Detectors for the Atacama B-mode Search experiment

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Abstract

Inflation is the leading theory for explaining the initial conditions that brought about our homogeneous and isotropic Universe. It predicts the presence of gravitational waves in the early Universe, which implant a characteristic B-mode polarization pattern on the Cosmic Microwave Background (CMB). The Atacama B-mode Search (ABS) experiment is a polarimeter observing from Cerro Toco (located in the Atacama desert of Chile at an altitude of 5190 m), searching for the yet undetected B-mode signal. ABS carries 480 superconducting Transition Edge Sensor (TES) Bolometers that couple 150 GHz radiation via planar Ortho-Mode Transducers (OMTs) mounted at the output of corrugated feedhorns. The feedhorn beam is projected onto the sky through crossed Dragonian reflectors, a set of reflective and absorptive filters, and a rotating Half Wave Plate (HWP) that modulates any polarized sky signal at 10.2 Hz. The bolometers are cooled to 300 mK by a He3-He4 adsorption fridge system backed by pulse tubes. The reflectors are located within the 4 K cavity of the cryostat, while the HWP is mounted on frictionless air bearings above the cryostat window. This thesis discusses the development and construction of the ABS detector focal plane, and presents results of its performance in the field through August 2012. The ABS detector array sensitivity of $31 \mu K s^{1/2}$, together with the experiment’s unique set of systematic controls, and expected multi-year integration time, could detect a B-mode signal with tensor to scalar ratio $r \sim 0.1$. 
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Chapter 1

Introduction

Cosmology is the scientific study of the large scale properties of the Universe, along with its origin, evolution and ultimate fate. The prevailing cosmological model for the early stages of the Universe is the “Big Bang theory.” In Big Bang cosmology the universe emerged about 13.8 billion years ago with a homogeneous and isotropic distribution of matter at a very high temperature and density, which has been expanding and cooling ever since. In 1929 Edwin Hubble observed this expansion by measuring the distance to, and velocity of, galaxies [56]. He found that in general galaxies were moving away from the Earth in all directions. Some close by galaxies with large peculiar velocity due to local gravitational effects may be moving towards the Earth, but further away these peculiar velocities become negligible, compared to the velocity of the expanding Universe. Hubble’s discovery disproved the idea of a static universe, a popular model at the time.

The dynamics of the Universe are governed by the Einstein field equations of general relativity [34]:

\[ G_{\mu\nu} = -8\pi G T_{\mu\nu}, \]  

where \( G_{\mu\nu} \) is the Einstein tensor describing the geometry of space-time, \( G \) is Newton’s constant, and \( T_{\mu\nu} \) is the energy-momentum tensor. The exact solution to these equations for a homogeneous, isotropic, expanding universe is given by the Friedmann-Lemaitre-Robertson-
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Walker (FLRW) metric \( g_{\mu\nu}(x) \) [39], written in spherical polar coordinates as:

\[
d\tau^2 = -g_{\mu\nu}(x)dx^\mu dx^\nu = dt^2 - a^2(t) \left[ \frac{dr^2}{1 - Kr^2} + r^2 d\Omega \right],
\]

(1.2)

where \( a(t) \) is the scale factor, and the constant \( K \) is the curvature of space. \( K \) is -1, 0 or 1, corresponding to hyper-spherical, Euclidean, and spherical space respectively. Measurements of the Cosmic Microwave Background (CMB) (explored in section 1.2) and Large Scale Structure (LSS) have shown that the Universe is flat to 0.5% \((K = 0)\)[106]. The wavelength of a photon emitted at some earlier time \( t_1 \) appears larger due to cosmological expansion by a redshift factor \( z \) when observed at the current time \( t_0 \). This redshift factor \( z \) is related to the scale factor \( a(t) \) by:

\[
1 + z = a(t_0)/a(t_1).
\]

(1.3)

The Hubble parameter is defined in terms of the scale factor as \( H \equiv \dot{a}/a \), and its current value as the Hubble constant \( H_0 \equiv H(t_0) \).

Applying the Einstein field equations to the FLRW metric yields the Friedmann equations that relate the expansion of the Universe, encoded in the scale factor, to the mass density \( \rho \) and pressure \( p \) of the matter in the Universe:

\[
\dot{a}^2 + K = \frac{8\pi G \rho a^2}{3},
\]

(1.4)

and

\[
\frac{3\dot{a}}{a} = -4\pi G(3p + \rho).
\]

(1.5)

In the Lambda-Cold Dark Matter (LCDM) cosmological model, the Universe is composed of radiation (photons and relativistic neutrinos), matter (leptons, baryons and dark matter), and a cosmological constant, each with its own equation of state \( p = w \rho \). For the first \( \sim 70000 \) years after the Big Bang the Universe’s energy density was dominated by radiation, which has an equation of state \( p = \rho/3 \). The Friedman equation then implies that \( \rho \propto a^{-4} \) (in addition to the dilution of photon density due to the expansion of the Universe, the photon wavelength is also stretched, hence reducing its energy). Matter on
the other hand, exerts no pressure, hence $w = 0$ and $\rho \propto a^{-3}$. This means matter density dilutes more slowly than radiation and eventually becomes the dominant component of the Universe energy content. This transition of matter-radiation equality occurs at $z \sim 3400$.

At redshift of $\sim 1100$, $\sim 380,000$ years after the Big Bang [55], “decoupling” takes place. This is the epoch when protons and electrons first became bound, reducing their interaction cross-section with photons, hence decoupling radiation from matter. Decoupling takes place when the temperature of the Universe decreases to $\sim 3300$ K (average photon energy of $\sim 0.3$ eV). Before this epoch the temperature was high enough that any hydrogen that formed would quickly become ionized by one of the many photons with energy greater than $13.6$ eV; when decoupling occurred most photon energies were below this ionization threshold.

After decoupling came a period called the “Dark ages” before the first stars and galaxies formed through gravitational collapse. During this period the Universe was transparent since electron and protons had formed bound states, but no radiation was emitted except for emission lines such as the 21 cm hydrogen spin line of neutral hydrogen.

The bound hydrogen eventually becomes reionized by the radiation emitted from the first stars. This “Reionization” epoch occurs near a redshift of $z \approx 6$ (700 million years after the Big Bang). This transition is observed in the CMB polarization spectra at large scales ($l < 10$) [70] and through the Gunn-Peterson trough [44] in the spectra of quasars. Quasars are very energetic and distant active galactic nuclei whose emission has been redshifted by cosmological expansion. The wavelengths of the quasar emission at some point match the Lyman-alpha absorption lines of the neutral hydrogen, causing a trough in the quasar spectrum. Only quasars that existed before reionization show this feature; the first such quasar was observed by the Sloan Digital Sky Survey in 2001, at a redshift of 6.28 [7].

Recently at a redshift of 0.5 ($\sim 5$ billion years after the Big Bang) the Universe has entered an epoch of “dark energy” domination. Dark energy exerts negative pressure, driving the accelerated expansion of space. The equation of state of dark energy has not yet been well measured. The data are consistent with $p = -\rho$ ($w = -1$) at this epoch. If $w$ is
constant at $w = -1$ then the dark energy may be equated with a cosmological constant $\Lambda$, which is the intrinsic energy density of vacuum. Initial evidence for the accelerated expansion of the Universe originated from redshift measurements of distant supernova [93] and [100], and was further cemented by large scale structure (LSS) [35] and Cosmic Microwave Background (CMB) measurements [106].

The Universe is currently $13.75 \pm 0.13$ Gyr, spatially flat, composed of 72% dark energy, 23% cold dark matter, and 5% baryons [61]. Most of the energy in the Universe is not well understood; only 5% is made of matter that we have observed. Cold Dark Matter (CDM) is a hypothetical form of matter that is slow moving compared to the speed of light, and interacts only very weakly (or not at all) with electromagnetic radiation. It is necessary to explain the rotational curves of galaxies [102], gravitational lensing measurements [74], and acoustic oscillations in CMB and LSS data [92]. Dark energy is even more mysterious; if understood as a cosmological constant in the form of vacuum energy, particle physics predicts its energy density to be 120 orders of magnitude larger than that observed through cosmological measurements.

1.1 Inflation

Inflation is a period of exponential expansion moments after the Big Bang, which increases the size of the Universe by a factor of $10^{26}$ or about 60 e-folds. It was initially proposed by Alan Guth [46] in 1980; his version (sometimes referred to as “old inflation”) used a scalar field trapped in a false minimum of the potential. The energy difference between the false minimum and the true vacuum acts much like a cosmological constant driving an accelerated expansion of the Universe. The expansion ends when the scalar field tunnels out of the false minimum on to the true vacuum state. It was soon realized [52][47] that this model was wrong, because of what has come to be called the graceful exit problem. The transition from the false vacuum to the true vacuum could not have occurred everywhere simultaneously, but in small bubbles which rapidly expanded. The problem is that the energy released in this transition would end up at the bubble walls, hence leaving the
interior of the bubble essentially empty. Initially Guth thought the bubbles would have merged leading to a homogeneous Universe, but this could not have happened; since the background of false vacuum continued to inflate, the bubble walls would have moved too fast from each other to ever coalesce.

Subsequent inflation models [73] [2] solved the bubble collision problem by instead of having the field trapped in a false vacuum, slowly rolling it down the potential to the true vacuum. This model is referred to as “slow-roll inflation.” When the field rolls slowly compared to the expansion of the Universe, inflation occurs. However, when the potential becomes steeper inflation ends and the scalar field decays to standard model particles via the “reheating” process [3]. There are other more complex models of inflation as well.

Inflation solves three puzzles of Big Bang cosmology; the flatness, horizon, and monopole problems [120].

- **Flatness problem.** Measurements show that the density of the Universe $\rho$ is very close to the critical density $\rho_c$, defined as:

$$\rho_c = \frac{3H^2}{8\pi G}. \quad (1.6)$$

Hence, the curvature of the Universe $K$ in Equation 1.4 is close to zero and the Universe appears to be flat. The ratio $\rho/\rho_c$ is referred to as $\Omega$. Its current value $\Omega_o$ is constrained to $0.9867 < \Omega_o < 1.084$ [67] (assuming a dark energy equation of state $w = -1$), by CMB, supernova (SN) and baryon acoustic oscillations (BAO) measurements.

Instead of considering the flatness of the Universe as a coincidence, inflation provides a possible explanation. Manipulating Equation 1.4 yields:

$$(\Omega^{-1} - 1)\rho a^2 = \frac{3Kc^2}{8\pi G}. \quad (1.7)$$

The Universe is flat if $\Omega = 1$, open if $\Omega < 1$, and closed if $\Omega > 1$. The right hand side of Equation 1.7 contains only constants. Therefore the left hand side must remain constant through the evolution of the Universe. As the Universe expands $a$ increases,
but $\rho$ decreases at a faster rate during radiation or matter domination, hence causing $\rho a^2$ to decrease, and $\Omega^{-1} - 1$ to increase by a factor of around $e^{60}$ [28]. In other words, soon after the Big Bang $\Omega$ must have been equal to 1 with a precision of 26 orders of magnitude, in order to explain today’s $\Omega_0$ measurement. A period of inflation driven by a field with constant energy density, makes $\rho a^2$ grow rapidly while suppressing $\Omega^{-1} - 1$ to zero. If $\Omega$ started at some arbitrary value near unity, then inflation would converge its value to one, to the precision required by our knowledge of $\Omega_o$.

- **Horizon problem** The Universe we can observe appears homogeneous even though without inflation, given the finite age of the Universe, many regions are not in causal contact. The distance between them is so large that light has not had enough time to travel from one to the other. The horizon problem is particularly evident from the uniformity of the CMB radiation. When it was emitted only 380,000 light-years diameter regions were in internal causal contact, while today the same black-body radiation (to $\sim 100$ ppm) is observed in all sky directions, over a sphere of radius 13.8 billion light-years. Inflation solves the horizon problem by taking a small patch of space that was initially in causal contact and expanding it to at least the size of the observable Universe; hence its homogeneity is no longer puzzling, since all the observable Universe was at one point in causal contact.

- **Monopole problem** Grand unified theories (GUT) that unite the electromagnetic, weak, and strong interactions at high energy ($10^{-16}$ GeV) predict the creation of particles with non-zero magnetic charge, roughly one per $10^9$ photons. Experiments searching for monopoles in iron ore, seawater, etc. put an upper limit of $10^{-39}$ monopoles per photon [90]. Inflation can explain this lack of monopole particles. If the GUT produces monopoles at higher energy than inflation, then the number of particles is diluted greatly by the expansion of space during inflation, compared to the number of photons that are produced after inflation by reheating.

To confirm or disprove inflation, we must search for other observable signatures predicted
by the theory. Once such inflationary signature is the generation of gravitational waves (tensor perturbation of the metric), which leave a unique imprint in the polarization pattern of the CMB [128] [63]. Tensor perturbations are small fluctuations of the space-time metric that stretch and compress space. Their amplitude is constrained by the energy scale of inflation and its potential. If their imprint on the CMB is small enough, then a large fraction of inflationary models will be ruled out.

1.2 The Cosmic Microwave Background

After about 379,000 [61] years the Universe cooled to a temperature of 10,000 K and the nuclei could combine with electrons to form neutral atoms. With fewer free electrons available to interact with photons radiation decoupled from matter. This radiation cooled by the continuous expansion of the Universe and is observed today as a black-body signal.
of 2.725 K. This signal is called the cosmic microwave background (CMB) radiation and was first observed in 1965 by Arno Penzias and Robert Wilson at Bell Laboratories [91] (see Figure 1.1). They measured an isotropic, unpolarized excess antenna temperature of 3.51\pm1.0 K at 4.08 GHz, which fitted the radiation predicted by Robert Dicke, Peter Roll, Jim Peebles, and David Wilkinson at Princeton University [26], who at the time were preparing their own search for microwave radiation from the Big Bang. In the 1990’s the Far Infrared Absolute Spectrometer (FIRAS) aboard the Cosmic Background Explorer (COBE) found the CMB radiation to have nearly an exact black-body spectrum, with a brightness temperature of 2.725\pm0.002 K from 60 to 600 GHz [77] (see Figure 1.2).
1.2 The Cosmic Microwave Background

1.2.1 CMB Temperature Anisotropies

The CMB is isotropic to one part in 10,000. The first experiment to observe the small temperature anisotropies was the Differential Microwave Radiometer (DMR), mounted on COBE. It measured differences in brightness temperature between pairs of points on the sky and detected anisotropies at the level of 30 $\mu$K with an angular resolution of 10 degrees [9].

Between 2001 and 2010 the Wilkinson Microwave Anisotropy Probe (WMAP) measured the anisotropy of the CMB at five frequency bands across the entire sky with a final sensitivity of 20 $\mu$K per 0.3° square pixel [61] (see Figure 1.3). The Planck spacecraft launched in 2009 has successfully made an all sky survey of the CMB at nine frequencies, with three times the resolution and ten times the sensitivity of WMAP [113]. Ground-based and balloon experiments that have detected the CMB temperature anisotropies include: ACT [111], SPT [17], TOCO [25], ACBAR [99], CBI [87], BOOMERanG [75], MAXIMA [110], VSA [27], DASI [50], Python [32], MSAM [43], Saskatoon [122], and ARCHEOPS [10].

It is convenient to expand the sky map of CMB temperature anisotropies $\Delta T(\hat{n}) = T(\hat{n}) - T_{CMB}$ (where $T_{CMB} = 2.725$ K) in the direction of the unit vector $\hat{n}$ into spherical

Figure 1.3: On the left is a map of the CMB anisotropies as measured by WMAP. This map is a linear combination of maps obtained from the instrument’s five frequency bands, such that the galactic emissions are subtracted. On the right is a plot of the angular power spectra extracted from the data in the image on the left. Note that the characteristic size of patches in the map is near 1 degree, where the power spectrum peaks. (Figures courtesy of WMAP science team.)
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harmonics $Y_l^m(\hat{n})$:

$$\Delta T(\hat{n}) = \sum_{lm} a_{lm} Y_l^m(\hat{n}),$$

with the sum over multipole $l$ running over all positive-definite integers, and the sum over $m$ running from $-l$ to $l$. Since $\Delta T(\hat{n})$ is real $a_{lm}^* = a_{l-m}$. We cannot make any cosmological predictions about any particular $a_{lm}$ since they depend on our unique position in the Universe. Hence we look at the angular power spectra $C_l$ defined as:

$$\langle a_{lm} a_{l'm'}^* \rangle = \delta_{ll'} \delta_{mm'} C_l,$$

(1.9)

to understand the distributions from which the $a_{lm}$'s are drawn. Note that for a given $l$ each $a_{lm}$ has the same variance. Hence for large $l$ we measure many samples from the same distribution, while for small $l$ we can only measure a few. Thus, there is a fundamental uncertainty in the knowledge we may get about the $C_l$'s called “cosmic variance.” It scales with the inverse of the square root of the number of possible samples:

$$\frac{\Delta C_l}{C_l} = \sqrt{\frac{2}{2l+1}}.$$  (1.10)

At large angular scales ($l < 50$) the CMB anisotropies are sourced by perturbative modes that entered the horizon much after decoupling, and hence were not subject to much evolution. At scales near 1 degree ($l \approx 100$) the first peak of the temperature anisotropy power spectrum (proportional to $C_l$) is observed (see Figure 1.3). This first peak is driven by a mode that entered the horizon and grew due to acoustic oscillation in the matter-radiation distribution until it reached an apparent maximum at the time of recombination. Subsequent peaks are sourced by modes that enter the horizon earlier, and which have undergone multiple oscillations. At smaller scale (larger $l$) the anisotropies are suppressed by diffusion damping and projection effects due to the finite duration of “last scattering.” Photons scatter off multiple electrons within some mean free path distance. This distance increases as the ionization fraction decreases during recombination, causing cold and hot regions to mix, and therefore washing out anisotropies at small scales.

How did the initial spectrum of perturbations that seeded the anisotropy of the CMB get there? Here again inflation provides an answer. Inflation expanded quantum fluctuations
in an initially small region of the Universe into a nearly scale invariant spectrum of density perturbations.

### 1.2.2 CMB Polarization Anisotropies

The perturbations that lead to temperature anisotropies also produce polarization anisotropies in the CMB. The polarization signal is small because it is only produced through Thomson scattering when quadrupole temperature anisotropies are present (see Figure 1.4). When Thomson scattering is frequent, the quadrupole anisotropies are weak because the high rate of interactions thermalize the system. When photons decouple from matter the quadrupole anisotropies are strong but there is very little Thomson scattering to produce polarization. This is why the polarization signal is an order of magnitude smaller than the temperature anisotropies. The quadrupole anisotropies that lead to polarization can be produced by velocity gradients in the photon-baryon fluid. These velocity gradients are due to density inhomogeneities. Quadrupole anisotropies can also be generated by primordial gravitational waves.

The polarization pattern on the sky can be decomposed into B-modes and E-modes [127](see Figure 1.5). Only gravitational waves produce primordial B-modes, while E-modes are also produced by the density fluctuations. E-mode polarization was first detected by DASI [68] in 2002. Other experiments that have detected E-modes include WMAP [70], CAPMAP [11], QUAD [15], MAXIPOL [124], CBI [104], BOOMERanG [81], BICEP [19], and QUIET [98]. B-modes have not yet been observed because of the exquisite sensitivity required to see them. If standard inflationary theory is right then the ratio of the tensor perturbations (gravitational waves) to scalar perturbations (density fluctuations) should be greater than 0.01 [14]. This ratio is referred to as \( r \). The current upper limit on \( r \) is 0.2 [67], inferred from the WMAP low-\( l \) angular power spectra together with baryon acoustic oscillation (BAO), and supernova data (SN). To date, the BICEP experiment upper limit on \( r \) of 0.72 is the most stringent based solely on CMB polarization measurements [19]. Figure 1.6 shows the E-mode power spectrum observations, as well as the upper limits on
Figure 1.4: An electron sitting at the origin is surrounded by a quadrupole anisotropy, with hot unpolarized radiation (high intensity) incoming from the left and right, and cold unpolarized radiation (low intensity) incoming from the top and bottom. Only the vertically polarized component of the hot radiation will scatter off the electron in the direction of ABS. At the same time only the horizontally polarized component of the cold radiation will scatter in the same direction. The radiation observed by ABS will have a net vertical polarization, since that component has a higher intensity, hence unpolarized light from a quadrupole temperature anisotropy generates a polarized signal through Thomson scattering.
1.2 The Cosmic Microwave Background

Figure 1.5: The polarization pattern on the sky can be characterized in terms of a scalar $(E)$ and a pseudo-scalar $(B)$ field [63, 128]. $E$ and $B$ differ in their behaviour under parity transformation: $B$ changes sign while $E$ does not. This can be seen in the figure above. The $E$ patterns when reflected about the vertical axis do not change while the $B$ patterns change handedness. The $E$-$B$ decomposition is a linear transformation of the $Q$-$U$ (Stoke’s parameters [60]) fields on the sky. This transformation is invertible, and makes $E$ and $B$ invariant under translation or rotations of the sky coordinate system. These conditions imply the transformation must be non local [127]. The values of $E$ and $B$ at a point $\Theta$ in the sky are computed by averaging the $(Q_r, U_r)$ Stoke’s parameters defined based on the radial directions about $\Theta$. The averages of $Q_r$ construct $E$, and averages of $U_r$ construct $B$. The weight along circles centered at $\Theta$ should be constant, but each circle is typically weighted by the inverse of its radius. The CMB polarization is expected to generate zero circular polarization; hence the Stoke parameter $V$ is zero.
Figure 1.6: Summary of published CMB polarization measurements by [72], [81], [104], [124], [11], [15], [19], [70], and [98]. On the top is the CMB $EE$ power spectrum observations, with the prediction from $\Lambda$CMB cosmology over-plotted in black. The bottom plot shows the 95% C.L. upper limits on the $BB$ power spectrum from CMB polarization measurements. Over-plotted is the predicted primordial $BB$ power spectrum if $r = 0.1$, and the contributions to $BB$ at higher multipoles from gravitational lensing of $E$-modes. (Figure courtesy of Akito Kusaka for the QUIET collaboration [98])
the B-mode power spectra from polarization measurements.

Three sources of foregrounds that contaminate the B-mode signal generated by primordial gravitational waves are: lensing [129], synchrotron radiation, and dust [88]. Structures in the form of galaxies and galaxy clusters between the Earth and the surface of last scattering bend the path of the CMB photon through gravitational lensing, converting E-mode polarization patterns to B-mode at small angular scale. To find primordial B-modes, uncontaminated by lensed E-modes, searches must concentrate at angular scales near a multipole of 100 and below\(^1\). Synchrotron radiation is generated by relativistic electrons spiraling through magnetic fields. It is polarized and has negative spectral index (intensity decreases at higher frequencies as \(I \propto f^{-2.7}\), approximately). Thermal emission from dust grains may be polarized depending on the size and shape of the grains. Dust has a positive spectral index (intensity increases with frequency as \(I \propto f^2\), approximately) and becomes relevant compared to the CMB polarization at frequencies above 100 GHz. To reduce dust, free-free, and synchrotron contamination, B-mode experiments point away from the Galaxy to the north and south galactic poles, and also rely on multi-frequency measurements to subtract out any contamination that remains.

Detecting or placing more stringent upper limits on a B-mode signal would constrain the various theories of inflation. Furthermore, a detection of primordial gravitational waves would be the first ever probe of physics at the inflation energy scale near \(10^{16}\) GeV. The Atacama B-mode Search (ABS) experiment has joined the effort to measure the CMB B-mode signal along with experiments such as BICEP2 [85], EBEX [86], SPIDER [22], the Keck array [108], Polarbear [114], CLASS [20], and the Planck spacecraft [113]. The ABS instrument brings forth a unique set of systematic controls, a large focal plane populated with the latest generation of cryogenic polarimeters, and the capability of a multi-year observation campaign.

\(^1\)Experiments such as ACTPol [83] and SPTpol [12] target multipoles larger than a few hundred, to study the lensing of the CMB polarization due to the structure of matter in the Universe.
1.3 Overview

Chapter 2 discusses the ABS instrument, including cryogenics, optics, detectors, and the instrument’s expected sensitivity to a B-mode signal. Chapter 3 addresses the detector’s SQUID multiplexing readout, and the tuning steps taken to reduce the readout noise. Chapter 4 presents the results of dark detector tests in the development of single pixels, and in the process of array construction. These tests include Johnson noise measurements, I-V curves, dark noise measurements, responsivity estimates, and complex impedance measurements and modeling. Chapter 5 examines detector optical results from single detector tests in Princeton, array tests in the Princeton high-bay and array performance in the Atacama Desert, including single detector and array optical efficiency measurements, bandpass measurements, beam maps, and observations of half-wave plate systematics. Chapter 6 summarizes the array assembly from mounting single chips on feedhorns, to pods of detectors, to the construction of the focal plane. Chapter 7 discusses the ABS detector array performance, suggests improvements for the instrument, and improvements for the next generation of detectors.
Chapter 2

The ABS experiment

The Atacama B-mode Search (ABS) experiment aims to detect the B-mode polarization pattern imprinted on the CMB by primordial gravitational waves. The ABS instrument is a ground based polarimeter optimized to map the CMB polarization on degree angular scales at which the expected primordial B-mode power spectrum peaks. The instrument is sensitive to microwave photons of $\sim 2$ mm wavelength, close to the intensity peak of the CMB blackbody spectrum. The nominal bandpass is centered at 145 GHz with a 20% bandwidth to maximize sensitivity to the CMB while reducing atmospheric noise by avoiding the wings of the atmosphere’s oxygen line at 117 GHz and its water line at 183 GHz. The noise of each ABS detector operating at $\sim 500$ mK is comparable to the photon shot noise, hence the sensitivity of a single detector cannot be improved much further. The ABS instrument sensitivity is enhanced by closely packing 240 polarimeters in the focal plane of the telescope. The telescope is a Dragonian crossed Cassegrain reflector pair that is mounted inside the cryostat with its primary mirror observing the sky through an aperture defined at the 4K stage. This well defined cold stop reduces systematics from feedhorn sidelobes by mapping them onto the cold 4K cavity. The stop also provides an ideal location to place optical elements like infrared (IR) filters, bandpass filters, a neutral density filter, or a half wave plate (HWP). Ideally any element placed on the stop illuminates equally all the detectors on the focal plane. Filters are located on the 4K stop and a few centimeters above
Figure 2.1: Plotted above are theoretical predictions for the CMB E-mode (EE blue), and B-mode (BB green) power spectra as functions of multipole $l$. Plotted in cyan is the BB power spectrum generated through lensing of E-modes. This is the dominant source of B-modes at high multipoles. The red line shows the estimated power spectrum polarization error per multipole of ABS following the Knox formula \cite{Knox}. This error depends on the scan strategy, the beam profile, and the array sensitivity. The grey boxes bin the above polarization error across various multipoles, and show that ABS can easily detect the BB spectrum if $r$ greater than 0.1.
on the 40K stage. A 300 K vacuum window translucent to microwaves is located above the 40K stage. The HWP is mounted on top of the vacuum window and operates at ambient temperature. It is suspended on an air bearing that permits frictionless rotation up to five revolutions per second. The HWP modulates the polarization signal from the sky at four times its rotational frequency, but does not modulate any intrinsic polarization produced by the instrument below. This modulation allows the signal band to be distinguished from any 1/f noise.

Given that for ground based telescopes the atmosphere is the dominant source of photons at 145 GHz, it is crucial to observe from a site that has low water vapor content to minimize atmospheric loading. ABS observes from the Atacama Desert in the north of Chile. The site is located on Cerro Toco next to the Atacama Cosmology Telescope (ACT) [111] at an altitude of 5190 m. Its altitude combined with the dry air of the desert provides excellent conditions for measuring the CMB. The site’s latitude of 22°S allows access to the entire southern sky and part of the northern sky.

ABS is designed for fast deployment. The experiment lives within a shipping container that holds the cryostat, the motion base, the cryogenic compressors, and all related read out electronics. At the site the base and cryostat are raised to the container’s roof. Once raised ABS can scan 360° in azimuth; and in latitude from 40° elevation to the zenith. A small linear drive coupled to the azimuth gear provides fast scanning of up to 2° s⁻¹, with a maximum throw of 10° around any azimuth direction. Two Cryomech pulse tubes [119] cool the cryostat’s 4K and 40K stages continuously. The focal plane is further cooled to 300 mK by a He₄-He₃ adsorption system with a hold time of up to ~ 42 h, and a fridge recycling period of ~ 6 h. A second He₄ system provides heatsinking for the detector array wires.

The telescope will target three main patches: one intersects the plane of the Milky Way galaxy, while the other two avoid it to its north and south. It also targets calibration sources, such as the Moon, Jupiter, Venus, TauA, and others. Observations are divided into scans that last between half an hour to two hours, where the telescope is pointing at one
The key characteristics of the ABS experiment are summarized in Table 2.1. Figure 2.1 shows theoretical predictions for the E-mode and B-mode power spectrum with $r = 0.1$. The error bars in grey are computed from the characteristics of ABS, and depict how well it can measure the E-mode and B-mode power spectrum excluding foregrounds. With 8760 hours of observations ABS could detect the primordial B-mode power spectrum if $r = 0.1$ to $2\sigma$.

### 2.1 Cryogenics

The ABS cryostat was constructed by Precision Cryogenic Systems, Inc [96](see Figure 2.2). It is a roughly 40" diameter, 41" high cylinder. The vacuum shell consists of two lids and two cylindrical sections held together by O-ring flanges. The lids are 1" thick Aluminum disks and the two open cylindrical walls are 0.27" Aluminum. The top lid has a 14" circular hole where the 11.75" diameter window is mounted. The bottom cylindrical section has two circular tube extensions at 45 degrees that hold the pulse tube heads. It also has multiple feed through flanges for thermometer and detector read out.
2.1 Cryogenics

Figure 2.2: Above is a section view of the ABS cryostat, including rays as computed by CodeV. A) Vacuum Vessel, B) Secondary Mirror, C) 40K Radiation Shield, D) 4K Cryoperm Radiation Shield, E) Primary Mirror, F) 1st Stage G10, G) 2nd Stage G10, H) Pulse Tube (1 of 2), I) 3 He Sorption Fridge, J) 4 He Sorption Fridge (1 of 2), K) Detector Array.
ABS uses two Cryomech pulse tubes, a PT407, and a PT410 ([119]). Each has two cold stages with nominal cooling powers of 0.75 W [1 W] at 4.2 K and 20 W [35 W] at 45 K for PT407 [PT410]. A generic pulse tube consists of a closed volume of high-purity helium, a compressor and a cold head. Within the compressor, compressed helium gas is cooled by passing through a heat exchanger. Cooling is achieved at the cold head by the adiabatic expansion of the high pressure helium. The expanded helium is then returned to the compressor. In particular the Cryomech cold heads have no moving parts, and therefore produce few mechanical vibrations that could affect the instrument.

Inside the vacuum vessel there are three cylindrical shells. All inner shells are composed of two sections that bolt together at horizontal planes close to the middle of the cryostat. This arrangement permits access to the working volume of the cryostat by only removing the top shell of each section. The first shell is made out of mu-metal, operates at 300 K and follows the inside of the vacuum vessel. Inside the mu-metal shell is the 40 K shield made out of Aluminum. This stage is held and thermally isolated from the 300 K vessel through a glass fiber reinforced epoxy typically referred to as G-10. G-10 is a strong material (tensile strength 817 MPa [64]) with low thermal conductivity of 0.08 W m$^{-1}$ K$^{-1}$ at 4 K and 0.2 W m$^{-1}$ K$^{-1}$ at 40 K [123] [121]. The 40 K stage is covered with a multi-layer insulation blanket (MLI) composed of 30 aluminized mylar layers. These low emissivity surfaces reflect the surrounding 300 K blackbody radiation, hence reducing the stage’s radiative loading by a factor approximately equal to the number of layers plus one [65]. A set of infrared filters reduce the radiation through the window aperture; see section 2.3.3.

Load curve measurements of both pulse tubes' first stage predict a cooling power of $\sim 27$ W at 40 K. On the other hand, the expected loading from G-10 is $\sim 0.5$ W, the radiation loading on the MLI covered walls of the stage is $\sim 1$ W (assuming an MLI power reducing factor of 30), the radiation loading through the window and the first reflective IR filter onto the 40K aperture is $\sim 10$ W, and the thermal loading from the detector wires (copper) is $\sim 1$ W. Inside the 40K stage is located the 4K stage which is held in place and thermally isolated by G-10 posts. This stage is made out of cryoperm [23] to act as
Figure 2.3: Above is a schematic of the ABS $^4$He/$^3$He adsorption fridge system backed by a 407 and a 410 cryomech pulse tube. The second pulse tube stage provides a 4 K heatsink that pre-cools the gas in the $^4$He and $^3$He fridges. When the heat switches are turned on, the fridge charcoals are also cooled to 4 K, and behave as adsorption pumps that lower the pressure and thus reduce the temperature of the condensate. The $^4$He1 pot provides a 0.7 K cold finger that condenses the $^3$He gas, while the $^4$He2 pot cools the 1 K hex. The $^3$He fridge cools the 300 mK hex, which holds the detector array.

magnetic shielding and covered with MLI to reduce 40K optical loading. The pulse tubes’ cooling power at 4K is $\sim 2.0$ W. The expected loading from the G-10 support is $\sim 0.04$ W, from radiation on the MLI is $\sim 0.025$ W, from radiation on the 4 K filter is $\sim 0.2$ W, and from detector wires (manganin) is $\sim 0.001$ W. Since the expected loadings are less than the cooling powers at each stage, the operating temperatures settle below 40 K, and 4 K.

The 4K stage holds two helium-4 and one helium-3, adsorption fridges [71]. One of the helium-4 fridges cools the 1K stage of the focal plane, while the other backs the helium-3 fridge that cools the focal plane below 300mK. The helium-4 fridges have 4 mW of cooling power at 700 mK, and the helium-3 fridge has about 36 $\mu$W of cooling power at 300 mK.
The ABS experiment

Both the helium-4 and helium-3 fridges operate by closed cycle adsorption. Each fridge has a charcoal cryopump, a condensation plate, and a reservoir or pot, all connected by a narrow low-conductivity tube. When cooled to \( \sim 4 \) K the charcoal surface area traps large amounts of helium gas, eventually adsorbing all the helium that is evaporated at the pot. The helium is expelled from the pump (regenerated) by heating it to \( \sim 45 \) K. The fridge cycle begins by having most of the gas trapped at the pump. The hot gas released by heating the charcoal is cooled at the condensation plate. The plate’s temperature is below the condensation temperature \( T_{\text{con}} \) of the gas, hence liquid helium falls into the pot. At atmospheric pressure \( T_{\text{con}} = 4.2 \) K for helium-4 and \( T_{\text{con}} = 3.2 \) K for helium-3; in the fridge this temperatures are higher since the gas is pressurized. Once the majority of the gas has been liquefied the charcoal is cooled to 4 K. At this point the charcoal acts as a pump, decreasing the pressure of the helium gas, which lowers its boiling temperature and therefore the temperature of the condensed liquid. This step continues until all the helium is adsorbed by the pump. See Figure 2.3 for a description of the ABS helium-3/4 adsorption fridge system.

A hexagonal kevlar support system suspends the 1 K stage from the 4 K, and the 300 mK stage from 1 K stage. The stages are connected to the adsorption fridges through copper cold fingers. Additionally, the 1 K stage is thermally connected to the 4 K stage through the heat sinking of the electrical wires from the SA board to the focal plane. These wires have a width of 0.004" and are made out of Niobium and Titanium (NbTi) cores that become superconducting below 9.2 K, with thin CuNi cladding. Superconducting wires have very small thermal conductivities since thermal energy is not mediated by electrons but only by phonons. Table 2.2 shows the properties of kevlar and NbTi, as well as the expected thermal loading on the sub-kelvin stages from their kevlar support and wire heat sinking arrangements.

Radiation that reaches the 4K volume is a major source of loading on the 300 mK and 1 K stages. Optical power from the atmosphere that passes through all filters into the 4K cavity amounts to \( \sim 135 \mu W \), while the emission from the last nylon filter at 6 K leads to
### 2.1 Cryogenics

<table>
<thead>
<tr>
<th>Property</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>NbTi thermal conductivity below 1 K (mW cm(^{-1}) K(^{-1}))</td>
<td>0.21T(^{2})</td>
</tr>
<tr>
<td>NbTi thermal conductivity between 1 K and 4 K (mW cm(^{-1}) K(^{-1}))</td>
<td>0.151 + 0.058T(^{1.9})</td>
</tr>
<tr>
<td>CuNi thermal conductivity between 0.05 K and 4 K (mW cm(^{-1}) K(^{-1}))</td>
<td>0.9366T(^{1.23})</td>
</tr>
<tr>
<td>Temperature of 4K stage (K)</td>
<td>3.8</td>
</tr>
<tr>
<td>Temperature of 1K stage (K)</td>
<td>0.7</td>
</tr>
<tr>
<td>Temperature of 300mK stage (K)</td>
<td>0.3</td>
</tr>
<tr>
<td>Number of wires between 4K and 300mK</td>
<td>258</td>
</tr>
<tr>
<td>Length of wire from 4K to 1K heat sink (cm)</td>
<td>17.8</td>
</tr>
<tr>
<td>Length of wire from 1K heat sink to 300mK board (cm)</td>
<td>7.6</td>
</tr>
<tr>
<td>Cross-sectional area of wires (cm(^2))</td>
<td>81 \times 10^{-6}</td>
</tr>
<tr>
<td>Wire heat sinking power from 4K to 1K ((\mu)W)</td>
<td>9.7</td>
</tr>
<tr>
<td>Wire heat sinking power from 1K to 300mK ((\mu)W)</td>
<td>0.63</td>
</tr>
<tr>
<td>Kevlar thermal conductivity below 4 K (mW cm(^{-1}) K(^{-1}))</td>
<td>0.039T(^{1.7})</td>
</tr>
<tr>
<td>Kevlar diameter (inch)</td>
<td>0.025</td>
</tr>
<tr>
<td>Kevlar length (inch)</td>
<td>1</td>
</tr>
<tr>
<td>Number of kevlar strings per stage</td>
<td>12</td>
</tr>
<tr>
<td>Kevlar power from 4K to 1K ((\mu)W)</td>
<td>8</td>
</tr>
<tr>
<td>Kevlar power from 1K and 300mK ((\mu)W)</td>
<td>0.07</td>
</tr>
</tbody>
</table>

Table 2.2: The thermal properties of NbTi, and Kevlar were extracted from [94] and [118] respectively. The length of the 4K to 300mK cables was determined from electrical read out considerations, while the location of their 1K heat sink was based on geometrical constraints and reduction of loading on the 300mK and 1K stages.
<table>
<thead>
<tr>
<th>Stage</th>
<th>Temperature at Princeton (K)</th>
<th>Temperature at Cerro Toco (K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>PT1 40K</td>
<td>42</td>
<td>36</td>
</tr>
<tr>
<td>PT2 40K</td>
<td>39</td>
<td>35</td>
</tr>
<tr>
<td>40K Filter Ring</td>
<td>59.8</td>
<td>60.5</td>
</tr>
<tr>
<td>PT1 4K</td>
<td>3.17</td>
<td>3.2</td>
</tr>
<tr>
<td>PT2 4K</td>
<td>3.14</td>
<td>3.1</td>
</tr>
<tr>
<td>4K Filter plate</td>
<td>3.97</td>
<td>3.6</td>
</tr>
<tr>
<td>NDF/Nylon middle</td>
<td>14.5</td>
<td>6.05</td>
</tr>
<tr>
<td>4He1 pot</td>
<td>0.681</td>
<td>0.661</td>
</tr>
<tr>
<td>4He2 pot</td>
<td>0.72</td>
<td>0.73</td>
</tr>
<tr>
<td>3He pot</td>
<td>0.243</td>
<td>0.239</td>
</tr>
<tr>
<td>Focal plane</td>
<td>0.310</td>
<td>0.290</td>
</tr>
<tr>
<td>Focal planes servo temperature</td>
<td>–</td>
<td>0.308-0.330</td>
</tr>
</tbody>
</table>

Table 2.3: Summary of ABS temperatures achieved while testing in Princeton, and observing from Cerro Toco. In Princeton an NDF was mounted below the 4K cold stop, while at Cerro Toco it was replaced by PTFE and Nylon filters, and a multilayer metal mesh filter was removed. See section 2.2.3.

4µW. Even a small percentage of this radiation power coupling to the focal plane stages easily dominates other contributions.

Table 2.3 summarizes the expected and achieved performance of the four temperature stages of the cryostat while operating in the highbay of Princeton with a neutral density filter, and while looking at the sky in Chile. The operating temperatures are all within range to optimize the detector behavior and minimize the optical loading from plastic filters at 40 K and 4 K.

## 2.2 Detectors

The detectors used in ABS were developed by the TRUCE collaboration [126], and fabricated at NIST Boulder. Each detector chip is 5 mm in diameter and 275µm thick. About one hundred detectors are fabricated on a 75 mm diameter silicon wafer. The fabrication process adds Niobium (Nb), Silicon dioxide (SiO₂), Silicon nitride (SiN), Copper (Cu),
Figure 2.4: Picture of a CMB5 prototype pixel for ABS. The OMT at the center couples radiation through CPW-MS transmission lines to TES bolometers at the edges. The transmission lines contain bandpass defining filters. Bias lines for the detectors and heaters include filters to reduce RF pickup. The heaters are small gold resistors located on the TES island.
Molybdenum (Mo), and Gold (Au) layers. Most signal carrying traces are made out of Nb to reduce loss. Niobium has a superconducting critical temperature of 9 K, much higher than the 300 mK operating temperature, and a gap frequency of 740 GHz well above the high frequency edge of the ABS bandpass ($\approx 170$ GHz). SiO$_2$ and SiN form membranes that separate or suspend metal traces. The TES sensor is made out of a MoCu bilayer, whose thickness can be tuned to set the critical temperature $T_c$ of the detector. The Au layer forms heater resistors on the bolometers and creates the lossy structure where the incoming optical power is dissipated. The detector performance depends strongly on the purity of the Nb, the recipe used to create the SiO$_2$ and SiN, and the thicknesses of the Cu, Mo and Au layers.

Figure 2.4 shows a picture of an ABS detector chip. Microwaves couple to the triangular probes in the middle of the chip. The photon power is then transferred to a microstrip transmission line [95] through a co-planar waveguide (CPW)-microstrip (MS) transition. The microstrip then carries the signal through stub filters that define the frequency band. Finally, the signal is carried on to the TES island where it is dissipated on a lossy gold microstrip meander. The signal power is measured by a TES bolometer read out by a Superconducting Quantum Interference Device (SQUID).

### 2.2.1 Ortho-Mode Transducer

The four triangular antenna probes are suspended on a silicon-nitride (SiN) membrane located in between the output of a corrugated feedhorn and a quarter wavelength backshort. They act as a planar ortho-mode transducer (OMT) that couples vertically polarized light to the top and bottom probes and horizontally polarized light to the right and left probes. Historically, waveguide devices performing the OMT function have been inherently three-dimensional. The planar OMT design permits close packing of a large number of detectors in the limited area of the focal plane. Each triangular probe serves as an impedance transformer between the end of the corrugated feedhorn and a CPW trace. The OMT is designed to have optimal transmission at 145 GHz, the center of the ABS band. Nevertheless the
### Chip OMT dimensions

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Diameter of detector chip</td>
<td>5 mm</td>
</tr>
<tr>
<td>OMT waveguide diameter</td>
<td>1.6 mm</td>
</tr>
<tr>
<td>Base and height of the four triangular OMT probes</td>
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### CPW and MS dimensions

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<tr>
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<tbody>
<tr>
<td>CPW center width</td>
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</tr>
<tr>
<td>CPW gap</td>
<td>4.5 μm</td>
</tr>
<tr>
<td>MS width</td>
<td>5 μm</td>
</tr>
<tr>
<td>Number of sections in CPW-MS transition</td>
<td>10</td>
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#### Stub filters

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<tbody>
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<td>MS width</td>
<td>7 μm</td>
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<tr>
<td>Distance between stubs</td>
<td>183 μm/210 μm</td>
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<tr>
<td>Outer stub length</td>
<td>176 μm/203 μm</td>
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<tr>
<td>Outer stub width</td>
<td>32 μm</td>
</tr>
<tr>
<td>Inner stub length</td>
<td>175 μm/201 μm</td>
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<tr>
<td>Inner stub width</td>
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#### Gold meander

<table>
<thead>
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<tbody>
<tr>
<td>Thickness</td>
<td>100 nm</td>
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<td>Length</td>
<td>3170 μm</td>
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### TES dimensions

<table>
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</tr>
</thead>
<tbody>
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<td>Island</td>
<td>305 μm × 300 μm × 2 μm</td>
</tr>
<tr>
<td>Each leg</td>
<td>350 μm × 9 μm × 2 μm</td>
</tr>
<tr>
<td>TES bilayer</td>
<td>77 μm × 70 μm × 0.3 μm</td>
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</table>

### TES properties

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Heat Capacity ((C))</td>
<td>(\sim 2) pJ K(^{-1})</td>
</tr>
<tr>
<td>Thermal conductivity ((G))</td>
<td>(\sim 100) pW K(^{-1})</td>
</tr>
<tr>
<td>TES normal resistance ((R_n))</td>
<td>(\sim 5) mΩ</td>
</tr>
<tr>
<td>Critical Temperature ((T_c))</td>
<td>(\sim 500) mK</td>
</tr>
</tbody>
</table>

Table 2.4: Dimensions and basic characteristics of an ABS pixel. The ABS array contains detectors with two different stub filter design. Dimensions for both designs are given in the table. See Figure 2.8 for more details. The TES properties section holds typical values of key detector characteristics; see also Chapter 4.
Figure 2.5: Plot of the OMT performance from HFSS simulations. Across the target band the OMT power coupling efficiency is greater than 90%. The majority of the power loss is due to reflections, but a small fraction is radiated away through the waveguide gap caused by the OMT. Figure previously published by McMahon, et al. (2009) [78].
2.2 Detectors

2.2.2 CPW-Microstrip Transition

In the ABS chip design the observed radiation must couple eventually to a microstrip line, which contains the bandpass filters and guides the power through the TES legs on to the

OMT transmission begins above the waveguide cutoff (110 GHz for ABS) and ends above the 165 GHz to 170 GHz high edge of the filter defined bandpass. The high OMT bandwidth provides the possibility of multichroic detectors where multiple frequency bands are sampled through the same OMT. Figure 2.5 shows the OMT power coupling efficiency between 110 GHz and 180 GHz and the target bandpass.

Figure 2.6: The design of the CPW-MS transition is shown on the top of the figure. When there is a hole on the ground plane the transmission line acts as a CPW, otherwise it is a microstrip. The reflection of the transition across the target band is less than $-15$ dB from HFSS simulations, Figure previously published by McMahon, et al.(2009) [78].
TES island. The OMT directly couples to a CPW instead of a microstrip to improve the impedance match between probes and transmission line. The CPW and microstrip dimensions given in Table 2.4 imply impedances of $\approx 140 \, \Omega$ and $\approx 10 \, \Omega$ respectively. A stepped transition is used to match these two components, reducing reflections between them to less than 20 dB across the band (see Figure 2.6).

### 2.2.3 On-Chip Filters

The frequency bands of CMB instruments based on bolometers have typically been defined using optical elements such as metal mesh filters [1]. For ABS the bandpass is defined on the chip by a 5-pole Chebychev $\lambda/4$-shorted stub filter [13]. The five pole design provides a swift roll-off at the band edges. Stub filters have a simple rectangular geometry that makes them easy to fabricate, but generate undesirable harmonic passbands at frequencies of 3, 5, 7, etc, times the design frequency. To suppress these resonances two simple stepped impedance low-pass filters with roll-offs at 190 GHz and 310 GHz are implemented. Simulations of the filter stack in Sonnet EM [105] showed that out-of-band leakage should be below $-30 \, \text{dB}$ up to the Nb superconducting gap frequency. Figure 2.7 shows a diagram of the stub filter design and its simulated response including dielectric loss. Its dimensions are summarized in Table 2.4.

### 2.2.4 Gold Meander

After passing through the stub filters the signal is guided on Nb microstrip lines across the TES legs. Once on the island the Nb transitions into Au. The signal power is then dissipated on the TES island through the lossy Au microstrip. HFSS simulations indicate that the characteristic impedance of the superconducting Nb microstrip is 13.7 $\Omega$. On the other hand, the expected impedance of the 100 nm thick gold microstrip is $13.8 - 3.1 i \, \Omega$ assuming a conductivity of $45 \times 10^6 \, \text{S/m}$ (the room temperature bulk value). Therefore the reflected power is $\left|\frac{(13.7 - (13.8 - 3.1i))}{(13.7 + (13.8 - 3.1i))}\right|^2 = 0.013$ or 1.3%. For comparison a 50 nm thick gold microstrip would reflect as much as 10% of the power. Conductivity
Figure 2.7: Plotted above is a diagram of the ABS microstrip stub filters. The dimensions quoted on the diagram apply to the first two wafers of detectors. On the bottom is the frequency bandpass simulated from the stub filter design. The filters attenuate by at least 30 dB radiation at frequencies below and above the passband. At the 740 GHz Nb gap frequency, the transmission lines become lossy, eliminating any coupling from the probes to the TES. The simulation includes dielectric loss assuming a loss tangent of 0.005 for the SiO₂ insulator.
tends to increase as temperature decreases hence the previous estimates are a worse case scenario. We have estimated the cold conductivity of the gold microstrip to be $72 \times 10^6 \text{S/m}$ by measuring the resistance of TES heaters made from the same gold layer. With higher conductivity the impedance mismatch between microstrips becomes smaller and the power reflected by the transition also decreases. The Au meander length is 3452 $\mu$m, long enough to guarantee that any power reaching it will be dissipated on the TES island. Table 2.4 summarizes the characteristics of the lossy microstrip. A drawing of the meander is found in Figure 2.8. There is one meander for each probe, and therefore two for each polarization channel.

### 2.2.5 TES Bolometer

The power dissipated by the gold meanders on the island is measured by a transition edge sensor (TES) thermometer. The TES bolometer consists of a superconducting film biased on its transition and thermally decoupled from the bath. The TES film used in ABS is a rectangular MoCu bilayer. The proximity effect [79] predicts that by varying the thicknesses of the Mo superconducting layer and the Cu normal layer the bilayer transition edge temperature ($T_c$) can be adjusted. ABS detector targets a $T_c$ of $\approx 500 \text{mK}$. The bilayer is located in the middle of the TES island, and electrically connected to the TES bias lines through Nb pads at two of its edges. Figure 2.8 shows the location of the bilayer, the lossy meander, the gold heater and the thin legs that thermally isolate the island from the bath. It also shows a cross-sectional view of the island and its multiple layers. The Nb ground plane sits on a SiN membrane and is separated from a top Nb wiring layer by a SiO$_2$ dielectric layer. The top Nb layer holds traces for the microstrips and bias lines. The ground plane has a hole in the middle of the island such that the MoAu bilayer sits on Nb pads. One of the bias traces connecting to the bilayer pads is etched from the Nb ground plane while the other connects through a via to the top Nb layer. The gold for the lossy meander and heater is deposited on top, above a SiO$_2$ layer. A second SiO$_2$ layer is added to the bottom side of the SiN, to relieve stress. The four legs that hold the island
Figure 2.8: Diagram of an ABS TES bolometer. The square island and legs are composed from two layers of Nb and SiO$_2$, and one layer of SiN. The superconducting MoCu bilayer is mounted on Nb traces in the middle of the island. It is surrounded by two lossy Au microstrip lines, and a small Au heater. (Figure courtesy of Cho, Hsiao-mei.)
A TES bolometer consists of a thermometer attached to a heat capacity \( C \), that is weakly coupled to a thermal bath. A Radiation signal adds power \( P_{\text{opt}} \) on the heat capacity block, while the electrical read out bias of the thermometer adds \( P_{\text{bias}} \) power. The temperature of the heat capacity block increases until it reaches an equilibrium temperature \( T_0 \). At this temperature the power added to the block equals the power that flows to the bath through the weak thermal link of conductivity \( G \). If \( P_{\text{bias}} \) is constant, then changes in \( P_{\text{opt}} \) lead to variation of \( T_0 \) that are measured by the thermometer. In a TES, the thermometer consists of a superconducting film operating at critical temperature where small changes in temperature cause large changes in resistance. The TES resistance is typically voltage biased, hence increases in \( P_{\text{opt}} \) are followed by decreases in \( P_{\text{bias}} \). This leads to negative electro-thermal feedback that keeps the TES on its transition at \( T_0 \equiv T_c \) and speeds up its response.

are composed of five layers: two made of Nb, two made of SiO\(_2\) and one SiN. Two of the legs carry microstrip lines from the probes to the meander, while the other two legs carry bias lines for the TES and heater. The legs set the thermal conductivity \( G \) from the island to the bath. Thermal conductivity \( G \) and \( T_c \) determine the detector noise as well as the maximum power each detector can measure, referred to as saturation power (\( P_{\text{sat}} \)).

A simple TES bolometer model [76] [59] is depicted in Figure 2.9. In the absence of optical loading the system is at equilibrium temperature \( T_o \) when the power deposited on the island by the read out circuit \( P_{\text{bias}} \) is equal to the power flowing to bath (\( P_{\text{bath}} \)). Now consider the case where optical power \( P_{\text{opt}} \) is deposited on the island of heat capacity \( C \) through the lossy meander. This increase in power causes the island temperature to increase,
hence the TES moves up its transition becoming closer to normal and its resistance $R_{tes}$ increases. The superconducting film voltage, $V_{tes}$, is held constant by a small shunting resistor in parallel, with resistance typically a factor of ten less than the operating TES resistance. The voltage bias translates a change in resistance to a change in current through the detector; this current is amplified and recorded as the detector response to the optical signal. This voltage bias scheme also provides electrothermal feedback. When optical power increases the TES resistance, the constant voltage bias guarantees that the TES bias power ($P_{bias} = V_{tes}^2 / R_{tes}$) decreases; therefore the total power on the island $P_{total} = P_{bias} + P_{opt}$ is constant and the detector stays roughly at the same operating point on its transition. The electro-thermal feedback described here is negative, making the detector response stable and faster than the intrinsic $C/G$ thermal time scale. Table 2.4 summarizes the physical dimensions, as well as typical electrical and thermal properties of an ABS TES bolometer.

### 2.2.6 Detector Readout

The current through the TES is amplified by a Superconducting Quantum Interference Device (SQUID) whose input coil is coupled to a small inductor in series with the TES. When the TES current changes the small inductor causes a change in magnetic flux through the input coil of the SQUID. The impedance of the SQUID is extremely sensitive to the magnetic flux through it. This impedance change is amplified by two more stages of SQUIDS at temperature below 4K and it is finally read out with 300K electronics. These last two stages are arranged such that 22 first stage squids can be multiplexed for every second and third stage SQUID. Multiplexing reduces the number of wires going to the cold stages, and therefore the associated heat load. A detailed review of the SQUID read out is given in Chapter 3.
2.3 Optics

2.3.1 Feedhorns

Each ABS polarimeter couples free space microwave radiation onto the OMT waveguide through a corrugated feedhorn with a full width half maximum (FWHM) beam of 16°. Corrugated feedhorn provide symmetric beams, small cross-polar response, and low side-lobes [21] [130]. Fabricating the small corrugations necessary for a 150 GHz feedhorns is a technical challenge. The machine shop at Princeton developed special tools to mass produce the ABS feedhorns in two sections. Figure 2.10 shows a side view of the top section with one quarter of the wall cut out. In total 240+ feedhorns were machined with greater than 96% yield. The feedhorn design was optimized for high transmission and single moded performance across the band using CCORHRN, a finite-element simulation software that solves Maxwell’s equations in a set of rings. A beam measured using a 145 GHz source is overplotted on a simulation of the final design in Figure 2.11. A simulation of the voltage standing wave ratio (VSWR) of the feedhorn versus frequency is shown in Figure 2.12. The
2.3 Optics

Figure 2.11: Beam profile of an ABS feedhorn at 145 GHz. The plot shows very good agreement between the measured and simulated beams. The beam width at −10 dB is 29° while the first sidelobes start at ± 20°. This is an E-plane beam profile; the H-plane beam profile is similar to 4%.

VSWR is the ratio between the voltage at a node and a voltage at an anti-node of standing waves in the feedhorn. A lack of standing waves and therefore maximum power transmission is indicated by a VSWR of 1. The reflection coefficient of the feedhorn $R$ is given by:

$$R = \frac{VSWR - 1}{VSWR + 1}. \quad (2.1)$$

The average power reflected across the target passband (127 GHz to 163 GHz) is 1.5%.
2.3.2 Telescope Mirrors

Light rays impinging at the same angle from the normal of the aperture are mapped onto the same positions in the focal plane of the ABS crossed Dragone telescope. This telescope design was chosen because of its low cross-polarization leakage [30] [31] [116] and large focal plane for a given reflector size. The primary and secondary reflectors are 57.1 cm and 58.5 cm in diameter and fit in the 4K volume of the cryostat. These reflectors form a focal plane big enough to house 240 feeds. Figure 2.13 shows a picture of the mirror focal plane setup, as well as a Code-V ray tracing diagram of the ABS reflector design with the location of the focal plane and sky aperture.

The cross-Dragonne telescope maps the 16° feedhorn beam on to the 25 cm aperture at the 4K stage. This mapping was explored using the DADRA program \(^1\). The intensity of the feedhorn beams at the edge of the aperture is \(\sim 7\) dB below their maximum, for detectors located near the middle of the focal plane. This edge taper translates to a 20% decrease in optical efficiency. The detectors located at the edge of the focal plane, may loose up to 33% efficiency due to the cold stop edge taper. DADRA simulations at 145 GHz including feedhorns, mirrors and stop predict detector beams on the sky with a FWHM angle of 0.68°, beam ellipticity of \(\sim 3\)%, and a cross polar response of \(\sim 1.5\)% [97].

2.3.3 Optical Filters

The cold stop holds the 4K filter stack composed of a 1" slab of PTFE and a 3/8" slab of Nylon, each of which is AR coated with Zitex [8]. The PTFE acts as an absorptive IR filter above 1 THz, while the Nylon absorption has a steep frequency dependence that cuts \(\sim 2\)% of power in band and more than 99.99% at 1 THz. The PTFE absorbs the majority of the IR power on the 4K, permitting the Nylon below to equilibrate to a lower temperature. Nylon inband emission is much higher than PTFE’s, therefore it is more important to cool Nylon as much as possible. A cryogenic test while observing from Cerro Toco showed that

\(^1\)Diffraction Analysis of a Dual Reflector Antenna, Rahmat-Samii, Y., Imbriale, W., & Galindo-Israel, V., YRS Associates
Figure 2.12: Plotted in green is the simulated VSWR for the ABS feed design, and in blue the atmospheric transmission. The VSWR within the band implies a 1.5% loss due to feedhorn reflections. The large spike at 165 GHz corresponds to the frequency where the second mode begins to propagate through the feedhorn. This mode is symmetric hence it deposits the same amount of power on both polarization channels. For the ABS detectors with a bandpass centered at 145 GHz [160 GHz], the second mode adds $\sim 10\% [\sim 19\%]$ the power of the first mode. The vertical black lines at 127 GHz and 163 GHz mark the edges of the targeted detector bandpass.
Figure 2.13: The left figure is a CodeV ray trace of the ABS telescope. Light rays originate from left to right, first passing through the 25 cm stop, then reflecting off the primary and secondary mirrors on to the focal plane. The picture on the right shows the ABS cryostat opened with the mirrors installed, but without the focal plane or the helium adsorption fridges. In the picture the aperture is located above the mirrors. The detectors will be mounted perpendicular to the rectangular slab protruding from the primary.

The middle of the PTFE reached a temperature of 16 K while the middle of the Nylon settled at 6 K.

A second 0.63" slab of AR coated PTFE is mounted on the 40K stage to reduce IR loading on to the 4K stage. Multiple single layer IR metal-mesh reflective filters are mounted above and below the PTFE [36]. As opposed to the absorptive filter these metal-mesh IR blockers radiate negligible amounts of power while reflecting away IR radiation. Mounted on the 300K vacuum shell above the 40K stage, is a 3.2 mm thick, 30 cm in diameter, ultra high molecular weight polyethylene (UHMWPE) vacuum window, which is AR coated with porous teflon.

Table 2.3.3 summarizes the reflection and absorptions losses due to the various optical filters between the bolometers and the sky signal.
Table 2.5: On-chip reflective losses are estimated from Truce collaboration measurements and simulations [78]. The stub filter absorptive loss is chosen to match the detector optical efficiency measurements described in chapter 5 (corresponds to an SiO$_2$ loss tangent of $\sim 0.01$). The feedhorn loss is derived from the VSWR simulation. The cold stop edge taper comes from DADRA simulations of feedhorns located in the middle of focal plane. PTFE and Nylon parameters are extracted from tan $\delta$ measurements by [48], [33],[49], and [69]. The frequency response between 300 GHz and 20 THz for a single IR-blocker was measured at NASA [36], and then extrapolated to 150 GHz. “IR blockers” represents the response of the twelve IR blockers positioned along the ABS optical path. The loss of the UHMWPE window was measured in [33]. The HWP losses correspond to its two optical axes [69], and include the RTduroid AR coating. The HWP reflective loss for the detectors with the shifted bandpass centered at 160 GHz is 0.029/0.035 Multiplying the transmission of all loss elements, predicts a total transmission of 30% (28% for those with the shifted bandpass).
2.3.4 Half Wave Plate

A half wave plate (HWP) suspended on a frictionless air bearing system is positioned on top of the vacuum shell above the window. The HWP is a 33 cm diameter disk of birefringent sapphire. Birefringent materials have different indices of refraction along their primary and secondary axes: for the ABS sapphire HWP the two indices are 3.07 and 3.40 [16]. The sapphire thickness is chosen such that 2 mm radiation aligned with the secondary axis is phase shifted by $180^\circ$ with respect to radiation aligned with the primary axis. The HWP is AR coated with RTduroid6002 ([101]) on both sides to improve transmission. This material was carefully chosen to have low emission in band since it operates at 300 K. Linearly polarized light traversing a rotating HWP is modulated at four times its rotational frequency. The electric field components of the polarized light are phase shifted by different amounts as the sapphire rotates, effectively reflecting the electric field vector about the HWP primary axis as shown in Figure 2.14. Modulating the polarized light at the aperture helps control instrument systematics. Any polarized light generated by the optical elements in the cryostat is not modulated by the HWP, and therefore can easily be rejected. Atmospheric and detector noise have a 1/f spectrum at low frequency; modulating the signal above this 1/f tail improves the sensitivity of the instrument. For more detail on the ABS optics and a complete treatment on the HWP non-ideal effects see ([36]).

2.4 Observing Strategy

ABS plans to target three fields: a large 4000 deg$^2$ patch that overlaps with the “southern galactic hole” a smaller 1200 deg$^2$ patch toward the north that overlaps the “northern galactic hole” and a third patch that targets the galactic center. The first two are expected to have low dust contamination since they point away from the plane of the galaxy. The patch shapes are determined by the large 400 deg$^2$ field of view of the detector array, the 10 deg peak-to-peak throw of the linear scanning motor, and the availability of the patches as they rise above the low elevation limit of the telescope ($\sim$35 deg). At least one of the
2.4 Observing Strategy

Figure 2.14: Above is a diagram depicting the modulation of a polarized signal by a rotating HWP. The electric field vector of the polarized signal is plotted in yellow, and is held constant in this example. The black line represents the direction of the primary axis of the HWP. The red arrow represents the electric field vector that reaches the detector after the polarized signal has transversed the HWP. The top diagram shows the resulting electric field on the detector when the angle between the polarized signal and the primary axis of the HWP is 0°, 45°, 90°, 135°, and 180°. Note that the electric field component of the incoming signal that is perpendicular to the HWP primary axis is phase shifted by 180° with respect to the parallel component, effectively reflecting it about the primary axis. The bottom figure plots the electric field amplitude on a polarimeter that is aligned vertically (same as incoming signal), as well as the square of this electric field, which represents the power signal measured by a bolometer. The figures show how rotating the HWP by 180° leads to two periods of power modulation, hence a HWP rotating at a frequency $f_{hwp}$ will lead to a modulated power signal at $4f_{hwp}$. 
Figure 2.15: Mollweide projection of FDS [38] sky dust model. The dusty regions near the galactic plane shows up in yellow and red. The three ABS fields are outlined with boxes. One overlaps the plane of the galaxy while the other two try to avoid it. The largest patch is 4000 deg$^2$ while the other two are 1200 deg$^2$. 
three patches is available 99% of the time, on top of which calibration sources, such as the moon, planets, or Tau A become available throughout the day. Figure 2.15 plots the locations of the fields over a simulated map of dust emission and the patch coordinates are given in Table 2.6. The ABS fields will overlap with observations made by QUIET and BICEP, and may also intersect with future measurements by BICEP2, PolarBear, ACTpol and others.

<table>
<thead>
<tr>
<th>Patch</th>
<th>Coordinates (RA, DEC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>North</td>
<td>180,-5</td>
</tr>
<tr>
<td>South</td>
<td>25,-42</td>
</tr>
<tr>
<td>Galactic</td>
<td>265,-29</td>
</tr>
</tbody>
</table>

Table 2.6: Coordinates of the patches targeted by ABS, in degrees.
Chapter 3

SQUID Multiplexing Readout

ABS takes advantage of DC Superconducting Quantum Interference Devices (SQUID) to amplify the detector signal. SQUIDs are composed of two Josephson junctions in parallel. Each junction consists of two superconducting electrodes separated by a narrow insulator, a normal metal, or a simple short. In 1962 Josephson [62] predicted that a zero voltage current would flow between the two electrodes of these junctions, and that this current would depend periodically on the phase difference between the electrodes’ Ginzburg-Landau wave functions [115]. The difference of the junction’s phases is sensitive to the magnetic flux through the loop that they form; therefore changes in magnetic flux cause changes in the zero voltage current. In the presence of an externally applied bias a more complete description is provided by the RCSJ (resistively and capacitively shunted junction) model [115], in which the ideal Josephson junction is shunted by a resistance and a capacitor. If the shunt capacitance of the junction is small, the SQUID acts as a voltage to flux transducer with a periodic response described by:

\[ V = \frac{R}{2} |I|^2 - [2I_c \cos(\pi \Phi / \Phi_0)]^2]^{1/2}, \]

(3.1)

where \( V \) is the voltage across the SQUID, \( R/2 \) the SQUID resistance, \( I \) the current through the SQUID, \( I_c \) the SQUID critical current, \( \Phi \) the magnetic flux through the SQUID loop, and \( \Phi_0 \) one magnetic flux quanta.\(^1\) Equation 3.1 predicts that the maximum voltage signal

\(^1\)One flux quanta equals \(2 \times 10^{-15} \text{ Wb}.\)
SQUID Multiplexing Readout

is obtained when the SQUID is biased at $I = 2I_c$.

The signal amplification from 300 mK to 300 K consists of three stages of SQUIDs one of which is multiplexed in the time domain. This SQUID time domain multiplexing (TDM) scheme was developed by NIST [58], and has been implemented with the bolometer arrays of SCUBA [54], ACT [5], BICEP2 [85], and the Keck Array [109]. The electronic readout system at room temperature was developed by the University of British Columbia (UBC) and is named the “Multi Channel Electronics” (MCE) [6]. At Princeton Norm Jarosick developed an analog electronics box to read out individual channels of the three-stage SQUID scheme. The ABS focal plane is read out with an MCE, while the SRDP and test dewar are read out with those analog electronics. Figure 3.1 shows a diagram of the NIST time domain SQUID multiplexing scheme for a two column by two row system.

For each detector there is a corresponding DC SQUID which measures the current through the detector by converting it into magnetic flux using a small coupling coil. This first stage SQUID (SQ1) is located on a micro-fabricated silicon chip operating at 300 mK, typically referred to as a ‘mux’ chip. ABS uses 24 NIST mux06A chips which contain 33 SQ1 channels. The set of stage one SQUIDs in a mux chip form what is typically referred to as a column. Each SQ1 has a set of bond pads located on the lower edge of the chip, that correspond to the SQUID bias, and a set of bond pads on the upper edge to connect the primary SQUID input coil ($\approx 20$ nH) to the bolometer circuit.

As in Figure 3.1, all channels have secondary input feedback coils connected in series across the mux to the so-called S1FB line. Current applied to this line leads to a magnetic flux signal on the SQ1 SQUIDs. The S1FB line is part of a feedback loop designed to apply flux opposite to the detector signal and hence keep the SQUID response linear. Variations in the applied S1FB feedback under standard operation are equivalent to variations in the detector response.

On the mux chip exists a second stage SQUID (SQ2) with its own feedback line (S2FB), whose input is coupled to the response of the 33 SQ1 channels. The SQ2 serves as an amplification stage between the SQ1 SQUIDs of a column and a series array (SA) located
Figure 3.1: Schematic of a two column by two row, time-domain SQUID multiplexing scheme from NIST. The shaded areas represent distinct TES detectors, whose signal is inductively coupled to a stage one SQUID (SQ1). Two SQ1 channels on the same column are transformer-coupled to a second stage SQUID (SQ2) read out by a SQUID series array (SA). SQ1s on the same row share a bias line that turns them on and off, such that at any given moment only one row per column is being read out. ABS has the same scheme except with 24 columns and 22 rows. (Figure courtesy W. B Doriese, previously published by Doriese, et al. (2007)[29]).
SQUID Multiplexing Readout

at 4K. The SA is an array of 100 SQUIDs that provides the final amplification of the signal before it is measured at room temperature. Each SA channel has its own feedback line (SAFB) used to tune the DC magnetic flux on the SA SQUIDS. The input coil of the SA is connected in series with the SQ2, and this circuit is voltage biased by a 0.1 Ω resistor. Because of this voltage bias, variations in the SQ2 impedance generate a current signal through the SA input coil.

Stage one SQUIDs are multiplexed: only one channel per chip is read at the same time. Multiplexing reduces the number of wires reaching the cold stages, hence decreasing the heat load on the cryogenics. Multiplexing with these mux chips permits having one S1FB line per 33 SQ1 channels, and current biasing one SQUID on each of multiple mux chips with one line. SQUIDs biased with the same electrical line form what is typically referred to as a row. Each ABS row consists of 24 SQUIDs. Twenty rows of SQUIDs couple to bolometers, and two rows couple to isolated SQUIDs, also referred to as dark SQUIDs. In total ABS multiplexes 528 SQUIDs to read out 480 bolometers and 48 dark SQUIDs.

3.1 SQUID Biasing

Initializing a three stage SQUID system requires setting five voltages ("biases") for each mux chip and one bias for each row. Therefore the ABS SQUID read out requires setting 142 biases, while for single detector tests in the SRDP or test dewar only six biases are set. The SQUID tuning begins by applying a bias to the SA and sweeping the SAFB current. Sweeping the SAFB exposes the typical SQUID sinusoidal response to magnetic flux changes, also referred to as a V-Φ. Each V-Φ period corresponds to a change in flux equivalent to a flux quanta. By exploring the SA bias range and observing the amplitude of the corresponding SA V-Φ, one can determine the bias value that maximizes the SA amplitude. This value corresponds to twice the SQUID critical current.

Since the SQ2 SQUID is read out through the SA, its biasing should be chosen after optimally tuning the SA. The SQ2 bias sets the amplitude of the SQ2 V-Φ when the S2FB bias is swept. The sinusoidal response of SQ2 is convoluted with the SA response, hence
Figure 3.2: The plots on the left show the V-Φ response of the SQUID stages in the NIST time domain multiplexing scheme. From top to bottom are the V-Φ of the SA, SQ2 servo, and SQ1 servo for one channel in the ABS array. The SA V-Φ is obtained by sweeping the SAFB and recording the SA output. The SQ2 servo V-Φ is obtained by sweeping the S2FB and locking the SA response at one point on its V-Φ through an SAFB feedback loop. In this case the recorded signal is the SAFB. Analogous to the SQ2 servo, the SQ1 servo is obtained by sweeping the S1FB, and running a feedback loop on the SQ2 stage, while recording S2FB. During typical data acquisition, a feedback loop servos the convolved V-Φ from the three stages through the S1FB, while using the SA output signal for the error signal upon which the feedback is based. The convolved V-Φ is shown in the plot on the right. To have an appropriate lock point in the middle of the convolved V-Φ, it is important to map the SQ1 V-Φ on a linear region of the SQ2 V-Φ as over-plotted in red and to map the SQ2 V-Φ on to a linear region of the SA V-Φ as over-plotted in green. This is achieved through an appropriate choice of SAFB and S2FB biases.
it is desired that the SAFB is chosen such that the SQ2 V-Φ falls on the linear region of the SA V-Φ as shown in Figure 3.2. The SQ2 bias is chosen such that the SQ2 amplitude is as large as possible without surpassing the SA amplitude. The SQ2 desired amplitude is less than the SA amplitude to avoid wrapping of the V-Φ. Wrapping occurs when part of a V-Φ is mapped onto a non-linear region of the previous stage V-Φ. The SQ2 V-Φ acquired through the MCE servo script will not wrap. This is a result of the MCE’s ability to run a feedback loop through the SAFB (not the usual S1FB) and hold the SA on one of its linear slopes. The signal in this case is the SAFB, hence the amplitude of this V-Φ is given in SAFB units.

Once the SA and the SQ2 are tuned, the SQ1 stage can be biased. Much like the SQ2, the SQ1 bias and S2FB are set to eliminate stage one wrapping and maximize V-Φ amplitude, with the added complication of having many SQUIDs per SQ1 bias line and S2FB line. An SQ1 bias may be chosen for each row of SQUIDs, hence for array assembly it is important to use mux chips with similar properties so that all stage one SQUIDs can operate close to optimally at one bias value. Although there is only one S2FB line for all SQ1s on a mux chip, the MCE can change the S2FB bias at the multiplexing rate, such that an optimal S2FB is applied to each SQ1 channel every time it is read out. This MCE feature is refer to as fast SQ2FB.

### 3.2 Multiplexing

The MCE can multiplex up to 41 channels, but the mux06 chips carry 33 SQUIDS, out of which only 25 are connected in an ABS Pod. Each pod carries 20 TES bolometers and 5 dark SQUIDs. To improve the multiplexing rate only two dark SQUIDs are read out. Hence ABS multiplexes 22 rows of SQUIDs.

The multiplexing begins by turning on the SQ1 bias of the first row, waiting a ‘row_len’ number of 50 MHz \( f_{\text{clock}} \) clock cycles, and then turning this first row off while biasing the next row. This cycle continuous 22 times before the first row is biased again. Therefore
3.2 Multiplexing

Each row is sampled at a multiplexing frequency \( f_{\text{mux}} \) of:

\[
f_{\text{mux}} = \frac{f_{\text{clock}}}{n_{\text{row}} \cdot \text{row\_len}},
\]

where \( n_{\text{row}} \) is the number of multiplexed rows. For ABS the \text{row\_len} was set to 150 clock cycles, which makes the ABS multiplexing rate \( f_{\text{mux}} \) equal to 15151 Hz. The \text{row\_len} parameter is chosen based on two factors: the bandwidth of the SQ1s which limits how fast they can be turned on/off, and aliasing of high frequency noise. A smaller \text{row\_len} reduces noise aliasing by increasing the multiplexing frequency, but if set too low, SQ1s will not have enough time to turn off, hence multiple channels will be read out simultaneously causing large correlations between detectors.

3.2.1 SQ1 Settling Time

The SQ1 settling time limits the multiplexing rate. The SQUID switch-on and switch off time constants have been measured to be less than 100 ns [24]. In ABS we expect the settling time instead to be dominated by the SQ2 bandwidth. The bias loop that couples the SQ2 to the series array forms an R/L circuit that slows the SQ2 response. The inductance of this loop is large because the SA at 4K is connected to the SQ2 at 300mK through a ten inch long twisted pair NbTi wire. The SQ2 dynamic impedance of a couple ohms together with an estimated SQ2 loop inductance of 0.33 μH lead to an f3db bandwidth of approximately 0.95 MHz.

The time required to switch SQ1 channels can be measured using the MCE raw data mode acquisition. The raw data mode samples the SQ2 response of one mux at 50 MHz. Figure 3.3 shows a set of raw MCE data with SQ1 multiplexing turned on. The fast sampling rate allows us to resolve the behavior of the mux chip during SQ1 switching. Each SQ1 row switch corresponds to a dip in the data stream, and the time it takes to settle sets the maximum multiplexing rate. A histogram of ABS SQ1 switching time delays for all columns and rows is shown in Figure 3.3. Some of the counts on the left hand side of the histogram come from misidentification of dips, usually very fast noise artifacts that the analysis code cannot distinguish from row switching. A few outliers above 110 clock
cycles are appropriately identified row switches. These long period dips do not repeat every 22 row_len cycles as expected if it were an anomalously slow SQ1, rather these outliers are likely due to a slowly drifting baseline that appears to the analysis algorithm as a continuation of the SQ1 switch on cycle.

The row_dly parameter sets the length of the idle period between row switches when no data is acquired. This time delay should be longer than the slowest SQ1 settling time in the array. For ABS row_dly is set to be 110 clock cycles, given that only three outliers fall above this threshold, and all are explained by artifacts in the data. The row_len is equal to row_dly plus the number of 50MHz samples sample_num measured every time a row is visited. For ABS sample_num is set to 40 clock cycles due to noise aliasing considerations addressed in the following sections. A trade off can be made to multiplex faster and permit a little more cross talk by lowering row_dly to 100 or even 90 clock cycles. ABS noise measurements show no clear improvement when making this trade off. The measured array median settling time of SQ1s is \( \sim 50 \) clock cycles, but depends strongly on the column. The distribution of settling times end at \( \sim 100 \) clock cycles. This end of the distribution is dominated by SQ1s on columns 0, 1, 2, and 17. The median of the distribution is consistent with the \( \sim 1 \) MHz bandwidth limit of the SQ2 loop estimated earlier in this section, while the high end of the distribution of settling times is a factor of two worse. The slow SQ1 channels may be related to increased inductance in the cryogenic wiring, tuning of the SQ2 biases, or fabrication related properties of each SQUID chip.

### 3.2.2 Noise Aliasing

Noise aliasing occurs when a signal is sampled with a Nyquist frequency \( f_{Nyq} \) lower than its bandwidth. The Nyquist frequency is defined as half the sampling frequency. For example consider a system where the signal is sampled every second. The highest frequency sine wave for which all peaks and troughs are sampled is 0.5 Hz. Sine waves with frequencies larger than 0.5 Hz are indistinguishable from corresponding sine waves at lower frequencies: see Figure 3.4. Similarly noise components above \( f_{Nyq} \) are indistinguishable from corresponding
3.2 Multiplexing

Figure 3.3: The top plot shows raw 50 MHz data from the MCE with SQ1 multiplexing on. A dip in data corresponds to switching SQ1 channels. The latency of SQ1 multiplexing is determined by first finding the lowest point of every dip (marked with a blue diamond on the plot), and then the two inflection points to the left and right (marked by red circles.). The latency is the number of clock cycles between inflection points. Notice how spurious noise can lead to misidentification of very short period dips, as seen in the middle of the plot. The bottom plot shows a histogram of the latency of dips for the ABS array. Short period dips are likely mis-identifications, while the few dips above 110 are due to artifacts in the data.
noise components in the zero to $f_{Nyq}$ range, hence noise throughout the signal bandwidth contributes to the measured noise in the Nyquist frequency range. This source of noise is referred to as aliasing. The noise power spectra $N_{total}$ at frequency $f$ when the bandwidth ($f_{bw}$) is larger than $f_{Nyq}$ may be expressed as:

$$N_{total}(f) = N_{bw}(f) + \sum_{i=1}^{k-1} [N_{bw}(2if_{Nyq} - f) + N_{bw}(2if_{Nyq} + f)] + N_{bw}(2kf_{Nyq} - f), \quad (3.3)$$

where $N_{bw}(f)$ is the noise power spectra at frequency $f$ with no aliasing, or equivalently the noise spectra when the Nyquist sampling frequency equals the signal bandwidth.

For simplicity Equation 3.3 makes the assumption that $f_{bw} = 2^{N_{folds}}f_{Nyq}$, where $N_{folds}$ is the exact integer number of times the signal bandwidth can be folded on to the sampling Nyquist bandwidth and then $k$ is defined in terms of $N_{folds}$ as $k = 2^{N_{folds}} - 1$. The previous assumption can be satisfied by padding with zeros the high frequency end of the non-aliased noise power spectrum. If the noise power spectrum is constant above the Nyquist frequency, equation 3.3 simplifies to:

$$N_{total}(f) = N_{bw}(f) + \sum_{i=1}^{k-1} [N_{white} + N_{white}] + N_{white}, \quad (3.4)$$

Equation 3.4 simplifies to:

$$N_{total}(f) = N_{bw}(f) + (2k - 1)N_{white}, \quad (3.5)$$

where $N_{white}$ is the constant noise power above the Nyquist frequency. An ABS detector signal, multiplexed through the MCE, suffers from two types of noise aliasing: detector noise aliasing, and SQUID noise aliasing.

**SQUID Noise Aliasing**

SQUIDs have large bandwidths; for example the SA at 4 K has an intrinsic bandwidth of at least 100 MHz [57]. The SA response is low pass filtered by the RC time constant of the wires from 4 K to 300 K. The SA dynamic resistance of $\sim 100 \Omega$ together with the wire capacitance of 100 pF imply an $f_{3db}$ of $\sim 10$ MHz. For ABS the read out cards have been modified to roll-off the SA signal at 1.3 MHz (see Figure 3.5). This roll-off together with a $sample\_num$ of 40 clock cycles (Nyquist frequency of 0.6125 MHz) defines a $\sim 1$ MHz bandwidth for
Figure 3.4: Consider sampling the green, blue, and red mode every second as denoted by the ‘o’, ‘x’, and ‘+’ markers. The frequency of the green mode matches the 0.5Hz sampling Nyquist frequency ($f_{nyq}$), the blue mode is 10% slower at 0.45Hz, and the red mode is 10% faster at 0.55Hz. By sampling every second we can determine the frequency of the blue and green modes, but we would not be able to distinguish the red mode from the blue. Similarly signal noise at frequencies above $f_{nyq}$ would map to frequencies below $f_{nyq}$ since they are indistinguishable, and this is what we call noise aliasing.
the SA read out. Each detector is sampled at 15 151 Hz, much slower than the SQUID bandwidth, therefore SQUID noise is aliased on to frequencies below $f_{Nyq} = 7575$ Hz. The SQUID noise power spectrum ($N_{SQ}$) is close to flat above and below $f_{Nyq}$, hence equation 3.5 can be used to calculate the total SQUID read out noise ($N_{\text{readout}}$) that includes aliasing. The SA bandwidth of $\sim 1$ MHz folds approximately seven times on to the frequency range below $f_{Nyq}$, making $k = 64$ and $N_{\text{readout}} = 128 N_{SQ}$.

Multiple SQUID channels in the ABS array are not connected to bolometers: these are called dark SQUIDs. Their noise power spectra near the signal band are white, and dominated by aliasing. The histogram in Figure 3.6 shows the distribution of dark SQUID power spectrum noise levels in the ABS array. Its median at $\sim 2 \times 10^{-20} \text{ A}^2 \text{ Hz}^{-1}$ is two orders of magnitude larger than the corresponding measurement in the SRDP (see high frequency noise data in Figure 3.7), where no aliasing occurs because its readout does not multiplex. This result is in agreement with the estimated seven fold SQUID noise aliasing contribution.

**Detector Noise Aliasing**

The ABS TES bolometers are responsive to optical signals modulated at frequencies up to a couple hundred hertz. On the other hand the detector noise bandwidth extends into the many kilohertz range. The detector bandwidth is set by the thermal characteristics of the TES island while its extended noise bandwidth is due to the Johnson noise from the TES film. The $f3db$ of this Johnson noise is determined by the R/L circuit formed by the TES operating resistance and the inductance of the TES loop. A typical operating resistance for an ABS TES is 2.5 mΩ, which is connected in series with a loop inductance $L$ that can be divided into the following three components:

$$L = L_{Nyq} + L_{SQ} + L_{tr},$$  \hspace{1cm} (3.6)

where $L_{Nyq}$ is a microfabricated inductor located on the shunt chip, $L_{SQ}$ is the inductance of the input coil to the SQUID, and $L_{tr}$ is the inductance of the traces and bonds.
Figure 3.5: Plotted above is the power spectrum of an ABS SA, sampled at 50 MHz with the MCE. The SQ1 multiplexing and fast SQ2FB bias switching are turned off. The SA signal is attenuated around 1 MHz by the MCE output amplifiers as shown in blue. The wide peak near this frequency appeared when the MCE amplifier bandwidth was reduced from its original value of 3 MHz. During multiplexing forty 50MHz-samples are summed at each channel visit, averaging out all noise above 0.61 MHz as plotted in green. The narrow line near 0.5 MHz may be RF pick up, or due to grounding between the MCE and its power supply.
Figure 3.6: The ABS array has multiple SQUIDs not connected to bolometers (dark SQUIDs), which provide a measure of the read out noise. The histogram above gathers the average noise power spectrum between 5 Hz and 15 Hz for this type of channels. Their median noise is $\sim 2 \times 10^{-20}$ A$^2$/Hz, at least a hundred times larger than the non-multiplexed SQUID noise. As expected, aliasing due to a much larger SQUID bandwidth than MCE multiplexing rate, dominates the white noise of dark SQUIDS in the ABS array.
connecting the TES to the shunt chip.

With no \( L_{Nyq} \) present, the TES loop inductance is between 40 nH to 80 nH, given that the \( L_{SQ} \) is \( \sim 20 \) nH and \( L_{tr} \) for ABS ranges from 20 nH to 60 nH. Under these conditions the f3db Johnson noise frequency occurs at \( \sim 8 \) kHz, which leads to \( \sim 50\% \) more detector noise in the signal band due to aliasing. To reduce noise aliasing an \( L_{Nyq} \) of 620 nH is added to the TES loop such that \( L \) is \( \sim 680 \) nH and the Johnson noise f3db frequency falls to \( \sim 600 \) Hz. With \( L_{Nyq} \) installed, aliasing of detector Johnson noise accounts for only \( \sim 1\% \) of the total noise in the signal band. Figure 3.7 shows the detector noise for an ABS detector, with and without a Nyquist inductor.

### 3.3 SQUID Readout

The SQUID read out chain can operate in open or closed loop mode once all three stages are biased. When running in open loop mode a change in SQ1 input signal becomes a voltage change across the SA. In this case large variations of input current will lead to a non-linear SQUID response that follows the V-Φ shape depicted in Figure 3.2. On the other hand, closed loop mode is designed to maintain the SQUID response linear. It consists of applying current on the S1FB line such that the SA voltage is maintained constant on the middle of the V-Φ curve, where the linear slope is greatest. The majority of data is acquired through closed loop mode, including observations, IV curves, bias steps, complex impedance and calibrated noise spectra. The open loop mode is useful for measuring V-Φ periods, tuning the three stages of SQUID parameters, and acquiring uncalibrated noise spectra that are not affected by the feedback bandwidth.

In closed loop mode the SQ1 input current is proportional to the S1FB feedback voltage, and the proportionality constant depends only on the room temperature S1FB bias resistor \( R_{fb} \) and the \( M_{ratio} \). The \( M_{ratio} \) is defined as:

\[
M_{ratio} \equiv \frac{M_{in}}{M_{fb}} = \frac{I_{fb}/\Phi_o}{I_{det}/\Phi_o}.
\]

(3.7)

Here, \( M_{in} \) and \( M_{fb} \) are the mutual inductances from the SQUID to the TES input coil and
SRDP Noise spectra of an ABS detector with and without the Nyquist inductor

Figure 3.7: Plotted above are noise power spectra acquired in the SRDP, with the TES operating at 20%, 50%, and 80% its normal resistance. Two sets of spectra were acquired, one with and one without a 620 nH Nyquist inductor installed in the TES loop. The TES loop by itself has an inductance of \( \sim 60 \) nH. A black vertical line marks the sampling frequency 15 kHz. Noise at frequencies near this line will be aliased into the signal band during MCE multiplexing. Without the Nyquist inductor the amount of power that would be aliased into the signal band is approximately 50% of the DC detector noise level. On the other hand with the Nyquist inductor this is reduced to a few percent.
S1FB coil respectively. $I_{fb}$ is the S1FB current, and $I_{det}$ the current through the detector. For ABS $M_{ratio} \approx 8.5$. The detector current is related to the MCE-supplied feedback voltage $V_{fb}$ by:

$$I_{det} = \frac{V_{fb}}{M_{ratio}R_{fb}}.$$  

(3.8)

The simple correspondence of equation 3.8 is correct up to a particular frequency determined by the bandwidth of the S1FB loop. The SRDP analog electronics has a 100 kHz bandwidth while the MCE is limited by the multiplexing frequency, the feedback gain, and the SQ1 slope, to $< 10$ kHz.

When using the MCE in closed loop mode, feedback is applied in discrete amounts every time the detector is sampled [51]. At time index $n$ feedback $f_n$ is applied and a SA error signal $e_n$ is measured. The error signal is equal to the difference between the measured SA voltage ($V_{SA}$) and the SA lock setpoint voltage ($V_{sp}$). The purpose of the feedback loop is to keep $V_{SA}$ close to $V_{sp}$, or equivalently keep $e_n$ close to zero. The SQ1 input signal $y_n$ in S1FB units may be expressed in terms of the error and feedback signals as:

$$y_n = \frac{e_n}{m} + f_n,$$  

(3.9)

where $m$ is the slope of the V-Φ response at the lock setpoint.

On the other hand the MCE servo applies a feedback based on the observed error in the previous frame. For a given servo gain $I$ we have

$$f_{n+1} = f_n + \frac{Ie_n}{4096},$$  

(3.10)

Let $m' = 4096/I$, and then equation 3.10 may be expressed as:

$$e_n = m'(f_{n+1} - f_n),$$  

(3.11)

were $m'$ carries the same units as the slope $m$.

To analyze the frequency response of the servo we take the discrete Fourier transform
of each signal \((e_n, f_n, y_n)\) on some interval of length \(N\) samples:

\[
E_k = \sum_{n=0}^{N-1} e^{-2\pi i kn/N} e_n; \quad (3.12)
\]

\[
Y_k = \sum_{n=0}^{N-1} e^{-2\pi i kn/N} y_n; \quad (3.13)
\]

\[
F_k = \sum_{n=0}^{N-1} e^{-2\pi i kn/N} f_n. \quad (3.14)
\]

Note that \(k\) represents a frequency \(f\) such that \(k/N = f/f_{\text{samp}}\) with \(f_{\text{samp}}\) the sampling rate. Re-writing equation 3.9 and 3.11 in Fourier space leads to:

\[
E_k = m(Y_k - F_k), \quad (3.15)
\]

\[
E_k = m'(F_k e^{-2\pi ik/N} - F_k), \quad (3.16)
\]

where \(k\) is an integer in \([-N/2, N/2]\). Substituting for the error signal in the above equations leads to the effective feedback gain of the servo loop:

\[
\frac{F_k}{Y_k} = [1 + g(e^{-i\alpha} - 1)]^{-1}, \quad (3.17)
\]

where:

\[
\alpha \equiv 2\pi k/N = 2\pi f/f_{\text{samp}}, \quad (3.18)
\]

\[
g \equiv \frac{m'}{m}. \quad (3.19)
\]

The ratio of the amplitudes between the feedback and the SQ1 input signal in the frequency domain is:

\[
\left| \frac{F_k}{Y_k} \right| = [1 + 4g(g - 1) \sin^2(\alpha/2)]^{-1/2}. \quad (3.20)
\]

A flat gain curve corresponds to \(m = m'\), and over- or underestimation of \(m'\) will suppress or inflate the signal at higher frequencies.

The feedback gain may be constant up to the sampling frequency if the \(I\) gain is chosen such that \(m\) and \(m'\) match for each channel. In practice a single \(I\) gain is chosen for each column, for the entire season. Since the SQUID V-\(\Phi\) slopes vary over a column and on a nightly basis, a flat feedback response is not possible for all channels. It is preferable to set
the $I$ gain low ($m' > m$) to attenuate high frequency noise that may be aliased into the signal band when the data are down sampled to 200 Hz. On the other hand if the $I$ gain is too low the frequency dependence of the feedback gain can affect the ABS band around 10 Hz. For ABS we set the $I$ gains following the same procedure used with ACT [51]. First we find the $I$ gain value that would generate a flat feedback response for each SQ1 channel in a column. This is done by measuring the slope each V-Φ curve at the lock point. Then we find the median of these $I$ gain values for each column and divide by 2, such that for most channels the feedback gain falls at high frequencies. Figure 3.8 plots $m'$ versus $m$ for of each ABS SQUID. Except for a few outliers the SQUID feedback loop is tuned to reduce high frequency noise ($m' > m$) while its gain decreases by less than 1% from DC to 10 Hz ($m > m'/35$).

3.4 MCE Butterworth Filter

The MCE samples each ABS channel at a rate of 15 151 Hz. If each sample for all detectors were saved, the array data rate would be $\sim 32$ Mb/s, eight times larger than the maximum 4 Mb/s MCE data transfer rate; hence the data timestreams are down sampled to 200 Hz. To prevent noise aliasing the data acquired at the sampling rate must be filtered before they are down sampled. The MCE uses a low pass 4-pole Butterworth filter designed in firmware that provides a flat signal response at low frequencies and reduces the signal amplitude by orders of magnitude at high frequencies. The low-pass filter is described by six MCE parameters ($b_{11}, b_{12}, b_{21}, b_{22}, k_1, k_2$) that can be modified to generate filters with different knee frequencies and gains. The transfer function for this filter is:

$$H(z) = \frac{1 + 2z^{-1} + z^{-2}}{1 + b_{11}z^{-1} + b_{12}z^{-2}} \frac{1 + 2z^{-1} + z^{-2}}{1 + b_{21}z^{-1} + b_{22}z^{-2}},$$  

(3.21)

where $z$ is the complex frequency. Figure 3.9 shows the amplitude and phase response of the ABS Butterworth filter. It was chosen to have a 60 Hz $f_{3db}$ frequency, which is far enough from our signal band at 10 Hz to reduce the signal gain by less than $-32$ db. This Butterworth filter permits a detector sampling rate of 200 Hz without adding any noticeable
Figure 3.8: Above is a scatter plot for all SQUIDs in the ABS array. The x-axis is $m'$, which is proportional to the inverse of the feedback I gain. The slope $m$ of the SQUID V-Φ curve at the lock point is plotted on the y-axis. Note that each column uses one term, hence all its SQUID channels have the same $m'$ value. The I gain terms for the ABS array were chosen to minimize high frequency noise, and still have a flat in-band response from the feedback loop. This can be seen in the plot by the fact that almost all detectors fall between the blue ($m = m'$) and the red ($m = \frac{m'}{35}$) lines, where the latter indicates $> 99\%$ flatness between 0 and 7.5 KHz.
Figure 3.9: Plotted above are the amplitude and phase of the ABS 4-pole Butterworth filter. The f3db knee of the filter is at 60 Hz, and its gain changes by less than $-32 \text{ db}$ between DC and 10 Hz. Noise aliasing from down sampling is less than 0.01% of the measured noise at 10 Hz due to the implementation of the Butterworth filter.

A low sampling rate leads to a smaller data set that is manageable in term of current hard drive storage capacity. Saving 200 samples per second for each detector translates to an MCE data rate of $\sim 0.42 \text{ Mb/s}$, which corresponds to 36 Gb a day or 13 Tb a year. Table 3.1 shows the parameters, DC gain, and f3db frequency for three different MCE filters including the one used in ABS.
### Table 3.1: Table of the parameters that characterize the digital Butterworth filters available to the MCE data acquisition.

- **b<sub>11</sub>** and **b<sub>21</sub>** should be multiplied by $-1$ before inserting in the filter Equation 3.9.

<table>
<thead>
<tr>
<th>Filter type</th>
<th>Parameters ($b_{11}, b_{12}, b_{21}, b_{22}, k_{1}, k_{2}$)</th>
<th>DC gain</th>
<th>Sampling Freq (Hz)</th>
<th>f3db (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABS</td>
<td>32022, 15648, 32449, 16075, 2, 13</td>
<td>1311</td>
<td>15151</td>
<td>60</td>
</tr>
<tr>
<td>ACT</td>
<td>32092, 15750, 31238, 14895, 0, 11</td>
<td>1218</td>
<td>15151</td>
<td>122.2</td>
</tr>
<tr>
<td>Other</td>
<td>32295, 15915, 32568, 16188, 3, 14</td>
<td>2048</td>
<td>30000</td>
<td>75</td>
</tr>
</tbody>
</table>
Chapter 4

Dark Detector Tests

The ABS TES bolometers measure the power of the impinging light. Their properties must be chosen carefully to minimize the detector noise given the expected power loading and modulation speed of the polarized signal. Many of the detector properties may be determined from testing the bolometers’ behavior in the dark, i.e. not loaded with optical power. Measurements of Johnson noise with the TES in the superconducting state yield the resistance of the shunt and the inductance of the TES loop. Johnson noise data at multiple bath temperatures can be used to find the critical temperature of the TES. Current versus voltage curves (I-V curves) may be analyzed to extract the TES normal resistance, its saturation power, the bias voltages to place it on the transition, and its responsivity. Multiple I-V curves at various bath temperatures can be used to estimate the thermal conductivity from the TES island to the bath. Noise power spectrum measurements calibrated with responsivities from I-Vs provide estimates of the detector’s noise equivalent power (NEP). Complex impedance measurements together with current noise spectra are used to constrain the internal thermal and electrical properties of the TES. Dark detector tests are critical for optimizing the TES design, and are also useful for quick detector screening before installation. With the limited number of detectors that fit in the focal plane it is of great importance to have as many working channels as possible.
4.1 Superconducting Johnson Noise

Johnson noise is electrical noise generated by thermal fluctuations of charge carriers. When measuring the current through a resistor $R$ at temperature $T$, Johnson noise shows up as current white noise fluctuation given by:

$$\text{Noise}[A/\sqrt{\text{Hz}}] = \sqrt{\frac{4k_B T}{R}},$$

(4.1)

where $k$ is Boltzmann’s constant. When the resistor is connected in series with an inductor $L$, the Johnson noise is attenuated by the circuit’s $R/L$ time constant. The power spectrum follows a one pole filter response described by:

$$\text{Noise}^2(f) = \frac{\text{Noise}_{dc}^2}{1 + (f/f_{3db})^2},$$

(4.2)

where $f$ is frequency in Hertz and $f_{3db} = R/L$.

4.1.1 Shunt and Nyquist Values

The TES electrical circuit simplifies to a shunt resistor in parallel with the TES resistance that is in series with an inductor as shown in Figure 4.1. This inductor represents the addition of a Nyquist inductor and the mux chip’s input inductance for the readout SQUID. When the TES is in the superconducting state and the bias current is off, the SQUID measures the current noise through the shunt (and any parasitic resistance in series with the TES) in series with the total inductance. Therefore the superconducting noise spectra may be fitted to Equation 4.2, to obtain the zero frequency noise level and the $f_{3db}$ frequency. If the temperature of the TES loop is known, then Equation 4.1 may be inverted to obtain the resistance value of the shunt, $R_{sh}$. The inductance of the loop may be deduced from the $f_{3db}$ frequency and $R_{sh}$; alternatively if the inductance of the loop is known a priori, then $R_{sh}$ can be obtained from the $f_{3db}$ frequency without knowing the temperature.

Measuring the shunt value through the DC Johnson noise level can be difficult because of three factors. First, any measurement of shunt Johnson noise includes a white spectrum of amplifier noise, which is included in the spectrum fit as a third parameter. The percentage error on the inferred resistance value scales with the ratio of the amplifier noise to...
4.1 Superconducting Johnson Noise

A bias current ($I_{bias}$), is applied to the TES loop, that has a shunt resistor ($R_{sh}$) in parallel with the TES resistor ($R_{tes}$). In series with $R_{tes}$ are two inductors: $L_{nyq}$ that is designed to limit the bandwidth of the detector’s Johnson noise, and $L_{SQ}$ that couples the current through the TES to the readout SQUID. Note that $R_{bias} \gg R_{tes}$ and $R_{tes} \gg R_{sh}$, and that $R_{bias}$ is a room temperature resistor.

Figure 4.1: A bias current ($I_{bias}$), is applied to the TES loop, that has a shunt resistor ($R_{sh}$) in parallel with the TES resistor ($R_{tes}$). In series with $R_{tes}$ are two inductors: $L_{nyq}$ that is designed to limit the bandwidth of the detector’s Johnson noise, and $L_{SQ}$ that couples the current through the TES to the readout SQUID. Note that $R_{bias} \gg R_{tes}$ and $R_{tes} \gg R_{sh}$, and that $R_{bias}$ is a room temperature resistor.

A second issue is the long integration time required for measuring the noise spectra below the f3db frequency, especially when the Nyquist inductor is large. Long data files required for low frequency data analysis are susceptible to jumps in the squid feedback loop or 1/f noise from drifts in the bath temperature. The third issue is knowing the temperature of the shunt which is one of the parameters required by equation 4.1. Accurately measuring the temperature of the shunts is difficult because typical thermometers can not be mounted on the small shunt chips and must be placed further away on larger structures, leaving the possibility of thermal gradients. To measure temperatures below 1K we use Ruthenium 0xide Thermometers (ROX) whose resistance have a strong temperature dependence. The resistance of the ROX is converted to a temperature following a resistance versus temperature calibration curve. Errors in the calibration curve propagate to our shunt estimates, as well as any temperature offset caused by self heating of the ROX. ROX self heating is caused by the current used to measure the resistance of the thermometer. This systematic can be mitigated by appropriate heat sinking of the ROX and decreasing the...
readout current as low as possible without suffering from electrical readout noise.

On the other hand, measuring the shunt value using the f3db frequency of the Johnson noise instead of the DC noise level exchanges the requirement of an accurate temperature value, for an accurate TES loop inductance value. The true temperature of the shunt can vary depending on the cryogenics, and ROX setup, while the inductance of the TES loop is expected to be the same for all tests. The known size and number of loops of the Nyquist inductor and SQUID input coil can be used to produce good estimates of the achieved inductance; harder to predict is the inductance of the traces of the TES loop.

Table 4.1 contains the average shunt and inductance values of all pairs of MUX and shunt chips used to populate the ABS pods. These values were extracted from Johnson noise measurements made with the SRDP on a test board that connected all 32 channels of SQUID and shunts, and shorted their outputs to simulate a closed TES loop. The inductance of the traces in this setup is the same for all channels, as opposed to an ABS pod where the traces connecting the shunts to the detectors have different lengths and therefore different inductances. These SRDP tests were also used to pick the best 20 SQUID-Shunt channels in each chip set. The selected channels are later connected to the 20 TES detectors in a pod.

Two types of shunt estimates are included in Table 4.1. Both calculate $R_{sh}$ for each channel based on the f3db frequency of the Johnson noise and then average across all good channels. They differ in the inductance value used to extract the shunt resistance from the f3db frequency. The first method uses the average of the measured inductances across a chip set. Each channel’s inductance estimate depends on the temperature of the stage, the DC Johnson noise level, and the f3db frequency. The second method uses the inductance value calculated from the geometrical design of the Nyquist inductor and SQUID input inductor.

4.1.2 $T_c$ from Johnson Noise Measurements

Johnson noise measurements at various temperatures are an effective method for obtaining the transition temperature $T_c$ of a TES. The normal resistance of the TES ($R_n$) is at
### 4.1 Superconducting Johnson Noise

Pod | Column | Chip-set | Wafer | $L_{avg}$ (nH) | $R_{sh}(\mu\Omega)$ at $L_{avg}$ | $R_{sh}(\mu\Omega)$ at $L=645$ nH | Error (%) |
---|---|---|---|---|---|---|---|
1 | 8 | 5 | 1 | 678 | 153 | 146 | 3 |
2 | 6 | 2 | 1 | 637 | 148 | 149 | 3 |
3 | 1 | 20 | 1 | 629 | 153 | 157 | 4 |
4 | 23 | 16 | 1 | 677 | 156 | 149 | 4 |
5 | 2 | 22 | 1 | 640 | 152 | 153 | 3 |
6 | 4 | 19 | 1 | 639 | 150 | 151 | 3 |
7 | 11 | new1 | 2 | 671 | 177 | 170 | 4 |
8 | 13 | 3 | 2 | 614 | 162 | 170 | 3 |
9 | 12 | 6 | 1 | 636 | 146 | 148 | 3 |
10 | 15 | 15 | 1 | 699 | 166 | 153 | 4 |
11 | 9 | 12 | 1 | 634 | 141 | 144 | 4 |
12 | 10 | 18 | 1 | 619 | 153 | 160 | 4 |
13 | 21 | 17b | 2 | 727 | 186 | 165 | 6 |
14 | 22 | 14 | 1 | 625 | 145 | 150 | 3 |
15 | 20 | 4 | 1 | 663 | 153 | 149 | 3 |
16 | 19 | 1 | 1 | 624 | 143 | 148 | 3 |
17 | 16 | 17 | 1 | 675 | 169 | 161 | 4 |
18 | 17 | 8 | 1 | 642 | 145 | 146 | 3 |
19 | 18 | 9 | 1 | 607 | 138 | 147 | 4 |
20 | 0 | 10 | 1 | 645 | 148 | 148 | 3 |
21 | 3 | 11 | 1 | 606 | 139 | 147 | 4 |
22 | 5 | new2 | 2 | 676 | 176 | 168 | 5 |
23 | 7 | 21 | 1 | 657 | 168 | 165 | 4 |
24 | 14 | 24 | 2 | 613 | 164 | 172 | 3 |

Table 4.1: Table of average shunt value for each ABS pod. Two shunt wafers were used in ABS. The first yielded shunts resistances of $\approx 150\mu\Omega$ and the second resistances of $\approx 170\mu\Omega$. The average inductance across a pod from fitting Johnson noise data ($L_{avg}$) is susceptible to systematic errors in the test temperature (primarily due to gradients in the heat sinking of the 300mK test board), noise lines contaminating the spectra, and the SQUID white noise level. The expected SQUID input inductance is 20 nH while the expected ABS Nyquist inductance is 620 nH. Since the TES loops are shorted during these tests, the trace inductance is small ($<10$ nH). The expected total inductance is 645 nH, assuming a trace inductance of 5 nH. The statistical error is derived from the Johnson noise fit of individual channels.
Figure 4.2: The noise spectra above belong to an ABS prototype detector, and were acquired in the SRDP. No current bias was applied to the TES during these measurements. The detectors were driven superconducting and normal by setting the bath temperature below and above $T_c$. Plotted in cyan and yellow are the Johnson noise measurements when the TES is superconducting, and normal, respectively. When the TES is superconducting the TES loop resistance equals the shunt resistance, which is far smaller than the normal resistance of the TES. Johnson noise is inversely proportional to the resistance, hence in the superconducting case it is much larger than in the normal case. $T_c$ may be determined by looking for the sharp change in Johnson noise spectra as the bath temperature is varied slowly from below to above $T_c$. 
least an order of magnitude greater than $R_{sh}$, therefore the Johnson DC noise level and f3db frequency change by an order of magnitude when the temperature is raised from below to above $T_c$. By observing this Johnson noise transition, $T_c$ may be determined to a few mK. The accuracy of this method is limited by the temperature width of the TES superconducting transition and the thermometer calibration. Figure 4.2 shows the dramatic difference between superconducting and normal Johnson noise.

### 4.2 I-V Curves

Current versus voltage curves are constructed by driving the TES normal by increasing the bias current that flows through the parallel circuit of the shunt and TES, and then decreasing that current in small steps. When the TES is normal most of the bias current flows through the shunt, hence applying a constant voltage across the TES. As the bias current is stepped down, the TES current measured with the SQUID decreases. This regime of the I-V is called the normal branch, and is shown in red in Figure 4.3. The point marked with a green circle in Figure 4.3 is the turn around point where the TES begins to move on to the transition. On the transition the TES behaves opposite to a resistor; when the voltage across it decreases the current through it increases. At the point marked with a yellow circle in Figure 4.3 the TES has gone superconducting. After this point all the bias current is flowing through the superconducting TES and not the shunt, this sometimes causes the SQUID to unlock because changes in bias current cause large changes in magnetic flux through its input coil that the lock feedback loop cannot compensate.

The SQUID amplifiers measure changes in current. To get the absolute current through the TES an arbitrary offset must be subtracted from the feedback readout signal. This arbitrary offset can be determined in multiple ways. The simplest one is to use the normal branch of the I-V. The current through a normal resistor equals zero when the voltage is zero, therefore we can fit a line to the normal branch and find the point were the fitted line intersects the current axis (y-axis). The current value at this intersection is exactly the arbitrary offset that must be subtracted. The same procedure works for the superconducting
Figure 4.3: The ABS I-V curve on the left was acquired with the SRDP. The TES voltage ($V_{tes}$) is plotted on the x-axis and the TES current ($I_{tes}$) is plotted on the y-axis. The standard I-V acquisition starts by setting $V_{tes}$ high and then decreasing it slowly until it reaches zero while measuring $I_{tes}$. When $V_{tes}$ is high the TES behaves as a normal resistor; this region of the I-V curve is called the normal branch. The normal resistance of the TES ($R_n$) is equal to the inverse of the normal branch slope. The green point marks where the TES enters the superconducting transition and starts to behave opposite to a normal resistor, with current increasing as voltage decreases. The yellow point marks the end of the TES transition, where the TES has become completely superconducting. The region after this point is called the superconducting branch. Note how the superconducting branch does not intersect the origin as expected in the ideal case. This is the result of the readout’s inability to keep up with the large changes in $I_{tes}$ when the TES is superconducting. On the right is a power versus resistance curve derived from the I-V. Note the uniformity of the normal resistance, and the narrow width of the transition in power units.
Figure 4.4: The I-V curve on the left was acquired with the SRDP following the third I-V calibration method described in the text. It belongs to an ABS prototype detector whose normal resistance varies with $V_{\text{tes}}$. This variation is clear on the resistance versus power plot on the right. Using the constant normal resistance calibration on this I-V, would yield the wrong calibration offset as shown by the red dashed line. With the I-V already calibrated, the normal branch fit intersects the y-axis above the origin, hence implying that $I_{\text{tes}}$ is smaller than the true value. A wrong current offset has dramatic effects on the shape of the R-P curve. If the offset from the normal branch calibration were applied, the R-P curve would fall entirely outside the range of the current plot.
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branch, but it is less reliable because of SQUID vphi jumps.

The normal branch of the TES may not have a constant resistance as shown in Figure 4.4. In the case where the calibration of the superconducting branch is not feasible, like in the SRDP system, there is a third, more complicated, method for obtaining the I-V offset. Begin by raising the temperature above $T_c$ to drive the TES normal, then lock the SQUIDs with no TES bias applied. At this point the readout signal corresponds to zero current. Next increase the current slowly to prevent vphi jumps, up to a current that will keep the TES normal at any bath temperature. Decrease the bath temperature below $T_c$ and wait for it to settle before stepping down the bias current. This method determines the zero current point at the beginning of the I-V acquisition, and then keeps track of current changes about this point.

Analysis of I-V curves yields a wealth of information derived from the measured current through the TES ($I_{tes}$) when applying a particular bias current $I_{bias}$. The voltage across the TES ($V_{tes}$) is related to the bias and TES currents by:

$$V_{tes} = (I_{bias} - I_{tes})R_{sh}, \quad (4.3)$$

where $R_{sh}$ is known from Johnson noise measurements. The power dissipated on the TES ($P_{tes}$) and the resistance of the TES ($R_{tes}$) are related to the bias and TES currents by:

$$P_{tes} = I_{tes}V_{tes} = (I_{tes}I_{bias} - I_{tes}^2)R_{sh}, \quad R_{tes} = \frac{V_{tes}}{I_{tes}} = \left(\frac{I_{bias}}{I_{tes}} - 1\right)R_{sh}. \quad (4.4)$$

At high bias currents the TES resistance is equal to the resistance extracted from the normal branch of the I-V. Once the bias current is low enough that the TES is on its superconducting transition, the TES resistance becomes a percentage of the normal resistance ($R_n$) defined as

$$%R_n = \frac{R_{tes}}{R_n}. \quad (4.4)$$

This is a useful parameter to track, since it shows where on the transition the detector is biased. For example, low on the transition ($%R_n < 0.3$) the detector is fast but more likely to oscillate, high on the transition ($%R_n > 0.7$) the opposite is true, while the middle of the transition ($0.3 < %R_n < 0.7$) tends to be the optimal operating region.

To limit the number of bias lines going to the 300mK focal plane, groups of three pods (60 TES bolometers) share one bias line. An optimal current is chosen for each bias line, such
that all its detectors are placed as close as possible to a target \( R_n \) \((%R_{\text{target}} \approx 0.5)\). This is achieved by analyzing the I-V for each detector in a group, finding the bias current that would place it at \( %R_{\text{target}} \), and then choosing the median of all such bias currents in a detector group.

The TES island operates at a critical temperature \( T_c \), while being thermally isolated from the colder bath by thin legs. To maintain this thermal equilibrium state, power must be dissipated on the TES island and flow through the legs to the bath. The amount of power \( P \) required to keep the island at \( T_c \) with the bath at temperature \( T_{\text{bath}} \) is modeled as:

\[
P = \kappa(T^n_c - T^n_{\text{bath}}),
\]

where \( \kappa \) and \( n \) are thermal conductivity constants that depend on the geometry and material composition of the TES legs.

There are four sources of power on the ABS TES island: the TES bias power (\( P_{\text{tes}} \)), the optical signal (\( P_{\text{opt}} \)) dissipated through the lossy microstrip, the heater bias power (\( P_{\text{heater}} \)), and the direct coupling of stray radiation on to the pixel (\( P_{\text{stray}} \)). \( P_{\text{heater}} \) is typically zero because the heaters are off or disconnected. \( P_{\text{stray}} \) for ABS is small, because its coupling efficiency to the TES island is low. Furthermore stray radiation near the TES pixel is low intensity since it comes from within the 4K cavity or it has been attenuated by the Nylon filter, Teflon filters, IR blockers, and/or the high pass waveguide cutoff at 110 GHz. This accounting of power on the TES leads to the following equation for \( P \):

\[
P = P_{\text{tes}} + P_{\text{opt}}.
\]

For I-Vs taken in the dark the optical power is zero, hence the TES power only depends on the thermal properties of the island. In this case the thermal parameters \( \kappa \), \( n \), and \( T_c \) are extracted by measuring \( P_{\text{tes}} \) at various \( T_{\text{bath}} \) temperatures. Figure 4.5 shows \( P_{\text{tes}} \) versus \( T_{\text{bath}} \) data as well as a fit following Equation 4.5.

Another quantity that is extracted from dark I-V curves is the saturation power of the detector (\( P_{\text{sat}} \)). The saturation power is the maximum power that can be dissipated on the
Figure 4.5: The data points in red are TES saturation powers extracted from I-Vs of an ABS1-4 detector at different bath temperatures. They are fitted to Equation 4.5 with \( n = 2.8 \), to extract \( \kappa \) and \( T_c \). \( G \) is computed from Equation 4.7. The values quoted on the legend are typical of ABS detectors.
TES island before its resistance becomes normal. In the ideal case where the transition is infinitely sharp and the critical temperature is single valued, \( P_{sat} = P \). A more complete model of the TES includes a reduction of \( T_c \) and \( P_{sat} \) lower on the transition due to the increasing TES current. In this case the power on the TES is not constant along the transition, hence the saturation power must be defined differently. For the analysis done here \( P_{sat} \) was defined as the power at 95\% of \( R_n \). This value was chosen because higher on the transition the current through the TES is smaller, and \( P_{tes} \) does not vary much between 95\% and 99\% of \( R_n \). Measuring the saturation power with dark I-V data is important for determining how the detector will operate in the field. If the expected optical loading from the sky is larger than the saturation power, the detector will always be normal and not work. On the other hand if the saturation power is too high the detector may be unstable and will have worse noise performance.

The parameters \( \kappa, n, \) and \( T_c \) can be used to calculate the thermal conductivity \( G \) of the TES legs, given by:

\[
G = \frac{dP}{dT} \bigg|_{T_c} = n \kappa T_c^{n-1}.
\]

The value of \( G \) is directly related to the expected thermal noise of the detector (see section 4.3), and its thermal time constant.

### 4.2.1 I-V Curve Responsivity

One of the most important results from I-V analysis is responsivity. The responsivity \( (S) \) of a detector is the ratio of the signal, a change in optical loading \( \left( dP_{opt} \right) \), to the measurement, a change in the current through the TES \( \left( dI_{tes} \right) \), while holding the bias current \( \left( I_{bias} \right) \) constant as is typically the case during observations:

\[
S = \frac{dP_{opt}}{dI_{tes}} \bigg|_{I_{bias}}.
\]

Equation 4.6 and 4.8 imply:

\[
S_{dc} = \frac{dP}{dI_{tes}} \bigg|_{I_{bias}} - \frac{dP_{tes}}{dI_{tes}} \bigg|_{I_{bias}}.
\]
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In the ideal case where $P$ is constant across the transition; the first term on the right hand side of Equation 4.9 is zero, and using Equation 4.4 to substitute for $P_{tes}$ leads to:

$$S = \left. \frac{dP_{tes}}{dI_{tes}} \right|_{I_{bias}} = (I_{bias} - 2I_{tes})R_{sh}. \quad (4.10)$$

For ABS $P$ is not constant across the transition, it depend on $I_{tes}$ but not directly on $I_{bias}$, hence:

$$\left. \frac{dP}{dI_{tes}} \right|_{I_{bias}} = \left. \frac{dP}{dI_{tes}} \right|_{I_{tes}}. \quad (4.11)$$

Note $dP = dP_{tes} + dP_{opt}$, so $\frac{dP}{dI_{tes}} = \frac{dP_{tes}}{dI_{tes}} + \frac{dP_{opt}}{dI_{tes}}$ and equation 4.9 becomes:

$$S = \left. \frac{dP_{tes}}{dI_{tes}} - \frac{dP_{tes}}{dI_{tes}} \right|_{I_{bias}}. \quad (4.12)$$

The derivative of $P_{tes}$ with respect to $I_{tes}$ starting from Equation 4.4 is:

$$\frac{dP_{tes}}{dI_{tes}} = (I_{bias} + I_{tes} \frac{dI_{bias}}{dI_{tes}} - 2I_{tes})R_{sh}. \quad (4.13)$$

Equations 4.10, 4.12, and 4.13 lead to:

$$S = R_{sh}I_{tes}\frac{dI_{bias}}{dI_{tes}}. \quad (4.14)$$

The slope of I-V curves may be used to calculate $\frac{dI_{bias}}{dI_{tes}}$. Equation 4.14 can be checked by comparing its prediction to the responsivity estimate using heaters. This check is done by acquiring two I-V curves from the same detector, one with the heaters on (adding $\sim 0.5pW$ of power) and one with the heaters off. The amount of heater power $dP_{heater}$ is the difference between the $P_{tes}$ powers from the two I-V curves evaluated at the same $I_{tes}$ current. To obtain the responsivity we divide the heater power by the change in TES current $dI_{tes}$ between the two I-V curves at the same $I_{bias}$ value. Figure 4.6 plots the heater responsivity, the responsivity from Equation 4.14 using the I-V with heaters off, and the responsivity from the ideal case of Equation 4.10.

4.3 Noise Spectra and NEP

Once the bias current is set after analyzing an I-V, the detector is ready for observation. In the case of dark tests it is only sensitive to intrinsic thermal and electrical fluctuations.
4.3 Noise Spectra and NEP

Figure 4.6: The plot above includes the ideal responsivity (red crosses) when the saturation power is constant across the TES transition, the responsivity estimate using the heater (blue circles), and the responsivity estimate from Equation 4.14 using a single I-V curve (green triangles). The ideal responsivity estimate only depends on $R_{sh}$, $I_{tes}$, and $I_{bias}$ as described by Equation 4.10. The responsivity from heaters may be extracted through two I-V curves one with the heater on and the other with it off. These I-V curves yield the ratio of the heater power loading to the change in $I_{tes}$ at a particular $I_{bias}$. The single I-V responsivity depends on $R_{sh}$, $I_{tes}$ and the slope of the $I_{bias}$ versus $I_{tes}$ curve as described in Equation 4.14. The single I-V estimate agrees with the more robust heater measurements. In the field ABS relies on single I-V curves for calibration, since the heaters were disconnected to reduce RF pickup.
Noise spectra of these fluctuations provide key characteristics of the detector. At low frequencies, where the signal band is placed, thermal noise dominates. At high frequencies close to the TES R/L time constant, detector Johnson noise dominates. High frequency noise may affect the signal band through aliasing if the Nyquist inductor is too small or the multiplexing rate is too slow. Noise pick up at particular frequencies show up as lines in the noise spectra measurement. These lines originate from electrical grounding issues (60 Hz), RF pickup on the wires, magnetic pickup by SQUIDS, or thermal fluctuations driven by mechanical vibrations.

Noise power spectra are obtained by computing the square absolute value of the Fourier transform of the TES current time series. They are functions of frequency typically reported in units of $A^2/\text{Hz}$ or $A/\sqrt{\text{Hz}}$. Consider an $I_{\text{tes}}$ time series of $N$ samples acquired at a rate of $f_{\text{samp}}$. The discrete Fourier transform $\mathcal{F}$ of the time stream at frequency $f = f_{\text{samp}}k/N$ where $k$ is an integer between $-N/2 < k < N/2$, is defined as:

$$\mathcal{F}(I_{\text{tes}})_f = \frac{N-1}{\sum_{n=0}^{N-1} I_{\text{tes}n} e^{-i2\pi f_{\text{samp}} n}},$$

and the noise power spectra $\text{Noise}(I_{\text{tes}})[A^2/\text{Hz}]$ is given by:

$$\text{Noise}(I_{\text{tes}})[A^2/\text{Hz}] = \frac{\mathcal{F}(I_{\text{tes}})_f \mathcal{F}^*(I_{\text{tes}})_f}{N f_{\text{samp}}}. \quad (4.16)$$

Figure 4.7 shows the detector noise when biased at 80%, 50% and 20% of $R_n$. High on the transition the current noise is small, while it gets much larger as you move lower on the transition. Larger current noise does not imply worse detector performance, since the responsivity also increases when biasing lower on the transition. The correct figure of merit is the Noise Equivalent Power (NEP) defined as the noise divided by the responsivity:

$$\text{NEP}[W/\sqrt{\text{Hz}}] = \frac{\text{Noise}(I_{\text{tes}})[A/\sqrt{\text{Hz}}]}{S[A/W]}. \quad (4.17)$$

The detector G-noise contribution to the NEP ($\text{NEP}_{G}$) at low frequencies is determined by its thermal properties, $T_c$, $G$, and $F_{\text{link}}$ through:

$$\text{NEP}_{G}[A/\sqrt{\text{Hz}}] = \sqrt{4F_{\text{link}}kGT_c^2}. \quad (4.18)$$
Figure 4.7: Plotted above are current noise spectra of one ABS detector biased on the transition at 20\% R_n (blue), 50\% R_n (green), and 80\% R_n (red). Notice how the current noise increases as you move lower on the transition, to a lower TES resistance. This is analogous to Johnson noise, where lower resistance leads to a higher noise level. The increase in noise at a few hundred Hertz is generally due to the internal thermal architecture of the TES (see also section 4.5).
The factor $F_{\text{link}}$, is between 0.5 and 1, depending on the temperature gradient across $G$ and the details of the heat transfer (specular or diffuse). ABS pixels are expected to have specular heat transfer and hence an $F_{\text{link}}$ factor given by:

$$F_{\text{link}} = \frac{1 + (T_{\text{bath}}/T_c)^{n+1}}{2} \approx 0.6. \quad (4.19)$$

Table 4.2 on page 100 shows results from a dark detector test carried in our SRDP. It lists current noise, responsivity and NEP measurements, as well as predicted NEP from $G$, $T_c$, and $F_{\text{link}}$ estimates. The NEP measurements are higher than the expectations by about 20\% for wafers 4, 14, and 15, likely due to underestimating $F_{\text{link}}$. (If $F_{\text{link}} = 1$ then the NEPs are consistent). The wafer 11 detectors have lower $R_n$, and appear to be faster and more unstable than the other wafers. We could not acquire in the SRDP noise data or I-Vs below 50\% $R_n$ because the wafer 11 detectors would fall off the transition and become superconducting. In chapter 5 we find that wafer 11 NEPs improves substantially in the field, when optical loading helps stabilize the electro-thermal feedback loop. The NEPs appear to decrease at lower $R_n$, in agreement with the hypothesis that $T_c$ is suppressed lower on the transition where $I_{\text{tes}}$ is much larger.

### 4.4 Heater Responsivity

Each TES pixel has a heater, through which power can be dissipated, mimicking an optical signal. These heaters were connected for dark tests in the SRDP but disconnected in the ABS cryostat to reduce RF pickup. By modulating the input power to the heater we can probe the response of the detector at various frequencies. To carry out this measurement we used the analog electronics of the SRDP together with the sweep sine utility of an HP3562A spectrum analyzer. The sweep sine output goes to the heater and one of the input channels of the spectrum analyzer. The second input channel reads the detector response. The spectrum analyzer then finds at each frequency the phase difference and relative amplitude of the two input signals. This relative amplitude is the responsivity of the detector as a function of frequency.
Figure 4.8: Plotted above is the responsivity of one ABS detector at 20, 50, and 80 percent of \( R_n \). Quoted on the legend are the f3db frequencies at the three set points. In general the detectors are fastest biased in the middle of the transition while they become much slower high in the transition. The responsivities were measured by applying a constant voltage plus a sine wave at various frequencies to the heater on the TES island, and recording the TES response.
For this measurement to work a DC voltage offset must also be applied to the heater. In the case with no DC offset the swept sine output applies a sine modulation in voltage, that the heater converts to a sine squared modulation in power signal. The measurement would then fail because the detector response is modulated at twice the heater input signal \((2\sin^2(x) = 1 - \cos(2x))\). By adding a DC offset to the sweep sine voltage the heater power is modulated by both a sine and a sine squared term \(((1 + \sin(x))^2 = 1 + 2\sin(x) + \sin^2(x))\). Figure 4.8 shows the responsivity as a function of frequency at three locations on the TES transition. The highest detector bandwidth is achieved in the middle of the transition. The median 3db frequencies from heater responsivities for wafers 4 and 11 are found in Table 4.2

4.5 Complex Impedance

Complex impedance is a mathematical construct to deal with amplitudes and phases of AC electrical signals using imaginary numbers. The complex impedance \((Z)\) at some angular frequency \((\omega)\), of some element is equal to the ratio of the voltage \(V(\omega)\) across it divided by the current \(I(\omega)\) that flows through it. Here \(V(\omega)\) and \(I(\omega)\) are the Fourier transforms of the time dependent voltage and current. These complex quantities depends on frequency and carry both amplitude and phase information. For example, for a simple resistor the complex impedance \(Z_{\text{res}}\) is real because the voltage across it is in phase with its current, but for a capacitor the voltage across it lags the current through it by \(\pi/2\), so that its complex impedance is purely imaginary. The TES impedance when the temperature is above \(T_c\) is that of a simple resistor, but when the TES is on transition its complex impedance is determined by the interaction between the electrical circuit and the thermal architecture (see Figure 4.9). Information is extracted from \(Z(\omega)\) data by comparing it to a TES electro-thermal model motivated by the layout of the pixel’s components. This model is considered successful if it fits the data with a small number of free parameters.

The complex impedance measurement on the SRDP is done with an HP3562A spectrum analyzer. A schematic of the electronic test setup is shown in Figure 4.10. The spectrum
4.5 Complex Impedance

Figure 4.9: Figure on the left shows the Thevenin equivalent circuit of the TES circuit, that corresponds to Figure 4.1 when stray impedances are negligible. $V_{th}$ is the Thevenin equivalent voltage, $Z_{th}$ the Thevenin equivalent impedance, $Z_{tes}$ the TES impedance, and $L$ the sum of the Nyquist inductor, SQUID input coil, and trace inductance. The figure on the right depicts the thermal architecture of a two block model, where the photon power is dissipated on the island of heat capacity $C_1$, and the readout bias power is dissipated on the TES of heat capacity $C_2$. 
Figure 4.10: Above is an electrical diagram of the complex impedance measurement. It includes the TES bias circuit on the left, the SQUID readout in the middle, and the spectrum analyzer on the right driving the TES bias and reading in the TES current. Figure adapted from [131]
4.5 Complex Impedance

The analyzer drives $V_{\text{bias}}$ with sine waves and at the same time reads out the resulting $I_{\text{tes}}$ signal (converted to a voltage by the warm-temperature SRDP electronics). At each frequency it calculates the real and imaginary values of the voltage ratio $dV_{\text{in}}/dV_{\text{out}}$ where $dV_{\text{out}} \approx dI_{\text{tes}}R_{fb}M_{\text{ratio}}$ (inaccurate at high frequencies where the gain of the readout loop decreases) and $dV_{\text{in}} = dV_{\text{bias}} \approx dI_{\text{bias}}R_{\text{detb}}$. $R_{\text{detb}}$ and $R_{fb}$ are resistors specific to the SRDP readout electronics and have been measured to better than a tenth of a percent. The $M_{\text{ratio}}$ is the constant factor that converts TES current to the measured feedback current, and has been calibrated to better than one percent.

A good understanding of the transfer function for the TES electrical readout and bias circuits is necessary to extract $Z_{\text{tes}}$. The ideal circuit in Figure 4.1 may not be an adequate model because of possible stray capacitance or stray inductance in the TES loop [131]. Thankfully the transfer function may be determined by taking complex impedance data with the TES in both the normal and superconducting states.

The total impedance $Z_{\text{total}}$ of the Thevenin equivalent circuit shown in Figure 4.9 is the sum of:

$$ Z_{\text{total}} = Z_{\text{tes}} + Z_{\text{th}} + i\omega L, \quad (4.20) $$

where $Z_{\text{th}}$ is the Thevenin impedance. If the ideal circuit of Figure 4.1 is correct, and no stray capacitances or inductances exist, then:

$$ Z_{\text{th}} = \frac{R_{sh}R_{\text{detb}}}{R_{sh} + R_{\text{detb}}} \approx R_{sh}. \quad (4.21) $$

The complex impedance measurement is related to $Z_{\text{total}}$ by:

$$ Z_{\text{total}} = \frac{dV_{\text{th}}}{dI_{\text{tes}}} = \frac{dV_{\text{in}}}{dV_{\text{out}}}C(f), \quad (4.22) $$

where $C(f)$ is a function of frequency. In the case of having ideal bias and readout circuits, $C(f)$ is equal to:

$$ C(f) = \frac{R_{sh}}{R_{sh} + R_{\text{detb}}R_{fb}M_{\text{ratio}}}. \quad (4.23) $$

The function $C(f)$ encapsulates the conversion from the measured impedance to $Z_{\text{total}}$ including any frequency dependence of the bias and readout circuits. From now forward it
is assumed that $C(f)$ is the same for all complex impedance measurements made with the SRDP setup. This means that $C(f)$ may be obtained by comparing complex impedance measurements with the TES normal ($dV_{in}^{nc}/dV_{out}$), superconducting ($dV_{in}^{sc}/dV_{out}$), or in the transition ($dV_{in}^{tr}/dV_{out}$). When the TES is superconducting $Z_{tes} = 0$, and Equation 4.20 becomes:

$$Z_{total} = C(f) \frac{dV_{in}^{sc}}{dV_{out}} = Z_{th} + i\omega L.$$  \hspace{1cm} (4.24)

When the TES is normal its impedance is constant and equal to $R_n$. Equation 4.20 then becomes:

$$Z_{total} = C(f) \frac{dV_{in}^{nc}}{dV_{out}} = R_n + Z_{th} + i\omega L.$$ \hspace{1cm} (4.25)

With the TES on the transition Equation 4.20 translates to:

$$Z_{total} = C(f) \frac{dV_{in}^{tr}}{dV_{out}} = Z_{tes} + Z_{th} + i\omega L.$$ \hspace{1cm} (4.26)

Equations 4.24, 4.25, and 4.26, can be combined to solve for $Z_{tes}$ in terms of the measured impedances and $R_n$:

$$Z_{tes} = R_n \left( \frac{dV_{in}^{nc}}{dV_{out}} - \frac{dV_{in}^{sc}}{dV_{out}} \right)^{-1} \left( \frac{dV_{in}^{tr}}{dV_{out}} - \frac{dV_{in}^{sc}}{dV_{out}} \right).$$ \hspace{1cm} (4.27)

The extracted $Z_{tes}$ data are fitted to the two block thermal model in Figure 4.9. This model is motivated by the physical layout of the bolometer, where the TES bilayer sits in the middle of the TES island. The bilayer heat capacity $C_2$ is thermally decoupled from the island’s heat capacity $C_1$ through a thermal conductivity $G_2$, the island is thermally linked to the bath through a thermal conductivity $G_{bath}$, and the electrical power is dissipated on $C_2$ while the optical power is dissipated on $C_1$. Note that the thermal conductivity $G$ from the bilayer to the bath satisfies $1/G = 1/G_2 + 1/G_{bath}$.

The two block model can be described with: one equation for the Thevenin equivalent circuit of the TES, and two equations for the TES thermal block architecture shown in Figure 4.9. These three differential equations are coupled and can be solved as ordinary equations in the frequency domain by making a small signal approximation [40]. The whole
system is described by $M \Delta X = \Delta Y$ where $M$ is the matrix:

$$
M = \begin{pmatrix}
i \omega L + R_{\text{tes}}(1 + \beta) + R_{\text{sh}} & L_0 G/I_{\text{tes}} & 0 \\
-I_{\text{tes}} R_{\text{tes}}(2 + \beta) & i \omega C_2 - L_0 G + G_2 & -G_2 \\
0 & -G_2 & i \omega C_1 + G_{\text{bath}} + G_2
\end{pmatrix}; \quad (4.28)
$$

$\Delta X$ is the vector:

$$
\Delta \begin{pmatrix} I_{\text{tes}} \\ T_2 \\ T_1 \end{pmatrix} 
$$

and $\Delta Y$ is the vector:

$$
\Delta \begin{pmatrix} V_{\text{th}} \\ P_2 \\ P_1 \end{pmatrix} 
$$

Here $\omega$ is angular frequency, $\Delta V_{\text{th}}$ a change in Thevenin voltage, $\Delta P_1$ and $\Delta P_2$ are changes in power dissipated on the thermal blocks, $\Delta T_1$ and $\Delta T_2$ are changes in temperature of the thermal blocks, and $\Delta I_{\text{tes}}$ is a change in TES current. The parameter $\beta$ is defined as $(I_{\text{tes}}/R_{\text{tes}})(\delta R_{\text{tes}}/\delta I_{\text{tes}})$, while $L_0$ is the loop gain and is defined as:

$$
L_0 = \frac{I^2_{\text{tes}} R_{\text{tes}} \alpha}{G T_o}, \quad (4.31)
$$

where $T_o$ is the temperature of the TES on transition and $\alpha = (T_o/R_{\text{tes}})(\delta R_{\text{tes}}/\delta T)$.

By inverting $M$ we can predict the responsivity and complex impedance of the TES ([37]). The responsivity is given by the matrix element $M^{-1}_{12}$ and the complex impedance is given by:

$$
Z_{\text{tes}} = (M^{-1}_{11})^{-1} - R_{\text{sh}} - i \omega L. \quad (4.32)
$$

4.5.1 MCMC

A Markov Chain Monte Carlo (MCMC) implementation to explore high dimensional probability distributions was developed in collaboration with P. Mosteiro, T. Morrell, and X. Du. One of its goals was to fit complex impedance models to data, and extract model parameters. MCMC methods were first introduced in the 1950s [80] to sample an unknown
probability distribution efficiently. They are described in detail in [82] and in [41]. Instead of calculating the probability density at sites on a regular grid spanning the entire parameter space, one draws samples sequentially according to a probabilistic algorithm. Briefly one begins the algorithm at one set of parameters $a, b, ...$ and a cost $f(a, b, ...).$ From a specified interval one takes a random step to a new set of parameters $a', b', ...$ which generates a new cost $f(a', b', ...).$ This new state is accepted with a probability of

$$P(a, b, ...) \rightarrow (a', b', ...)) = \begin{cases} 1 & f(a', b', ...) < f(a, b, ...) \\ e^{-f(a', b', ...) - f(a, b, ...)} / T & \text{if } f(a', b', ...) > f(a, b, ...) \end{cases}$$

where $T$ is the temperature of the simulation. Hence a state of lower cost will always be accepted, while a state of higher cost will be accepted depending on the temperature of the simulation. The sequence of states visited forms a Markov chain. Rather than scaling exponentially with the number of parameters varied, the time needed to sample a distribution using MCMC grows approximately linearly with dimension.

The MCMC implementation provides a flexible framework where various probability distributions can be explored. The user provides data and a model, and the MCMC algorithm explores the parameter space of the model and returns the values that minimize some cost function. A default cost function is defined, but the user has the option of providing his own. The implementation allows for the user to specify a temperature schedule to more accurately determine the minimum. In addition to the location of the minimum energy parameters, the implementation provides standard deviations and the model evaluated at the computed parameters.

Applying this MCMC implementation to complex impedance data requires translating the model from the previous section into a C++ class that feeds into the MCMC driver program. To reduce the number of free parameters, the heat capacities and thermal conductivities are required to remain constant at all points on the transition. Typically, $Z$ data are acquired at three different setpoints, and then concurrently fitted to the model. The $Z$ model for the three setpoints contains 9 static parameters, 9 free parameters, and the matrix form of $M$ given in Equation 4.28. The static parameters include $R_{tes}$ and $I_{tes}$ at
4.5 Complex Impedance

Figure 4.11: The top figure shows the Markov chain for each of the free parameters in the model. Notice that after the chain stabilizes, the internal thermal conductivity $G_2$ (from the TES to the island) is more than an order of magnitude larger than the usual thermal conductivity to the bath of $100\text{pW} \text{K}^{-1}$, and that the TES heat capacity $C_2$ is much smaller than the $\sim 1\text{pJ} \text{K}^{-1}$ heat capacity of the island. We see $L_g$ and $\beta$ increase as the TES operating resistance decreases. The bottom plot shows the measured complex impedance data, as well as the fitted model at three setpoints. The cyan crosses mark the impedance data point at 100 Hz.
each setpoint, and $R_{sh}$, $L$, and $G$, all of which are extracted from I-V curves and Johnson noise measurements. The free parameters that will be extracted are: $\beta$, and $L_g$ at each setpoint, and $C_1$, $C_2$, and $G_2$. Figure 4.11 shows the Markov chain of the fitted parameters as well as the best fit model result over plotted with the data.

4.6 Dark Test Result Summary

Initially single detector SRDP tests were used to probe the performance of detector wafers, but eventually all pods of detectors were screened. For some of the pods, I-Vs, Noise, and Heater responsivity data sets were acquired during screening. The measured dark detector parameters are grouped in terms of fabrication wafer and summarized in Table 4.2. Results quoted in the table are the median of the available detector measurements for a wafer. For wafers 14 and 15 the heater lines were never connected, hence the $f3db$ frequency was not measured. Wafer 11 has a lower normal TES resistance, that translates into faster less stable detectors. The instability prevented the measurement of $0.2R_n$ noise. Another side effect appears to be worse NEP performance.

The thermal conductivity increased substantially after the fabrication of wafer 4, possibly related to variations in the SiN furnace deposition. The general trend of NEP decreasing as the detector is biased lower on the transition may be due to two factors: lower on the transition the detector noise increases and makes the amplifier noise less important, and/or the TES current increases effectively reducing $T_c$, one of the parameters of NEP$_G$. Complex impedance data were only acquired for pod 15, which is composed of wafer 4 detectors. The parameters extracted from complex impedance: $C_1$, $C_2$, $G_2$, $\beta$ and $\alpha$, are Pod averages from [125] and are summarized in Table 4.3. The total heat capacity implied by the $Z$ model is four times larger than the bulk estimate, given the geometry and composition of the TES [4]. This excess heat capacity hints at two-level systems in the amorphous SiN and/or SiO$_2$ layers.

Dark detector testing was critical in developing the ABS TES bolometer design. The low normal resistance of the TES reduces the amplifier noise contribution to the NEP, which
is important when multiplexing because of aliasing. The TES constant normal resistance permits straightforward calibration of I-Vs, and hence responsivity. The achieved $T_c$ values were close to the 500 mK target, while the achieved $G$ was higher than hoped for the last three wafers. The speed of the detectors is adequate compared to the HWP 10 Hz signal modulation. The measured detector NET is higher than that estimated from $G$ and $T_c$, especially for Wafer 11, but is still comparable to the photon shot noise of the atmosphere and the optics, described in the following chapter. The screening of SQUID, shunt, and detectors chips through dark tests, led to an ABS detector array with 458 operating TES bolometers out of a maximum of 480.
### Table 4.2: All available dark SRDP measurements are grouped based on the detector fabrication wafer. For each group, the median value of extracted dark TES parameter is recorded in the table above. Detectors fabricated on the same silicon wafer are expected to have similar properties, and similar issues, that may be addressed in future fabrications.

This table holds the median values of dark TES parameters across each wafer. I-V measurements yield $R_n$, $P_{sat}$, $T_c$, $\kappa$, $G$, NEP$_G$, and $S$. For $G$ measurements we assumed $n = 2.8$, while $F_{link}$ was estimated to be 0.6 from Equation 4.19 using $T_c$ and $T_{bath}$. Noise spectra yield Noise$_{0.2R_n}$ at 10 Hz, that together with I-V responsivity $S_{0.2R_n}$ leads to NEP$_{0.2R_n}$ at 10 Hz. Heater responsivity measurements are fitted to extract the f3db frequency of the detector response. The errors quotes are the standard deviation of all the detector measurements.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Wafer 4</th>
<th>Wafer 11</th>
<th>Wafer 14</th>
<th>Wafer 15</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_n$ (mΩ)</td>
<td>5.2 ± 0.14</td>
<td>3.6 ± 0.09</td>
<td>4.9 ± 0.3</td>
<td>5.0 ± 0.07</td>
</tr>
<tr>
<td>$P_{sat}$ at 0.33 K (pW)</td>
<td>11 ± 2</td>
<td>15 ± 2</td>
<td>16 ± 3</td>
<td>17 ± 2</td>
</tr>
<tr>
<td>$T_c$ (K)</td>
<td>0.510 ± 12</td>
<td>0.513 ± 32</td>
<td>0.520 ± 21</td>
<td>0.518 ± 17</td>
</tr>
<tr>
<td>$\kappa$ (pW/K$^n$)</td>
<td>107 ± 16</td>
<td>141 ± 25</td>
<td>145 ± 19</td>
<td>149 ± 21</td>
</tr>
<tr>
<td>$G$ at $T_c$ (pW/K)</td>
<td>88 ± 15</td>
<td>117 ± 17</td>
<td>123 ± 17</td>
<td>127 ± 15</td>
</tr>
<tr>
<td>NEP$_G$ ($10^{-17}$ W/√Hz)</td>
<td>2.7 ± 0.25</td>
<td>3.2 ± 0.21</td>
<td>3.3 ± 0.26</td>
<td>3.3 ± 0.21</td>
</tr>
<tr>
<td>Noise$_{0.2R_n}$ at 10 Hz ($10^{-10}$ A/√Hz)</td>
<td>6.2</td>
<td>-</td>
<td>3.6</td>
<td>4.2</td>
</tr>
<tr>
<td>Noise$_{0.5R_n}$ at 10 Hz ($10^{-10}$ A/√Hz)</td>
<td>3.0</td>
<td>5.8</td>
<td>2.2</td>
<td>2.5</td>
</tr>
<tr>
<td>Noise$_{0.8R_n}$ at 10 Hz ($10^{-10}$ A/√Hz)</td>
<td>2.0</td>
<td>3.4</td>
<td>1.6</td>
<td>1.7</td>
</tr>
<tr>
<td>$S_{0.2R_n}$ ($10^7$ A/W)</td>
<td>1.6</td>
<td>-</td>
<td>0.9</td>
<td>1.0</td>
</tr>
<tr>
<td>$S_{0.5R_n}$ ($10^7$ A/W)</td>
<td>0.8</td>
<td>1.0</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>$S_{0.8R_n}$ ($10^7$ A/W)</td>
<td>0.5</td>
<td>0.5</td>
<td>0.3</td>
<td>0.3</td>
</tr>
<tr>
<td>NEP$_{0.2R_n}$ at 10 Hz ($10^{-17}$ W/√Hz)</td>
<td>3.3</td>
<td>-</td>
<td>4.0</td>
<td>4.2</td>
</tr>
<tr>
<td>NEP$_{0.5R_n}$ at 10 Hz ($10^{-17}$ W/√Hz)</td>
<td>3.4</td>
<td>5.5</td>
<td>4.4</td>
<td>4.7</td>
</tr>
<tr>
<td>NEP$_{0.8R_n}$ at 10 Hz ($10^{-17}$ W/√Hz)</td>
<td>3.9</td>
<td>6.5</td>
<td>4.9</td>
<td>5.4</td>
</tr>
<tr>
<td>f3db$_{0.2R_n}$ (Hz)</td>
<td>40 ± 22</td>
<td>67 ± 29</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>f3db$_{0.5R_n}$ (Hz)</td>
<td>56 ± 12</td>
<td>69 ± 22</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>f3db$_{0.8R_n}$ (Hz)</td>
<td>39 ± 7</td>
<td>53 ± 17</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>
Table 4.3: These complex impedance results are extracted from measurements of Pod15 detectors (wafer 4) [125]. The data is fitted to a two-block TES model yielding heat capacities $C_1$, $C_2$, the internal thermal conductivity $G_2$, and the TES film parameters $\alpha$, and $\beta$ at $0.2R_n$, $0.5R_n$, and $0.8R_n$. 

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Wafer 4</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_1$ (pJ/K)</td>
<td>1.31 ± 0.03</td>
</tr>
<tr>
<td>$C_2$ (pJ/K)</td>
<td>0.26 ± 0.01</td>
</tr>
<tr>
<td>$G_2$ (pW/K)</td>
<td>6370 ± 200</td>
</tr>
<tr>
<td>$\alpha^{0.2R_n}$</td>
<td>117 ± 12</td>
</tr>
<tr>
<td>$\alpha^{0.5R_n}$</td>
<td>46 ± 1.2</td>
</tr>
<tr>
<td>$\alpha^{0.8R_n}$</td>
<td>21 ± 0.4</td>
</tr>
<tr>
<td>$\beta^{0.2R_n}$</td>
<td>4.3 ± 0.41</td>
</tr>
<tr>
<td>$\beta^{0.5R_n}$</td>
<td>0.96 ± 0.05</td>
</tr>
<tr>
<td>$\beta^{0.8R_n}$</td>
<td>0.22 ± 0.01</td>
</tr>
</tbody>
</table>
Chapter 5

Optical Detector Tests

The optical performance of ABS detectors was characterized through measurements of their bandpass, optical efficiency, beam response, and cross polar response. Beam and bandpass measurements were carried out in the “test dewar” and in the ABS cryostat using a neutral density filter with have the array installed, while cross polar leakage and optical efficiency were measured with the SRDP cold load. In the field, sky dip data provide estimates of optical efficiency, intensity to polarization leakage, and NEPs for the entire detector array. The HWP motion induces the second largest detector response after atmospheric fluctuations. This systematic is explored in section 5.4. The last section discusses estimates of the detector sensitivity derived from previous optical and dark test results, and compares them to the estimates extracted from sky dip measurements.

5.1 Detector Bandpass

A Fourier Transform Spectrometer (FTS) was utilized to measure the range of microwave frequencies that the ABS polarimeters detect; this frequency range is referred to as the detector bandpass. An FTS consists of a version of a Michelson interferometer, whose output can be Fourier transformed to obtain the frequency spectrum of the detected radiation. The Princeton FTS is a Martin-Puplett interferometer (MPI) designed to operate from 80 GHz to 3 THz with a resolution of \( \sim 1 \text{ GHz} \). A schematic of its design is shown in Figure
Figure 5.1: Diagram of a Martin-Puplett FTS. The light source is on the right, the adjustable mirror on the left, the static mirror on the top, the detector on the bottom, and the beam splitter in the middle. The path of a light ray emanating from the source is depicted by a dashed line. The adjustable mirror moves away and towards the beam splitter through a linear stage that precisely controls its position.
5.1. To understand how the MPI operates consider a beam of unpolarized incoherent light emanating from a blackbody source. This beam enters the FTS through a grid of closely spaced wires (wiregrid) that polarizes the light vertically. The polarized beam then encounters a beam splitter oriented at a 45° angle from the direction of the incoming beam. The beam splitter is a horizontal wiregrid rotated 54.7° vertically, such that its projection on the plane perpendicular to the incoming beam is at 45°. Half the light, now polarized at 45°, continues straight towards the adjustable arm while the other half, polarized at −45°, is reflected toward the static arm located at a 90° angle from the incoming beam. The two beams are reflected by roof top mirrors that invert polarization about their vertical axis, hence 45° polarization reflects to −45°, and vice versa. The beam returned from the adjustable arm is polarized at −45°, hence the beam splitter reflects it to the opposite side from the static arm, where the detector is placed. The beam returned from the static arm, now polarized at 45°, passes straight through the beam splitter toward the detector. When the distances between the beam splitter and the two arms are the same, the electric fields of the two beams will be in phase and the resulting polarization at the detector will be vertical. When the adjustable arm is moved such that the electric fields are out of phase by 180°, the resulting polarization will be horizontal. The modulation of the polarization angle as the adjustable arm moves will cause polarimeters aligned vertically and horizontally at the detector location to measure an interference pattern. Polarimeters aligned at 45° and −45° will not measure an interference pattern since they are only sensitive to light from either the fixed or the adjustable arm but not both. Adding a vertical polarizer at the output of the FTS converts polarization modulation into intensity modulation of vertically polarized light. In this case a polarimeter at any angle other than horizontal, or a bolometer with no polarization sensitivity, detects the intensity modulation of the vertically polarized light that leads to an interference pattern.

The FTS source is a black body at ∼ 1000 K and the ABS detectors are single moded, hence the expected power at frequencies close to the ABS band is constant. Since the source provides a flat spectrum, the measured interference pattern is determined by the detector
stub filters and any optical element between the FTS output and the detector feedhorn. An example of a measured ABS detector interferogram is shown in Figure 5.2. The y-axis is the measured intensity while the x-axis marks the location of the adjustable arm as it moves away from the beam splitter. Converting the interferogram to a frequency bandpass requires a few data processing steps.

- First, the data are fitted to a 4th order polynomial and the result subtracted to eliminate any drifts.

- Second, the white light point is found. This is the data point of maximum intensity about which the interferogram is symmetric. It corresponds to the FTS configuration where the adjustable and the static arms are at the same distance from the beam splitter. The ABS FTS setup has a white light point near the 12 cm mark of the 60 cm linear stage that translates the rooftop mirror of the adjustable arm. Most measurements extend the linear stage past 24 cm, hence the resulting interferogram is asymmetric about the white light point, since it has more data to its right.

- Third, the data are symmetrized following one of the following methods: the data to the left of the white light point are replaced by a reflected copy of the data to its right; the missing data to the left is completed with a copy of the data to the right; or excess data on the right is removed until the interferogram is symmetric. The last method greatly reduces the frequency resolution of the resulting bandpass, because many data points are lost. Caution must be taken when interpreting the accuracy of the resulting bandpass when data are copied from the right to the left of the interferogram. For example, when the interferogram is symmetrized by copying the right side over the left, it is guaranteed that the imaginary part of the derived bandpass is exactly zero, while if the full symmetric interferogram were measured, errors would propagate into the imaginary component.

- Fourth, the symmetrized interferogram must be translated such that its white light point is at coordinate zero. This translation is equivalent to a phase shift in Fourier
Figure 5.2: Interferogram from an ABS detector (wafer 15). The top plot shows the measured interferogram after subtracting a 4th degree polynomial fit. The average value of the raw interferogram is non zero, and decreases slightly as the adjustable mirror moves further away due to diverging rays. This effect together with moisture build up in the cryostat window, leads to long period drifts in the data that are subtracted by the fit. The x-axis of the first plot is in units of absolute distance of the linear servo motor that drives the adjustable mirror. In the bottom plot, the interferogram has been shifted such that white light point is at zero (red). The interferogram is symmetrized by assuming the missing data on the left (green) is the reflection of the data on the right.
Optical Detector Tests

space, given by:

\[
F(\text{shifted interferogram})_k = F(\text{raw interferogram})_k e^{(\pi i d_{\text{white}} k)/(d_{\text{step}} N)},
\]

(5.1)

for \(0 < k < N\), where \(N\) is the number of data points in the interferogram, \(d_{\text{step}}\) is the distance the adjustable arm of the FTS moves between each data point, \(d_{\text{white}}\) is the location of the white light point, in distance units from the linear stage encoder, \(F\) refers to a Fourier transform.

- Fifth, since the interferogram is finite, it must be apodized appropriately to reduce ringing when computing its Fourier transform. A Welch window is chosen to apodize the data in the analysis below.

- Finally, the bandpass is the real part of the Fourier transform of the apodized, symmetrized, and translated interferogram. The standard deviation of the imaginary part about zero is an estimate of the error of the measurement.

The amplitude of the resulting bandpass is arbitrary, and may be normalized by different methods. A common method is to set the largest response data point equal to one; another is to make the area under the bandpass equal to the square of that same area such that it satisfies:

\[
\frac{\int B(\nu) d\nu}{\int B(\nu)^2 d\nu} = 1,
\]

(5.2)

where \(\nu\) is frequency and \(B(\nu)\) the normalized bandpass. Optical efficiency data, explored in the following section, may be used to calibrate the bandpass into absolute transmission units.

Figure 5.3 shows bandpass measurements made with the test dewar for detectors from wafers 4, 11, 10, and 15. Bandpass measurements were also made in the ABS cryostat with a neutral density filter in the Princeton high bay, and the results are summarized in Figure 5.4. The bandpasses for wafer 4 and 11 are centered appropriately at 145 GHz with a low frequency cut off at 127 GHz, and a high frequency cut off at 162 GHz. The bandpass of wafer 10 (not used in ABS) is centered low at 129 GHz with cut-offs at 110 GHz and 148 GHz,
Figure 5.3: Above are the measured absolute bandpasses for ABS wafers 4, 10, 11, 15. Optical efficiency measurements were used to calibrate the bandpass transmissions. The measurements for wafer 4 and 11 have lower signal to noise, while the wafer 15 bandpass has lower frequency resolution. Wafers 10 and 15 miss the 145 GHz center frequency target. All bandpasses show a gradual roll off at the high frequency range. It is not clear if it is related to the FTS setup, or the detectors.
while the bandpass of wafer 15 is centered high at 160 GHz with cut-offs at 138 GHz and 182 GHz. The fabrication issues that lead to the shifts in the bandpass centers are explored in the following chapter. The aggregate spectra in Figure 5.4 are similar to the simulated CMB5 bandpass, hinting that the in-band wiggles and slopes of single FTS measurements are due to the experimental setup (for example, non-uniform illumination of the detector beam by the FTS output, or unaccounted frequency dependence of the intervening optical filters), and are not intrinsic to the detector response.

FTS measurements of early prototype detectors showed that the TES pixel could couple radiation at frequencies higher than the target bandpass [13]. This source of optical power on the detectors is referred to as the “blue leak.” Its coupling efficiency is low, but its bandwidth is large as shown in Figure 5.5. The “blue leak” was mitigated in ABS by adding an absorbing moat above the TES pixels, and through high frequency attenuation provided by the free space filters, in particular the Nylon low pass. The ABS array has a few detectors that respond weakly to sky dips (see section 5.5). Assuming the blue leak affects all detectors pointing at the same elevation equally, then the signal observed by these low efficiency detectors, corrected by elevation, provides an upper limit on the blue leak amplitude. The blue leak should not generate polarized signals since it is not modulated by the HWP. The measured median blue leak contributing to the sky dip signal (data acquired 04/15/2012) for detectors in wafers 4, 11, 14, and 15 is 10%, 10%, 17%, and 10% of the total signal respectively.

5.2 Detector Optical Efficiency and Crosspolar Isolation

The optical efficiencies of a few ABS detectors were measured with a polarized cold load. The cold load (CL) and ABS pixels were enclosed in the SRDP. The ABS pixel is housed in a brass module so that it couples to a rectangular waveguide through a standard flange. The cold load consists of a waveguide section filled with absorbing material with a conical space carved out in the middle to reduce reflection. This load is heated with a normal wire that is wrapped around it, and its temperature is measured with a diode that is located at
5.2 Detector Optical Efficiency and Crosspolar Isolation

Figure 5.4: On the left are interferograms of ABS detectors while installed in the ABS cryostat. These measurements were made during testing in the Princeton high bay, when only half the pods (wafers 4, and 11) were installed, a neutral density filter (NDF), not AR coated, was placed at the 4 K stop, and a highpass (cutoff $\sim 100$ GHz) thick grill filter was mounted on top of the window. The Martin-Puplett FTS was mounted such that its output pointed directly at the ABS window (the HWP was not installed). The size of the FTS mirrors is much smaller than the cryostat window, and the rays emanating from detectors on the edge of the array exit the window up to $12^\circ$ off its normal vector, therefore only the center of the detector array is illuminated by the FTS. The pods in the middle of the array that are most sensitive to the FTS signal were populated by ABS1-11 detectors. High signal to noise measurements were made for 32 ABS1-11 and 9 ABS1-4 detectors. The figures on the right are the spectra derived from the interferograms on the left. The thick blue line represents the average of the plotted spectra, with its maximum normalized to one. The edges of the bandpasses are consistent across detectors, but each has different peaks and valleys in the middle of the band, attributed to non-uniform illumination of the feedhorns by the FTS signal. The average band is much closer to the ideal top-hat band, and more so, a band comprised of the peaks of each spectra.
Figure 5.5: Above is an FTS measurement of a CMB4 (had square waveguide OMT and larger bandwidth than CMB5) prototype detector, with a thick grill filter in the optical path. The thick grill filter acts as a high pass filter, with a cutoff frequency of 200 GHz. In this setup only the “blue leak” coupling is measured, since the detector bandpass is below the filter cutoff. Note that “blue leak” power extends well above the niobium cutoff of 740 GHz, indicating that some fraction couples directly to the TES island, and not through the antenna probes. In black is the spectrum for TESA, while in red is the spectrum for the dark TES. Structure in the spectra comes from the thick grill filter, atmospheric lines, and roll off from the high frequency attenuation from plastic filters. Figure courtesy of TRUCE collaboration, previously published by Bleem, et al. (2009) [13].
the top end of the load. The cold load bolts to a 10 cm long stainless steel (SS) waveguide that thermally isolates the cold load, which can be heated to 30 K. The SS is heatsunk to 1 K in the middle and to 0.3 K at the other end, which connects to a copper waveguide with a 90° bend to align it with the detector input flange. This waveguide determines the polarization and restricts the field to a single mode in the target band. The rectangular output of the waveguide is converted to the circular input at the detector, via a stepped transition waveguide section.

The response of the detectors to changes in the cold load temperature \(T_{CL}\) is determined by measuring the bias power \(P_{\text{tes}}\). At a constant bath temperature, the sum of \(P_{\text{tes}}\) and the optical loading \(P_{\gamma}\) is constant, so that when the cold load temperature is changed, \(\Delta P_{\text{tes}} = -\Delta P_{\gamma}\). The optical loading can also be predicted by considering the power a blackbody at \(T\) emits in the ABS frequency band onto a single-mode detector. This estimate must be corrected for the SS waveguide loss as well as its emission. The power from the cold load, \(P_{\text{CL}}\), may be expressed as:

\[
P_{\text{CL}} = \int \int A_{e}(\theta, \phi) \frac{B_{\nu}}{2} d\Omega d\nu, \tag{5.3}
\]

where \(B_{\nu}\) is the blackbody spectrum for unpolarized radiation:

\[
B_{\nu} = \frac{2h\nu^{3}}{c^{2}} \frac{1}{e^{\frac{h\nu}{kT}} - 1}, \tag{5.4}
\]

and \(A_{e}(\theta, \phi)\) is the effective area of the detector. \(h\) is Planck’s constant\(^1\) and \(k\) is Boltzmann’s constant\(^2\). For a detector that perfectly couples to a single mode, \(\int A_{e} d\Omega = \lambda^{2}\). Therefore the maximum power loading from the cold load at a temperature \(T\) in the frequency band between \(\nu_{0}\) and \(\nu_{1}\) is:

\[
P_{\text{CL}} = \int_{\nu_{0}}^{\nu_{1}} \frac{h\nu d\nu}{e^{\frac{h\nu}{kT}} - 1}, \tag{5.5}
\]

where the integral is performed over a uniform band from 127 to 165 GHz. This choice means that the efficiency is defined with respect to a detector with an ideal top hat bandpass of

\(^{1}\)Planck’s constant \(h = 6.626068 \times 10^{-34} \text{ m}^{2} \text{ kg s}^{-1}\)

\(^{2}\)Boltzmann’s constant \(k = 1.3806503 \times 10^{-23} \text{ m}^{2} \text{ kg s}^{-2} \text{ K}^{-1}\)
the same effective width as the ABS simulated bandpass. Alternatively, the integral may be performed over a measured bandpass with arbitrary normalization. In this case the ratio of the measured power to the predicted power may imply optical efficiencies that are much higher or lower than the true band average efficiency. Dividing the arbitrarily normalized bandpass by the efficiency it predicts, leads to the “Absolute Bandpass,” which gives the transmission at each bandpass frequency. (Absolute bandpasses are presented in Figure 5.3.)

The SRDP CL loading estimate is corrected for a 51% loss across the 10 cm stainless steel waveguide. The loss was estimated by comparing cold load power loading measurements using the same detector, but with 10 cm and 15 cm SS waveguides. The power emitted by the 5 cm warm section of the waveguide between the cold load and the 1K heat sink, is estimated assuming an emissivity of 0.3, the same as the fractional power loss through it.

![Figure 5.6: Optical efficiency measurements for CMB4 prototype detector.](image)

Figure 5.6: Optical efficiency measurements for CMB4 prototype detector. In the left panel are the measured optical power differences from 2.5 K for the cross polar channel, TES A (stars), the copolar channel, TES B(crosses), and the dark TES (circles), which is considered a measure of the out of band power. The statistical errors on the measurement are smaller than the size of the markers. Not included are the 5% uncertainty in the shunt resistance or the small uncertainties in the loading from the warm waveguide. The wide solid swatch shows the predicted optical loading with uncertainties. In the right panel are measured optical efficiency (stars) and crosspolar coupling (circles) as a function of internal cold load temperature.

The predicted optical loading as a function of cold load temperature referenced to the
predicted loading at \( T = 2.5 \) K is given by the wide solid line in the left panel of Figure 5.6. The width represents the errors in estimating the loading, described below. The measured power differences are plotted as stars (crosspolar channel), crosses (co-polar channel) and circles (“dark” channel not connected to the microstrip). Notice that all of the channels see a signal even though we only expect to see power in the co-polar channel (TES B). The majority of the power that couples to TES A and the dark bolometer is believed to come from radiation at frequencies higher than the band (i.e, the blue-leak, see section 5.1). At these higher frequencies the waveguide becomes multi-moded and radiation from both polarizations can propagate.

Assuming that the out of band power is distributed equally amongst the three channels, and that the dark TES can only see out of band radiation, the in-band power is equal to the power on TES B minus the power on the dark TES. Comparing this in-band power to the predicted power load yields an optical efficiency of 54% for the CMB4 prototype detector. The right panel of Figure 5.6 shows the estimated optical efficiency versus \( T \). The errors include the I-V measurement error, an uncertainty of 5% on the electrical conductivity of SS, and a 3% uncertainty in the length of the waveguide. Not included are the 5% uncertainty in the shunt resistance and the small uncertainties in the emission from the warm waveguide. Under the same assumptions the crosspolar response is obtained by subtracting the power on the dark TES from TES A. The resulting upper limit on the cross-polar leakage is 3%; these limits are plotted as circles in the right panel of Figure 5.6. See [78] for a discussion of the expected OMT cross-polarization properties. Band average optical efficiencies for detectors from different wafers are given in Table 5.1.

5.3 Beam Maps

Power radiated by the antennas of an ABS detector is projected on the sky through the feedhorn, reflective mirrors and cold stop, forming a beam pattern. By time reversal symmetry the detectors will be sensitive to only the sky region encompassed by this beam. The detector beam may be measured by observing a point source with the telescope pointing
Table 5.1: Table summarizing the detector efficiency of prototype and ABS wafers. All measurements are from single pixel test with the SRDP cold load. Efficiencies are calculated from the ratio of the measured power to that expected by the ABS simulated bandpass. In bold are the wafer used to populate the ABS array.

<table>
<thead>
<tr>
<th>Wafer</th>
<th>Detector</th>
<th>% Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>CMB4</td>
<td>4106</td>
<td>54</td>
</tr>
<tr>
<td>CMB5</td>
<td>5113</td>
<td>50</td>
</tr>
<tr>
<td>ABS2-5</td>
<td>57</td>
<td>20</td>
</tr>
<tr>
<td>ABS1-4</td>
<td>24, 43, 80</td>
<td>45</td>
</tr>
<tr>
<td>ABS1-8</td>
<td>41</td>
<td>22</td>
</tr>
<tr>
<td>ABS1-10</td>
<td>59,82</td>
<td>11</td>
</tr>
<tr>
<td>ABS1-11</td>
<td>80</td>
<td>45</td>
</tr>
<tr>
<td>ABS1-14</td>
<td>54</td>
<td>43</td>
</tr>
<tr>
<td>ABS1-15</td>
<td>53</td>
<td>43</td>
</tr>
</tbody>
</table>

at various azimuth and elevation angles. Since the ABS telescope is designed to scan in azimuth, beam maps typically consist of multiple sweeps in azimuth as the telescope bore-sight is stepped in elevation. Beam maps of ABS detectors were measured during testing of the ABS cryostat in the Princeton high bay with the NDF mounted. The beam testing setup is shown in Figure 5.7. A broadband polarized source with a center frequency near 130 GHz is mounted on the high bay crane, 11 m from the ABS cryostat window. With the top half the array installed, the cryostat scanned the source in azimuth, at various elevation steps. One of the measured beams is shown in Figure 5.8, along with the expected beam from simulations. The far field of ABS is much further than the position of the source, hence the simulated beam is computed in the near field limit. This simulation predicts a $1 ^\circ$ FWHM as opposed to the far field expectation of $0.6 ^\circ$ FWHM. Simulations agree well with the measured beams in the high bay, adding confidence that the expected performance would be achieved in the field. In Chile beams are measured by looking at point sources like the moon, Jupiter or Venus. For obtaining a polarized beam, tauA is useful, but to get higher signal to noise it may be necessary to duplicate the beam map measurements made with the incoherent broadband source.
Figure 5.7: Picture of the ABS cryostat while scanning a broadband polarized source in Princeton high bay. The source was mounted on the high bay crane, approximately 11 m away.
Figure 5.8: The left plot shows a beam pattern measurement of the row 0, column 16, ABS detector. The x-axis and y-axis are the azimuth and elevation pointing coordinates of the telescope. The measurements were made in the Princeton high bay, by observing a polarized source 11 m away from the ABS cryostat. The HWP was rotating, hence the amplitude color scale in decibels (dB) corresponds to the amplitude modulation caused by the HWP while pointing at a particular elevation and azimuth away from the source. Azimuth coordinates are corrected for the apparent decrease in scan width as the elevation pointing is increased. On the right is the simulated near field beam pattern 11 m away (ABS far field is at least 100 m away). The simulated beam was computed by propagating the field pattern predicted by DADDRA at the window. Measured and simulated beams have \( \sim 1^\circ \) FWHMs, and as expected, they are larger than the predicted 0.6° far field FWHM.
5.3 Beam Maps

Figure 5.9: Crosses mark each detector’s beam center on the sky based on DADRA simulations. Detector pairs are labeled with their row and column numbers as well as their pod and feed numbers. Rows range from 0 to 21, columns from 0 to 23, pods from 0 to 23, and feeds from 0 to 9. The y-axis angle is such that the detectors are plotted at their physical location on the focal plane but keep in mind that their projection on the sky is inverted. Hence the detectors that are located at the top of the array look lower on the sky. These beam center locations are analogous to the feedhorn’s locations on the focal plane when looking at it from its back side (Figure courtesy of Mike Nolta).
120 Optical Detector Tests

The beams of all the detectors make up the array’s field of view shown in Figure 5.9. This figure provides the mapping from the readout row and column numbers of the MCE, to the offset from boresight of the detectors generated by DADRA simulations. Note that the y-axis is the angle from zenith, hence the detectors on the top of the plot look low on the sky.

5.4 HWP Systematics

The rotating HWP generates a detector response at its rotational frequency \( f \) and harmonics \( (1f, 2f, 3f, 4f, ...) \). The \( 1f \) signal is similar to a sky temperature modulation, in that an orthogonal detector pair observes the signal in phase. This signal may be due to imperfections in the HWP thickness or AR coating, leading to spots of higher transmission and/or emission. Its median amplitude across detectors is \( \sim 2 \times 10^{-15} \) W. The \( 2f \) signal, generated by the difference in reflection coefficients between the two HWP axes, is typically an order of magnitude larger than the \( 1f \) signal. The ABS HWP AR coating is optimized for the average index of refraction of the two HWP optical axes. Designing an AR coating that matches the index of each axis would mitigate the \( 2f \) signal. The \( 3f \) signal is an order of magnitude smaller than the \( 1f \) signal, and is likely a higher harmonic of the \( 1f \), and/or generated by beating of the \( 1f \) and \( 2f \) signals. The \( 4f \) signal is five times larger than the \( 1f \) signal, making it half the \( 2f \) signal. Its relatively large amplitude suggests it is not just a harmonic of the \( 1f \) and/or \( 2f \). Any polarized signal that is not modulated by the azimuth scan shows up as \( 4f \). One such polarized signal is generated by unpolarized light from the warm baffle or the atmosphere impinging on the HWP at an angle. The HWP-AR coating-air interface has a transmission coefficient determined by Fresnel’s equations. These equations predict different transmissions for the two polarization components of the unpolarized light, hence generating a polarized signal. Another source of polarized light could be reflections off the baffle.

Figure 5.10 shows the power spectrum of one ABS detector while the HWP is rotating and the base is doing scans in elevation. The HWP harmonics are large signals comparable
Figure 5.10: The top plot shows the power spectrum of a detector while the HWP is rotating and the telescope is performing sweeps in elevation called sky dips. The sky dip signal (green) is at 0.04 Hz, while the HWP generates large signals at 1 (red), 2 (cyan), 3 (purple), and 4 (yellow) times the HWP rotational frequency $f$ (2.3 Hz). Below 1 Hz the detector noise is dominated by the atmosphere 1/f signal. The bottom plot zooms into the power spectra around the 4$f$ harmonic, where the modulated CMB polarization signal is located. The HWP harmonics have two sets of side bands, one due to the servo control 0.23 Hz away (gray), the other due to leakage of the sky dip temperature signal into polarization, 0.04 Hz away (green) .
Optical Detector Tests

to the atmosphere 1/f signal. Thankfully they are narrow, on the order of 0.05 Hz, which means even slow scanning of the sky will place the signal band away from the HWP 4f. To compare, the linear azimuth motor scans up to 2 deg/s with a maximum throw of 10°, hence the azimuth scans frequency can reach 0.1 Hz. The HWP harmonics have side bands due to the servo that tries to maintain the HWP at a constant speed. The side band amplitudes are at least two orders of magnitude smaller than the amplitude of the corresponding harmonic, and may be easily demodulated using the HWP encoder channel.

5.5 Sky Dips

The largest signal in the top plot of Figure 5.10 is due to scans in elevation, also referred to as sky dips. When the telescope moves in elevation the amount of airmass in its line of sight changes, hence changing the loading on the detectors. The detector loading is proportional to \( \sec(90° - \Theta_{el}) \) where \( \Theta_{el} \) is the elevation angle from the horizon. The least loading occurs when pointing at zenith and then it increases as the telescope points lower. Sky dips from 55° to 50° elevation generate a detector response on the order of 0.25 pW when the atmosphere precipitable water vapor (PWV) is \( \sim 3.0 \) mm; when the PWV is low (\( \sim 0.27 \) mm) the amplitude drops to 0.05 pW. The sky dip signal is calculated as the peak of the 0.04 Hz line in each detector power spectra. Figure 5.11 shows the measured sky dip signal for all ABS detectors, plotted at their focal plane location, as well a histogram of all results. The top of the array, which looks lower on the sky, observes a larger change of airmass in its line of sight than the bore sight, and hence a larger sky dip signal. The opposite holds for detectors in the bottom of the array looking high on the sky. The top half of the array also holds the detectors with the highest sensitivity, further enhancing the difference of sky dip response from top to bottom.

The expected power observed by each detector during a sky dip can be estimated from the Chajnantor atmosphere model [18] that predicts the atmosphere opacity (\( \tau \)) at frequencies where the ABS bandpasses lie. The model depends on the atmospheric precipitable water vapor (PWV), defined as the amount of water vapor contained in a vertical column...
Figure 5.11: The figure on the left shows an array plot where each detector is represented by an ellipse placed at its beam center with respect to the bore sight. The position of the beam center follows the same convention discussed in Figure 5.9. The color of each ellipse represents the amplitude in pW of the sky dip signal measured by a detector, when changing elevation from $55^\circ$ to $50^\circ$. This particular set of data corresponds to an atmosphere PWV of 0.27 mm. The loading on the detectors is proportional to the secant of the angle from zenith, which implies that sky dips about lower elevation lead to larger signals. Therefore detectors on the top of the array looking lower on the sky observe a larger signal than detector on the bottom of the array. This effect coupled with the higher efficiency of the detectors on the top half of the array (wafers 4 and 11) lead to the observed pattern of sky dip amplitudes. On the right is a histogram of the sky dip amplitudes. The large number of detectors with a low sky dip signal at 0.01 pW corresponds to columns 4 and 11, while most of column 7 is superconducting at the applied bias. These columns hold detectors from wafer 14 with low efficiency and high saturation power, that in the extreme case of column 7 prevents appropriate detector biasing.
Figure 5.12: The optical efficiency is estimated from sky dip measurements made on April 15 2012, with the PWV at 0.27 mm, and assuming a sky temperature of 250 K. On the left is an array plot of the detector efficiencies. Note that most detectors efficiencies are near 0.3, within expectations, but for a few pods they are much lower. The low efficiencies belong to the edge of wafer 14. Some of the very high efficiencies outliers (> 0.4) are due to the readout SQUID unlocking, while others may be due to increased susceptibility to the blue leak. On the right is a histogram of the detector efficiencies.
of unit cross-sectional area above the site. The Atacama Pathfinder EXperiment (APEX) weather monitor [45] measures PWV from the “Llano de Chajnantor”, located a few miles from the ABS site. The PWV measurement together with the atmospheric model, and an estimate of the atmosphere temperature \(T_{atm}\) predicts the atmospheric power \(P_{atm}\) loading on the detectors when pointing at zenith. This power is corrected by the secant airmass law mentioned earlier, to obtain the power loading at the two elevations \(\Theta_{zen1}, \Theta_{zen2}\) measured from zenith) of the sky dips. These powers are then subtracted to calculate the expected sky dip signal. The expected power from a sky dip \(P_{sd}\) on a detector is given by:

\[
P_{sd} = P_{atm}(e^{-\tau \sec \Theta_{zen1}} - e^{-\tau \sec \Theta_{zen2}}).
\]  

(5.6)

From the raw sky dip power measurements, we subtract the blue leak contribution (see section 5.1) modeled as the response of detector C8r18, corrected for the elevations pointing of each detector. We also subtract the expected contributions of the second mode (see Figure 2.12). Each of these corrections is on the order of \(\sim 10\%\) of the signal. Not accounted for is any ground pick up signal. Also note that the bandpasses of the lower half of the array are not yet measured; instead they are estimated to track the simulated bandpass shifted up in frequency by 15 GHz, as suggested by single detector FTS measurements. The ratio of the corrected power signal to the expected power signal from the sky dips, yields the combined in-band optical efficiency of the HWP, filters, cold stop, feedhorns, and detector chips. The combined optical efficiency of each detector in the ABS array is plotted in Figure 5.12.

The atmosphere is unpolarized, therefore changes in elevations should be pure intensity modulations that lead to equal responses from detectors pairs. Nevertheless the instrument leaks some intensity power to polarization (I→Q) leading to side bands around the HWP 4f frequency, as shown in the bottom plot of Figure 5.10. The I→Q leakage is estimated for every detector and plotted in Figure 5.13. Most detectors leak less than 0.1\% of a pure intensity signal to polarization.
Figure 5.13: On the left is an array plot for the sky dip estimated $I\rightarrow Q$ leakage. On the right is a histogram of all such estimates. The leakage is computed by dividing the power observed at the side bands of the $4f$ signal (9.15 Hz and 9.23 Hz) by the sky dip power observed at 0.04 Hz. For most detectors less than 0.1% of the sky dip intensity is converted to a polarized signal. Detectors in the middle of the array appear to suffer higher levels of $I\rightarrow Q$ leakage.

5.6 ABS Sensitivity

The array polarization sensitivity is described by the noise equivalent $Q$ (Stoke’s parameter), which is calculated from:

$$\text{NEQ} \equiv \mu_{\text{pol}} \text{NET},$$

(5.7)

where $\mu_{\text{pol}}$ includes the detector polarization efficiency and the HWP modulation efficiency. The detector sensitivity will be quantified in terms of noise equivalent temperature (NET) for the majority of this section, hence isolating it from the polarization efficiency of the optics. NET is the detector noise level converted to units of small temperature fluctuations about the CMB blackbody temperature (2.725 K). It can be calculated from the detector NEP through:

$$\text{NET}(K\sqrt{s}) = \frac{\text{NEP}}{\mu_{\gamma} \frac{dP}{dT_{\text{cmb}}} \sqrt{2}},$$

(5.8)

where $\mu_{\gamma}$ is the optical efficiency that converts power fluctuations on the TES to power fluctuation from the sky. It includes the efficiency of the on chip elements as well as the
transmission of the free space filters, the HWP and the atmosphere (0.98 when pointing at zenith with a PWV of 0.27). The factor \( \frac{dP}{dT_{\text{cmb}}} \) is the amount of optical power per small CMB temperature fluctuation, given the bandpass of the detectors. It is equivalent to the derivative of Equation 5.5 (replacing \( P_{\text{CL}} \) by the in-band CMB power \( P \)) with respect to temperature evaluated at \( T_{\text{cmb}} \):

\[
\frac{dP}{dT_{\text{cmb}}} = \frac{h^2}{kT_{\text{cmb}}^2} \int \frac{B(\nu)\nu^2 d\nu}{(e^{\frac{h\nu}{kT_{\text{cmb}}}} - 1)^2},
\]

integrated over the detector bandpass \( (B(\nu)) \). The \( \frac{1}{\sqrt{2}} \) factor converts \( 1/\sqrt{\text{Hz}} \) to \( \sqrt{\text{s}} \). The NET for the top half of the array is computed assuming the simulated CMB5 bandpass, while the bottom half is computed assuming the CMB5 bandpass shifted up in frequency by 15 GHz. The NEP of each detector is the average measured at a side band (usually take 9 Hz±0.5 Hz) of the 4f signal. The optical efficiency is estimated in section 5.5 from sky dips given the Chajnantor atmosphere model [89] and the PWV measured by APEX [45]. Not included in the NET plots is the HWP modulation efficiency \( (\mu_{\text{pol}}) \), that varies from 95% in the center to as low as 88% for detectors located at the edge of the array.

The NET of each detector is shown in an array plot in Figure 5.14. The best detectors are located in the top half of the array, and belong to wafers 4 and 11. Columns 4, 7, 11, and 8 show the worst performance, and belong to wafer 14. Their high NET is due to low efficiency and high \( G \) and \( T_c \) values. The total array NET is the weighted average of the individual detector NETs:

\[
\text{NET}_{\text{array}} = \left( \sum_{\text{all, det}} \text{NET}_i^{-2} \right)^{-0.5}.
\]

Figure 5.15 places each detector NET at its location on the fabrication wafer. Wafers 4, 11, and 15 show uniform distributions, but pixels on the edge of wafer 14 have much higher NET than those in the middle. Only a single pixel from the middle of wafer 14 was tested for optical efficiency in the SRDP; the bad edge pixels were only found in the field. These low efficiency pixels may be useful for observing bright sources such as the moon.

Table 5.2 summarizes the expected NET performance given the dark and optical test carried out in Princeton, combined with the Chajnantor atmosphere model. These NET
Figure 5.14: On the left is an array plot of the ABS detector NETs using the blackened baffle, and on the right is a histogram of the NETs when using the blackened versus the shiny baffle. The blackened baffle NET data set was acquired with an atmosphere PWV of 0.27 mm. In this case the NET of the majority of detectors in the top half of the array is between $450\,\mu\text{K}\sqrt{s}$ and $900\,\mu\text{K}\sqrt{s}$. Column 7 is off, and columns 4, 11, and 8 have NET greater than $1500\,\mu\text{K}\sqrt{s}$. The total array NET with the blackened baffle is $38.6\,\mu\text{K}\sqrt{s}$ when pointing at elevation $55^\circ$. The shiny baffle NET data set was acquired with an atmosphere PWV of 0.33 mm. The total array NET with the shiny baffle is $28.7\,\mu\text{K}\sqrt{s}$ when pointing at elevation $67^\circ$. 
Figure 5.15: Above are NET plots where the detector are arranged in their wafer locations. Only detectors installed in the array are included. Wafer 14 shows the worst performance, particularly at its edge. A single pixel test from the middle of the wafer indicated a similar NET to wafer 15; only after the full array was assembled and tested in the field were the low efficiency edge detectors found. The data plotted above was acquired on March 15 2012 with the blackened baffle installed.
Table 5.2: The expected (exp.) NEP were obtained from SRDP measurements discussed in the previous chapter, together with an estimated photon noise NEP of $3 \times 10^{-17} \text{W}/\sqrt{\text{Hz}}$. The extracted NEP came from noise power spectra measurements (blackened baffle:shiny baffle) of the 4f side bands calibrated into power with I-V responsivity. The ABS1-11 NEP is much better in the field, it is likely that the additional optical loading help stabilize these fast detector, improving their noise performance. The expected optical efficiency is computed from the reflection and absorption of each element discussed in Table 2.3.3. The measured (meas.) efficiency was extracted from sky dips (04/15/2012) given the Chajnantor atmosphere model at PWV of 0.27 mm, and correction factors from the blue leak and second mode. The conversion factor from power to CMB temperature fluctuations ($dP/dT_{\text{cmb}}$) was computed from the simulated bandpass for wafers 4 and 11, and from simulated bandpass shifted up 15 GHz for wafers 14 and 15. With the blackened baffle on, the measured NEP was much higher than the expectation for all wafers, indicating an excess of optical loading. The sky dip efficiency of wafers 4 and 11 are only a few percent lower than the expectation, while for wafer 14 is $\sim 50\%$ lower, a result of bad detectors in the wafer’s edge. The table’s measured values are the median of all detectors in the array from a particular wafer. All individual detector NET estimates include an atmosphere transmission of 0.98.

<table>
<thead>
<tr>
<th>Wafer</th>
<th>NEP exp./meas. $10^{-17} \text{W}/\sqrt{\text{Hz}}$</th>
<th>Eff. exp./meas.</th>
<th>$dP/dT_{\text{cmb}}$ $10^{-13} \text{WK}^{-1}$</th>
<th>NET exp./meas. $\mu\text{K}\sqrt{\text{s}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABS1-4</td>
<td>4.5/(6.3:5.1)</td>
<td>0.3/0.28</td>
<td>2.82</td>
<td>387/(576:466)</td>
</tr>
<tr>
<td>ABS1-11</td>
<td>6.3/(6.6:5.2)</td>
<td>0.3/0.26</td>
<td>2.82</td>
<td>534/(650:512)</td>
</tr>
<tr>
<td>ABS1-14</td>
<td>5.3/(6.6:4.9)</td>
<td>0.29/0.14</td>
<td>1.82</td>
<td>738/(1860:1387)</td>
</tr>
<tr>
<td>ABS1-15</td>
<td>5.6/(7.7:5.4)</td>
<td>0.29/0.26</td>
<td>1.82</td>
<td>762/(1174:823)</td>
</tr>
</tbody>
</table>
5.6 ABS Sensitivity

results are divided based on wafer and compared to those obtained in the field from sky dips. The optical efficiencies are close to expected, given the known issues with wafer 14. The NEP values measured in the field on April 15, 2012 are higher than expected, indicating an excess of optical loading. The main excess loading contribution was found to be from the blackened baffle enclosure. The baffle is a conical shaped aluminum piece that is located above the cryostat window and HWP to eliminate ground pick up. When blackened its interior is painted with Stycast mixed with carbon chunks, when shiny the Stycast has been removed, and the shiny aluminum surface is visible. Comparing I-V curves acquired with the blackened baffle (04/15/2012) and a shiny baffle (08/21/2012), shows that the blackened baffle added $\sim 1.7 \text{ pW}$ of excess loading. Replacing the black baffle by a shiny baffle improved the array NET from 38.6$\mu K\sqrt{s}$ to 28.7$\mu K\sqrt{s}$. When using the shiny baffle the ABS array NEQ is $\sim 31\mu K\sqrt{s}$, assuming a polarization efficiency of $\mu_{\text{pol}} > 0.9$, estimated from simulations of the HWP and detector OMT.
Chapter 6

Array Assembly

The ABS focal plane is populated with 240 polarimeters grouped in 24 pods of ten. Pods are individually assembled and tested before being mounted on the focal plane support (FPS). Each pod holds ten feedhorns with detector chips attached to their waveguide outputs. The detector chips are bonded to an aluminum circuit board that connects them to the shunt and SQUID chips. The SQUID output and bias lines are bonded to a tin-copper circuit board that connects to the “300 mK backplane” through ZIFs ([53]). The 300 mK backplane is a ten layer FR4 circuit board located behind the FPS and pods. It routes all bias lines from the 24 pods to 6 MDMs ([42]). The 300 mK backplane is connected to two “4 K readout boards” through MDM cables with 10” long, 0.004” thick NbTi wires. These wires are heatsunk at 1 K to reduce thermal loading on the 300 mK stage. The 4 K boards hold 24 series array modules that are the last stage of cryogenic signal amplification, and also route all electrical lines from the focal plane on to five 100-pin MDM connectors. These boards are connected to the 300 K MCE electronics through 100-pin MDM cables with 0.004” diameter wire. The cable from the 4 K stage to the 40 K stage is 1 m long and made of Constantin to reduce thermal loading. The cable from the 40 K to the 300 K stage is 2.3 m long and made of copper. Starting with the 4 K boards and out to the MCE, the electrical wiring for ABS is the same as that used for ACT [84].
Figure 6.1: Above is a picture of an ABS backshort. The backshorts are etched from silicon and coated with gold. The middle cavity defines the quarter wavelength termination behind the OMT. The U-shaped cavity is filled with absorbing material to reduce stray radiation. L-shaped SU posts are located on four corners to provide 50μm clearance between the OMT and the backshort.

6.1 Detector Assembly

ABS detectors are fabricated on silicon wafers. Each wafer yields a maximum of 110 individual pixels. For ABS, NIST fabricated eight wafers that yielded detector chips. Their characteristics are summarized in Table 6.1.

Pixels from the working wafers were screened based on broken TES pixels, defects on the OMT area, scratches, and cleanliness. The best pixels were then glued to backshorts. A backshort as seen from the detector’s perspective is shown in Figure 6.1. Each backshort forms a quarter-wavelength waveguide cavity behind the OMT of a pixel, and holds black absorbing material above the TES bolometer to reduce coupling to stray radiation. The backshorts are mounted on the top side of the detector chip. To avoid touching the OMT membrane the backshorts have four L-shaped SU posts that are 50μm thick. The space between the chip and the backshort should be minimized to decrease loss of signal, but at the same time the backshort must be far enough away to avoid ripples in the OMT. These ripples are caused by stress in the OMT SiN membrane and require at least 25μm vertical clearance.
Table 6.1: The table above summarizes the key characteristics of the detector wafers fabricated by NIST for ABS based on SRDP tests of 1-2 detectors per wafer. CMB5 was the prototype detector developed in the TRUCE collaboration. Eight detector wafers were fabricated for ABS, out of which four provided detectors to the focal plane. These were ABS1-4, ABS1-11, ABS1-14, and ABS1-15. ABS2-5 and ABS1-9 were discarded for missing the $T_c$ target by a large amount. The efficiency of ABS1-8 and ABS1-10 was poor, and their bandpass center was lower than the target by 15 GHz. Their poor performance was attributed to impurities in the Nb target used to make the detector’s superconducting traces. These two wafers, were the first to use a new SiO2 recipe to help decrease loss in the microstrip. The SiO2 recipes used in ABS labeled A and B correspond to settings: SiH4:118.2 sccm, O2: 18.7 sccm, RF: 220 W; and SiH4:131 sccm, O2: 10.2 sccm, RF: 180 W. The bandpass for ABS1-14 and ABS1-15 was redesigned to compensate for the shift observed in ABS1-8. Unfortunately, the redesign was unnecessary and the last two wafers ended with band passes centered higher, at 160 GHz, making them slightly sensitive to the atmospheric water line at 180 GHz. Efficiencies quoted in the table above are from single detector tests with the SRDP cold load, hence they do not include wafer uniformity issues, as observed on wafer 14 through sky dips.
Figure 6.2: Above is a diagram of the jig used to align the detector chip with the backshort. On the left is a top view, while on the right is a side view. The black lines are the edges of the acrylic jig, while the red lines emphasize the alignment surfaces. Holes are made near the mounting location to provide easy access for the pins and tweezers used to handle the detector components. The detector chip in gray (5 mm diameter), is placed on the jig, and two of its sides are pressed against the alignment surfaces. The backshort (orange) is placed above the detector following the same alignment features. Scotchk weld is added between the SU post and detector chip, but before it settles the jig is flipped to check and improve the alignment between the triangular probes and the backshort cavity. Note that the OMT is below the backshort and can only be observed through the bottom side of the acrylic jig.
6.2 Pod Assembly

It is critical for good optical and polarization efficiency ([78]) to align the backshort and OMT to better than 25\,\mu m. To achieve this goal ten acrylic jigs as shown in Figure 6.2 were used. The detector chip is mounted on the jig and its side walls are aligned with the two reference surfaces of the jig. These surfaces are also used to guide the backshort while it is lowered on to the chip. Once the backshort is resting on the chip, Scotchweld 2216 [103] is added to its four corners and spread gently between the backshort and the chip. Before the glue settles, the detector is flipped, and the alignment of the OMT with the quarter-wave length cavity is checked with a high magnification microscope. Small imperfections on the edges of the chip or backshort can cause misalignments; these are fixed manually by using small mil probes to manipulate the two pieces. The alignment of most pixels is better than 20\,\mu m, based on microscope observations of the overlap between the OMT and the backshort edges. Once aligned the glue dries for at least twelve hours before the chips are moved off the jigs into storage.

The final step of pixel assembly consists in mounting a chip with backshort onto an aluminum feedhorn. This step requires precision placement of the chip from the beginning, since the OMT-waveguide alignment is not visible and therefore cannot be adjusted. The OMT is obscured on one side by the backshort and on the other by the depth of the feedhorn. The positions of the OMT and waveguide are measured to micron accuracy with respect to the chip and feedhorn edges. These edges are then aligned such that the OMT is centered on the waveguide.

6.2 Pod Assembly

Each pod interface plate holds and determines the polarization alignment for ten feedhorns [36]. Feedhorns with detectors are introduced and bolted through the front of the plate, while an FR4 circuit board is glued on the back, such that its traces are coplanar with the bond pads of the detectors. The electrical connection between circuit board and detector chips is made through Al wirebonds. Multiple iterations of the circuit board were carried out to improve the reliability of the bonds, traces, and ZIF contacts. Some of the important
characteristics of the ABS circuit boards are:

1. Tin-copper traces with gold bonds pads were used where possible: an oxide layer forms quickly on top of aluminum traces, making their contact with ZIF connectors less reliable than that achieved with tinned traces. Also the bending radius and trace reliability of Tin-Copper is much better than Aluminum due to industry standards. Nevertheless superconducting aluminum traces are necessary because the resistance of Tin-Copper traces is on the order of mΩ, appropriate when compared to the impedances of SQUIDS (∼1 Ω), but too high to form part of a TES loop where the operating resistance is ∼ 2 mΩ. In the cases where traces need not be superconducting, the bond pads are made of gold. Bonds to gold tend to be stronger than those to aluminum.

2. Aluminum traces were used in the TES loop. The TES bolometers are connected to the shunt and mux chips through flexible aluminum-on-kapton circuit boards that add essentially zero parasitic resistance. The flexible section allows the mux and shunt to be sandwiched between Nb sheets on a plane above the detector chips. This provides great magnetic shielding and permits the close packing of pods.

3. Bond pads were made as large as possible. Larger bond pads provide a stable surface for bonding. This is particularly important for the aluminum circuit boards, where the Al layer is glued to the FR4. Here the bond pads lay on a soft substrate that easily moves when the bonding head applies pressure.

4. Bending radius of aluminum flex was increased. Most of the open TES channels in ABS are due to broken traces on the aluminum flex. This was mitigated by increasing the length of the aluminum flex section, and limiting the number of times it was bent.

5. Bond pad surfaces were stabilized. All circuit boards, detectors, shunts, and mux chips, were glued down at one point with Scotchwell 2216. It held the bond pads firmly in place leading to strong bonds with little hassle. Early experiments with rubber cement, showed that after multiple cryogenic cycles it would loosen up making
it impossible to re-bond without re-gluing. Rubber cement is useful for single tests, after which components are replaced and re-bonded.

6. We avoided mounting silicon components on FR4. The ABS circuit board is susceptible to thermal contraction due to its thin layers of FR4 and Kapton. The side of the board not mounted on the interface plate has a niobium sheet glued on its back. This sheet stiffens the board. The shunt and mux chips are placed onto the Nb sheet through a hole cut in the circuit board. The chip and board bond pads are approximately level with each other with this arrangement.

A circuit board mounted on a pod assembly jig can be seen in Figure 6.3. Feedhorns with detectors are mounted on the right while the shunt and mux chips are mounted on the left. All 178 bonds between pod components and circuit board are made in this jig configuration.

After bonding, brass rods are introduced on the corners of the interface plate. Washers are then dropped over the rods to set a 1.6 mm vertical spacing between the interface plate and the first niobium sheet. These spacers are necessary to hold the niobium above the detector bonds and backshorts. A second set of washers are placed on the rods above the first niobium sheet. These set a spacing of 0.8 mm between the mux-shunt section of the circuit board and the niobium sheet below. This spacing was chosen to clear the height of the mux-shunt wire bonds, while minimizing the space between niobium sheets to improve the magnetic shielding of the SQUIDS.

The stack of spacers, niobium sheets and circuit boards is secured with brass nuts on the top end of the rods. These brass nuts provide pressure to keep all components in place and generate good thermal contact between components. Next an aluminum lid is bolted to the interface plate forming an almost complete (except for the feedhorn waveguide holes) superconducting box around the detectors and SQUIDs. The lid also has a slit that allows the two flex lines from the circuit board to connect to ZIFs on the 300 mK board.

Once a pod is assembled, all its connections are continuity checked, to make sure they have the right resistance, and no shorts to ground. Next it is cooled in the SRDP and undergoes a dark cryogenic test. During this test all SQUIDS and detectors are biased,
Figure 6.3: The picture above shows a pod mounted on a pod assembly jig. On the left is the tin-copper circuit board with the flex that connects the pod to the 300 mK backplane. This circuit board has a rectangular hole in the middle where the mux and shunt chips are mounted to the Nb sheet below. Aluminum wirebonds connect the gold bond pads of the tin-copper traces to the bond pads on the chips. The shunt chip is connected to the detectors through the aluminum circuit board. This circuit board is composed of two stiff sections connected via a flex section. The stiff section on the left is glued to the tin-copper circuit board, and holds the aluminum bond pads that are bonded to the shunt chip. The stiff section on the right carries the traces and bond pads to the ten detector chips. This section is glued to the pod interface plate where the feedhorns mount. Three brass rods on the corners, with spacers, set the vertical spacing of the niobium sheets and the tin-copper circuit board, when it folds over above the detectors. Nuts on the end of the brass rods clamp the vertically spaced component for mechanical stability and good thermal contact. Finally, an aluminum lid not shown above, mounts to the interface plate and forms a superconducting shield around the pod components. All electrical bonds are made in the configuration shown above.
while checking for electrical continuity, good thermal heat sinking, and faulty detector channels. If issues are found the pod is opened, the faulty components are replaced, and it is retested. If the cryotest is successful the pod is mounted on the ABS focal plane.

6.3 Focal Plane Assembly

The ABS focal plane is composed of 24 pods mounted in a hexagonal pattern. The pods are introduced through the backside of the hexagonal FPS structure (opposite to the optics), but are bolted from the front side. Three 4-40 screws connect the corners of the pod aluminum interface plate to the gold coated copper FPS. Once the 24 pods are mounted, the 300 mK backplane is placed behind the pods, and the pod flex are routed to their appropriate ZIF connectors. This backplane carries out three main tasks:

1. Connects in series all SQ1 channels with the same row selects from different pods. This is part of the multiplexing scheme that reduces the number of wires to the 300 mK stage.

2. Divides the pods in eight groups of three, each group having its detector bias lines connected in series to one of the eight bias lines provided by the MCE electronics.

3. Converts all electrical lines originating from the pods, in form of traces on flex, to pins on an MDM connector.

To achieve these tasks a ten layer board was designed, with special care taken to reduce electrical pickup by keeping signal and return traces as close to each other as possible, and including multiple layers of copper for shielding and good thermal heat sinking. The board has large cutout regions to allow access to the flex from the pods.

Once the 300 mK board is set in place, six MDM cables going from the focal plane to the 4 K board are mounted. These MDM cables are made with 10" long, twisted pair, superconducting NbTi wires, designed for low thermal conductivity and low RF pickup. The middle of each cable is permanently attached to a copper heat sink that clamps the
Figure 6.4: Four photos labeled one through four are shown above. 1. The back of the focal plane mounted on the 300 mK, 1 K and 4 K hex structures. All 24 pods have been bolted from the front side. 2. The 300 mK backplane is installed and the two flex lines from each pod are connected to their corresponding ZIF connectors. 3. The cables from 300 mK to the 4 K board are mounted with their 1 K heatsinks. Six cables are heatsunk through three copper clamps coated with white teflon on one side. All wires are 10” long NbTi twisted pairs, soldered to MDM connectors. 4. The assembled array from the front side mounted on its carrying case.
wires between two copper pieces, one of which is wrapped in Teflon tape to provide a soft contact point. This helps prevent shorts to ground and accidental mangleing of the wires while clamping. The clamps are mounted first on the 1 K stage, before connecting the cable to the 300 mK stage. Figure 6.4 shows the steps of the focal plane assembly process. A special carrying case was purchased to transport the array so it could be installed in the field.

At this point the focal plane hex structure is complete, and the next step is to mount it on the mirrors inside the cryostat. The 4 K hex mounts to a set of posts that place the detector array at the focal plane of the optics. Behind these posts lies the 4 K readout board with SQUID series arrays, and the MDM mates for the 300 mK connectors. For assembly the hex structure is set in front of the posts, while the MDMs are connected to the 4 K board. Once connected, the 4 K hex is bolted to the posts, arriving at the final detector array configuration shown in Figure 6.5.

The ABS 4 K SA board is the same as the one used in ACT [84]. It mounts up to four SA modules with eight channels each. ABS uses three SA modules to readout the 24 pods, and takes the biases originally assigned to the fourth module to bias the TES bolometers. Five manganin/copper cables with 100pin MDM connectors made by Tekdata [112] connect the 4 K board to the MCE electronics. The manganin section is placed between the 4 K board and the 40 K stage, while the copper section starts at the 40 K stage and ends at the 300 K MCE electronics. The ABS array assembly was successful; the few silicon components that broke were replaced, and all electrical lines from 300 K to 300 mK are continuous and without shorts.
Figure 6.5: The picture above shows the ABS detector array mounted on an aluminum bracket comprised of triangular posts, that places it at the focal plane of the optics. Note that the copper structure behind the array holds the 4K series array board.
Chapter 7

Conclusions

As of September 2012 the ABS experiment was observing from the Atacama desert of Chile (see Figure 7.1), with an array sensitivity of $31\mu\text{K}\sqrt{s}$. Multiple design steps were taken to reduce systematics. For example, we cryogenically enclosed the telescope to reduce beam spill-over contamination and also cooled the telescope to reduce its emission; added a continuously rotating HWP to modulate the polarized signal away from the atmosphere $1/f$ spectrum; designed a very quiet air bearing system for the HWP; used corrugated feedhorns and planar OMTs to couple the sky signal with low cross-polar leakage; installed a linear motor for fast azimuth scanning; defined a cold stop at 4K, and placed all optical elements near it; and built two baffles around the cryostat window and HWP to minimize ground pick up.

The raw sensitivity of the ABS experiment, not including systematic effects, is driven by three main factors.

- **Optimizing the individual detector sensitivity.** Development of the ABS single pixel design yielded bolometers with NEP comparable to the expected photon shot noise, even while only achieving 50% optical efficiency.

- **Building and operating the array of ABS detectors near the optimal condition.** Many tests were carried out to validate the performance of the SQUID multiplexing read out, the shunt biasing chips, the superconducting circuit boards, and all
Figure 7.1: Picture of the ABS telescope deployed in Chile with the HWP mounted but the baffle and ground screen off
read out wiring. This led to only one broken SQUID channel out of 528, and a measured amplifier noise, including aliasing, of only a few percent of the detector signal. Up to 458 TES bolometers may be biased on the transition, and on average 435 are biased during observations, which is equivalent to a 90% yield of operating detectors. This high percentage is a result of the fabrication and testing of eight detector wafers (100 pixels each), along with individual TES pixel screening through dark cryogenic pod tests before final installation. The cryogenic performance of the ABS array in the cryostat is excellent; it reaches temperatures as low as 290 mK, and is actively servoed during observations no higher than 330 mK, 20 mK below the design target. Three circumstances that negatively effected the array sensitivity are: four pods of detectors have low efficiency due to fabrication issues at the edge of wafer 14; the bottom half of the array (wafer 14 and 15) holds detectors with bandpasses that are shifted 15 GHz above the 145 GHz target, and the blackened baffle added excess optical loading until August 2012, when it was replaced by a shiny baffle.

• Maximizing the instrument’s observing time. The amount of data collected will depend on the year round observing conditions in Chile, and more importantly on the stability of the experiment. The SQUID read out and detectors can be re-biased in less than a minute with V-Φ and I-V auto tuning scripts. These help channels recover from falling off the transition or having their SQUIDs unlock, and are typically performed every hour and always after elevation changes. The adsorption fridges have a hold time of nearly 42 hours, and require 6 hours to cycle, permitting the telescope to observe 88% of the time. Down time due to software glitches or mechanical issues with the HWP rotation and azimuth scan, was high early in the deployment period. The stability and monitoring of the instrument greatly improved after the first few months of deployment.

Given the final configuration of the ABS array, Table 7.1 summarizes the possible avenues for improving its sensitivity. Three cases are explored: when using the blackened baffle, when using the shiny baffle, and the expected sensitivity from early test results. The
other sensitivity improvements considered are:

- **Constructing a uniform array of detectors.** Multiple fabrications dedicated to improving uniformity could be used to yield detector wafers that closely approach the targeted $T_c$, $G$, optical efficiency and bandpass values. Equally important would be to check for and reject wafers with unexpected fabrication problems such as impurities in the superconducting layers, breaks in the OMT area, stressing of the TES film, dirt or scratches on the pixels, etc. The improvements in fabrication coupled with single pixel tests, would yield the desired array uniformity by taking advantage of the modular design of the ABS focal plane, where every detector may be replaced individually. A large number of fabricated and tested wafers, could compensate for unpredictable variations in the detector fabrication. If all detectors in the ABS array were equal to the median of wafer 4, then the array NET would be $\sim$30% better.

- **Increasing the detector optical efficiency.** The ABS pixels have detector optical efficiencies of $\sim$45%. Doubling this would improve the array NET by $\sim$40%. Many of the design changes made (modifications to the SiO$_2$ recipe, the CPW-MS transition, the stub filter design, and the meander Au thickness) during the ABS detector fabrication were done in hopes of improving the detector optical efficiency; unfortunately this goal was not achieved and the array uniformity suffered. Improving the efficiency of the ABS pixel design will require further development, but could pay off handsomely. A quick design fix, is to remove the on chip filter and rely on a free space filter to define the band. SPTpol and ACTpol made this choice, with early tests indicating the desired optical efficiency improvement. This choice was not initially possible for ABS, because the cold stop aperture would require larger free space filters than were available at the time. Now 30 cm filters are available from the Cardiff group [117].

- **Increasing the optical efficiency between the feedhorn and the sky.** This efficiency is driven by the beam edge taper at the cold stop and the multiple absorptive and reflective losses from the various filters. Increasing the cold stop size, and hence
the size of all the filters and HWP would lead to much of the desired improvement. This is not a likely change for the ABS cryostat, but should be considered for the design of a similar experiment. Re-designing the telescope mirrors may improve by a small amount the edge taper. Improving the efficiency of the optical elements from 65% to 85% would decrease NET by \(\sim13\%
\).

- **Have all array detectors work.** About 435 ABS TES bolometers work during observations; if all 480 worked with the same average sensitivity, then the array NET would decrease by \(\sim5\%\).

The other improvements considered in Table 7.1 are a combination of the ones just described. Fixing the baffle excess loading provided a substantial improvement in sensitivity. In the long term, populating the array with a uniform set of high efficiency pixels may double the sensitivity of ABS.

ABS with the achieved array sensitivity of \(31\mu\text{K}\sqrt{s}\), individual detector beams of \(35\arcmin\), and 8760 hours of observations, could detect the primordial B-mode power spectrum to \(2\sigma\) if \(r \geq 0.1\).
### Table 7.1

<table>
<thead>
<tr>
<th>Improvement condition</th>
<th>NET with blackened baffle $\mu K \sqrt{s}$</th>
<th>NET with shiny baffle $\mu K \sqrt{s}$</th>
<th>Expected NET $\mu K \sqrt{s}$</th>
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</thead>
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<tr>
<td>0</td>
<td>39</td>
<td>29</td>
<td>26</td>
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<tr>
<td>1</td>
<td>Uniform detector array, all equal to ABS1-4 median</td>
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<td>21</td>
</tr>
<tr>
<td>2</td>
<td>Improve detector efficiency to 90%</td>
<td>24</td>
<td>18</td>
</tr>
<tr>
<td>3</td>
<td>Improve optical efficiency of filters and cold stop to 85%</td>
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<td>26</td>
</tr>
<tr>
<td>4</td>
<td>All detectors work</td>
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<td>28</td>
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<td>1 and 2</td>
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<td>1, 2, and 3</td>
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</tr>
<tr>
<td>7</td>
<td>1, 2, 3, and 4</td>
<td>17</td>
<td>13</td>
</tr>
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</table>

Table 7.1: Table listing the array NET with the blackened baffle, the shiny baffle, and expected from estimates. Here NET does not take into account the HWP modulation efficiency. The blackened baffle and shiny baffle NETs come from sky dip measurements, while the expected NET is derived from Table 5.2. The array improvements considered are: building a uniform array of detectors, improving the detector optical efficiency, improving the efficiency of the optical elements, having all the detectors in the array functional. Each of these cases is considered individually given the characteristics of the built ABS array. The NET calculations from efficiency improvements include both the gain in signal amplitude and the increase of photon noise, assuming that half the array (wafer 4 and 11) is photon noise limited, and the other half is detector noise limited. The three main factors to target for improving the array NET are: replacing blackened baffle by shiny baffle, increasing detector optical efficiency, and making the detectors in the array uniform.
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