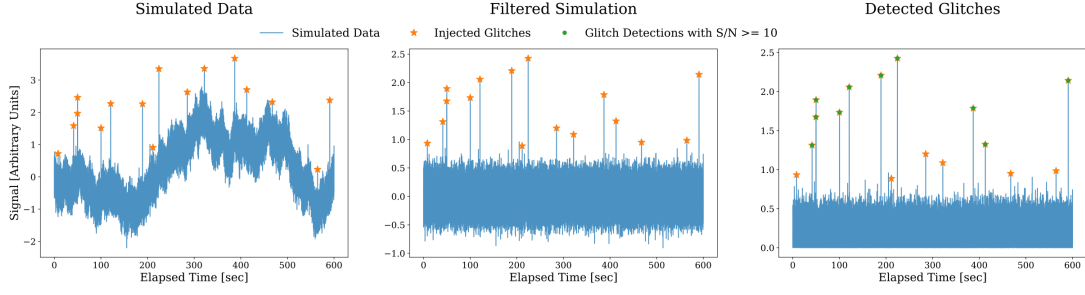


Figure 5.1: Demonstration of glitch detection scheme. (Left) simulated TOD with white noise, $1/f$ noise, and glitches. The location of the added glitches are highlighted with orange stars. (Center) highpass and gaussian filtered data. (Right) the absolute value of the center plot (blue) and the signal to noise > 10 detected glitches identified with green points.

of 0.002 seconds. The dashed black line shows the signal-to-noise detection threshold (set to 10 in these plots).

5.2.2 HWPSS Estimation and Subtraction

Background

The HWPSS is a spurious signal due to imperfections in the HWP and optics on the sky side of the HWP. These imperfections cause differential (between the two polarization axes) transmission, reflection, and emission, which generate signals at harmonics of the HWP rotation rate. The HWPSS, $A(\chi)$ is described by a harmonic expansion of the HWP angle (χ —which under normal operation should be increasing

or decreasing as a function of time depending on rotation direction.):

$$A(\chi) = \sum_{n=1}^{\infty} (A_n + a_n I) \cos(n\chi + \phi_n) \quad (5.1)$$

Theoretically, there can be harmonics at all multiples (n) of χ . However, given the construction of the SAT HWP, the 2nd and 4th harmonics are expected to be the largest[92]. Each harmonic has two contributions, one that is static over an observation (A_n) and one that is a function of the input intensity (I), which we denote with a lowercase here (a_n). The 4th harmonic of the half-waveplate contains the modulated polarization from the sky (atmosphere, foregrounds, and CMB), so a_4 we refer to as the intensity-to-polarization (I-P) leakage for each detector. In the presence of detector non-linearity, other harmonics can get mixed into the 4th harmonic, and gain drift in the detectors can also leak into the 4th harmonic[72]. For these reasons, we want to monitor the HWPSS over the experiment’s lifetime. Anomalously high $n = 2$ signals can indicate poor observing conditions, and ABS has used it for data quality cuts[95]. The $n = 4$ signal vs. the mean loading per observation estimates a_4 ; ABS and POLARBEAR used this a_4 estimate to remove intensity to polarization leakage[72, 96]. The noise in our measurement of χ limits the fidelity with which we can estimate $A(\chi)$, in [45] Hill *et al.* demonstrate that we have achieved the required precision and accuracy in measuring *chi*.

Simulated HWPSS TOD

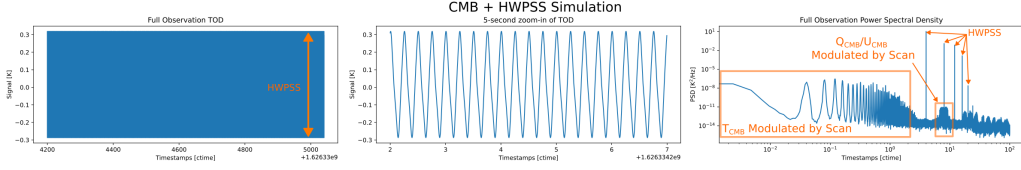


Figure 5.2: TOAST simulation of a timestream including HWPSS, CMB temperature, and CMB polarization modulated at four times the HWP rotation rate (8 Hz). The HWPSS dominates the TOD and is much larger than the CMB signal (left). A zoom-in on 5 seconds of the TOD (center) shows the HWPSS more clearly. The harmonics of the HPWSS, the CMB temperature modulated by the scan, and the CMB polarization modulated by the scan and the HWP are visible in the spectral density of the TOD (right).

To test the TOD processing, we use the Time Ordered Astrophysics Scalable Tools (TOAST) software framework to simulate SAT-like time-ordered data[97]. In TOAST, we configure an experimental setup that includes details of the telescope (number of optics tubes, number of wafers, number of detectors, the layout of the telescope, scanning axes, etc.), the site (which determines available astrophysical objects, and the atmospheric conditions), and then use it to "scan" an input map and project the sky signal which can include CMB, foregrounds, planets, and atmosphere into a timestream at a specified sample rate (200 Hz for our purposes). Optionally, we can inject several instrument-specific systematics at the timestream level, such as HWPSS, SSS, and gain mismatch. To test the SAT FLP, we generate TODs for 860 detectors (1 wafer) of a single frequency band for an SAT-like telescope, including

only the CMB. Then depending on what aspect of the pipeline we are testing, we also inject glitches, HWPSS, SSS, noise, and atmosphere. The flexibility to turn on and off various astrophysical signals and data quality issues in TOAST is invaluable for isolating and testing individual aspects of the pipeline.

For testing the HWPSS fitting, we use a simulation that includes only CMB and HWPSS or CMB, HWPSS, and glitches. Figure 5.2 shows an example TOD for one of the detectors. The left plot shows the entire observation (~ 10 minutes), which is dominated by the HWPSS, the center plot shows a zoom-in of a few periods of the HWPSS, and the right plot is a power spectrum that shows the CMB temperature and polarization modulated by the scan and the HWP with an amplitude smaller than the harmonics of the HWPSS.

HWPSS Estimation Implementation

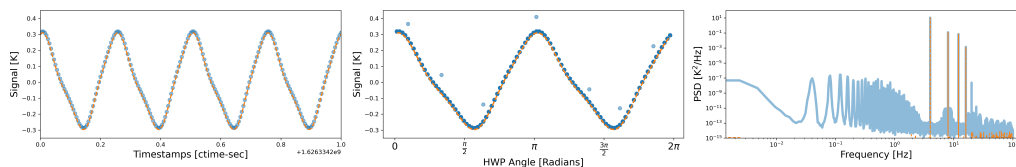


Figure 5.3: Fit to TOAST simulation described in section 5.2.2. Simulation is in blue points, and the fit is in orange points. Zoom-in on 1 second of the TOD (left) and the full observation plotted in χ -vs-signal space (center) shows some small glitches not detected with the glitch detector before HWPSS subtraction as discussed in section 5.2.3. PSD shows HWPSS estimating the amplitude of the 2nd, 4th, 6th, and 8th harmonics well (right).

We implement the HWPSS estimation in function `sotodlib.hwp.hwp.get_hwpss` which uses an Ordinary Least Squares (OLS) regression[‡]:

$$y = \beta_1 \sin(n_0\chi) + \beta_2 \cos(n_0\chi) + \beta_3 \sin(n_1\chi) + \beta_4 \cos(n_1\chi) + \dots$$

$$+ \beta_{2n+1} \sin(n_{N_{\text{harmonics}}}\chi) + \beta_{2n+2} \cos(n_{N_{\text{harmonics}}}\chi) + \epsilon \quad (5.2)$$

where y is the $N_{\text{dets}} \times N_{\text{samples}}$ matrix of HWPSSs, ϵ is the $N_{\text{dets}} \times N_{\text{samples}}$ matrix of differences between the signal and the fit, and n takes on $N_{\text{harmonics}}$ many values of modes (the default is $n = 1, 2, 3, 4, 6, 8$ but is an optional argument). Let X be the $N_{\text{samples}} \times 2N_{\text{harmonics}}$ matrix of trig functions from the above equation. Then

$$\hat{\beta} = (X^T X)^{-1} X^T y \quad (5.3)$$

Before performing equation 5.2, χ and y_d are masked to avoid the most significant glitches in the TOD from biasing the HWPSS estimation. The HWPSS estimation $\hat{y}_d = \hat{\beta}_d X$ can then be subtracted from the data y_d for each detector (with a separate function `subtract_hwpss`). Since we compute $A(\chi)$ for every observation on all detectors, we have implemented a binned version of the fitting to speed up the computation time. In this case, we solve the same OLS problem with χ and y

[‡]We also have implemented a non-linear least square curve fitting routine using `scipy.optimize.curve_fit` for the same data model given in equation 5.2 but since the HWPSS is a linear equation we found that the curve fit solution gave equivalent fit results with a larger computation time. Therefore our default plan is to fit using OLS.

binned, excluding flagged samples.

5.2.3 Iterative Glitch Flagging and HWPSS Estimation

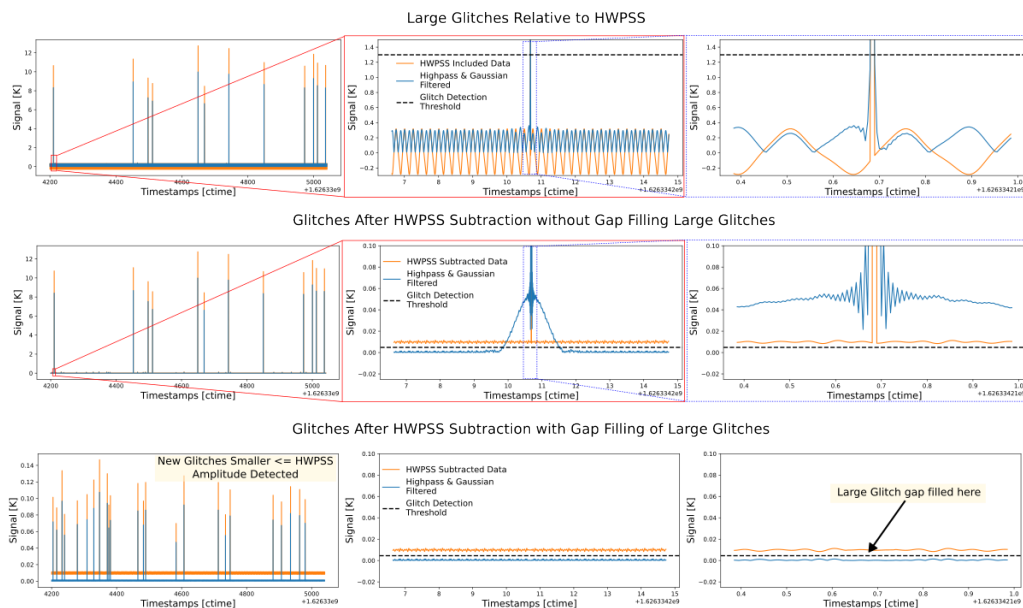


Figure 5.4: A demonstration of iterative glitch flagging in a simulation with HWPSS. Specifically, we highlight the importance of gap-filling in between iterations to minimize false detections from ringing of the Fourier space filters used for glitch detection. The raw TOD (orange), the absolute value of the high-pass and Gaussian-filtered signal used for glitch detection (blue), and the glitch detection threshold of signal-to-noise = 10 (black dashed line). Full observation TOD (left column). Zoom-ins on the samples around the first glitch (center and right columns). Top row: the largest glitches flagged during initial glitch detection. Middle row: glitch detection on the HWPSS subtracted signal without gap filling the pre-HWPSS subtraction detected glitches. Large ringing from the high-pass filter is prominent in blue. The ringing produces spurious glitch flags many samples away from glitches. Bottom row: after gap filling the initial set of glitches, smaller glitches are visible in the full tod (left), not visible before gap-filling and HWPSS subtraction. There is no ringing around the large glitches due to gap-filling shown clearly in the center and right columns.

Since the glitch finder relies on the signal-to-noise ratio of the glitch detection relative to the variance and the HWPSS is quite large, it dominates the variance. Some glitches with amplitude comparable to or less than the HWPSS amplitude can be missed in the initial glitch detection and still bias the HWPSS estimation. To avoid this bias, we rerun the glitch detection on the data after subtracting the initial estimation of the HWPSS and the gap-filling of the large initial glitches. Then we add the initial HWPSS estimation back in and re-estimate the HWPSS using equation 5.2 but with the new set of flags masked. In theory, we can repeat this process until we detect no additional glitches and the residual of the fits are minimized; however, in practice, this is achieved in two iterations. Figure 5.4 shows an example of the iterative glitch flagging and HWPSS estimation procedure.

5.2.4 Detrending

Detrending refers to fitting a first-order polynomial (mean and slope) from each detector TOD. Detrending removes the arbitrary offset in the detector TODs and the atmospheric drift over the observation. We calculate the mean and slope from the difference in the samples' average values at the TOD's beginning and end. The number of samples used to estimate the trend is an optional argument that defaults to 10 (0.05 sec). This estimation method can fail for TODs with significantly large

oscillating signals if the number of samples used to calculate the mean is insufficient to average over the oscillating component.

5.2.5 Apodization

Apodization refers to smoothly tapering the ends of the data to zero so that the whole TOD obeys periodic boundary conditions. Apodization is important because Fourier filters (used in glitch detection, demodulation, and downsampling) applied to data without periodic boundary conditions produce ringing at the edge of the data similar to that shown in the center row of figure 5.4. These filter oscillations at the edge of the data are challenging to flag and produce stripey artifacts when projected to maps. We transition to zero using one quarter-period of a cosine, and the number of samples apodized is an optional argument that defaults to 1600 samples (8 seconds).

5.2.6 Demodulation: Dephasing and Low-pass filtering

Demodulation consists of two steps: 1) dephasing and 2) low-pass filtering. In the dephasing step, the TOD is multiplied by the complex phasor $e^{i4\chi}$, which shifts the information at four times the HWP rotation rate to zero frequency. Any signal shifted to negative frequencies reflects across 0 to a positive frequency. Following

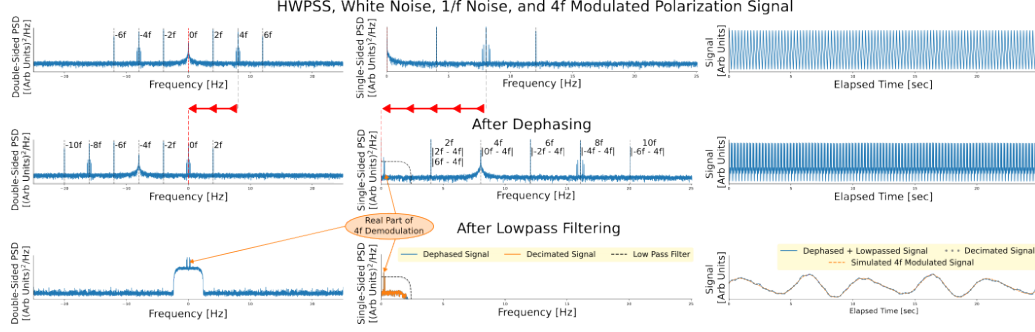


Figure 5.5: Demodulation steps, (left column) the double-sided (including negative frequencies) power spectral density, (center column) the single-sided power spectral densities where negative frequencies reflect across 0, and (right column) the corresponding TOD zoomed in to the first 25 seconds. Top Row: The input data includes the 2f, 4f, and 6f harmonics of the HWP rotation rate, 1/f noise visible around 0f, a white noise floor, and a small 4f modulated signal at 0.2 and 0.3 Hz, which we aim to recover with the demodulation procedure. Middle row: dephasing shifts the 4f signal to 0f, producing extra spurious signals at higher harmonics after reflecting the negative frequency components about 0f. Bottom row: the low-pass filter removes the higher harmonics and recovers the input-modulated signal at 0.2 and 0.3 Hz. The signal is then decimated to the Nyquist rate of 2x the filter cutoff frequency (orange in the center row and grey dots in the right row).

dephasing, a low-pass filter is applied to remove all of the neighboring harmonics of the HWPSS and any signals reflected across 0 frequency. Our resulting timestream is only sensitive to signal and noise in a narrow bandwidth $\pm$ the low-pass filter cutoff around the HWP 4f frequency. We set the default low-pass filter to $0.95 \times$ the rotation rate of the HWP, which maximizes the bandwidth over which we are sensitive while still filtering out all neighboring HWPSS harmonics (3f and 5f in particular). Figure 5.5 illustrates the demodulation steps.

Decimation/Downsampling

After the low-pass filtering, signals above the filter cutoff frequency are suppressed to 0 so we can reduce the sample rate to twice the filter cutoff without incurring any noise aliasing penalty. That means we can reduce the sampling rate from ~ 200 Hz to ~ 4 Hz. We perform this downsampling by interpolating the 200 Hz data onto a new grid of timestamps evenly spaced over the same time range but with a 50 times larger spacing between the time points. Decimation dramatically reduces the size of the data held in memory and the size of the matrix inverted when projecting the TODs into map space. The bottom row of figure 5.5 shows the downsampled TOD, which shows no signal loss since the data below the low-pass filter cutoff is still Nyquist sampled.

5.2.7 Polarized $1/f$ Noise Suppression

The following operations are needed to further suppress residual correlated noise in the demodulated polarization timestreams due to non-uniform HWP spin rate, detector non-linearity, gain drifts, polarized ground pickup, and polarized atmospheric emission. When none of these instrumental non-idealities are present, we do not require further filtering to the polarized timestreams because the demodulation operation removes the dominant correlated noise component from the atmospheric

temperature fluctuations. A detailed analysis of these systematics will be the focus of the commissioning and early science observation period and will inform the exact filter implementation. However, the filters described here have a heritage in similar Chilean B-mode experiments ABS[94] and POLARBEAR[23], so we are preparing the software infrastructure to implement them in our pipeline from the start of telescope operations.

Polynomial Az Signal Subtraction

Our telescope scan strategy consists of sitting at a fixed elevation for a single observation ($\sim 10 - 40$ min) and scanning left-to-right in azimuth to cover the entire width of one of our two designated patches in declination. The patch's sky rotation over the observation traces out a sub-width in right ascension (RA). We then continue successive constant elevation scans, each tracing a sub-width until we have fully observed the patch or the remainder becomes unobservable due to dropping below the horizon or overlapping with the sun or the moon. Polynomial azimuthal signal subtraction is fitting and subtracting a low-order (< 3) polynomial from each azimuth sub-scan. And a sub-scan is one scan from left-to-right or right-to-left of the azimuth range for the given observation. Polynomial Az signal subtraction removes any slow drift within a sub-scan due to atmospheric I-P leakage.

SSS Estimation and Subtraction

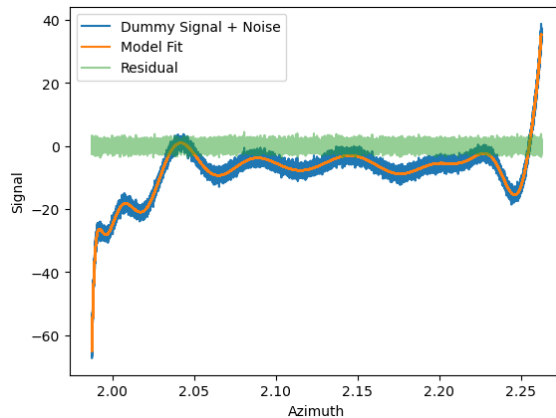


Figure 5.6: Demonstration of fitting a 20th order Legendre series to a simulated timestream with white noise and an az-synchronous signal. Input data (blue), fit (orange), and residual (green) in signal vs azimuth.

Scan-synchronous signal estimation and subtraction address the polarized emission from the ground. The SAT ground screen is designed to significantly decrease pickup from the surrounding ground signals from the surrounding mountains and terrain on the horizon; what remains (i.e., due to gaps in panels and non-uniform diffraction off the edge of the ground screen) will produce a polarized signal at particular azimuthal scan angles. The ground template should be coherent in azimuth for a given elevation over time up to any long-timescale variations in the ground structure (i.e., changes in the ground screen over time and diurnal or seasonal variation in the ground temperature). In order to construct an estimate of the ground signal, we bin the signal in azimuth bins and normalize the azimuthal scan range to ± 1 so

that the Legendre polynomials are orthogonal in this basis. We then fit a high order (~ 20) Legendre series to the binned signal averaged over each observing elevation in a season. The fitted polynomial series is then subtracted from the data to remove noise-correlations in azimuth between detectors due to the coherent ground signal.

Counter 1/f weighting

Counter 1/f weighting is a high-pass filter whose roll-off follows the inverse of the noise model:

$$N = w \left(1 + \frac{f_{\text{knee}}}{f^\alpha} \right) \quad (5.4)$$

where N is the noise power spectral density, w is the white noise component, f_{knee} is the "knee" frequency at which the $1/f$ component is equal to the white noise component, and α is the spectral index of the $1/f$ component. We perform a fit to each detector's PSD to equation 5.4 to extract w , f_{knee} , and α . From the fit to equation 5.4, we construct a Fourier space filter with a transfer function $= w/N$ and apply it to the TOD. The top plot in figure 5.7 shows an example timestream with a white noise component and $1/f^\alpha$ component and the same timestream after applying the counter 1/f filter. The bottom plot shows the PSD of the timestream before and after filtering with the fit to equation 5.4 and the counter 1/f transfer function overplotted.

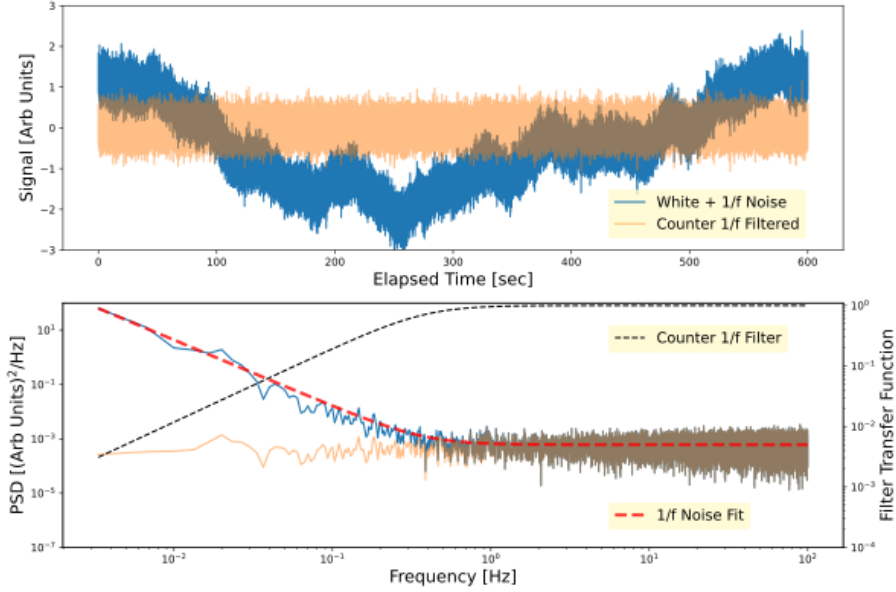


Figure 5.7: TODs (top) and PSDs (bottom) of data with white and 1/f noise (blue) and the same data after applying a counter 1/f filter (orange). PSD plot also shows a fit to the 1/f noise model described by equation 5.4 (red dashed) and the corresponding counter 1/f filter (black dashed).

5.3 Preprocess Pipeline

The preprocess module in `sotodlib` implements a pipeline that is automatically run on any data written to disk and can be run on data when loaded from disk into memory. The pipeline is defined using a configuration file that specifies a series of processes that are run sequentially and can pass a modified version of the TOD, calculated values from the TOD, or both to the following processing step. Each process within a preprocess configuration can include up to two types of operations: 1) `calc`, 2) `save`, or 3) `process`. A `calc` operation takes in the TOD and

estimates some derived information from it (i.e., noise levels, glitch flags, HWPSS harmonic amplitudes) and passes it on to the following process. A **save** operation saves derived information in a **SQL** database organized by observations tied to specified TODs. A **process** operation modifies the TOD using parameters specified in the configuration file or parameters derived from one of the **calc** operations. For example, **cut_detectors** could be a **process** that cuts detectors with the number of glitches detected in a previous **calc** step exceeding a threshold specified in the preprocess pipeline configuration file. The **calc** operations can be time intensive, so they are only performed on the data once unless otherwise specified. If the derived data product is in the **SQL** database, it will just load that data from disk instead of recalculating it. The preprocess module structure allows us to define our pipeline configuration and have all of the derived data products calculated as soon as the data write to disk, then load the data later with all of the flags, data cuts, and filters pre-calculated and applied. We can also change our preprocess pipeline configuration to include (or exclude) processing steps and automatically rerun all of the calculation steps on the data to generate a new **SQL** database.

5.4 Conclusion

We have presented a plan for time ordered data filtering, and data quality monitoring based on the experience of the ABS and POLARBEAR 1 experiments[?, ?] which will be a part of the SO-SAT’s first light pipeline. We reviewed each of the processing steps implemented in `sotodlib` and demonstrated their preparedness using simulated data from TOAST. Lastly, we presented the preprocess module which allows us to automate the tod processing, data quality, and data cuts steps.

We expect the processes presented here to be our starting point for data processing during commissioning. Through commissioning we will monitor the data quality and evaluate the need for modifications to this baseline plan. After commissioning ends and we move to initial science operations the tod processing presented here and finalized during commissioing will transition to our daily map-making pipeline to be used for monitoring the instrument health over the project lifetime. This sort of instrument health and data-quality monitoring will be a vital tool for maximizing observing efficiency.

Chapter 6

Small Aperture Telescope

Commissioning

The first SO SAT was constructed and tested at UCSD. The subsystems of the SAT were integrated in successively more complete phases and tests—often requiring a cryogenic cycle of the receiver—validated the integrated performance at each phase. The integration and testing effort began with the receipt of the empty SAT in June 2019 and ended with the full integration run in March 2023. A brief overview of the phases of integration and tests (I & T) during this period is summarized in section 6.1. We focus on the final full integration run and present the instrument configuration and a summary of all tests in sections 6.1.2 and 6.1.1. We present

detailed descriptions of the readout noise validation I led in section 6.2. Finally, we conclude with a discussion of the results and our site commissioning plans.

6.1 Integration and Testing Overview

The assembly of the SAT receiver required input from nearly every aspect of the SO project. The receiver, readout, detectors, optics, HWP, environmental enclosures, calibration hardware, data acquisition, and analysis software were all designed, built, and managed by different technical working groups within the SO and then came together at UCSD. Before integration, robust tests must be designed with clear success and failure criteria to address requirements derived from the SO science goals. Our analysis team members, which represent nearly half of the SO membership, produced a broad set of science goals[33] which were then flown down to instrument requirements and their associated tests through an interface with systems engineering. In many cases, subsystem tests were performed by their respective working group members, but many tests require an integrated receiver. These were performed by UCSD cosmology lab members and are the focus of this chapter. Due to the I & T effort at UCSD we had the increasingly rare opportunity in modern large-multinational physics experiments to be able to work on most aspects of the experiment. Because I believe this system-level view valuable, I will provide an overview of all of the I &

T efforts at UCSD but I will focus my detailed results on those areas of the fully integrated SAT testing which I led. Where appropriate in the overview I will direct the reader to other dissertations and articles which provide more detail on the I & T led by other UCSD cosmology lab members.

6.1.1 Early Integration Runs

All SAT I & T phases were tested with the SAT installed on a cart which we can configured to rotate the SAT along it's long or short axis. The SAT was always cooled pointing at Zenith and most tests were performed in this orientation unless specifically testing elevation dependent effects. The I & T phases successively integrated more components and included in order:

1. **Pulse Tube Validation** - The pulse tube cryocooler was cooled in a stand-alone test chamber with thermometers and heaters. We performed load curves at varying tilt angles to validate the PTC cooling capacity over (and beyond) our expected range of boresite and elevation angles. More details on these tests can be found in Tsan *et al.* 2021[98].
2. **Dilution Refrigerator Validation** - The dilution refrigerator was cooled in a stand-alone test chamber with thermometers and heaters installed. We performed load curves to validate the DR cooling capacity. A vendor provided set

of helium lines as well as longer lines designed for installation on the telescope platform were both tested to evaluate performance degradation from the longer lines.

3. **Vacuum Validation** - The SAT vacuum vessel was evacuated to confirm the vacuum integrity upon receipt from the manufacturer.
4. **Pulse Tube Integration** - The pulse tube cryocooler was installed into the SAT receiver along with heaters and thermometers. Load curves were performed to confirm cooling capacity and loading relative to PTC validation. Thermal gradients were compared to simulations to analyze heat strap and radiation shield thermal conductivity.
5. **Dilution Refrigerator Integration** - The dilution refrigerator was installed into the SAT receiver. Load curves were performed with the DR at 0° tilt angle to confirm cooling capacity and loading relative to DR validation. We also measured loading at a tilt angle of 27.5° corresponding to the maximum tilt over the planned observing elevation range.
6. **300-4K Readout Integration** - We installed the harness that holds the input/output RF-chain and DC wiring for the readout system between 300-4K. We performed load curves and thermal gradient analysis as well as RF-

transmission measurements. We additionally performed RF transmission measurements of the RF input/output chains looped back at 4K.

7. **4-0.1K Readout Integration** - We installed the 4-0.1K RF chain components and their associated support structure as well as the focal plane assembly with the location used for mounting the UFM's blanked off with copper masses emulating their thermal mass. We performed load curves and thermal gradient analysis as well as RF-transmission measurements.
8. **HWP Integration** - We installed the half-wave plate assembly with a dummy mass in place of the birefringent sapphire stack. We also installed two μ mux multiplexer chips in separate copper boxes on the focal plane one of which was coupled to a combination of optically coupled and dark bolometers. This validated the mechanical fit, cryogenic operation, and thermal performance of the HWP. The readout was used to test for excess pickup from the HWP and other sources of environmental pickup in the SAT.
9. **Window and IR Blocking Filter Integration** - We installed the infrared blocking filters and vacuum window on the sky side of the HWP assembly. We performed load curves to estimate the added thermal loading from the 300, 40, and 4K IR blocking filters.

10. **Optics Tube Integration** - All baffling, filters, and lenses in the 1K optics tube were installed opening the FPA to the 4 K volume through all of the optics. A blackened dummy mass was installed in place of the birefringent sapphire so the infrared environment was close to a full optical configuration but the detectors were not open to mm-wave signal from the room. We tested the added loading from the optics to the 1 K and 0.1 K stages[40].

A number of subsystems did not meet compliance during the first test and required a remediation and subsequent cooldown to validate. A few examples include:

1. **Out of Compliance:** During the 300-4 K integration we found that the cryogenic coaxial cables were manufactured with their connector tolerances out of specifications.

- **Remediation:** We replaced the out of spec coaxial cables and implemented a quality control process to measure the connector interface dimension tolerances upon receipt from the manufacturer. For more detail refer to Moore *et al.* 2022[99].

2. **Out of Compliance:** The support truss which thermally isolates the 40K shells from the 300K vacuum chamber failed.

- **Remediation:** We performed a reanalysis of the truss stresses and iden-

tified a stress from the vacuum chamber wall on the truss which was not included in the initial analysis. We updated the specification on the truss elements based on the new analysis, redesigned the truss elements, and undertook a through truss element pre-validation test which included failure testing individual truss elements and load testing the assembled trust after iterative cryogenic cycling. For more detail refer to Crowley *et al.* 2022[100].

A more complete review of the performance during the above I & T phases can be found in Galitzki *et al.*[101].

6.1.2 Full Integration Run Configuration

The final I & T phase included a full implemented SAT receiver including all optics with anti-reflection coatings, the HWP, and 7 UFM's on the focal plane. In addition two coherent receivers attached to horns with the same design as the horns integrated into the UFM's were installed on the exterior of the focal plane for holography measurements[102]. Figure 6.1 shows the SAT during the full integration run from the backside with the vacuum and radiation shields removed to view the focal plane with with seven UFM's, readout, and associated cooling equipment. The right of the figure shows the outside of the vacuum shells with the telescope pointing at

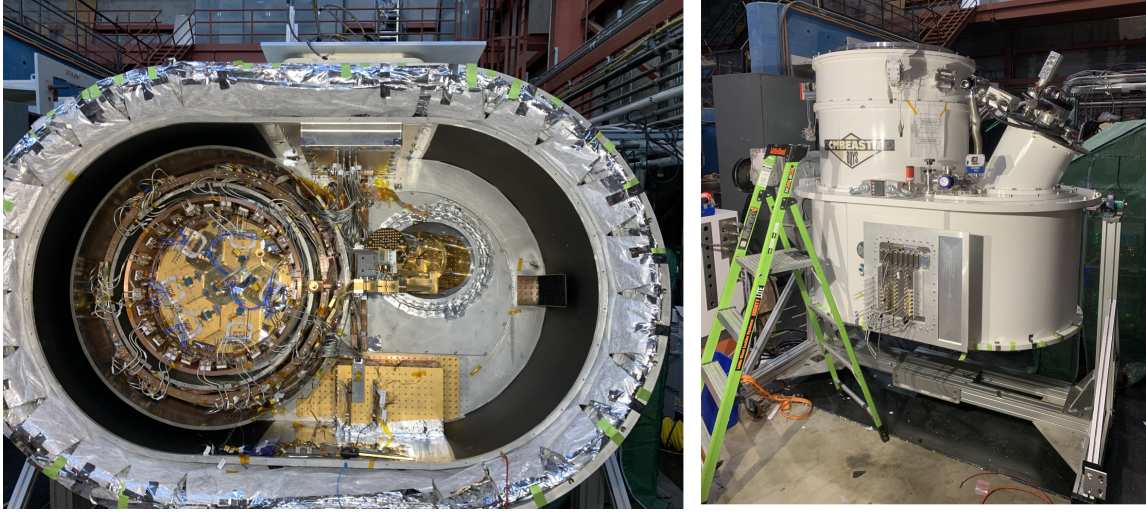


Figure 6.1: SAT during the full integration run. (left) the back of the SAT with the vacuum and radiation shell covers removed to view the seven UFMs, readout electronics, and associated cooling equipment. (right) the outside of the SAT vacuum shells with the telescope pointing at zenith on its cart.

zenith.

The testing included:

1. Thermal Performance - Load curves, thermal gradients, and base temperatures.
2. Optical Beam Response - Holographic and thermal beam maps. See Chesmore [102] 2022 and Seibert 2023[103].
3. Spectral Response - Bandpass measurements with a fourier transform spectrometer and out of band leakage measurements with thick grill (low-pass) filters in front of a thermal source. See Seibert 2023[103].
4. HWP Performance - Spin up/down tests, spin stability, encoder readout sta-

bility, demodulated 1/f noise measurements, and HWPSS analysis.

5. Polarization performance - Dense wiregrid rotated in front of chopped thermal source and beam filling sparse wiregrid rotated in front of vacuum window to check cross polarization response. See Seibert 2023[103].
6. End-to-end efficiency - Measurement of the efficiency of all elements in the receiver and the detectors.
7. Readout Performance - Measurement of the noise performance of the readout. See section 6.2.
8. Detector Performance - Yield, time constant, normal resistance, and noise performance of the detectors. See Lashner 2023[104].
9. Magnetic Performance - Solenoid placed under receiver in direction perpendicular to detectors and SQUIDs to test for magnetic sensitivity with full shielding in place relative to previous runs where shields were not in place. See Randall 2023[105]

6.2 Full Integration: Readout Noise Performance

*Our nominal scan rate of $1^\circ/s$ maps signals detected at 0.1 Hz to 10 degrees on the sky. However, how a 1-D angle maps onto the 2-D pattern generated by the $Y_{\ell,m}$'s is poorly defined but

Table 6.1: Readout requirements derived from the SO science goals. Crosstalk was not measured during SAT I & T and was verified during component-level testing[2]. The remaining requirements relate to the noise model described by equation 6.1. The white noise requirement ensures that the readout noise does not raise the detector noise by more than 10% and the low-frequency noise parameters are chosen to access a minimum multiple of $\ell_{\min} \simeq 20^*$.

Readout Requirement	Unit	Baseline	Goal
Crosstalk	%	0.3	-
White Noise - w	pA/ $\sqrt{\text{Hz}}$	65	45
Knee Frequency - f_{knee}	Hz	0.1	-
$1/f$ Index - α	-	1	-

The science goals set requirements on the CMB map noise which in turn sets requirements on the detector properties including the normal resistance, saturation power, transition temperature, time constants, and yield. For a review of the detector characterization in the final integration run refer to Jacob Lashner’s Ph.D. Dissertation[104]. Here we focus on the requirements on the readout summarized in table 6.1. Crosstalk was not measured in the full integration run but was studied in component-level verification testing and is summarized in Dober *et al.* 2020[70] and Groh *et al.* 2023[2]. The noise requirements in table 6.1 require that the readout white noise adds $< 10\%$ to the detector noise and the power spectral density falls

we make the assumption $\theta = \frac{360^\circ}{2\ell}$ which is that sampling one hot or cold spot between two nodes of angular size θ is sampling $1/2$ cycle of a signal at multipole ℓ and $\ell=1$ is the CMB dipole which has one period over the full sky (360°). So 10 degrees maps to $\ell = 18$ which is sufficiently low to probe the recombination peak at $\ell \simeq 100$.

below the curve between 0.01 and 100 Hz:

$$\text{NEI}_{\text{Readout}} = w \sqrt{1 + \left(\frac{f_{\text{knee}}}{f} \right)^\alpha} \quad (6.1)$$

where w is the white noise component, f_{knee} is the frequency at which the $1/f^\alpha$ component has equal amplitude to the white noise[†], and α is the exponent of the low-frequency noise component in the power spectral density (NEI^2).

6.2.1 Measurement Summary

The noise we measure in the full integration run consists of a combination of sensor noise and readout noise:

$$\text{NEI}_{\text{Measured}} = \sqrt{\text{NEI}_{\text{TES}}^2 + \text{NEI}_{\text{Readout}}^2} \quad (6.2)$$

where NEI_{TES} includes contributions from the Johnson noise of the shunt resistor and TES, photon noise, and thermal carrier noise[106]. The readout noise comes from the contributions described in chapter 4 but also the noise contributed by the TES bias circuit. We performed two measurements to isolate the noise contributed by the readout system from the detector noise and isolate the noise contributed by

[†]The knee frequency is dependent on the white noise level which implies a larger overall amplitude of the $1/f$ component when comparing the "baseline" to "goal" requirements.

the TES bias circuit relative to the other readout sources:

1. **Overbiased (OB) Dataset:** First we set the detector bias to zero and raise the bath temperature to operate the detectors in their normal resistance state. We stream flux-ramp demodulated data for 20 minutes and estimate the noise contributions by fitting the root of the PSD to equation 6.1. When normal the detector photon and thermal carrier noise are not present and the total detector noise is well described by the Johnson noise of a resistor with resistance R_n held at temperature T_b : $\text{NEI}_{\text{TES}}^{\text{Normal}} = \sqrt{4k_B T_b / R_n}$. We measure the detector's normal resistance from the normal branch of the I-V curves and the bath temperature is measured using the calibrated Ruthenium Oxide thermometer installed on the focal plane. At a typical $R_n = 8 \text{ m}\Omega$ and $T_{\text{bath}}=185 \text{ mK}$ the typical amplitude of the Johnson noise is $36 \text{ pA}/\sqrt{\text{Hz}}$. We subtract the detector's Johnson noise from the noise measured in this state to recover an estimate of the readout noise.
2. **Superconducting (SC) Dataset:** Next we drop the bath temperature below T_c so that the detectors are in their superconducting (SC) state. We stream flux-ramp demodulated data for 60 minutes and estimate the noise contributions by fitting the root of the PSD to equation 6.1. In the SC state, the photon and thermal carrier noise are zero and the noise is dominated by the

Johnson noise of the shunt resistor $NEI_{\text{TES}}^{\text{Normal}} = \sqrt{4k_B T_b / R_{sh}}$ at $T_{\text{bath}}=100$ mK and $R_{sh} = 400\mu\Omega$ the Johnson noise is 117 pA/ $\sqrt{\text{Hz}}$. We subtract the calculated Johnson noise from our measured noise to estimate the readout noise. Additionally, since $R_{\text{TES}} \ll R_{sh}$ when in the SC state, we can make the approximation that all current commanded down the detector bias line is coupled to the SQUID. We take advantage of this to maximize our sensitivity to the noise added by the bias circuit.

Expected Bias Circuit Noise

As described in section 3.7.3 and summarized in figure 6.2, the TES bias circuit consists of a positive and negative low-noise DAC, each buffered by an OpAmp and powered by a low noise voltage reference. The output from the OpAmps are converted to a bias current through low temperature coefficient ($\text{TC} = 10$ ppm/ $^{\circ}\text{C}$) resistors. The current is then split at each sensor between the sensor and SQUID on one branch and a shunt resistor on the other branch. The shunt resistors are $\simeq 400\mu\Omega$ however they're measured in a separate test cryostat and their exact resistance is a source of uncertainty when we are estimating their Johnson noise contribution.

We use the component datasheets for the voltage reference, DACs, and OpAmps to estimate their voltage and current noise contributions and add the Johnson noise

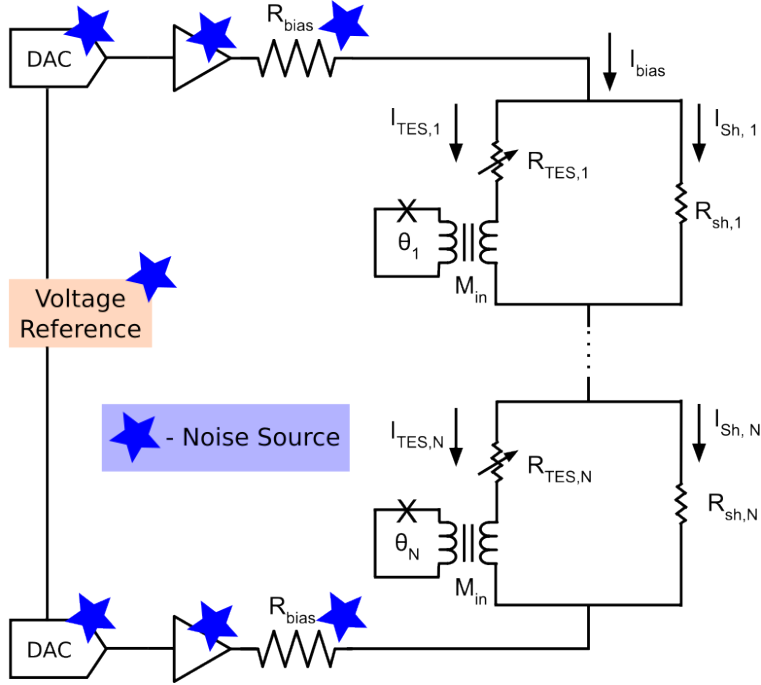


Figure 6.2: A schematic representation of the components in the TES bias circuit and the sources of noise. The DAC voltage reference is expected to be the largest contributor to the white noise and the current setting resistor temperature coefficient is expected to be the largest contributor to $1/f$ noise. When the detectors are biased in the superconducting transition at their resistance is $\simeq 0.5R_n \simeq 10R_{sh}$ so roughly 90% of the bias circuit noise we measure when the detectors are superconducting is shunted away from the detectors during operation.

of all resistors in their circuits as well as the current setting resistors on the cryocard to estimate the current noise into the cryostat. The ASDs of the noise sources identified in Figure 6.2 are plotted in Figure 6.3. All Johnson noise contributions are flat and the curves with $1/f$ included in their spectrum come from the component datasheets. The total current noise is $33 \text{ pA}/\sqrt{\text{Hz}}$ rising to $39 \text{ pA}/\sqrt{\text{Hz}}$ at 0.1 Hz due to the low-frequency noise in the DAC. The intrinsic component low-frequency noise

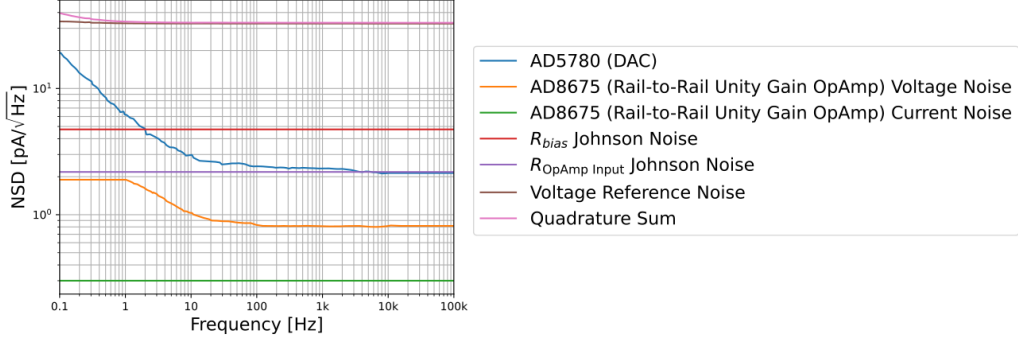


Figure 6.3: Current noise at the input to the cryostat from the TES current bias circuit (pink). All of the input contributions are enumerated in other colors (blue, orange, green, red, purple, and brown). We estimate a flat Johnson noise for all resistors and curve with $1/f$ come from manufacturer datasheets

is quite small (by design) and we expect the low-frequency noise contribution to be dominated by the current setting which induce a bias current fluctuation proportional to the environmental temperature change: $\delta I = \langle I \rangle \text{TC} \times \delta T_{\text{environment}}$.

We expect to measure the full $33 \text{ pA}/\sqrt{\text{Hz}}$ white noise when the detectors are superconducting but during operation the current noise is shunted away from the SQUID by $\frac{R_{\text{sh}}}{R_{\text{TES}} + R_{\text{sh}}} \simeq \frac{1}{10}$ so the bias current white noise contribution is expected to be $\simeq 3 \text{ pA}/\sqrt{\text{Hz}}$ much less than all of the sources discussed in chapter 4.

6.2.2 Results

In figure 6.4a we present a representative OB dataset TOD and ASD with the fit components overplotted. Similarly, figure 6.4b shows an SC TOD and ASD where we've also overplotted the $1/f$ component reduced by the shunt ratio and scaled by

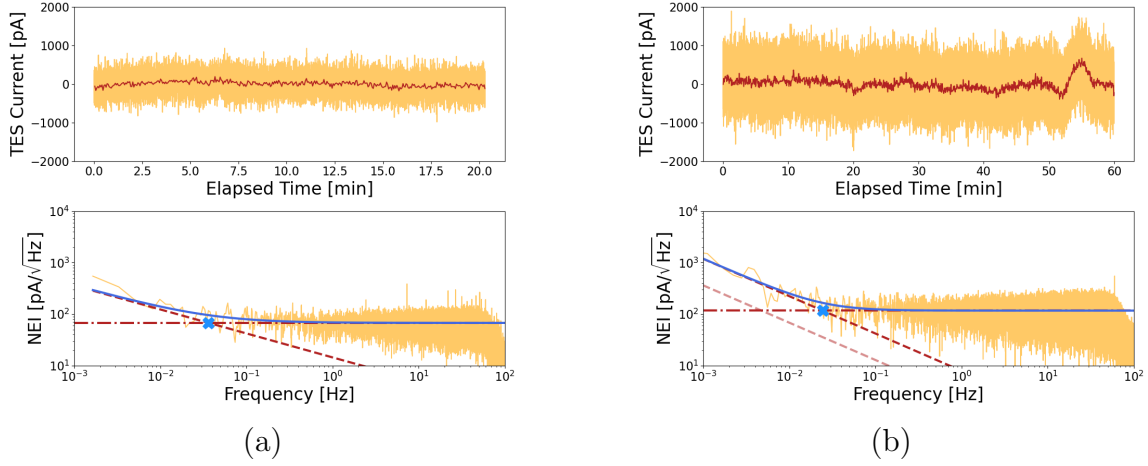


Figure 6.4: Typical TODs (top) and PSDs (bottom) measured from the OB dataset (a) and the SC dataset (b). Orange is the full sample rate data with y-scales equal to compare variance. The SC dataset has a much higher noise because of the larger Johnson noise from the. Red in the TOD is a downsampled version of the data to compare low-frequency drift in the TOD. Larger 1/f noise is present in the SC dataset compared to the OB as seen in the larger drift in the red line in the TOD, the higher 1/f index and higher 1/f knee in the PSDs. The PSDs show the total model fit to equation 6.1 (blue) and the white (dot-dashed red) and 1/f (dashed red) noise components separated from the fits. We have also scaled the 1/f noise measured in the SC PSD by the expect shunt ratio between the sensor and it's shunt resistor and scaled by the ratio of the bias current during the measurement relative to the expected bias current during a typical observation ($33/10\mu\text{A} = 3.3$).

the ratio of the bias current during the measurement relative to the expected bias current during a typical observation ($33/10\mu\text{A} = 3.3$). We performed fits on all of the 6 operating UFM's in the full integration run and distributions of the fit results for each wafer are shown in figures 6.5, 6.7, and 6.6.

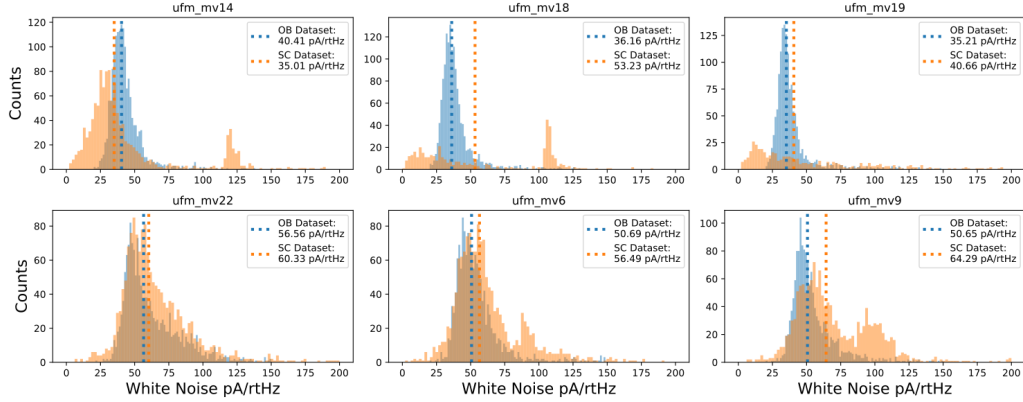


Figure 6.5: White noise fit distribution for OB (blue) and SC (orange) datasets after subtracted estimated Johnson noise from the TES normal resistance and shunt resistance respectively. The superconducting dataset has a larger error due to uncertainty in the shunt resistance we are using to calculate the Johnson noise to subtract.

White Noise Discussion

Looking at the OB dataset the median of all of the histograms are below our baseline noise requirement. The variation within a UFM is due to not being able to optimize the power (as discussed in chapter 4), and flux-ramp frequency for all channels simultaneously due to fabrication variations between channels and unequal loss across the readout band. Since the median is consistent with the dark resonator

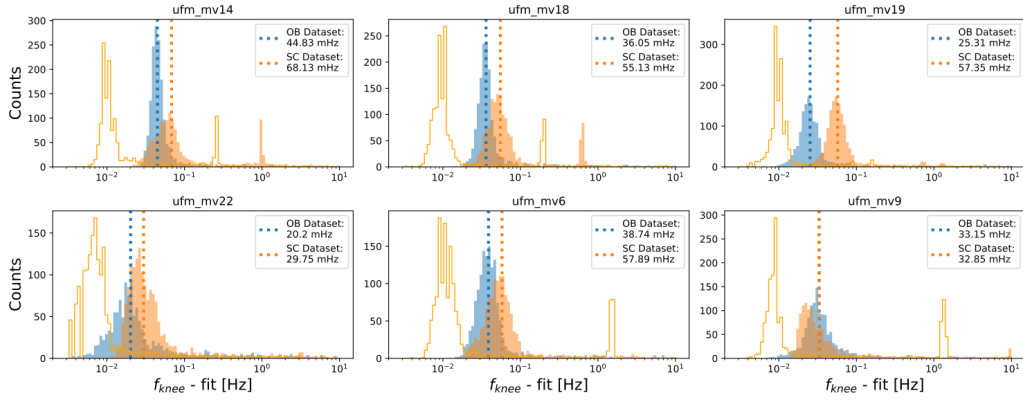


Figure 6.6: $1/f$ -knee frequency fit distribution for OB (blue) and SC (orange) datasets after scaling the knee frequency to the equivalent knee frequency for a white noise level = $65 \text{ pA}/\sqrt{\text{Hz}}$. In the unfilled orange histogram we show the knee if the $1/f$ component component were scaled down by the shunt ratio and the bias current as in figure 6.4b. This is valid if we assume that the $1/f$ noise when superconducting is a noise source fractional to the bias current.

(no TES or bias current coupled) channels we can conclude that we are reasonably estimating the Johnson noise and not measuring additional noise from the sensors above the expected Johnson noise. The fact that we are close to our expectation for the single tone noise as given in chapter 3 also shows that we are not suffering a multi-tone non-linearity noise penalty when running with 1000 tones simultaneously.

1/f Noise Discussion

The $1/f$ noise index and knee frequency meet our specifications in the OB dataset but have a higher index in the SC dataset. We attribute this excess $1/f$ noise to the temperature coupling of the resistors on the output of the DAC voltage references.

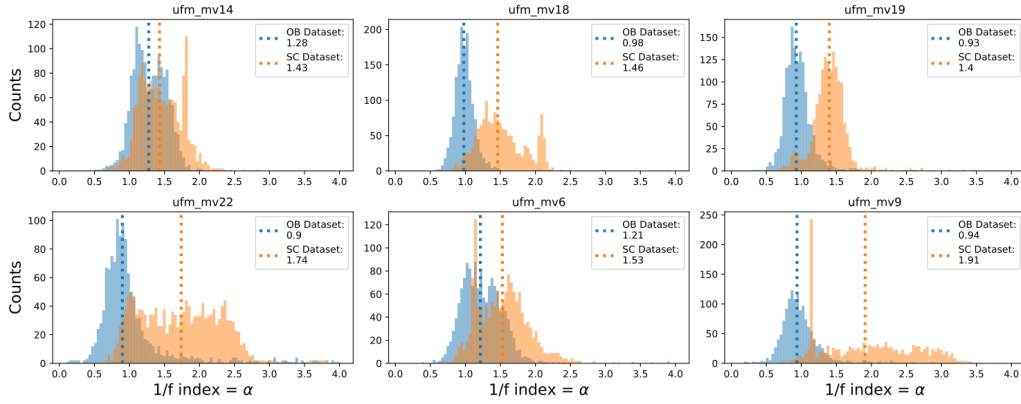


Figure 6.7: 1/f-index (α) fit distribution for OB (blue) and SC (orange) datasets. SC dataset shows a systematically higher index.

They are rated to <10 ppm/ $^{\circ}\text{C}$, and we have established based on analyzing historical weather data at the site that if they are coupling to the ambient temperature swings the 1/f will be sufficiently low to meet our requirements (we require < 17 ppm/ $^{\circ}\text{C}$ to make this true). At $10 \mu\text{A}$ output current we expect $250 \text{ pA}/^{\circ}\text{C}$ to show that we are seeing an excess 1/f noise consistent with this we have analyzed the common mode signal from all detectors and plotted against the temperature of the AMC faceplate in figure 6.9. To get the wafer-averaged common mode we perform the following operations:

1. Exclude outliers with large trends or high noise by making a cut channels with peak-to-peak signal greater than $0.15 \mu\text{A}$ (these are mostly channels that aren't tracking the flux-ramp properly and the number of channels cut are shown in the title of each plot in figure 6.9).

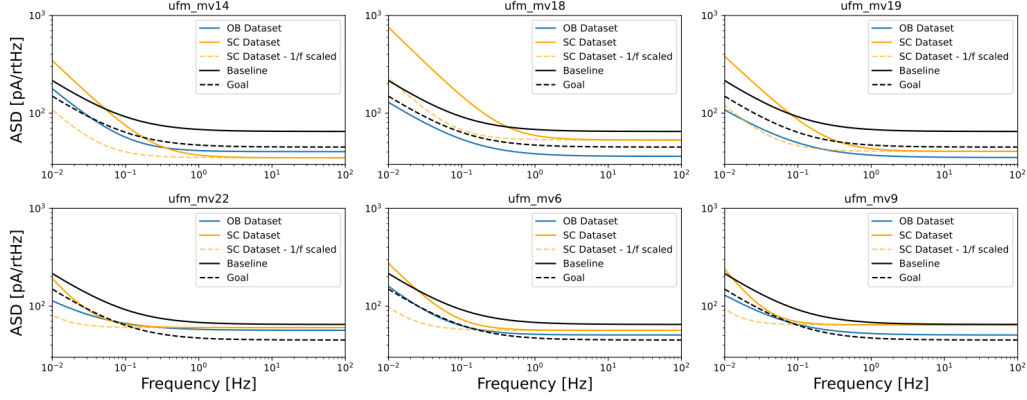


Figure 6.8: A characteristic noise spectrum derived from the median values of each of the fit parameters for each of the UFMs for the SC (orange) and OB (blue) datasets. Overplotted is the baseline (solid black) and goal (dashed black) noise spectra. We also shows the SC dataset with the 1/f component scaled by the shunt and measurement-to-observation bias current ratio as described in figures 6.4b and 6.6.

2. Subtract the DC-offset from each signal
3. Multiply by -1 to for timestreams that have $\text{abs}(\min) > \max$ - this accounts for thermal coupling with the opposite sign.
4. Divide each timestream by the maximum
5. For each sample take the median across all of the detectors
6. Multiply by the median of the maximum we divided by before to restore the scale

If all detectors are uncorrelated this should just average down the variance in each sample by $\sqrt{N_{\text{detectors}}}$, have zero mean, and no trend. A resulting trend represents a

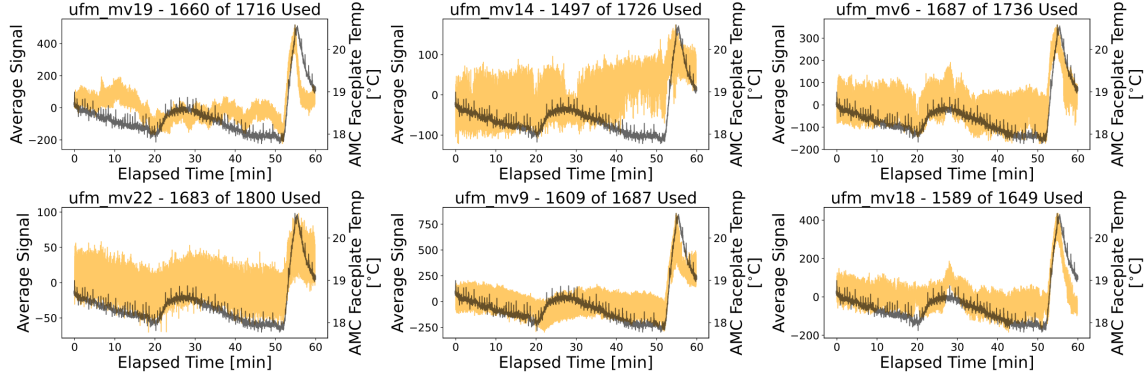


Figure 6.9: Average signal across detectors after cutting outliers and flipping sign of detectors with $\text{abs}(\min) > \max$ for each UFM (orange). This pulls out the common mode signal between all detectors which correlates well with the AMC faceplate temperature on the SMuRF (black). This confirms our expectation that the excess noise in the SC dataset with $10\mu\text{A}$ current is due to thermal coupling at the $\simeq 10$ ppm/ $^{\circ}\text{C}$ level to the resistors in the voltage reference circuit. The temperature rise here is due to the on-SMuRF-crate fans cycling which we will disable in the field so that the temperature change will follow the more gradual change of ambient or slower which is acceptable for our $1/f$ noise spec.

common mode that could have either a positive or negative sign. We see that there is a resulting trend that is correlated with the faceplate temperature confirming that at least some of the excess can be accounted for by thermal coupling. The resulting signal amplitude is $170\text{-}1151 \text{ pA}/2.5^{\circ}\text{C} = 7\text{-}46 \text{ ppm}/^{\circ}\text{C}$ which is slightly higher than the $10 \text{ ppm}/^{\circ}\text{C}$ we expect from the voltage regulator resistors however this temperature profile is following the SMuRF crate fans cycling due to their fan cooling protocol. We know that there is a large gradient from the on-board temps to the AMC faceplate and that the on-board temperature respond more strongly to the fans than the ambient temperature. This means that this was likely an even larger

on-die temperature change corresponding to a smaller temperature coefficient more inline with our expectations. We plan to disable the on-crate fan policy, run the fans at a fixed high speed, and keep the board temperatures low enough by adjusting the flow and inlet water temperature of a water-to-air heat exchanger attached to the inlet fans in their weather sealed enclosure. This should ensure that the temperature rise on-die is much closer to ambient.

6.3 Discussion and Future Outlooks

We have validated the performance of the SAT integrated at UCSD meets the requirements for science operations to the extent we can in the sea-level lab environment. The SAT has since been packed and shipped to Chile and will eminently undergo the on-site commissioning process which involves assembly and cooldown in the site highbay to ensure no degradation during transit. Then the SAT will be moved to its platform and we will begin a series of instrument characterization measurements exercising the range of platform motion, scanning planets, and calibrators while under science-quality atmospheric loading conditions. Upon completion of instrument commissioning which is planned to last a few months we will start initial science observations where we will exercise the full TOD-to-map pipeline and produce initial CMB maps to check data quality. The other 2 SATs and LATR will join

the first SAT in Chile within a few months so that the full observatory will begin full capacity CMB observations in the next $\sim$ year. The SO will be a powerful probe of Cosmology enabled by the expanded readout capabilities of μ mux, the well planned TOD processing pipeline, and rigorous integration and test program discussed in this dissertation.

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