Probing the Cosmic Microwave Background Radiation Using Large Superconducting Detector Arrays

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Abstract

The discovery and measurements of the Cosmic Microwave Background (CMB) radiation have laid the foundation of the current cosmological model, in which the universe is primarily composed of dark energy and dark matter. Higher-precision measurements of the CMB temperature and polarization anisotropies, enabled by installation of larger numbers of superconducting detectors, will unveil more information about the universe.

This thesis will present my work in the Advanced Atacama Cosmology Telescope Polarimeter (AdvACT) and the Simons Observatory (SO). Both the AdvACT and SO are located at ~5200 m altitude on Cerro Toco in the Atacama Desert in Chile. The AdvACT is the second and final instrument upgrade of the Atacama Cosmology Telescope (ACT), and is one of the most sensitive ground-based instruments for measuring the CMB anisotropies at small angular scales. The total number of detectors that the SO plans to deploy is one magnitude higher than the AdvACT. The major content of this thesis is related to the development, design, integration and test of the multichroic detector arrays of these two instruments.
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\textsuperscript{1}From Chinese ancient text “Discourse on Teachers” by Han Yu.
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Related Works

Part of the work in this thesis has been presented in conferences and published. Here I describe the presentation and publications involved. My contribution to each publication will be also described alongside.

- Oral presentation, July 2018, “Performance of the Advanced ACTPol Low Frequency Array”.


- Proceeding “Performance of the advanced ACTPol low frequency array” [15].

- Proceeding “Assembly and Integration Process of the High-Density Detector Array Readout Modules for the Simons Observatory” [16].

I wrote the bulk part of the proceedings as the first author and include much of the content in this thesis. In Chapter 2 and Chapter 3, the content describing the assembly of the first detector array for AdvACT is based on [11], while the content related to the characterization and performance of the last AdvACT array in lab is based on [15]. In Chapter 4, the integration process of prototype UMMs are based on [16].
Chapter 1

Introduction

The discovery of the Cosmic Microwave Background [17] has laid the foundation of the modern cosmology, i.e. the $\Lambda$CDM model, in which the universe is mainly composed of dark energy and dark matter. As we entered the era of precision observation, the focus of CMB research has been shifted to probing the $\sim \mu K$ level anisotropies in CMB temperature and polarization using superconducting detectors, which are densely packed in detector arrays. Chapter 1 gives an overview of the CMB science, including the purpose of the observation and an introduction of the current CMB telescopes. Chapter 2 and Chapter 3 present the integration process and the performance of the Advanced ACTPol (AdvACT) detector arrays, which is the current camera of the Atacama Cosmology Telescope (ACT) [18]. The test performance is focused on the last detector array, which is the low frequency array, because these detectors are operated with unprecedentedly low bias energy. Chapter 4 describes a prototype detector array for a next generation instrument called the Simons Observatory (SO) [19]. The SO detector array will maintain the same sensitivity while only occupying half of the focal plane area as in the ACT. Since the detector array is still under developing, Chapter 4 will focus on the part of the innovative cold readout system.
1.1 CMB and Cosmology

1.1.1 CMB

The CMB is the oldest radiation we receive from the universe (detailed description can be seen in [20]). Shortly after the Big Bang, the universe was filled with nearly homogeneous hydrogen plasma. The plasma was so hot and dense that nearly all electrons were free and due to the Thompson scattering from those electrons, the mean free path of photons was short compared to the Hubble radius $c/H(t)$, which is the observable distance. Hence the universe was opaque at that time. The universe expanded and got cooler, until 380,000 years after the Big Bang when the temperature dropped to around 3000 K. Once electrons combined with protons into neutral hydrogen atoms, which couldn’t scatter the protons anymore, the photon’s mean free path was then long enough to travel through the universe. Therefore the universe became transparent and the thermal radiation was released, which is known as decoupling or sometimes recombination epoch. After billions of years of expansion, the universe has cooled and the wavelength has stretched such that this relic thermal radiation is observed today in the microwave range, and is referred to the Cosmic Microwave Background. Measurements of the CMB nowadays reveal that it consists of extremely homogeneous and isotropic blackbody radiation at a temperature around $2.72548\pm0.00057$ K [21].

1.1.2 CMB Temperature Anisotropies

The CMB is highly homogeneous and isotropic, but it still contains small anisotropies, which store rich information. After removing the dipole pattern due to the relative movement of the Earth, close examination reveals smaller-scale anisotropies shown as a few $\mu$K variation in the CMB temperature maps of the sky. These anisotropies reflects the unevenness of energy density distribution in the primordial universe. In
the primordial plasma, the over dense regions pulled the matter inward by gravity, while the photons emitted by the plasma exerted pressure to push away, forming acoustic oscillations. The CMB map is similar to a snapshot of the universe’s energy density distribution at the moment of decoupling. The acoustic oscillations import features that are uniquely determined by the energy composition of the universe.

Figure 1.1: The CMB temperature power spectrum measured by Planck [1], showing acoustic peaks in the early universe. Figure courtesy of ESA and Planck Science Team [1].

Studies of the statistical patterns in the CMB maps have laid the foundation for the current standard cosmological model. Details are described in [22]. Since the CMB is observed on the celestial sphere, it is useful to decompose the anisotropies of CMB temperature map into spherical harmonics:

\[
\Delta T(\theta, \phi) = \sum_{l=1}^{\infty} \sum_{m=-l}^{l} a_{lm} Y_{lm}(\theta, \phi),
\]

(1.1.1)

where the \( Y_{lm}(\theta, \phi) \) are Laplace’s spherical harmonics; \( \theta \) and \( \phi \) are the spherical coordinates; \( l \) is the multipole of the harmonics and indicates angular scales by:

\[
\theta \sim \frac{\pi}{l};
\]

(1.1.2)
while $m$ indicates the orientation of the harmonics for each $l$. The complex harmonic coefficient $a_{lm}$, which specifies the amplitude and phase of the corresponding harmonic, can be calculated by:

$$a_{lm} = \int Y_{lm}^*(\theta, \phi) \Delta T(\theta, \phi) d\Omega.$$  \hspace{1cm} (1.1.3)

The temperature anisotropies are commonly assumed to comprise a Gaussian random field, which has been confirmed by the Planck [23] and WMAP [24] satellites, hence the resulting harmonic coefficients are also Gaussian random variables with $\sigma$ only related to $l$. We therefore explore the statistical pattern associated with $l$ only, and define the angular power spectrum $C_l$ with a given $l$ as:

$$C_l \delta_{ll'} \delta_{mm'} = \langle a_{lm} a_{l'm'}^* \rangle,$$  \hspace{1cm} (1.1.4)

with the brackets denoting the average over a set of $m$ values. The measured $C_l$ can be calculated as:

$$C_l = \frac{1}{2l+1} \sum_{m=-l}^{l} |a_{lm}|^2.$$  \hspace{1cm} (1.1.5)

Note that the sum has $2l+1$ terms. Each $C_l$ has a fundamental precision limit:

$$\Delta C_l = \sqrt{\langle |a_{lm}|^2 \rangle - C_l} = \sqrt{\text{Var}(|a_{lm}|^2)}.$$  \hspace{1cm} (1.1.6)

For a variable $X$ with a Gaussian distribution centered at zero and standard deviation $\sigma$, $E[X^4] = 3 \sigma^4$. Eq. (1.1.6) then becomes:

$$\Delta C_l = \sqrt{\langle |a_{lm}|^4 \rangle - 2C_l^2} = C_l \sqrt{\frac{2}{2l+1}}.$$  \hspace{1cm} (1.1.7)

which is called “the cosmic variance”.

5
Figure 1.1 shows the CMB power spectrum from Planck measurements [1]. There are three major effects that determine the shape of the power spectrum curve. The first one is the photon diffusion effect [25]. The recombination era has a finite period, during which the photon scattering tended to smooth the anisotropies at small scales, resulting in the damping tail at high $l$. The second one is the gravitational redshifts (the ordinary Sachs-Wolfe effect [26]) when the photons are escaping from potential wells. This effect dominates at low $l$ features. Last but the most important are the peaks and troughs in the middle region of the power spectrum, resulting from the aforementioned acoustic oscillations. These peaks are the modes which completed an integer number of oscillations at the moment of recombination and therefore were reaching the peak amplitude, while the troughs are ones that returned to the mean amplitude. The first peak represents the modes that went through only one compression. The relation between the angular position of the first peak and the calculated sound horizon is consistent with flat geometry, indicating the universe is spatially flat (Figure 1.2), and the total energy density is equal to the critical density. The second peak is for the modes that completed one rarefaction and one compression cycle. Since the baryonic matter itself also contributed to the gravitational potential, the compressions are stronger than rarefactions and the amplitudes of the odd peaks are higher than the amplitudes of the even peaks. Thus, the ratios of the ampli-
tudes reveal the abundance of the baryonic matter. In summary, the peaks’ positions and amplitudes establish the current cosmological model: the ΛCDM model. In the ΛCDM model, the universe is primarily composed of dark energy (\(\Lambda\)) and cold dark matter, with less than 5% baryonic matter.

1.1.3 CMB Polarization Anisotropies

The focus of recent decades’ CMB research has shifted to the CMB polarization patterns, which were generated from Thomson scattering in the presence of temperature quadruple anisotropies [27]. As shown in Figure 1.3, the scattered photon acquires a linear polarization with amplitude corresponding to the different intensities of incident radiation coming from two perpendicular directions, while it would remain unpolar-
ized without the quadruple anisotropies. The polarization is usually described by Stokes parameters $I, Q, U, V$ (shown in Figure 1.4). For radiation propagating along the $z$ axis, its electric field along the $x$ and $y$ axes can be expressed as:

$$E_x = \hat{x}E_x e^{ikz-\omega_0 t + \phi_x},$$

$$E_y = \hat{y}E_y e^{ikz-\omega_0 t + \phi_y},$$

(1.1.8)

where $E_x$ and $E_y$ are the electric field amplitude components, $\omega_0$ is the angular frequency, and $\phi_x$ and $\phi_y$ are the phases. The Stokes parameters can be expressed as:

$$S = \begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} E_x^2 + E_y^2 \\ E_x^2 - E_y^2 \\ 2\text{Re}(E_x E_y^*) \\ 2\text{Im}(E_x E_y^*) \end{bmatrix} = \begin{bmatrix} E_x^2 + E_y^2 \\ E_x^2 - E_y^2 \\ 2E_x E_y \cos \Delta \phi \\ 2E_x E_y \sin \Delta \phi \end{bmatrix},$$

(1.1.9)

where $\Delta \phi \equiv \phi_x - \phi_y$ is the phase difference between $E_x$ and $E_y$. The intensity is $I$, while $V$ is the circular polarization component and therefore is usually omitted. The parameters $Q$ and $U$ describe the linear polarization. The values of $Q$ and $U$ at a point on the sky can be directly measured by a polarimeter, but are dependent on the coordinate system defined. We would like to connect $Q$ and $U$ with coordinate-independent quantities, with which we can study the spatial patterns as in the case for the CMB temperature. When rotating the $xy$ axes by an angle $\alpha$, the $Q'$ and $U'$ in the new coordinates become:

$$Q' = Q \cos 2\alpha + U \sin 2\alpha,$$

$$U' = -Q \sin 2\alpha + U \cos 2\alpha.$$  

(1.1.10)

If we construct two complex field $Q + iU$ and $Q - iU$, in the new coordinates they becomes:

$$Q' \pm iU' = e^{\pm i2\alpha}(Q \pm iU).$$

(1.1.11)
According to [28] and [29], a quantity $\eta$ that transforms under rotation $\alpha$ as

$$\eta' = e^{\pm is\alpha} \eta$$  \hspace{1cm} (1.1.12)

has spin weights $s$. Fields $Q \pm iU$ therefore can be expanded by spherical harmonics with spin weights $\pm 2$:

$$(Q + iU)(\theta, \phi) = \sum_{lm} a_{\pm 2,lm}(\pm 2Y_{lm}(\theta, \phi)),$$ \hspace{1cm} (1.1.13)

where as in Equation 1.1.1, $\theta$ and $\phi$ are the spherical coordinates on the sky. The signal can be further decomposed into two independent quantities– the gradient-like $E$-mode and the curl-like $B$-mode (shown in Figure 1.4):

$$a_{lm}^B = \frac{i}{2}(a_{2,lm} - a_{-2,lm}),$$
$$a_{lm}^E = \frac{1}{2}(a_{2,lm} + a_{-2,lm}).$$  \hspace{1cm} (1.1.14)

Similar to the case of temperature anisotropies, $C_{EE}^l$ and $C_{BB}^l$, the power spectra of the $B$-mode and $E$-mode patterns, are defined as:

$$C_{EE}^l \delta_{ll'} \delta_{mm'} = \langle a_{lm}^E a_{lm}^{E*} \rangle,$$
$$C_{BB}^l \delta_{ll'} \delta_{mm'} = \langle a_{lm}^B a_{lm}^{B*} \rangle.$$  \hspace{1cm} (1.1.15)

As was pointed out in [28] and [29], sound waves in the early universe produce any $B$-modes but gravitational waves can.
1.1.4 Primary and Secondary Anisotropies

In the paradigm of inflation \[30\] \[31\], primordial quantum fluctuations are the sources for the primary anisotropies in the CMB polarization. Under the current standard cosmological model, there are two types of fluctuations: the scalar perturbations and the tensor perturbations seeded by inflation. We model both types of perturbations to have power-law power spectra in terms of the 3D wave vector amplitude $\kappa$:

\[
\begin{align*}
P_s &= A_s \left( \frac{\kappa}{\kappa_0} \right)^{n_s-1}, \\
P_t &= A_t \left( \frac{\kappa}{\kappa_0} \right)^{n_t},
\end{align*}
\]  

where $A_s$ and $A_t$ specify the amplitudes of the scalar and tensor perturbations, $n_s$ and $n_t$ are the scalar and tensor spectral indices, and $\kappa_0$ is the pivot scale. Similar to the temperature anisotropies, the $E$-mode pattern was generated by the scalar perturbations in the primordial plasma; it therefore is an independent probe of the
cosmological model. As shown in Figure 1.5, the $EE$ power spectrum has a similar peak structure compared to the $TT$ spectrum but is $90^\circ$ out of phase [32] [33]. The reason for the phase difference is that $EE$ traces the velocity of the fluid instead of the density, since the primary sources of quadrupole anisotropies in the primordial plasma are Doppler shifts. Tensor perturbations are the source for the primary $B$-mode pattern, which is dependent on the tensor-to-scalar ratio $r$ [34]:

$$r = \frac{A_t}{A_s}.$$  \hspace{1cm} (1.1.17)

Note that gravitational waves are tensor perturbations. Thus, the $B$-mode power spectrum on large scales ($l \sim 80$) is the most promising tool for probing primordial gravitational waves produced during an inflationary epoch.

![Figure 1.5: Up-to-date summary of the $TT$, $EE$ and $BB$ power spectra measured and made by Planck [1], ACTPol [2], SPTPol [3] [4], BICEP2/KECK [5] and POLARBEAR [6] [7]. The $TT$ spectrum shows acoustic peaks with the first one located at $\ell \sim 200$, corresponding to $\sim 1^\circ$. The $EE$ spectrum also includes peaks that are $90^\circ$ out of phase with respect to $EE$.](image)
There are also secondary anisotropies in the CMB maps as the light propagates from the last scattering surface to the current time. The CMB essentially serves as a backlight, providing a powerful probe for the reionization era and for the structure formation. Secondary anisotropies were generated from two types of interactions. The first type consists of interactions between photons and gravitational potential wells, such as the Integrated Sachs-Wolfe (ISW) effect, and the gravitational lensing effect \[35\], which was first detected by ACTPol \[36\]. The ISW is similar to the aforementioned non-integrated Sachs-Wolfe effect, in which photons were blueshifted as they were escaping gravitational wells, but instead ISW occurred after the last scattering and occurred because gravitational collapse in an expanding universe means the potential wells evolve as the photons traverse them, leading to the gravitational redshifts or blueshifts. The ISM, together with the non-integrated Sachs-Wolfe effect, dominates the $TT$ power spectrum at large angular scales. The gravitational lensing effect is caused by matter overdensities changing the trajectory of photons, which rearranges parts of the $E$-mode patterns into $B$-modes at high $l$. The second type of secondary anisotropies is the interactions between photons and free electrons via inverse Thomson scattering during or after the reionization epoch. The interaction between photons and electrons generated during the reionization period results in a local peak in the $EE$ power spectrum at low $l$. Measurement of its amplitude is a probe for the reionization optical depth $\tau$. Two other prominent examples are the thermal Sunyaev-Zel’dovich (tSZ) \[37\] and kinetic Sunyaev-Zel’dovich (kSZ) \[38\]. The corresponding CMB distortions are respectively due to the temperature and the motion of clusters \[39\], which are filled with hot and ionized gas. A review of the SZ effects can be seen in \[40\].
1.2 CMB Experiments

Large superconducting detector arrays enable the precise measurement of the CMB anisotropies if the detectors achieve background-limited performance, among which the most mature technology is the Transition-Edge Sensor (TES) [41]. The overall sensitivity of CMB instruments is usually expressed as the noise equivalent temperature (NET):

\[ NET_{\text{tot}} \propto \frac{NET_{\text{det}}}{\sqrt{YN}}, \tag{1.2.1} \]

where \( N \) is the number of deployed detectors and \( Y \) is the detector yield. The sensitivity for each detector, \( NET_{\text{det}} \), depends on the optical loading, with a large portion contributed by the atmosphere. Another characteristic of the CMB instruments is the angular resolution:

\[ \Delta \theta \sim 1.2 \frac{\lambda}{D}, \tag{1.2.2} \]

where \( D \) is the telescope (mirror) aperture size and \( \lambda \) is the radiation wavelength. The angular resolution determines the range of \( l \) that an instrument can measure on the power spectrum.
There are three categories of CMB instruments, which are respectively installed on satellites, balloons or the ground. The satellite instruments, such as *WMAP* and *Planck*, are able to scan the full sky. The signal is not affected by the noise from the atmosphere, but the payload limits the telescope resolution and the focal plane area. As shown in Figure 1.6 for ground-based instruments, the sensitivity is degraded due to the atmospheric loading, while the telescope mirror could be larger, allowing measurements with resolution on the level of a few arcmins. In addition, without the payload limitations, the ground-based instruments compensate the low sensitivity by increasing the total number of detectors. The balloon-based instruments are considered to be an intermediate between the satellite and ground-based telescope with lower noise but shorter observation periods.
Ground-based CMB instruments has experienced rapid development during the past two decades. Among the areas with low PWV, two traditional sites are the Atacama Desert in Chile and the South Pole. In the Atacama Desert, the representative large-aperture telescopes include the ACT [42] and POLARBEAR [43], while the CLASS [44] probes the CMB polarization at large angular scales. In the South Pole, the representative large and small-aperture telescopes are the SPT [45] and BICEP/KECK [46] [47]. All of the instruments mentioned above are at the current frontier of CMB measurements with high sensitivities that have improved through generations of instrument upgrades with increasing total numbers of detectors. The detector counts of the current ground-based instruments, including the AdvACT, are on the scale of $10^4$. Figure 1.7 categorizes CMB instruments in three stages; the current instruments are Stage III. The second project that will be described in this thesis, the SO, plans to deploy around 60,000 bolometers in total, placing it closer to the class of Stage-IV CMB instruments [48].
1.2.1 Science Goals of Small Angular Instruments

As mentioned previously, one of the reasons that the ground-based instruments are indispensable is their capability of measuring CMB anisotropies at small angular scales, which fills out the details in the full-sky CMB maps measured by satellite and balloon-borne instruments. Since both of the two projects involved in this thesis have arcmin-level angular resolutions, this subsection is dedicated to the science goals of CMB measurements at high $l$.

First and foremost, the high $l$ measurements on the CMB temperature maps probe the primary anisotropies and constrain the cosmological parameters, such as the baryonic matter density $\Omega_b h^2$ and the dark matter density $\Omega_c h^2$. It also gives a precise constrain of the sound horizon by measuring more acoustic peaks. The CMB power
spectra at $1000 < l < 3000$ measures the damping tail and constrain the spectral index $n_s$, which breaks the degeneracy of the effect from $n_s$ and $r$ at large angular scales. Secondly, clusters and other high-density structures are on the scale of arcminutes; therefore the CMB anisotropies at high multipoles include secondary anisotropies that were generated by the tSZ, kSZ (first found by the ACT team [49]) and gravitational lensing effects [50]. These measurements map the distribution of the dark matter and the momentum field of galaxies [51], probing the growth rate of structure and providing constrains on the total mass of neutrinos [52]. In addition, combining with data from optical telescopes, cosmic infrared background (CIB) measurements or X-ray surveys such as eROSITA [53], strengths the power of the CMB in these areas; the CMB instruments at small angular resolutions can detect astronomical objects such as clusters and point sources (recent ACT cluster and source catalogs can be seen in [54] and [55]).

1.3 The Advanced ACTPol Instrument

The ACT is located at 5190 meter altitude on Cerro Toco in the Atacama Desert in Chile, sharing the same plateau with a few other telescopes such as the Cosmology Large Angular Scale Surveyor (CLASS) and the Simons Observatory. The design of the telescope is off-axis Gregorian [56], with a six-meter primary mirror and a two-meter secondary mirror. After being reflected by the secondary mirror, radiation enters the receiver cabin, which contains three optics tubes each housing one detector array [57].

There are three generations of receivers. The first generation is the Millimeter Bolometer Array Camera (MBAC) [58], observing the CMB temperature anisotropies from 2008 to 2010 using Transition-Edge Sensor (TES) bolometers operated at 300 mK. Then the instrument was upgraded to the Atacama Cosmology Telescope Polarimeter
Major changes include coupling TES bolometers to planar orthomode transducers to probe the CMB polarization and operating the TESes at 100 mk, by upgrading to a cryostat cooled by a He\textsuperscript{3}/He\textsuperscript{4} dilution fridge backed by a second set of pulse tube refrigerators. The last upgrade, operating from 2016, is the Advanced ACTPol (AdvACT), which is similar to ACTPol except for upgrades in the cold readout system and detector design that enable use of almost twice the number of detectors as in ACTPol. Details about other new parts, such as the cold optical elements and the structure of the cryostat, are described in [59], [60] and [42].

The detector arrays of the AdvACT consist of one high-frequency (HF) array sensitive to 220 and 150 GHz CMB signals, two middle-frequency (MF) arrays sensitive to 150 GHz and 90 GHz, and one low-frequency (LF) array sensitive to 27 and 39 GHz. They have been observing since June 2016, April 2017 and January 2020 respectively, with up to 6000 detectors deployed in total, which is nearly double the detector counts in ACTPol. Table 3.1 shows the parameters of the AdvACT detector arrays. The detector counts at different frequency bands are designed to optimize the foreground removal. Details of the detector arrays of the Advanced ACTPol will be described in Chapter 2.

Table 1.1: Parameters of the detector arrays

<table>
<thead>
<tr>
<th>Array</th>
<th>Freq (GHz)</th>
<th>Bandwidth (GHz)</th>
<th>TES</th>
<th>NET\textsubscript{CMB} (\mu K\sqrt{s})</th>
<th>Beam (arcmin)</th>
</tr>
</thead>
<tbody>
<tr>
<td>HF</td>
<td>220</td>
<td>100</td>
<td>1059</td>
<td>20</td>
<td>0.9</td>
</tr>
<tr>
<td>MF/HF</td>
<td>150</td>
<td>41</td>
<td>2816</td>
<td>5.4</td>
<td>1.3</td>
</tr>
<tr>
<td>MF</td>
<td>90</td>
<td>38.5</td>
<td>1760</td>
<td>5.8</td>
<td>2.2</td>
</tr>
<tr>
<td>LF</td>
<td>39</td>
<td>18</td>
<td>192</td>
<td>32</td>
<td>4.8</td>
</tr>
<tr>
<td>LF</td>
<td>27</td>
<td>12</td>
<td>192</td>
<td>35</td>
<td>7.1</td>
</tr>
</tbody>
</table>
1.4 Simons Observatory

The Simons Observatory is a joint effort of the ACT and Simons Array team, aiming to build a suite of telescopes in the Atacama Desert to measure the CMB temperature and polarization anisotropies. The SO plans to build three small-aperture telescopes (SAT) \(^{61}\) each with a 0.5 m primary mirror, and one large-aperture telescope (LAT) \(^{62}\), with a 6 m primary mirror, which is similar in size to the ACT. The LAT is able to accommodate 13 optics tubes with three detector arrays in each, while each SAT accommodates one optics tube, with seven detector arrays. A total of 49 multichroic TES bolometer arrays will be deployed, containing more than 60,000 detectors in total that span six frequency bands centered from 27 to 280 GHz.

The SO detector array unit is a hex-shaped six-inch universal focal-plane module (UFM), which will be described in detail in Chapter \(^\Box\). Each UFM consists of a cold multiplexing unit, called the universal multiplexing module (UMM), and \(\sim2000\) TES bolometers and associated optical coupling. The UFM implements an innovative microwave multiplexing system, which reads out all of the detector signals using only one or two pairs of coaxial cables.
1.5 High Density CMB Detector Array Structure

Although different projects may use different components in the detector array, they still share great similarities in detector array structure. As shown in Figure 1.8, a detector array usually consists of three major parts: radiation collectors, pixels, and a cold multiplexing system. The most critical parts are the pixels, comprising an array of antenna-coupled bolometers that couple radiation to transmission lines and measure the dissipated heat using superconducting thermistors such as TES. For polarimeters, lithographed optical elements (antennas or orthomode transducer probes) belonging to the same pixel are positioned at orthogonal orientations to measure the light polarization. The wave collector part, such as a feedhorn or lens, focuses the radiation to the pixel. Last but not least, since detectors are operated in cryostats,
a cold multiplexing system is needed to condition the output signals and reduce the cables connecting the inside and outside of the cryostats.
Chapter 2

Advanced ACTPol Detector Arrays

The Advanced ACTPol (AdvACT) is the second instrumental upgrade of the Atacama Cosmology Telescope. The upgrade consists of four multichroic detector arrays observing CMB intensity and polarization anisotropies at five frequency bands centered from 27 GHz to 220 GHz. Three major advances in the detector array technology are involved. First and foremost, the detector wafers, which are developed and fabricated at NIST, have been scaled up to 150-mm diameter silicon wafers. This improvement over the previous 75 mm wafers is possible because the uniformity across the wafer has been greatly improved by using new sensor and insulator materials (AlMn \cite{63,64} and SiN\(_x\) respectively). The resulting higher pixel density, along with the fact that all pixels are multichroic, doubles the number of detectors previously deployed in the ACT camera. Second, AdvACT uses a new time-division multiplexing (TDM) architecture \cite{65} with only two SQUID stages, which reduces the number of SQUID bias and feedback lines in the warm electronics. The freed channels in the warm electronics are then used for multiplexing extra rows of TESes, allowing the system to read out double the number of detectors without upgrading the hardware of the
warm electronics. Additionally the new TDM architecture reduces the number of channels multiplexed by each chip, allowing flexibility in the system design. Last but not least, AdvACT implements newly developed superconducting flexible cable (flex) with 70 \( \mu \text{m} \) bondpad pitch compared to 200 \( \mu \text{m} \) pitch which is commercially available, electrically connecting the densely packed detectors to the readout system and accommodating the difference in thermal contraction between the readout’s mount and the detector wafer.

To date, all arrays have been installed into the telescope, including one high-frequency (HF) array sensitive to 220 and 150 GHz CMB signals, two middle-frequency (MF) arrays sensitive to 150 GHz and 90 GHz, and one low-frequency (LF) array sensitive to 27 and 39 GHz. They have been observing since June 2016, April 2017 and January 2020 respectively, with up to 6000 detectors deployed in total. This chapter describes the assembly process, lab tests and on sky performance of the AdvACT detector arrays.

### 2.1 Basic Bolometric Unit

The basic bolometric unit of the AdvACT detector array is a dichroic single pixel containing a planar orthomode transducer (OMT) and four TESes [66], as seen in
Initially, the spline-profiled feedhorn optically couples radiation from free space to each single pixel. The incoming radiation is then radioactively coupled to the niobium transmission line on each pixel by the OMT, where there are four sword-shaped probes suspended on a SiN membrane, splitting the radiation into X and Y polarization. The signals are further split into two frequency channels in a diplexer. Before arriving to the TESes for detection, signals with the same frequency bands from probes separated by 180° (i.e. from either the vertical probes or the horizontal ones) are recombined by hybrid tees. Therefore the four TES bolometers respectively measure the signals in two frequency bands and two polarizations. Each AlMn TES is fabricated on the thermally isolated SiN membrane with only four legs weakly connecting to the bulk silicon providing the thermal bath. The width of the legs determines the thermal conductance of the TES. As shown in Figure 2.2, the signal enters the TES island and dissipates heat through a lossy gold meander, resulting in a change in temperature and sequentially in TES resistance. A layer of PdAu is deposited with the appropriate thickness to set the TES heat capacity, as needed to define the target time constants (see Section 3.1.2).
2.1.1 The TES and Negative Electro-thermal Feedback

The TES is a bolometric sensor that is able to measure small temperature variations resulting from absorbed optical power. When operating between the superconduct-
ing and normal states, the TES resistance has a strong dependence on temperature (shown in Figure 2.3). In the thermal equilibrium state, the thermal equation goes as:

\[ P_{\text{bath}} = P_\gamma + P_{\text{Joule}}, \quad (2.1.1) \]

where \( P_\gamma \) is the incoming radiation power, and \( P_{\text{Joule}} = I^2 R_{\text{TES}} \) is the heating power when electric current \( I \) passes through the TES with resistance \( R_{\text{TES}} \). See the left side of Figure 2.4 for a schematic representation of Equation (2.1.1). The heat transferred to the thermal bath is \( P_{\text{bath}} \) and is assumed to follow this well-tested model:

\[ P_{\text{bath}} = \kappa T^n - \kappa T^n_{\text{bath}}, \quad (2.1.2) \]

where \( T \) is the temperature of the TES, and \( \kappa \) and \( n \) are constant parameters related to the TES geometry design. The thermal conductance \( G \) to the bath is defined as:

\[ G \equiv \left. \frac{dP_{\text{bath}}(T)}{dT} \right|_{T_c} = n\kappa T_c^{n-1} \quad (2.1.3) \]

with \( T_c \) being the critical temperature of the TES. For small variations of the incoming radiation power, equation (2.1.1) becomes

\[ P_{\text{bath}} + C \frac{dT}{dt} = P_\gamma + P_{\text{Joule}}, \quad (2.1.4) \]

where \( C \) is the TES heat capacity.
As seen in the right side of Figure 2.4 for AdvACT, each TES is in parallel with a shunt resistor, the resistance (≈ 200μΩ) of which is much lower than the TES resistance (on the scale of several mΩ). Therefore with constant bias current $I_{bias}$, the TES is biased with nearly constant voltage $V \sim I_{bias}R_{shunt}$. When the TES is in unequilibrium state, a small increase of photon power will increase the temperature of the TES, resulting in an increase of $P_{bath}$ and thus the TES resistance. Since the TES is voltage-biased, the electro-heating power $P_{Joule} = V^2/R_{TES}$ is decreased. The term $C \frac{dT}{dt}$ in equation (2.1.4) becomes negative, so the TES temperature tends to drop down to its original value.
2.2 Time Division Multiplexing System

Figure 2.5: (Left): Circuit diagram for a single TES. (Right): Conceptual scheme of the TDM system consist of \( n \) rows and \( m \) columns, with each unit reading a TES from a circuit shown in left side.

To read out and amplify up to 2000 detector signals, the detector array uses a time-division multiplexing (TDM) system \[68\] \[69\], which greatly reduces the thermal loading by reducing the numbers of cables between the inside and outside of the cryostat. Each TES signal, which is in the form of \( I_{TES} \) variations, is inductively coupled to a DC superconducting quantum interference device (SQUID) \[70\]. Figure 2.5 shows the conceptual scheme of the TDM system, in which detectors and readout system form an \( n \times m \) matrix. Each unit of the matrix represents a single TES circuit inductively coupled to a SQUID.

SQUIDs belonging to the same column are put in series. SQUIDs belonging to the same row share a common control line. The system is then able to send currents to each row for activation and deactivation, and read signals that are the sums of the signals from each column. Therefore, the system multiplexes the TES signals by rapidly switching on rows of SQUIDs by turns and each time only reading out one
TES signal from each column. With the TDM readout system, the total number of required lines between warm electronics and the detector array is on the scale of $n + m$, while it is $n \times m$ if without.

### 2.2.1 mux-15b

AdvACT uses a new TDM architecture that contains two stages of SQUID arrays. The schematic is shown in Figure 2.6. Each multiplex chip (mux), which is called “mux15b”, contains 11 rows of the first stage series array, which is an array of four dc SQUIDs and is called SQ1. Each SQ1 is inductively coupled to a TES circuit and is part of the detector array operated at bath temperature. Within the mux chip, each SQ1 is also in parallel with a Zappe interferometer that act as a flux-activated switch (FAS). When applying different external magnetic flux induced by different addressing currents, the FAS switches between its superconducting and normal states. Sequentially, the SQ1 bias current will be either passing through the SQ1 or shunt...
by the superconducting FAS, activating or deactivating the corresponding the signal readout of the corresponding row of TESes. Each column of the SQ1 and FAS chain is in parallel with a 1Ω resistor. Since the dynamic resistance of the SQ1 and FAS chain is large compared to 1Ω, the column is close to voltage-biased. After coming out of the SQ1 circuits, the readout signal of each column is the form of the SQ1 bias current variation. It is then further amplified by the second stage SQUIDs series array (SSA), which is contained in SSA PCB boards that are mounted on 1K plate.

Since the SQUID response to magnetic flux ($V - \phi$ curve) has similar shape to a sine curve, feedback coils are respectively added to both of the SQ1 and SSA. By applying external magnetic flux to the SQ1 feedback coils that couple the variation in magnetic flux induced by the varying signal current, the circuit forms a flux-locked loop (FLL) that keeps the SQUID response at the linearity region. The variation of the SA output is then nearly proportional to the variation in the signal current. The signal is eventually readout by the warm multichannel electronics (MCE), which also serves to send control the aforementioned feedback, bias and addressing lines.
2.3 Array Component and Circuitry

2.3.1 Components Overview

Figure 2.7: Components of the AdvACT HF array [11]. The corner-to-corner distance of the copper support ring is around 300 mm. The large ring PCB is mounted to the copper support ring by 4-40 brass screws. Four large and four small wiring chips are glued on top of the ring PCB, with mux and shunt interface chips glued on top of it. The detector wafer stack sits on the copper wafer base, which will be replaced with the feedhorn array. Six flexes electrically connect the detectors to the readout system.

As shown in Figure 2.7, the AdvACT detector array consists of a single detector wafer stack containing up to 2000 Transition-Edge Sensors (TES), six flexible superconducting wiring cables (flex), multiplexing (mux) chips that contain the SQ1s and their coupling FASes, interface chips that provide shunt resistance and Nyquist inductance, wiring chips that assist in circuit routing, wire bonds, feedhorns, printed circuit boards (PCBs), and mechanical supporting parts including the copper support ring.
2.3.2 Detector Wafer Stack

As shown in Figure 2.8, the AdvACT detector wafer stack consists of several layers fabricated on 150mm-diameter silicon wafers stacking and gluing together. It includes one waveguide interface plate (WIP), one detector wafer, one or several backshort (BS) cavity wafer and one backshort cap wafer.

The WIP serves as an interface between the feedhorn and the detector wafer. The WIP mechanically couples to the detector wafer by inserting its ring-shaped features into the silicon cutout of each OMT, guiding the light to the pixel. For the HF and MF arrays, the edges of the WIP also extend out of the detector wafer, onto which one end of each flex is glued. For the LF array, however, there’s no extension part on the WIP, since the LF detector wafer is larger, and the WIP with extension would be larger than the 150 mm silicon wafer. LF flex is mounted on the feedhorn instead.
As aforementioned in this chapter, after switching the materials of the TES and dielectric to AlMn and SiN$_x$ from MoCu and SiO$_2$, the uniformity of the TES parameters across the wafer is greatly improved. In addition, NIST developed tools for 150 mm wafer fabrication. Thus all of the detectors in an array can be fabricated on a single 150 mm-diameter silicon wafer, instead of on multiple 100 mm-diameter tiles as in ACTPol (see below) [72] [73]. This change makes the detectors more densely packed. The hex-shaped detector wafer contains three large rhombuses that are rotational symmetric to each other (shown in Figure 2.9). Each large rhombus is a matrix of $n \times n$ rhombus-shaped pixels, with 2 to 3 pixels in total removed to allow for alignment holes. Within the large rhombus, each pixel has its probes orientation rotated 45 degrees with respect to the adjacent pixels. There are routing traces between the pixels, connecting the TES and the bondpads on the edge of the detector wafer.

On top of the detector wafer, the combination of the BS cavity and the BS cap wafers gives a quarter-wavelength backshort termination to the radiation, which improves the optical efficiency of the bolometer. The features on the BS cavity wafer include a circular opening for each pixel and two semi-ring shaped “moats” filled with CR-110\(^1\), which absorbs leakage light between adjacent pixels. The BS cap has blind

\(^1\)https://www.laird.com
circular cutout features, which together with the BS cavity circular openings form the backshorts. All of the related wafers are fabricated, aligned and eventually glued to become a wafer stack in NIST. Prior to gluing, the WIP, BS cavity and BS cap are also gold-plated to enhance heat sinking.

![Figure 2.10: Photographs of the ACTPol array (top left) and the AdvACT HF array (top right), and conceptual sketches of the the ACTPol array (bottom left) and the AdvACT array (bottom right), illustrating the readout part (blue) and the detector wafer (green). Compared to the ACTPol arrays, the detectors in AdvACT are more densely packed; the readout system can then be unfolded and on the same plane of the detector wafer, making the integration process easier.](image)

Fabricating each detector wafer out of a single 150 mm-diameter silicon wafer makes a profound impact on the physical design of the detector arrays. As shown in Figure 2.10, ACTPol used an assembly of hexagonal and semi-hex shaped detector stacks, which were fabricated out of 100 mm-diameter silicon. The irregular shape of the detector wafer assembly results in empty spaces wasted on the focal plane area. The readout system then has to be installed vertically to save space. For AdvACT,
not only is the detector density higher but also the wafer stack has the regular shape of a hexagon. The use of large wiring chips hosting mux and interface chips made enough space for the readout system to be on the same plane, which greatly simplifies the physical design and packaging process.

2.3.3 Superconducting Flexible Cable

Figure 2.11: (Top left): Photograph of the HF flex that has been installed in the HF array. (Right): Sketch showing the cross section of the flex and the adjacent parts. (Bottom left): Photograph of the LF flex that has extra features used for mounting and aligning.

Each AdvACT array has six flexes (shown in Figure 2.11), serving as electrical bridges between the detector wafer and the readout system. All flexes are fabricated in the PRISM cleanroom at Princeton, and the development and fabrication process is described in \cite{74}. Figure 2.12 shows the conceptual sketch of the cross section of the flex. The flexible part of the flex consists of 40 nm thick Al traces sandwiched by two layers of polyimide. The two ends of the sandwich structure partially land on two 500 µm-thick silicon stiffeners, which provide mechanical support for flex gluing and wirebonding. The Al layer extends out of the polyimide and forms bondpads

\footnote{Small details of the fabrication has changed but Pappas et. al., 2016 \cite{74} provides a good overview.}
on the stiffeners. For the flex of the LF array, additional features for alignment and mounting were added by a laser dicer after fabrication (shown in the bottom left side of Figure 2.11).

![Cross section of the flex](image)

**Figure 2.12:** Sketch of the cross section of the flex.

The Al traces will become superconducting at the bath temperature of the array, minimizing the parasitic resistance introduced into the TES circuits. The four lead testing at Cornell University shows that the flex resistance in the superconducting state is around 0.02 mΩ, which is significantly small compared to the TES operating resistance (2.5 mΩ). The critical current of the Al traces is above 10 mA, which is higher than $I_{\text{TES}}$. 
2.3.4 Readout Components and Array Circuitry

As shown in Figure 2.7 and Figure 2.13, the largest part of the detector array readout system is the large ring-shaped PCB with associated parts sitting on top of it. On the PCB, there are eight silicon wiring chip, four of which are large and trapezoid-shaped and four of which are small and semi-trapezoid-shaped. For the HF array, which has the largest number of pixels among the four arrays, each small or large wiring chip respectively houses two or six columns of mux and interface chips. Each column
contains six of the 11-channel mux chips and three of the 22-channel interface chips, resulting in 66 rows per column. Since the total number of pixels and TESes varies with different arrays, the number of readout components (mainly mux and interface chips) also varies. Table 2.1 shows the number of each component needed for each AdvACT array.

Table 2.1: Numbers of rows, columns, parts and readout lines in the four AdvACT arrays.

<table>
<thead>
<tr>
<th>Array</th>
<th>Pixels</th>
<th>Rows</th>
<th>Col</th>
<th>Mux</th>
<th>Interface</th>
<th>Det Bias</th>
</tr>
</thead>
<tbody>
<tr>
<td>HF</td>
<td>2000</td>
<td>64</td>
<td>32</td>
<td>192</td>
<td>96</td>
<td>24</td>
</tr>
<tr>
<td>MF</td>
<td>1800</td>
<td>55</td>
<td>32</td>
<td>160</td>
<td>80</td>
<td>24</td>
</tr>
<tr>
<td>LF</td>
<td>492</td>
<td>25</td>
<td>16</td>
<td>48</td>
<td>24</td>
<td>16</td>
</tr>
</tbody>
</table>

Figure 2.13 also shows the circuit diagram based on the physical position of every part. The diagram is simplified to have two rows and three columns, with each mux chip containing only one SQ1 channel and each interface chip containing two TES bias channels. The signal comes out of the detector wafer and reaches the interface chip via the metal traces on the flex and wiring chip. The interface provides shunt resistances and the Nyquist inductance that filters out the noise at high frequency. The signal then enters the mux chip and goes through the inductance that is coupled to the SQ1. The mux chip also contains a FAS to activate and deactivate the corresponding SQ1. By putting the mux chips right next to each other and connecting their SQ1 bias and feedback lines with wire bonds, SQ1s belonging to the same column are then in series with each other. On the right part of the diagram, the inductors, coupled to the FASes from the same row, are put in series to each other by the wiring chip traces; therefore the current through “row select” (RS) could control all of the corresponding SQ1s.
The RS, SQ1 bias, SQ1 feedback and TES bias are routed out to the edge of the wiring chip, and through wire bonds enter the PCB.

![Schematic of the ring PCB](image1)

![Photograph of the zoomed view of the PCB](image2)

![Photograph showing the interface PCBs mounted on top of the copper cover](image3)

![Sketch of the cross section of edge area of the detector array](image4)

Figure 2.14: (Top left): Schematic of the ring PCB, showing the PCB wiring that connects SQUIDs belonging to the same row but different wiring chips. (Bottom left): Photograph of the zoomed view of the PCB. (Top right): Photograph showing the interface PCBs mounted on top of the copper cover. (Bottom right): Sketch of the cross section of edge area of the detector array, showing the interface PCBs are connected to the PCB ring through ZIF flexes.

The ring-shaped PCB further assists in constructing the readout structure. As shown in Figure 2.14, the PCB routes the same row of SQ1s on each wiring chip in series. There are zero insertion force (ZIF) connectors soldered to the bottom of the edges of the PCB. The readout lines come out via the ZIF flex to the small interface boards that are on top of the copper cover, and then connect to Series Array (SA) and addressing PCB boards at 1K. Details are described in [12] and [75].
2.3.5 Feedhorn Array

Instead of using corrugated horns [76] as in ACT, AdvACT uses spline-profiled feedhorn arrays [13] to optimize the coupling efficiency and the symmetry of the beam profile. For the HF and the two MF arrays (shown in Figure 2.15), the feedhorn arrays are consists of multiple etched silicon layers with thickness of 0.333 or 0.250 mm, which are fabricated, assembled and electroplated in NIST. The silicon wafers are glued on the sides by Stycast 2850. In addition, choke features are added to the wafer side of the feedhorn to prevent light leakage between horns [77]. The LF feedhorn array [78] (shown in Figure 2.16) is machined out of a Si-Al alloy, called CE7, and then gold plated except for the bottom surface. For the LF array, the choke features are fabricated on a silicon wafer that is integrated to the detector wafer stack.
2.4 Integration

The content of this subsection is mainly based on the HF array assembly process \[11\] since the integration of the MF arrays is the same except for requiring fewer chips (See Table \[2.1\]). There are a few difference between the HF array assembly and the LF assembly, which will be described in separated paragraphs.

The integration methods of the AdvACT detector arrays have been developed to meet several requirements. First, different parts should be assembled with good alignment to avoid electrical shorts due to dense wire bonds. Second, the array needs to be cryogenically robust. This requirement necessitates development of gluing or mounting methods of different parts with thermal contraction differences taken into account. Additionally, the wire bonding parameters are optimized so that the bonds can survive through multiple cryogenic cycles.

\[3\]Much of this section closely follows Li et. al., 2016 \[11\].
As shown in Figure 2.17, there are six stages in the assembly process, which will be described in rich detail below. In the first stage, mux and interface chips are glued to the wiring chip. The second stage is mounting the PCB to the copper support ring and gluing the assembled wiring chips to the PCB. In the third stage, wire bonds connecting the mux and interface chips to the wiring chips are placed. The fourth stage includes mounting the 150mm detector wafer stack to a temporary copper base and aligning to the PCB. Flex cables are also glued with one side to the PCB and the other side to the edge of the detector wafer. The fifth stage is placing wire bonds between the flex and the wiring chip, and between the flex and the detector wafer. The final stage is mounting the feedhorn to the copper support ring and the detector wafer, replacing the copper wafer base used during wirebonding.
2.4.1 Pre-Assembly

![Pre-screening board for the mux chips](image)

Figure 2.18: The pre-screening board for the mux chips. One board contains four rows by eleven columns of mux chips. The mux chips are attached to the board with double sided tape, which remains functional for at least three cooldowns. With the double sided tape, the chips are easy to remove.

Before getting assembled into the arrays, mux and interface chips, as well as the flexes, are pre-screened cryogenically by the collaboration group at Cornell University. The mux chips are integrated into the 4 by 11 pre-screening board (shown in Figure 2.18) at Princeton, while the interface chips and flex cables are simply mounted to a test board with wire bonds placed. There are two purposes of the chip pre-screening. First, it is a sanity check to validate the functionality of the SQUIDs and the continuity of the traces. Second, several parameters are tested, including the shunt resistance of the interface chips, and the $I_{c}^{\text{max}}$ for the mux chips, which is the bias current for the SQ1s that maximize the amplitude of the $V - \phi$ curve. These parameters are needed to do chip mapping for the readout system. Interface chips with similar shunt resistance values will be grouped into the same bias line, while mux chips with similar $I_{c}^{\text{max}}$ values will be grouped into the same column. For the flex, a portion of traces
are randomly selected to test for the critical current. The criterion for the critical current is that it should be above 10 mA.

Apart from screening, several parts including the wiring chips, flex and interface chips are also visually inspected under a high-magnification microscope. It is difficult to screen and validate the continuity of the traces on the wiring chips since at least one end of the traces is at the middle of the chips. We then do visual inspection on all of the critical lines, including the SQUID bias and feedback, and the detector bias lines. For flex and interface chips, the purpose of visual inspection is to check for fabrication defects that cause failure of individual TES channels. Parts with high yields are picked for assembly.

2.4.2 Readout Assembly

![Flowchart showing the procedure of glue the mux and interface chips to the wiring chips.]

The readout assembly starts with mounting the mux and interface chips to the wiring chips. Previously in ACTPol, the mux and interface chips were glued to the PCB boards with rubber cement, which is ideal for gluing relatively large parts with different coefficients of thermal expansion (CTEs). However in AdvACT, the size of
the mux-15b chip is 3 mm by 6.6 mm. Rubber cement is then not strong enough to hold the mux chip while bonding, resulting in poor strength of the wire bonds. Since there’s no CTE difference between the wiring chip, mux and interface chip, the glue is therefore changed to Stycast 1266.

Stycast 1266 is a runny adhesive that is hard to remove, while the bondpads on the wiring chip are only 250 µm away from the mux and interface chips. Hence a tape mask is used to confine Stycast within certain area. Silicon templates are fabricated at Princeton with the same dimensions as the wiring chips and with the same chip outline patterns. As shown in Figure 2.19, the template is first placed on a flat metal base with several silicon alignment parts to secure the template’s position. A piece of low tack tape, which does not leave residue on the surface, is then put on top of the template. The next step is cutting rectangular windows according to the pattern of the template. The edge of the window cutouts is around 50 µm smaller than the edge of the chip. The tape mask is then partially peeled so that the wiring chip can replace the template, retaining the same position and same alignment with the tape mask. After the mask is put back, the Stycast, which is pre-cured at 55°C to increase the viscosity, is deposited into the windows. A die bonder, equipped with a vacuum pickup and a microscope, is used to place the mux and interface chips on the wiring chips. The chip alignment is adjusted according to the bond pads on the wiring chips. The whole assembled wiring chip is left in a dry box overnight to cure the glue.
Figure 2.20: (Left): Drawing of the top view of the detector array, showing the silicon L-brackets with wiring chips sitting next to them. (Top right): Photograph of the weights that press the wiring chip when the glue is curing, showing the surface in touch with the wiring chips. (Bottom right): side view of the weights. One set of weights consist of two parts. One part has alignment features and directly sit on the wiring chip, while the other part can be mated to the handler of the first part and increases the pressure.

During the wiring chip glue curing, the ring-shaped PCB is cleaned and mounted to the copper support ring. After the Stycast is cured, the wiring chips are glued to the PCB using rubber cement. To achieve the alignment, 500 µm-thick silicon L-brackets, which define the outline of the wiring chip position, are first placed on the PCB with 1/16”-diameter alignment pins. Several 200 µm-thick and 1 mm-wide tape strips are then placed on the PCB along the edge of the wiring chip position. Rubber cement is deposited to the space surrounded by the tape strips. Any extra glue is removed by swiping a flat board on top of the tape strips. After removing the strips, the wiring chip is place right next to the L-brackets. Finally, aligning with the same 1/16”-diameter pins, rectangular shaped copper weights are placed on top of the wiring chips for curing. There are cutouts on the bottoms of the weights to avoid directly pressing on the mux and interface chips. The bottom surface of the weight is
make out of Teflon, preventing scratching the traces of the wiring chips. Again, the entire assembly is left overnight for the rubber cement to cure.

To allow tests of the readout alone, and to minimize the handling near the detector wafer, wires bonds between the mux chip, the interface chip, the wiring chip and the PCB are performed before installing the detector wafer stack and flex. This step, together with the detector wire bonding, will be summarized in the Section 2.4.4.

### 2.4.3 Detector and Feedhorn Assembly

![Figure 2.21: Photographs of the top side view (top left), the bottom side view (bottom left), and the zoomed view around the corner of the copper base.](image)

AdvACT uses a copper wafer base to provide physical support for the detector wafer stack before installing the feedhorn. As shown in Figure 2.21, the wafer base has screw hole and alignment pin hole features that roughly align the detector wafer with the wiring chip. On the top surface of the wafer base, there are also vacuum slots,
which once plugged to the vacuum, can hold the wafer stack during bonding. The wafer base also has hole features at the corners, which assist in alignment during the flex gluing.

Figure 2.22: (Left): Photographs of the flex gluing jig. The black jig, made from Delrin, carries the flex by vacuum and transfers the flex between the detector array and the Al jig. The Al jig has a silicon indicator glued on top. (Top right): Zoomed view of the yellow dashed box region on the photo in the left. The indicator was diced so that the dicing markers are $\sim 50 \, \mu m$ from each other. (Bottom right): A flex glued to the detector array. The surround area is protected by the Al cover. Two copper clamps are screwed down to secure the flex stiffeners when the glue is curing.

After the detector wafer is installed on the wafer base that is mounted to the copper ring, six flex parts are will be glued, with one stiffener on the WIP right next to the detector wafer and the other on the PCB right next to the wiring chip\textsuperscript{4}. To achieve the alignment goals, two jigs were developed (shown in Figure 2.22), one of which is made of Delrin and the other is made of aluminum. Both of the jigs have vacuum features. On the Delrin jig, there are two 1.5 mm dowel pins that can couple to the alignment pin hole features on the Al jig and the wafer base. The Al jig also has a scale indicator with 50 $\mu m$ spaced dicing marks.

\textsuperscript{4}The flex assembly for LF is slightly different. See Section 2.4.4
The flex-gluing process starts with placing the flex right beside the indicator on the Al jig. The Delrin jig then lands on top of the flex with dowel pins coupled to the holes on the Al jig. After turning on the Delrin jig’s vacuum and turning off the Al jig’s vacuum, the Delrin jig is able to carry the flex and transfer it onto the extension area of the WIP. The side facing the detector wafer of the Delrin jig has two slots cut out, which reveal a small part of the flex bondpad area. With a microscope, the misalignment between the bondpads of the flex and the detector wafer can be estimated. The Delrin jig then transfers the flex back to the Al jig. With the Al jig vacuum partially turned down, the flex can be moved along the indicator by the estimated misalignment distance according to the dicing marks on the indicator. The flex is then transferred back to the WIP to check the alignment for a second time. This step is repeated several times until the misalignment is less than half of the bondpad. The next step is depositing rubber cement on the WIP using an electric glue dispenser and placing the flex stiffener on the glue using the Delrin jig. To achieve gluing uniformity across the stiffener and secure the stiffeners during the glue curing time, a copper clamp is placed on top of the stiffener and secured to the wafer base by two screws.

The distance between the edge of the detector wafer and the edge of the wiring chip is designed to be 0.23 mm shorter than the length of the flex to compensate the thermal expansion difference between the copper ring and the silicon wafers. Hence the flex needs to be glued with its polyamide part curved to fit into the space. To tackle this challenge, two cylindrical spacers are then placed underneath the flex polyamide. The size of the spacers is tuned so that by simply pressing down on the wiring chip side of the flex, its stiffener will land right beside edge of the wiring chip. The glue deposition at this side is similar and another copper clamp is placed on top of the stiffener and fixed down by screws. The two clamps are later removed in 12 hours when the glue is cured. The final wire bonding of the flex to the wiring
chips and to the detector wafer is performed after the flex glue is cured. This step is described in Section 2.4.4.

Figure 2.23: (Left): Photograph of the HF array with spokes mounted at the inner-corner areas of the copper ring. (Right): The brass L-brackets are installed to the feedhorn arrays using screws and nuts. Teflon washers are added to protect the silicon feedhorn wafer from being scratched by the metal screws and nuts. (Middle): Photograph of the Delrin base, which assists the alignment of the L-brackets on the feedhorn before we start mounting it to the detector array. There are cutouts on the bottom of the Delrin base, through which we can tighten the screws on the feedhorn and fix the orientations of the L-brackets.

So far, the detector wafer has been sitting on the wafer base, which needs to be replaced by the feedhorn array. Since there is a CTE difference between the silicon feedhorn array and the copper support ring, the feedhorn array can’t be directly mounted to the copper ring as the wafer base is. Instead, it is installed through BeCu L-brackets and OFHC support “spokes,” as shown in Figure 2.23. The softness of the BeCu L-brackets allows the feedhorn and the copper ring to contract differently without breaking any parts, while the spokes serve as a mechanical interface between the L-brackets and the copper ring.

The process starts with aligning the orientation of the L-brackets on the feedhorn according to the spokes, which are aligned to the copper ring using 1/16”-diameter dowel pins. It is difficult to adjust the L-brackets on the real array, hence a Delrin base was developed, which has the same cutout and spoke mating features as the copper ring. The L-brackets on the feedhorn can then be aligned to the spokes that are installed on the Delrin base.
After finishing the L-bracket alignment, the detector array is mounted on an Al ring that is supported by three vertical rods made of stainless steel (Figure 2.24). Above the detector array, the vacuum chunk is slowly lowered until it touches the top surface of the detector wafer stack (the backshort cap). The vacuum is then turned on and the wafer is hold by the vacuum chunk and no longer supported by the wafer base, which is subsequently dismounted. Below the array, the feedhorn with the aligned L-brackets is slowly raised by a jack. The position of the feedhorn is also adjusted by an XY stage. During the feedhorn lifting, two camera focus on two areas at the corners of the detector wafer to assist in the alignment between the feedhorn and the detector wafer stack. After the feedhorn is lifted up, the spokes and their mounting holes are visible above the plane of the PCB. The L-brackets are secured to the spokes and to the copper support ring by screws. The vacuum is turned off, releasing the detector wafer stack to the feedhorn array. Two alignment pins are inserted into both of them and the assembly is complete.
Figure 2.25: (Left): Photograph of two flexes mounted to the LF array. Each flex is held by two copper bars that are \(\sim\)1 mm thick. (Top right): Sketch showing the cross section of the flex region. For each flex, one stiffener is mounted on the PCB right next to the wiring chip, while the other side is mounted on the feedhorn array, with two of 0.5 mm thick silicon bars in between. (Bottom right): Photograph showing that the LF feedhorn is mounted to the copper ring via L-brackets on the side.

For the LF array, prior to mounting the flexes, the feedhorn array, carrying the detector wafer stack, is installed to the copper wring through the brass L-brackets on the side (shown in Figure 2.25). Since the WIP doesn’t extend out of the detector wafer, the flexes are mounted to the feedhorn with two silicon bars in between so that the height difference between the detector wafer and the flex is only \(\sim\)0.5 mm (shown in Figure 2.25). Each flex is held down by two copper clamps with low-profile brass screws. Wire bonds of the flex to the wiring chips and to the detector wafer are then placed and the LF array assembly is complete.
2.4.4 Wire Bonding

Figure 2.26: (Top left): Photograph of the HF array installed on the bonding jig. (Top right): The bonding jig set contains a cap, a cylindrical body and a base block. (Bottom): Photograph of the short checker.

This subsection describes the wire bonding of the detector assembly, which proceeds in two steps as described earlier. The readout is wire bonded after the completion of the gluing of the wiring chips, while the flex wire bonding complete the attachment of the detectors to the readout.

The AdvACT array needs up to 23,000 wire bonds to complete the array circuit. The wire bonds include the ones between detector wafer and flex, between flex and wiring chips, between wiring chips and interface chips, between wiring chips and mux chips, between interface chips and mux chips, and finally between wiring chips and
the PCB. All of the wire bonds are completed using an automatic wedge bonder. During the wire bonding process, the bonder’s motorized table brings the detector array to the position of the bonder head, which then places Al wire bonds. As shown in Figure 2.26, a group of bonding jigs has been developed to ensure that the mechanical connection between the detector array and the bonder stage is stable. The jigs connect to the detector array through the wafer base and allow the array to rotate in increments of 30 degrees. It is found that the uniformity and stability of the bonder’s performance is optimized when bonding in a certain orientation. The array is then rotated to the optimized angle as each edge is bonded. Figure 2.27 shows two zoomed-in pictures of the wire bonds.

![Figure 2.27: Zoomed view of wire bonds on the flex and on the wiring chips.](image)

A special-purpose short checker (designed by Norm Jarosik and shown in Figure 2.26) is also connected to the detector array to prevent shorts between different detector bias lines, which can prevent usage of ~100 detectors for one unmitigated short. The shorts are usually originate from shorts between flex traces or between traces in the detector wafer. The short checker contains an automated digital circuit that measures resistances between a set of 8 pins. Once plugged into a laptop, it gives an $8 \times 8$ matrix of resistances and updates every few seconds. If a short happens, the automatic wire bonding process is halted and the last placed bonds will be pulled off one by one until the short disappears.
2.4.5 Other Mechanical Assembly

There are other mechanical parts and circuit boards to further close the detector array and connect the array to the readout devices at higher temperature stages. After the feedhorn is mounted and the final continuity check is finished, a gold-plated OFHC “snake-tongue ring” is mounted on the six spokes (Figure 2.28). On the bottom surface of the snake-tongue ring there are BeCu tripods that provide mechanical support and heat sinking for the detector wafer stack. A gold-plated OFHC cover is then mounted on top of the copper ring and closes the detector array. During the mounting, two long 4-40 screws are used to guide the position of the cover and prevent the cover from crashing onto delicate parts of the detector array. A filter cell, carrying a metal mesh optical filter that filters out the radiation at higher frequency, is installed to the copper ring from below. As mentioned in Section 2.3.4 there are
small “interface PCBs” on the cover, serving as interface between the detector array and the boards at 1K stage that contains SSAs (mentioned in Section 2.2). ZIF flexes and NbTi cables are installed from the ring PCB to the interface PCB and from the interface PCB to the SA boards to complete the connection. The integration of the detector array is finished.

2.4.6 Rework

Although all of the mux and interface chips have been pre-screened, mishandlings during the integration process can damage the electrical components fabricated on the chips. As mentioned previously, we test the readout system of the detector arrays alone before integrating the detector wafer stack into the array, to detect failure channels that impact an entire row or column of SQUIDs. After warming up, the identified malfunctional chips will be replaced by stacking new ones on top.
Chapter 3

Validation and Characterization of Detector Arrays

Before entering the stage of CMB observation, each detector array undergoes a series of tests both in the lab and on the telescope. One major purpose of the tests is to validate the detector array’s functionality, which includes fitting for detector parameters and comparing them with the expected values, and checking the yields to determine whether to do integration rework. Another purpose of the tests is to characterize the SQUIDs and the detector bias level so that they are operated at the optimal conditions. Last but not least, several calibration measurements before and during the CMB observation are conducted to convert the detector data, which is in the form of TES currents, to CMB temperature.

The content and data presented in this chapter are primarily based on the tests undertaken on the AdvACT LF array.
3.1 Lab tests and parameters

Figure 3.1: (Top): Flowchart of the integration process, which is categorized into the readout (green), detector (yellow) and mechanical assembly (red). (Bottom): Flowchart of the three types of assemblies and validation tests.

The lab tests includes the SQUID tuning and the detector characterization, both of which are conducted when the detector array is cooled down to the bath temperature of $\sim 100$ mK. As shown in Figure 3.1, the first SQUID tuning takes place after the readout assembly. It not only validates the SQUID functionality but also optimizes input parameters such as the SA and SQ1 bias voltages. The detector characterization takes place after the detector assembly. It can be further separated into dark testing, during which a cover protects light so that intrinsic detector parameters can be measured, and the optical tests, in which the detectors measure radiation emitted by a near-blackbody source installed at the 4K stage of the cryostat. Both the SQUID tuning and the detector characterization may need to be repeated depending on whether integration rework needs to be done.
3.1.1 SQUID Tuning

The SQUID tuning process is described in [80] in detail. As shown in Figure 3.2, the process begins with taking second-stage series array (SSA) SQUID response curves by ramping the SA feedback current and measuring the SA output voltage $V_{SA}$. The measurement repeats multiple times, each with different values for the SSA bias. The optimized SSA bias, with which the peak-to-peak SSA response reaches the maximum, is found and saved in the configuration file. Using the SQUID curve with the optimized SSA bias, the system will choose an operating point, where the SA SQUIDs have linear response to the magnetic flux change. During FLL operation, the SA output voltage will be held at this point by constantly adjusting the SQ1 feedback current. The next step (labeled “RS servo” in Figure 3.2) is to get the SA...
feedback, which is required to keep the aforementioned $V_{SA}$ fixed while ramping FAS currents. On the SQUID response curve, the FAS current corresponding to the first maximum (minimum) SA feedback is chosen as the addressing current to activate (deactivate) the corresponding row of SQ1s. The third step (labeled “SQ1 servo sa” in Figure 3.2) is to ramp the SQ1 feedback and get the corresponding SA feedback for FLL, with individual channel’s FAS turned on. Similarly to step “SSA,” multiples curves are generated with different SQ1 bias values to find the optimized SQ1 bias. In addition, the SA feedback is also determined for running the SQUIDs at the SA operating point. The last step is ramping the SQ1 feedback (labeled “SQ1 ramp” in Figure 3.2) or the detector bias lines (labeled “SQ1 ramptes” in Figure 3.2) to get the open loop response from SA out. The system gain is obtained from the slope of the SA output at the operating point.
Figure 3.3: Examples of SQUID curves with red boxes labeling abnormal channel. (Left): Lack of SQUID response at row 2 column 3. If other channels from column 3 behaves similarly, it implies that the wire bonds connecting the bias or feedback lines of column 3 is broken. (Middle): Abnormal period of SQ1 rampes implying unexpected heat driving the corresponding detectors normal. (Right): SQ1 ramp plot when the addressing currents are off. The channel showing SQUID response is the persistence channel that can not be turned off.

Apart from optimizing the parameters for operating the readout system, the SQUID tuning also assists in detecting detector and SQUID failures. Figure 3.3 shows several examples of abnormal SQUID response curves. The most common type of abnormal SQUIDs curves are those showing pure noise, which is usually related to broken wire bonds or damage from fabrication and handling. If rework is possible, the array will be visually inspected and probed to find the specific malfunctioning channel. Rework involves repairing the wire bonds, or placing wire bonds to short out the bad SQUIDs and recover the rest. Another type is the SQ1 rampes curve with an atypical period, which is because the corresponding TES is normal due to unexpected thermal loading or bath temperature perturbations. No rework needs
to be done on the array in this case. In addition, an extra SQUID test is conducted to take SQ1 ramp curves with all FASes turned off. The purpose is to check “persistence,” which happens if a FAS is always in its normal state due to magnetic flux trapped in it during the process of cooling down. A column will then have a specific channel that cannot be turned off during multiplexing. If the persistence happens, some channels remain operational with the and having usual SQUID response instead of only noise. The array will be warmed up to 10 K and recooled down to expel the trapped flux. Still, there are rare situations that the persistence issue is unfixable. As mentioned in Section 2.4.6, in this case we replace the problematic mux chip by stacking another one on top and complete the related wire bonds.

3.1.2 Detector Dark Tests

Figure 3.4: (Top left): Photograph showing the bottom surface of the optical cell, covered with five to six layers of superinsulation films to reflect emission from other parts in the cryostat. (Right): Plot of the bath temperature log, with vertical lines illustrating the time stamps of IV measurements at each temperature stage. (Bottom left): Photograph showing the original setup when the dark test and optical test are conducted simultaneously with a silicon mask covering two thirds of the pixels.
The detector tests are described in detail in [81], [82], [83] and [15]. The setup of the dark parameter test is to either cover the bottom of the feedhorn with a copper-plated silicon wafer or to close the filter cell with a metal plate (Figure 3.4).

As mentioned in Section 2.1.1, the electro-thermal model of the TES follows the equation:

\[ P_{bias} + P_{\gamma} = \kappa T^n - \kappa T_{bath}^n. \]  

(3.1.1)

In the dark test, the detectors are isolated from radiation so \( P_{\gamma} \) is approximately zero. The saturation power \( P_{sat} \) is defined as the power that drives the detectors to 90% of the normal resistance \( R_n \). Equation 3.1.3 becomes:

\[ P_{sat} = \kappa T_c^n - \kappa T_{bath}^n, \]  

(3.1.2)

with thermal conductance \( G \equiv \frac{dP_{sat}(T)}{dT} \big|_{T_c} = n\kappa T_c^{n-1} \). By taking measurements of \( P_{sat} \) at different \( T_{bath} \), dark parameters such as \( T_c, G, n, \kappa \) can be fit (Figure 3.5).

![Figure 3.5: Example plots of \( P_{sat} \) versus \( T_{bath} \) for a 39 GHz (left) and a 27 GHz (right) detector with the fitted model.](image)

During each data acquisition, the TES is first driven to the normal state. The bias voltage then ramps down to measure the IV curve of the TES with about 2000
data points. The bath temperature ramps up with increments of 5 or 10 mK from 70 mK to 180 mK and the IV curves are taken at each $T_{bath}$. For the HF and MF arrays, the targeted $P_{sat}$ for the array’s two frequencies are on the same scale, so all of the detector’s IV curves are taken simultaneously. For the LF array, the targeted saturation powers for 27 GHz and 39 GHz detectors are significantly different because the atmospheric emission at 27 GHz is much smaller than that at 39 GHz. IV measurement of the 27 GHz detectors and the 39 GHz are then taken separately to avoid dissipating too much heat and causing significant $T_{bath}$ variation. Figure 3.6 shows the result of the saturation power measurements of the LF detectors.
Figure 3.6: Histogram plots of the dark parameters $P_{sat}$, $G$ and $T_c$ for the LF detectors. The light red color is used for the 27 GHz TESes while the darker blue is for the 39 GHz ones. As described in the text these values are the result of fitting IV curves at different bath temperatures to three parameters: $T_c$, $n$ and $\kappa$.

During the fitting process, the $P_{sat}$ uncertainties are assumed to be equal at each $T_{bath}$. The critical temperature $T_c$ is tuned by the Mn doping percentage in the AlMn target used during fabrication, so it is relatively independent of the other dark parameters. The parameters $n$ and $\kappa$ is both related to the TES legs’ geometries; hence there are strong degeneracy between them. One way to demonstrate this degeneracy is to first do a 3-parameter fitting and then fix $n$ to the median of $n$ values to do a 2-parameter fitting. When comparing the results from the two methods, there’s less
than 3% deviation in the $T_c$ and the calculated $G$ values. The distributions of the result from 3-parameter fitting is shown in Figure 3.6. Figure 3.7 reports the dark parameters in array plots, which shows pixel parameters with their physical locations on the detector wafer.

Figure 3.7: Array plots of the dark parameters $P_{sat}$, $G$ and $T_c$ for the LF detectors. The orientation of the array plots is looking through the detectors at the sky.
Apart from the bath temperature ramp tests, we also conduct bias step tests to measure each detector’s time constant, which represents the time delay of the detector response to a signal, by sending step functions to the bias lines. The time constant $\tau$ is fitted by:

$$I_{TES}(t) = I_{TES}(t_0) + \Delta I e^{-\frac{t}{\tau}},$$

(3.1.3)

where $\Delta I$ is the amplitude of the step, and $I_{TES}(t_0)$ is the mean of the TES currents at $0.5R_n$ for all detectors on the specific bias line. The time constant is usually expressed in frequency $f_{3dB}$:

$$f_{3dB} = \frac{1}{2\pi \tau},$$

(3.1.4)

representing the point, at which the detector response is half of what it is at $t \sim 0$. The time constant depends on the loading of the detector and increases if there is extra optical power or heat. The measurement is crucial since the time constant degrades the observation data by applying a low pass filter. For the LF array, we took bias step measurements in the lab at $T_{bath} = 130\, \text{mK}$, which simulated the thermal and optical environment in the telescope with 100 mK bath temperature and 1 mm PWV (shown in Figure 3.8).
Figure 3.8: Histogram of $f_{3dB}$ for the 27 GHz and 39 GHz detectors at 130 mK bath temperature, chosen to simulate conditions in the field with 1 mm PWV and $T_{bath} = 100$ mK.
3.1.3 Detector Optical Tests

Figure 3.9: (Top left): Photograph of the original setup of the optical test with a mask covering 2/3 of detectors. (Bottom left): Photograph of the final setup with a reflective frame covering the edge area of the feedhorn. (Top right) Photograph of the cold load, showing the coated pyramids. (Bottom right) Photograph of the bottom of the cold load, showing eight heaters in series with each other and four Teflon legs. The cold load is heat sunk to the 4K stage through the stainless steel wires on the legs.

As shown in Figure 3.9, the setup of the detector optical test is to use a gold-plated silicon wafer to cover two thirds of the feedhorns. For the LF detector array, a second test was done to cover the blank edge area using a gold-plated silicon wafer frame. The exposed detectors are illuminated by a cold load mounted at the 4K stage. The cold load is made of a 15 cm by 15 cm array of Al pyramids with a coating layer of Eccosorb-110, and can be treated as a blackbody radiation source. Between the cold load and the feedhorn, there are three metal-mash low pass filters with cutoff frequencies at 5.58 cm\(^{-1}\), 8.5 cm\(^{-1}\) and 12 cm\(^{-1}\), respectively heat sunk to the 100 mK, 1 K and 4 K stages.
The primary detector optical test is to measure the detectors’ optical efficiency. The process largely follows [82] and [15]. For the cold load at temperature $T_{CL}$, the radiation power $P_{\gamma}^{CL}$ that reaches a detector can be estimated by:

$$P_{\gamma}^{CL} = \frac{1}{2} \int \varepsilon(\nu)f(\nu)A_e(\nu)P(\theta, \varphi, \nu)B(\nu, T_{CL})d\Omega d\nu,$$

(3.1.5)

where the integration is taken over the radiation frequency $\nu$ and the solid angle $\Omega$. The $\varepsilon(\nu)$ in Equation 3.1.5 is the emissivity of the cold load, and $f(\nu)$ is the product of the three filters’ transmissions and the detector’s bandpass. The $A_e(\nu)$ is the feedhorn’s effective area, which is equal to the ratio of $\lambda^2 = c^2/\nu^2$ over the feedhorn beam’s total solid angle $\Omega_{tot}$. $P(\theta, \varphi, \nu)$ is the normalized feedhorn beam pattern simulated by the collaboration group at the University of Michigan. $B(\nu, T_{CL})$ is Planck’s law for blackbody emission. There is a fraction $\frac{1}{2}$ in front of the integral because the radiation power splits into two linear polarizations when coupling to the OMT. The beam and solid angle related part in Equation 3.1.5 is:

$$\frac{\int_{CL} P(\theta, \varphi, \nu) d\Omega}{\Omega_{tot}} = \frac{\int_{CL} P(\theta, \varphi, \nu) d\Omega}{\int_{all} P(\theta, \varphi, \nu) d\Omega} = \frac{\Omega_{CL}}{\Omega_{tot}}.$$

(3.1.6)

The numerator can be estimated by dividing the cold load into a 1000 by 1000 grid elements and taking the sum of the product of the beam pattern and the solid angle extended by each small element piece of the cold load. $\Omega_{CL}$ is then calculated by:

$$\Omega_{CL} = \sum_{n=1}^{1000^2} \frac{a^2 \cos\theta}{(x-x_d)^2 + (y-y_d)^2 + d^2} P(\theta, \varphi, \nu),$$

(3.1.7)

where $a$ is the side length of each small unit of the cold load, $(x, y)$ and $(x_d, y_d)$ are the positions of the cold load unit and the specific detector, and $d$ is the distance between...
the cold load and the bottom surface of the feedhorn. Equation 3.1.5 becomes:

\[ P_{\gamma}^{CL} = \frac{1}{2} \int \varepsilon(\nu) f(\nu) \frac{c^2 \Omega_{CL}}{\nu^2} \Omega_{tot} B(\nu, T_{CL}) d\nu, \]  

(3.1.8)

and the integration can be easily estimated over the frequency band of the detector.

The actual optical power absorbed by the detector \( P_{\gamma}^{abs} \) can be estimated by measuring the detector’s saturation power since:

\[ P_{sat} + P_{\gamma}^{abs} = \kappa(T^n_c - T^n_{bath}) \]  

(3.1.9)

During the measurement, the bath temperature of the detector array is fixed (at 100 mK for the HF and MF array, and at 75 mK for the LF array) by a thermometer mounted at the snake tongue ring and a heater mounted at the mixing chamber stage of the cryostat, through a PID regulator. The cold load temperature ramps from around 10 K to 20 K with increments of 0.5 K. At each \( T_{CL} \), \( P_{sat} \) is measured for each detector. Ideally, as the cold load temperature ramps up, the decrease in \( P_{sat} \) will be the increase in \( P_{\gamma}^{abs} \).
Figure 3.10: Concept plot illustrating the method of calibrating the optical power measurements for the HF and MF detectors. Each exposed detector (yellow) has corresponding dark detectors located within the annulus (red dashed lines).

However, there might still be local $T_{bath}$ variation when the cold load heats up. For the HF and MF arrays, two thirds of the pixels are covered by a silicon mask and these detectors can be considered as “dark detectors,” while the detectors exposed to the cold load are “optical detectors.” Assuming radial symmetry of the detector array, the optical power for each optical detector is:

\[ P_{\gamma}^{abs} = \langle P_{sat}^{dark} \rangle - P_{opt}^{sat}, \]

(3.1.10)

where $\langle P_{sat}^{dark} \rangle$ is the average $P_{sat}$ of all dark detectors located within an annulus, which is concentric with the detector wafer with two radii equal to $\pm 1$ cm plus the distance between the optical pixel and the origin of the detector wafer (Figure 3.10). The optical efficiency $\eta$ can then be fit from:

\[ P_{\gamma}^{abs} = \eta P_{\gamma}^{CL} + P_{DC}, \]

(3.1.11)
where $P_{DC}$ is a constant offset, resulting from $P_{\text{sat}}$ variation between individual detectors in the dark and thermal radiation power from other parts in the cryostat.

This method doesn’t work for the LF array, the feedhorn of which is machined out of an entire piece of Al-Si alloy called CE7, instead of comprising a silicon wafer stack as in the HF and MF arrays. Little is know about CE7’s properties, such as the thermal conductance and heat capacity at 100 mK, and the absorption spectrum below the cutoff frequency of the three low pass filters. The same optical test, i.e. covering 2/3 of pixels and subtracting each optical detector’s $P_{\text{sat}}$ from the average $P_{\text{sat}}$ of the covered detectors, was conducted on the LF array. The calculated $P_{\gamma}^{\text{abs}}$ is significantly larger than $P_{\gamma}^{CL}$ and the resulting optical efficiency is thus around 2 to 3 (where 1 indicates 100% efficiency), which indicates that the LF array experiences larger systematics from $T_{\text{bath}}$ variation than the HF and MF arrays. In addition, when comparing the $P_{\text{sat}}$ from this optical test and from the dark test at the same $T_{\text{bath}}$ (Figure 3.11), detectors located near areas that were not fully covered by the silicon
mask have higher $P_{sat}$ differences. This implies that the bottom surface (sky-facing) of the feedhorn, which is not gold-plated as are other surfaces or as are the bottom surfaces of the HF and MF feedhorns, absorbs significantly large radiation power from the coldload. The uncovered area includes corners and the two edges on the exposed rhombus; therefore the assumption of radial symmetry no longer holds.

To recover the radial symmetry, a second optical test was performed in which all pixels on the LF array were exposed to the cold load, and the gold-plated silicon mask was replaced with a gold-plated silicon frame covering the edge area to reduce heat absorption. For the LF array, each pixel has two extra TESes that do not couple to the OMT and these TESes are treated as the “dark detectors”. For each frequency, non-OMT detectors on only two rhombuses are read out. The calibrated bath temperature for an optical detector is:

$$T_{bath} = < T_{bath}^\text{dark} >= < \left( T_c^\text{dark} \right)^{n^\text{dark}} \frac{P_{sat}^\text{dark}}{\kappa^\text{dark}} \right>^{\frac{1}{n^\text{dark}}}, \quad (3.1.12)$$

where $< T_{bath}^\text{dark} >=$ is the average of the calculated bath temperature of all of the “dark detectors” (non-OMT detectors) that are located within the annulus, and $T_c^\text{dark}$, $n^\text{dark}$ and $\kappa^\text{dark}$ are the dark parameters of the dark detectors from Section 3.1.2. For each non-OMT detector, the bath temperature is estimated from its $P_{sat}$ using the inverse function of Equation 3.1.12. The calibrated saturation power of the optical detector without optical loading is:

$$P_{cali}^{\text{sat}} = \kappa^{\text{opt}} \left[ (T_c^{\text{opt}})^{n^{\text{opt}}} - < T_{bath}^\text{dark} >^{n^{\text{opt}}} \right], \quad (3.1.13)$$

where $n^{\text{opt}}$, $\kappa^{\text{opt}}$ and $T_c^{\text{opt}}$ are the dark parameters of the optical detector. The optical power absorbed is:

$$P_{\gamma} = < P_{cali}^{\text{sat}} > - P_{\text{sat}}^{\text{opt}}, \quad (3.1.14)$$
With the methods of calculating $P_{\text{abs}}$ (Equation 3.1.14), the optical efficiency $\eta$ for the LF detectors can finally be estimated as the slope from fitting a linear model between $P_{\gamma}^{\text{abs}}$ and $P_{\gamma}^{\text{CL}}$ at different cold load temperatures.

Figure 3.12: (Top left): Plot of the cold load temperature log. (Bottom left) Plot of the bath temperature log. Vertical lines indicate when the cold load data were acquired. The associated numbers on the plot are the cold load temperatures. (Top right): Histogram of the optical efficiency of the 39 GHz detectors. (Bottom right): An example plot of $P_{\gamma}$ versus simulated $P_{\gamma}^{\text{CL}}$.

Figure 3.12 shows the distribution of the 39 GHz optical efficiency (estimated using Equation 3.1.14) of the LF array and one example of the fitting. Although values of $\eta > 1$ are estimated for some detectors, the median of $\eta$ is still lower than 1. By contrast, for the 27 GHz detectors, the median of $\eta$ is larger than 1 and the majority of the data do not fit well to linear function of $P_{\gamma}^{\text{abs}}$ versus $P_{\gamma}^{\text{CL}}$. There are several possible reasons that could explain the unreasonably large 27 GHz $\eta$, and the overall non-uniformity for detectors at both frequencies. First, though the edge area of the feedhorn bottom surface is covered, there is still area between the horns.
that can absorb heat from the cold load. Second, the LF array has fewer pixels, and consequently only a few dark detectors are found in the annulus to calculate \( < T_{\text{dark}}^{\text{bath}} > \), thus its precision is low due to insufficient sample size. In addition, the heat sinking of the detector wafer stack might be uneven, which undermines the assumption of radial symmetry.

### 3.2 \textit{in-situ} performance

The LF array was deployed in February 2020 and the cryostat has been cold since March 2020. Because of a problem with the metal mesh filter stack, the mixing chamber of the cryostat only cools down to 120 mK, instead of 100 mK. During observations, there was also a column of SQUIDs and detectors dissipating extra heat, which was not noted in the lab tests. We successfully identified the column and turned it off. However, due to the global outbreak of Covid-19, which delayed the shipping of replacement parts for telescope maintenance, we lost the telescope for around two months. For the same reason, we had limited manpower to complete the cold continuity measurement, which gives the cable resistance for precisely calculating the TES currents. Up to now, we have been using the old resistance values from the MF2 array. This thesis therefore only gives partial results for the site performance of the LF array.

The site characterization includes two major parts. The first part is characterizing and validating the SQUIDs and detectors so that they are operated at optimal configurations, which is similar to Section 3.1. The second part is the detector calibration, which converts the raw data in DAC units to differential CMB temperature units.
3.2.1 Calibration

The raw data from a the detector is the TES current value in DAC units. It is first converted to an estimate of the optical power using the bias step measurement, in which step functions are sent to the detector bias lines and the detector response to a small change in bias current is measured (the same data are used to estimate the time constants as described in Section 3.1.2. Details are described in [84] and [72]. The measurement gives a responsivity $S$ for each TES:

$$S = \frac{\delta P}{\delta I},$$  \hspace{1cm} (3.2.1)

which converts the measurement of the change in TES current to the change in power.

The second step of the conversion is from the power to the CMB temperature. We conduct planet observations on Uranus, Saturn or Jupiter, whose brightness temperature spectra are well known. The solid angles of these planets are considerably smaller than the detector beam; they can then be considered as point sources and the antenna temperature $T_a$ is diluted by the detector beam:

$$T_a = \frac{\Omega_p}{\Omega_b} T_p,$$  \hspace{1cm} (3.2.2)

where $\Omega_p$ and $\Omega_b$ are the solid angles of the planet and detector beam, and $T_p$ is the brightness temperature of the planet. Under the assumption that the spectrum of the brightness temperature is relatively flat around the detector bandwidth $\Delta \nu$, the optical power measured by the detector becomes:

$$P_{\gamma} \sim \eta_T \Delta \nu k_B \frac{\Omega_p}{\Omega_b} T_p,$$  \hspace{1cm} (3.2.3)

where $k_B$ is the Boltzmann constant, and $\eta_T$ is the end-to-end efficiency of the telescope, which is the product of the detector optical efficiency, the detector’s coupling
to the Lyot stop, and the efficiencies of other optics in the telescope. During CMB observations, since the CMB fills the beam, the measured optical power is:

\[ P_\gamma \sim \eta T_\Delta \nu k_b T_{CMB}. \]  

(3.2.4)

The conversion factor can then be calculated from the planet observation by:

\[ \frac{\delta T_{CMB}}{\delta P} = \frac{\Omega_p T_p}{\Omega_b P_\gamma}. \]  

(3.2.5)

Figure 3.13: (Top): Maps of Jupiter respectively by the 27 GHz and 39 GHz detectors. The sizes of the maps are 15 arcmin by 15 arcmin. (Bottom): Preliminary beam profiles are made before fitting the beam solid angle using moby2. The purpose of this step is to provide the range of \( \theta \) to moby2 for fitting the wing part. For the 27 GHz detectors, it is from 15 to 17 arcmin, while for the 39 GHz detectors, the range is 8 to 10 arcmin.
Here we describe preliminary planet measurements for the LF array, mainly following the methods in [85]. Since Saturn and Uranus are relatively dim in the LF bands, Jupiter is used for the LF array calibration. The peak amplitude from Jupiter is measured for each detector during the planet scan. The peaks are then converted to power as mentioned previously to get $P_\gamma$. The maps of Jupiter are generated using the code moby2. The code is also used for fitting the map to compute the beam solid angle of the detector, which contains the Gaussian main lobe and “wing” part that decays as $1/\theta^3$. Figure 3.13 shows the maps of Jupiter by the 27 GHz and 39 GHz detectors. The brightness temperature of Jupiter is taken from the WMAP measurement [14]. Since the distance to Jupiter changes substantially, we scale the solid angle of Jupiter to $2.22 \times 10^{-8}$ sr with $D_J = 5.5$ AU [86]. Since the brightness temperature of Jupiter varies over the LF bandwidth, Equation 3.2.6 becomes:

$$P_\gamma \sim \eta T \frac{\Omega_p}{\Omega_b} \sum T_p k_B \times 1 \text{ GHz},$$

(3.2.6)

where $T_p$ values are interpolated results at frequencies from 24.5 to 29.5 GHz with 1 GHz step for the 27 GHz detectors; for the 39 GHz detectors, the frequency steps are from 30.5 to 47.5 GHz (shown in the left side of Figure 3.14). The right side of Figure 3.14 shows the result of $\eta T$ for the LF detectors. Since $\eta T$ include the coupling between the feedhorn beam and the Lyot stop, it tends to decrease and scatter more as the detector is further from the array center.
As mentioned previously, the estimated $\eta_T$ values can then be used to calculate the conversion factor (Equation 3.2.5). We use this conversion factor to estimate the detector noise equivalent temperature (NET), and eventually the array sensitivity. We first choose a TOD file taken when the PWV is close to 1 mm. For each detector, the raw TOD data in DAC units is converted to power using the responsivity from the bias step measurement. The noise equivalent power (NEP), i.e. the white noise level for each detector, is estimated as the mean value of the power spectrum centered at 20 Hz with a bandwidth of 4 Hz (for details about the noise spectrum of AdvACT TES bolometers, see [87]). We then calculate NET for each detector as:

$$NET = NEP \frac{\delta T_{CMB}}{\delta P}. \quad (3.2.7)$$

Figure 3.15 shows the histograms of NEP and NET for the LF detectors. The array sensitivity $NET_{array}$ at each frequency band is estimated as:

$$NET_{array}^2 = \frac{1}{\sum_i NET_i^2}, \quad (3.2.8)$$
where $NET_i$ is NET of each detector belonging to the frequency band. The final result values of $NET_{array}$ are $51 \mu K/\sqrt{\text{Hz}}$ at 39 GHz and $102 \mu K/\sqrt{\text{Hz}}$ at 27 GHz.

![Figure 3.15: (Left): NEP histogram of the LF detectors when the PWV is close to 1 mm. The dashed black lines illustrate the median values of NEP results, which are $6.21 \pm 1.58 \times 10^{-18} \text{W}/\sqrt{\text{Hz}}$ at 27 GHz and $1.67 \pm 0.49 \times 10^{-17} \text{W}/\sqrt{\text{Hz}}$ at 39 GHz. (Right): Histogram of the converted NET values using the conversion factors.](image)

Apart from the planet calibration, we also present the time constant of the LF detectors by the bias step measurement. There are two types of time constants, one from the bias step measurement as mentioned previously, and the other from measuring the detector response time when scanning a planet. The bias step time constant, which is taken every hour, will be calibrated by the planet time constant, which happens every other day. The calibrated time constants will be eventually used for correcting the filter effect on the corresponding TOD spectra. Here we only present the preliminary bias step $f_{3dB}$ results for a sanity check. Figure 3.16 shows the histogram of $f_{3dB}$ values.
3.2.2 Summary

The detector yield is mainly affected by an open column and an open row in the readout system, which respectively cause failure of 25 and 16 detectors. Table 3.1 shows the detector parameters in the field.

Table 3.1: Parameters of the LF array.

<table>
<thead>
<tr>
<th>Array</th>
<th>Num</th>
<th>$\Omega_b$/nsr</th>
<th>Beam (arcmin)</th>
<th>$f_{3dB}$ (Hz)</th>
<th>NET ($\mu K/\sqrt{s}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>27 GHz</td>
<td>123</td>
<td>4165±317</td>
<td>6.8±0.46</td>
<td>48.1±27.2</td>
<td>72</td>
</tr>
<tr>
<td>39 GHz</td>
<td>97</td>
<td>3064±1204</td>
<td>5.7±0.67</td>
<td>68.4±24.5</td>
<td>36</td>
</tr>
</tbody>
</table>

The estimated $NET_{array}$ of the LF array is higher than the target values, which are 32 $\mu K/\sqrt{s}$ at 39 GHz and 35 $\mu K/\sqrt{s}$ at 27 GHz. The 39 GHz $NET_{array}$ is still on the same scale with the target value, while the 39 GHz $NET_{array}$ is significantly higher. As shown in Figure 3.15, the NETs at 27 GHz are supposed to have similar distribution as the ones at 39 GHz, but instead more widely spread out. Hence the
issue is rather related to the conversion factor or the optical response than to the *NEP* level of the detectors.

![Array plot of $\frac{p_a - p_b}{p_a + p_b}$ for 27 GHz detectors](image)

![Array plot of $\frac{p_a - p_b}{p_a + p_b}$ for 39 GHz detectors](image)

Figure 3.17: Array plots of the AB ratio for the 27 GHZ (top) and 39 GHz detectors. Arrows in the top plot illustrate the orientation, in which the AB ratio increases.

We find out that the non-uniformity in the optical response of the detectors from the same pixel greatly undermines the array sensitivity. In each pixel, there are two detectors for each frequency band, which are categorized as “type A” and “type B”. The categorization is based on the polarization angle and the TES location within each pixel. We notice that the LF array has an overall “AB pattern”. Figure 3.17 shows
the array plots of a parameter called the “AB ratio”, defined as \((P_a - P_b)/(P_a + P_b)\), where \(P_a\) and \(P_b\) is the peak response (in power units) to Jupiter of detector A and B from the same pixel at the same frequency band. Within each rhombus of the detector wafer, the AB ratio goes from being negative to being positive from one 60° corner to the other, and the pattern is rotationally symmetric across the entire array. The issue for the 27 GHz detectors is especially worse, with some detectors having almost no optical responses at the corner area, which tremendously degrades the detector sensitivity. The HF array has similar issue but less severe, while the MF arrays have not shown such pattern. The cause of the AB pattern remains unknown. On candidate is the geometrical design with different routing lengths of the two types of detectors. For future work, we aim to identify or confirm the cause of the AB pattern and minimizing the issue for the future CMB detector arrays.
Chapter 4

Simons Observatory Detector Arrays

SO plans to deploy 49 dichroic TES bolometer arrays in total with six frequency bands centered from 27 GHz to 280 GHz [88]. Similar to the AdvACT arrays, each SO array unit, which is called a universal focal plane module (UFM), contains around 2000 TES bolometers measuring CMB at two frequencies and in two polarizations (characterization of the SO prototype TES can be seen in [89]). There are two major differences between AdvACT arrays and UFMs. First, the UFM’s cold multiplexing readout system, which is called the universal microwave multiplexing module (UMM), sits on top of the detector wafer stack, instead of on the same plane as in AdvACT. The size of the UFM is therefore smaller (six-inch-diameter hexagon), allowing denser packing of detector arrays on the focal plane. Second, SO implements an innovative multiplexing system based on the microwave SQUID multiplexer (µmux), which is able to read 2000 detector signals with one to two pairs of coaxial cables.

The UMM is still under development. So far, we have roughly finished the development of the packaging method, which will be described in details in Section 4.3. Section 4.4 provides preliminary results from one prototype UMM that is relatively
close to the final design, but only has a small number of $\mu$mux resonators installed. The design has been advanced in several ways since this early prototype, but we are still working on improving the packaging yield and possible crosstalk between the two transmission lines on the same UMM. These problems will be described in Section 4.5.

4.1 Microwave Multiplexing System

4.1.1 Microwave SQUID Multiplexer

The SQUID $\mu$mux system has been largely developed at NIST, with details in [90] [91] [92] [93]. The system contains multiple microwave resonators, each of which has a unique resonant frequency $f_0$ and is capacitively coupled to a common transmission line that is in the form of coplanar-waveguide (CPW). In the transmission coefficient $S_{21}$ spectrum, each resonator shows as a trough-shaped feature with the minimum at its $f_0$. A quarter-wave resonator is coupled to an rf SQUID [94] (as shown in Figure 4.4), so that when applying an external flux, the effective load inductance of the resonator is changed, which in turn changes $f_0$. Therefore, by coupling a TES to the dissipationless rf SQUID, the TES signal, which is in the form of the TES current’s variation, is transformed into the magnetic flux on the rf SQUID and then into effective inductance variations for the resonator, and eventually into changes of the complex $S_{21}$ around the $f_0$ region.
Figure 4.1: (Left): Circuit diagram of a single resonator coupled via an effective capacitance $C_c$ to a transmission line with impedance $Z_0$, and the corresponding TES circuit that is coupled to the resonator via rf SQUID (modeled here in terms of its junction inductance $L_j$ and a series inductance $L_s$).

The complex transmission $S_{21}$ as a function of frequency for a single resonator can be modeled as:

$$S_{21}(f) = ae^{i\alpha}e^{-i2\pi f\tau} \left[ 1 - \frac{Q/\tilde{Q}_c}{1 + 2iQ(f/f_0 - 1)} \right], \quad (4.1.1)$$

where $\alpha$ is the initial phase and $\tau$ is the cable delay. The $Q$ is the total quality factor, related to the resonator bandwidth ($BW$) by:

$$Q = \frac{2\pi \omega_0}{BW}, \quad (4.1.2)$$

where $\omega_0 = 2\pi f_0$ is the resonant angular frequency. Setting the proper frequency spacing between adjacent resonators, $\Delta f$, relative to the bandwidth helps reduce crosstalk between those resonators. The coupled quality factor $Q_c = Re\{\tilde{Q}_c\}$ de-
scribes the resonator’s energy loss to the feedline (transmission line). There is an extra internal quality factor $Q_i$, which describes the resonator’s internal loss, such that (neglecting the prefactor $ae^{i \alpha} e^{-i2\pi f \tau}$) the $S_{21}$ at the resonant frequency is:

$$\left| S_{21}^{\min} \right| = \frac{Q}{Q_i}. \quad (4.1.3)$$

The relation between the total quality factor $Q$, the coupled quality factor $Q_c$ and the internal quality factor $Q_i$ is:

$$\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_i}. \quad (4.1.4)$$

In short, $Q$ and $Q_i$ respectively determine the width and depth of the individual troughs in the $|S_{21}|$ frequency spectrum. The current $\Delta f$ for SO is $\sim 1.8$ MHz with $BW \sim 100$ kHz, allowing for $\sim 2000$ resonators in the 4-8 GHz band or 1000 resonators between 4 and 6 GHz.

### 4.1.2 Flux Ramp Modulation

The voltage response of a SQUID to an external magnetic flux is close to a sinusoidal curve. As mentioned previously in Section 2.2, the method used in TDM to linearize the SQUID response is to use a flux-locked loop, which implements a feedback loop to apply an oppositely-directed flux through the SQUID with respect to the flux induced by the TES signal to maintain the SQUID at a point in its response that is linear and has maximal slew rate (the sensitivity of the SQUID response degrades as the slew rate decreases). The flux-locked method is suitable for TDM systems since TDM systems only read out a single detector per time so that SQUIDs from the same column are able to share a common feedback loop. However, it is not suitable for frequency-division multiplexing, since the system simultaneously reads out all detector signals. In stead, the SQUID $\mu$mux system uses a innovative method called
flux ramp modulation \[95\], which applies a fast ramping flux to all of the SQUIDs with a single line passing in series through inductors coupled to every SQUID.

Figure [4.2] shows conceptual plots of how the flux ramp modulation works (see also \[92\]). With a flux ramping rate $\omega_{FR}$, the SQUID response $V(t)$ can be approximated by:

$$V(t) \propto \sin \left( \frac{2\pi \Phi_{total}(t)}{\Phi_0} \right) = \sin(\omega_{FR}t), \quad (4.1.5)$$

where $\Phi_{total}$ is the total magnetic flux of the SQUID, $\Phi_0 = \frac{h}{2e}$ is the flux quanta, where $h$ is the Planck constant and $e$ is the electron charge. The extra phase $\Delta \phi$ caused by an incoming signal can be approximated by $A \sin(\omega_s t)$. Equation \[4.1.5\] then becomes:

$$V(t) \propto \sin \left( \frac{2\pi \Phi_{total}(t)}{\Phi_0} \right) = \sin(\omega_{FR}t + A \sin(\omega_s t)). \quad (4.1.6)$$

The response becomes the sum of infinite harmonics $\sin(\omega_{FR}t + n\omega_s t)$. When the ramping rate $\omega_{FR}$ is considerably higher than the amplitude $A$ and the signal frequency $\omega_s$, the signal’s power is mostly stored in the first harmonic $\sin(\omega_{FR}t + \omega_s t)) = \sin(\omega_{FR}t) \cos(\Delta \phi) + \cos(\omega_{FR}t) \sin(\Delta \phi)$. With a piece of data $D(t)$ containing an integer number of SQUID response cycles, the phase difference $\Delta \phi$ can be calculated by Fourier transformation:

$$\Delta \phi = \arccot \left( \frac{\int D(t) \sin(\omega_{FR}t) dt}{\int D(t) \cos(\omega_{FR}t) dt} \right). \quad (4.1.7)$$

In practice, we can’t keep increasing the flux to infinity. Instead, we apply a sawtooth wave to the flux ramp line (shown as Figure \[4.3\]), with the amplitude of each period corresponding to a few flux quanta. Due to variation in the inductances coupled to flux ramp to the SQUIDs and the precision with which the sawtooth amplitude can be specified, there are discontinuities in the SQUID response near the beginning and the end of each sawtooth period; we therefore select a data window
Figure 4.2: Conceptual plots of FR with simulated data. Note that the data sampling rate and carrier frequencies are on different scales than are used in reality. (a) Power spectrum of simulated SQUID response, in which a TES signal $\Delta \phi = \sin \omega_s t$ is modulated by two different carrier frequencies $\omega_{FR}$, for each the main lobe is located at $\omega_{FR}$, and the adjacent two side lobes at $\omega_{FR} \pm \omega_s$. (b) Time series of SQUID response to the flux ramp modulation alone and in combined with the input signal with $\omega_s = 0.03 \omega_{FR}$. The vertical lines define windows over which the demodulation will be carried out. The signal is also plotted (and uses the right-side vertical axis). (c) Zoom view of (b) for $\omega_s = 0.03 \omega_{FR}$, in which the signal appears as the increasing phase shift between the SQUID response to the flux ramp with and without the signal. (d) The demodulated signal in comparison with the original signal.
within each sawtooth period. The length of the window is approximately four to five complete SQUID curve cycles. The set of data points within the window then goes through discrete Fourier transform to recover a single estimate of \( \Delta \phi \) induced by the signal during the time interval of the window:

\[
\Delta \phi = \arccot \left( \frac{\sum D(t_i) \sin(\omega_{FR} t_i)}{\sum D(t_i) \cos(\omega_{FR} t_i)} \right). 
\] (4.1.8)

The resulting time series for \( \Delta \phi \) will be eventually converted to TES current.

Figure 4.3: Conceptual plot of the SQUID response time series when applying a sawtooth current to the flux ramp line. For each sawtooth period, a window (shown as green blocks) defines a set of data and after demodulation, that window yields a single estimate of the TES signal.
4.2 UMM Design and Components

The original SO UMM is described here. It is called the silicon UMM. Currently a second design is being developed that uses /mu\mux\ chips, but handles routing of the DC and RF lines differently. It is called the Cu UMM. The original UMM design was mostly carried out by Patty Ho with Gene Hilton and Erin Healy. We have assembled a series of UMMs of the original type (see Table ??). I did most the die bonding and wire bonding of the first few UMMs. I will describe tests on two of these UMMs.

The UMM mostly consists of a DC wafer, an RF wafer, 28 /mu\mux\ chips, as well as associated metal mechanical parts and magnetic shields. Each /mu\mux\ chip contains CPW microwave resonators with corresponding rf SQUIDS that are able to read out 64 TESes. As shown in Figure 4.4, /mu\mux\ chips are glued in eight rows on top of the DC wafer, which contains resistive and inductive components for TES biasing and coupling. The RF wafer provides the CPW and ground planes connection for the
µmux chips, as well as flux ramp lines mentioned in Section 4.1.2. The RF wafer has eight rectangular cutouts and sits on the DC wafer, with µmux chips in the cutouts (see the bottom diagram in Figure 4.4). The wafers are aligned by two stainless steel dowel pins, which are respectively inserted into an alignment hole and a slot through the DC and RF wafers. The UMM electrical connections for completing the circuits and grounding are achieved by approximately 10,000 wire bonds between µmux chips and from µmux chips to the DC and RF wafers. Besides, there are two flexible circuits that respectively connect the CPW and DC lines (flux ramp lines and detector bias lines) on the wafers to the readout system at higher temperature stages. One side of each flex is glued on the RF wafer, while the other side, on which the 37-pin DC connector or the mini sub miniature push-on (SMP) RF connector is soldered, is mounted on the metal cover of the UMM. There are pogo pins or BeCu springs inserted on the bottom surface of the cover, pushing and securing the wafers during the UMM test. After completion, the UMM will be transferred onto the detector wafer stack and the detectors will be connected to the UMM through wire bonds on the edges.

4.3 UMM Integration

Figure 4.5: Flow chart of the integration process for UMM.

Figure 4.5 shows a brief overview of the assembly flow process. There are several criteria that the packaging methods have been developed to meet. First and foremost, similar to the case in AdvACT, the UMM needs to be electrically functional. Different
parts should be firmly mounted or glued to each other with good alignment to prevent opens and shorts from the dense wire bonds. Meanwhile, the bonding parameters, which define the shape of the wire bonds, need to be developed to optimize the resonator performance. Mechanically, the UMM ought to be cryogenically robust to undertake multiple cooldowns, and strong enough for shipping to other testing facilities or Chile. Besides, the integration process should allow the arrays to be assembled uniformly and simultaneously to enhance the overall efficiency of mass production.

4.3.1 Inspection

Before the integration formally starts, all parts including \( \mu \text{mux} \) chips, DC wafers and RF wafers are pre-checked. DC wafers are tested cryogenically to check for shorts and opens of the detector bias lines, while \( \mu \text{mux} \) chips and RF wafers are visually inspected for fabrication defects and probed for continuity of CPW and flux ramp lines, and the absence of shorts.
4.3.2 Gluing and Die Bonding

Figure 4.6: (Left): Nordson dispenser, which has an automated dispensing robot that carries a glue-loaded syringe. (Top right): Lines of dispensed glue on a DC wafer. (Bottom right): Placing and aligning \( \mu \text{mux} \) chips on a DC wafer with the die bonder.

The first stage of the integration process is gluing \( \mu \text{mux} \) chips to a DC wafer. The first few UMM prototypes used Stycast 1266, which has been proven in AdvACT to be cryogenically robust, to have its CTE matched with silicon, to have good thermal conductivity, and to be a strong adhesive for wire bonding. To make sure the gluing of all 28 \( \mu \text{mux} \) chips is uniform and consistent, the encapsulant and catalyst of Stycast 1266 were mixed by an automatic mixer, which uses a metal block to blend the mixture at 360 rpm for 1 min. The mixing process is conducted under vacuum to reduce air bubbles. Compared with mixing by hand, the mixing process with the automatic mixer produces glue mixture with higher consistency in viscosity as a function of heating time, which is crucial for later steps of depositing a consistent amount of glue on the DC wafers. Before glue disposition, there is an extra step of preheating the glue at 55\(^\circ\)C for 10 min to increase the viscosity, which prevents chips from moving
during die bonding due to excess surface tension of the glue. The glue is automatically
dispensed onto the DC wafer using an air-controlled Nordson EFD dispenser. The
deposited glue amount is tuned with the preheating time as well as the air pressure
so that once the chip is placed, the glue is enough to cover most of the chip area
but won’t be too much to flow out and run onto the bond pads. Afterwards, \( \mu \text{mux} \)
chips are placed on the DC wafer by a West Bond die bonder. There are cross-shaped
markers fabricated on the DC wafer, pointing out the corner positions of the chips to
assist the die bonding alignment. After all of the chips are placed, the DC wafer is
left on the die bonder stage for at least 15 min to prevent chips from moving relative
to the DC wafer. Finally, the fully loaded DC wafer is transferred to a hot plate at
65°C for 2 h to cure the Stycast. The final alignment between the chip and the DC
wafer is better than 60 \( \mu \text{m} \). Figure 4.6 illustrates some of the die bonding steps.

After the Stycast is cured, the DC wafer is transferred to a copper base. The RF
wafer is then mounted on top of it. The two wafers and the copper base are aligned
by two alignment pins. Since the alignment holes on the wafers are designed to be
around 100 \( \mu \text{m} \) larger than the pins, several drops of rubber cements are applied along
the edge of the RF wafer and the DC wafer, preventing horizontal movement of the
wafers with respect to the copper base. The UMM is then ready for wire bonding.

**Wire bonding development** Different from AdvACT, the shape of wire bonds
plays a crucial role since the related capacitance and inductance of the wire bonds
can affect the feedline impedance and the resonator performance [96]. Wire bond-
ing parameters are developed to optimize the performance of the resonators, with the
constraints from the geometric design of the \( \mu \text{mux} \) chips and the capacities of the wire
bonder taken into consideration. Two types of wire bonds are described. The first
type is used for the CPW bonds connecting CPW segments belonging to individual
chips or on the RF wafer. Connection wire bonds can cause impedance discontinu-
ities along the CPW line. To prevent sensitivity degradation, the impedance discontinu-

ities need to be minimized. SO uses 25-µm-diameter Al wire for wire bonding, the inductance of which is approximately proportional to the length of wire and inversely proportional to the number of wire bonds per connection [97]. The second type is used for the ground bonds connecting the different ground planes on the μmux chips, RF wafers, DC wafers and metal parts. Connections to ground planes keep the CPW single moded, reducing radiation from its antisymmetric mode. Each length of CPW is connected to a transition section (“launch”) to a bond pad at each end. The RF current density reaches the maximum near the CPW launch, therefore the ground bonds there need to be flat and dense. Due to limits from the bonding machine’s capability, the “flat” bonds are wire bonds with peak height less than 80 µm, and the minimal length of flat bonds without reduction in bonding consistence and strength is around 600 µm.

To develop the wire bonding parameters, I conducted a set of time domain reflectometry (TDR) tests at NIST to measure the impedance of various chip-to-chip bonding configurations. Figure 4.7 shows the test results. The configuration with minimal impedance mismatch has three 600 µm-long CPW bonds and seventeen ground bonds, among which six connect each chip to the RF wafer ground plane and five connect the two chips’ ground planes to each other. For the wire bonds between each chip and the RF ground plane, the result shows no difference between the configurations with three bonds per chip and six bonds per chip. Since the bonds are flat and subject to break due to relative movement between the RF wafer and the DC wafer, the configuration with six ground bonds per chip is chosen to be the final, using the three extra ones to provide mechanical protection for the inner three, and minimize the relative movement.
Figure 4.7: TDR results with different wire bonds configurations. (Top): photograph of the overall configuration. A copper box houses four silicon chips, carrying CPW lines between the input and output SMA connectors on the sides. (Middle left): Bonding schematics for individual configuration tests. The top surfaces of the copper (yellow) and silicon chips (grey) are at the same height. The purple bonds connect the center traces of the CPWs. The corresponding tests vary the number and length of the CPW bonds and the result is shown in the middle right and bottom left. The red bonds are variables in the ground bond test, the result of which is plotted in the bottom right. The red and purple boxes illustrate the steps of adding bonds between the tests. The blue bonds were placed at the beginning of all the tests and kept unchanged. For the tests of ground bond configuration, two 600 µm CPW bonds were placed. For the CPW bond tests, all of the blue and red bonds were present. The following three plots are the TDR measurement results, showing the impedance versus propagation time and having three sharp peaks in the middle due to the three chip-to-chip connections. (Middle right): TDR results of placing one, two, three or four flat 600 µm CPW bonds. The old bonds remained in place while one new bond per connection was added for the next round of test. (Bottom left): Result plot of TDR measurements that were done on a configuration that is the same as the middle right except that the lengths of the CPW bonds were 1.2 mm. (Bottom right): Plot of the TDR measurements of ground bond tests, during which ground bonds were added according to the red labels in the top left and one TDR measurement was taken after each bonding step. The optimal bonding configuration comes from examining the three plots and is summarized in the text.
4.3.3 Wire Bonding

Figure 4.8: (Top left): Photograph of the bonding mask that presses the RF wafer down to prevent it from moving during bonding. (Top right): Photograph of the bonding jigs. (Bottom left): Bonding schematic. (Bottom right): Photograph of the chip-to-chip connection, in the final bonding configuration.

UMM wire bonding Informed by the above TDR test described in Figure 4.7 and above, the final µmux bonding scheme is shown in Figure 4.8. Apart from the wire bonds related to µmux chips, there are perimeter bonds between the edge area of the RF wafer grounding plane and the copper base. These bonds, with pitch of 200 to 300 µm, connect the ground of the µmux chips to the (copper) ground of the UMM, which is eventually connected to the cryostat ground.

There are several mechanical jigs designed for improving the bonding uniformity across the whole array and the strength of wire bonds. These include a stainless steel bonding mask, a Thorlabs rotary stage and an Al holder (Figure 4.8). The bonding mask is developed for overcoming the curvature of the RF wafer, which greatly undermines the bonding strength and consistency since the bar area of the RF wafer bounces as the bonding tool touches it. The mask is mounted on top of the
RF wafer using screws with the assistance of alignment pins to prevent the mask from crashing onto any wire bonds during the mounting. The surface of the mask facing down is also covered by a layer of low tack tape to avoid direct contact between the RF wafer and metal. The Al holder serves as an mechanical interface between the Thorlabs rotary stage and the UMM, while the rotary stage allows 360° of rotation and adjustment of height by one inch. For each UMM, more than 10,000 wire bonds are placed with bonding strength larger than 7 g.

4.3.4 Flex Gluing and Mechanical Parts Installing

![Figure 4.9: Photographs of an UMM with flex and metal covers on.](image)

After the completion of wire bonding, the flexes are glued on top of the RF wafer with Stycast 1266. The process is similar to that used in AdvACT. Jigs are designed that hold the flex stiffener by vacuum and then are mounted to the copper base until the Stycase 1266 is cured. As shown in Figure 4.9, six sections of copper wall are then mounted on the copper base, while the Al cover sits on top. The connector side of the
flexes are secured to the cover by screws, and the flex forms a curve to compensate the CTE difference between the silicon, the copper base and the lid during cool downs.

4.4 \( \mu \text{Mux} \) Measurement and Characterization

There are two types of \( \mu \text{mux} \) measurements. The first type is to measure the transmission coefficient \( S_{21} \) spectrum with a Vector Network Analyzer (VNA). The purpose of this measurement is to quickly validate resonator parameters including the three quality factors. The analysis process starts with finding peaks in the frequency spectrum of \( |S_{21}| \) to detect resonators. We then extract segments of data centered at each detected \( f_0 \) and fit the piece of complex \( S_{21} \) to the model in Equation 4.5.1 to get \( Q, Q_i \) and \( Q_c \). Figure 4.10 shows the VNA measurement for a single \( \mu \text{mux} \) chip with histograms of the resultant parameters. VNA measurement can also be used for probing the crosstalk between two CPW lines from the same UMM. Figure 4.11 shows the VNA measurement of a prototype UMM called UMM-v0b, which has two sets of CPW chains. The chip-side CPW is populated with 13 CPW-only dummy chips and one real \( \mu \text{mux} \) chip, while the through-side CPW is an integrated CPW line fabricated on the RF wafer. The crosstalk measurement is achieved by connecting one portal of the chip side to the VNA output and one portal of the through side to the VNA input. The transmission coefficient spectra of both the crosstalk and the through side alone show trough-shaped features exactly at the resonant frequencies from the chip side. Discussions of the crosstalk will be continued in Section 4.5.
Figure 4.10: (a) The frequency spectrum of $|S_{21}|$ for a single V2.1 µmux chip. (b) Histogram plots of $Q_i$, $Q_c$, $Q$, $b_r$ (defined as $f_0/2Q$) and frequency spacing between adjacent resonators on the µmux chip.
Figure 4.11: (Left): Photograph of the UMM-v0b labeled with the CPW traces. The connectors of the chip-side CPW are 1 and 2, while for the through side they are 3 and 4. (Top right): Frequency spectra of $|S_{21}|$, $|S_{34}|$, $|S_{23}|$. (Bottom right): Zoomed view of the top right.

The second type of measurement is the $\mu$mux characterization using SLAC Microresonator Radio Frequency (SMuRF) Electronics, which are being developed in the SLAC National Accelerator Laboratory. Information about the electronic devices and characterization algorithms is given in [98] [99]. The SMuRF system will be used in the field for SO CMB observations. Similar to the VNA analysis, the $\mu$mux characterization with the SMuRF system starts with sending out broadband signals to the CPW line to locate individual resonators in frequency space. Then for each resonator, SMuRF makes a finer sweep of $S_{21}$ around the resonant frequency and plots the $S_{21}$ curve in the $IQ$ plane, where $I$ is the real part of $S_{21}$ and $Q$ is its imaginary part. The purpose of this step is to find the complex $\eta$, by whose angle $IQ$ is rotated into a new coordinate $I'Q'$ so that the $Q'$ component of $S_{21}$ equals zero at resonance. In other word, small deviations from resonance only result in the variation
of the $Q'$ component. We only need to measure $\eta$ once for each resonator since it is not related to $f_0$ shifts. During operation, the system sends a comb of narrow band probing tones to the UMM CPW, as well as a sawtooth flux ramp signal to linearize the rf SQUID response. Each tone corresponds to a single resonator and gives one complex $S_{21}$ value at a time. SMuRF keeps the tone on resonance by adjusting the tone frequency to make the $Q'$ component zero, and uses this feedback mechanism to get $\Delta f$, the deviation between the tone frequency and the resonance frequency. From $\Delta f$, the rf SQUID response can be calculated, which after flux ramp demodulation yields the phase shift $\Delta \phi$ caused by TES current.

![Figure 4.12: Time series of the tracked resonant frequency (top) and the frequency error for a resonator when the flux ramp current is turned on](image)

The $\mu$mux characterization with SMuRF includes setting up the tone tracking parameters, and adjusting the tone power with respect to the noise measurement (see Figure 4.12). For the tone tracking setup, the process start with adjusting the flux ramp amplitude so that each flux ramp period contains four to five cycles of rf SQUID response. The flux ramp reset rate, which is the inverse of each flux ramp period, is usually set to 4 kHz. The system then automatically finds the carrier frequency $\omega_c$, using which the flux ramp demodulation gives least mean square of
tracking error. The carrier frequency is therefore also called the \textit{lms} frequency, and is approximately the product of the flux ramp reset rate and the number of cycles of rf SQUID response. Figure \ref{fig:tone_tracking} shows an example of tone tracking for a single \(\mu\)mux channel. Due to discontinuities near the two ends of each flux ramp period, we also specify the fractions in each period where the feedback mechanism starts and stops. The tone tracking setup is finished.

The second part of the \(\mu\)mux characterization is optimizing the tone power. The optimal \(\mu\)mux tone power is around -80 to -70 dBm. However, the attenuation of coaxial cables and other cryogenic components varies with frequency, resulting in a \(\sim 10\) dB difference in \(S_{21}\) amplitude from 4 to 8 GHz. SMuRF divides 4-8 GHz into eight bands and is able to vary the tone power for each band by adjusting the internal step attenuators. We take a set of noise measurements on each channel with different step attenuation values. The noise data is a 60 s timestream that is converted to units TES current. The resulting power spectrum is fitted with a model to get the white noise level in \(pA/\sqrt{Hz}\). The optimal attenuation value can then be chosen for each band to minimize the resonators’ noise level (noise properties of microwave resonators are described in detail in \cite{100}.

\section{4.5 Ongoing UMM Development}

The UMM development is not complete. This section will describe the remaining problems to be solved, which can be categorized into problems related to packaging and resonator performance. We will continue to focus on the so-called “silicon UMM” though parallel work is ongoing with the “copper UMM”. The data presented in this section is mainly from the measurements of a prototype silicon UMM, called UMM-v3. Due to
4.5.1 Packaging and Designing

**Wafer securing.** Currently, the RF and DC wafer are aligned by two alignment pins. Since the size of the alignment holes has around 100 $\mu$m tolerance from fabrication, we also add a few drops of rubber cement on their edges, further preventing relative movement between the wafers, and between the wafers and the metal base. However, this method creates a challenge if we want to replace the RF wafer since rubber cement on the edges could seep into the tiny gap between the RF and DC wafer. Besides, there is usually a curvature to the RF wafers due to the silicon etching process in fabrication, which degrades the bonding quality. As aforementioned, we currently uses a thin metal mask to press the RF wafer down during wire bonding. This method is good enough for integration of individual UMMs, but inefficient for mass production of 49 arrays since the mask itself has non-uniformities and we need to spend a long time to adjust it every time we integrate a new array. There is a new method proposed, in which the wafer is held down by several beads, each with a string that is tied up at the edge of the array. By fastening the string, the pressure of the beads can be easily adjusted. So far, this method remains untested.

**Replacement of components.** The UMM design has two sets of CPWs with the corresponding flux ramp lines. Integration handling can possibly scratch the traces on the RF wafer, creating shorts or opens. A single electrical short or open unavoidably results in waste of all the $\mu$mux chips belonging to the CPW. We have proposed making a partial rf wafer, which contains a ground plane and a single segment of flux ramp or CPW trace, as long as corresponding bondpads. The partial piece can be glued on top of the RF wafer, covering the original trace. The ground planes will be connected by wire bonds along the perimeter of the partial pieces. Again, we still need to validate the practicality of this method.
4.5.2 Resonator Performance.

Figure 4.13: (Top): $|S_{21}|$ spectrum of UMM-v3 from 4 to 8 GHz. The region in the blue box contains resonators at higher frequency with degraded $|S_{21}|$ depths. (Bottom): $|S_{21}|$ spectrum of UMM-v3 from 5.2 to 5.8 GHz. The blue dashed line sketches the dip depths of the resonators.

Transmission coefficient. There are several types of undesirable features in the UMM-v3 $S_{21}$ spectrum. In comparison with the measurement of test boxes that each contain only a single $\mu$mux chip (shown in Figure 4.10), resonators in the larger format UMMs tend to have varying $S_{21}$ depths at resonant points (shown in Figure 4.13) exhibiting a quasi-sinusoidal envelop bounding the a lower limit on $S_{21}$ for resonators. This envelope is a hallmark of low $Q_i$ coupled with frequency variation of $Q_e$ because
$Q_e$ depends on the fixed geometry of the coupling capacitor between the resonator and the feedline. The geometry is not varied to match each resonator.

As the frequency gets higher, resonator performance is further undermined. At higher bands (6-8 GHz), the depth and quality factors are significantly lower, and the upper edge of the $S_{21}$ in Figure 4.13 is dominated by a standing wave. We also find similar properties in $|S_{21}|$ for UMM-v0b (see Section 4.4 and Figure 4.11), in which there are 27 dummy CPW chips with only a single real µmux chip. Therefore the cause is not merely too many resonators on the same CPW coupling to each other. We suspect the resonator performance at high frequency is influenced by box modes, in which radiation loss at certain frequencies arises from coupling to standing waves in the UMM metal enclosure. We therefore have decided to have two pairs of CPW lines, both having resonators only in the 4-6 GHz range.

1The relationship between the physical position of a resonator on the chip and its targeted resonant frequency is complex and varies according to which fabrication run these early chips came from.
Figure 4.14: (Left): Sketch showing the UMM-v3 chip mapping and the crosstalk measurement setup. The green boxes illustrate the dummy chips that only have a CPW segment, while the red ones represent µmux chips. The numbers labeled on the chips represent the subband where resonators in each chip are located on the frequency space. The 4 to 8 GHz frequency span is divided into 28 subbands; subbands 0-13 are in 4-6 GHz and called “low band”, while subbands 14-27 are in 6-8 GHz and called “high bands”. The four horizontal lines on the left indicate the two pairs of CPW lines. (Right three): $|S_{21}|$ spectrum of the UMM-v3 chip-side, and the thru-side with and without SMuRF sending tones to the chip-side.

**Crosstalk.** In a single UMM, since the resonator bands from two CPW lines overlap with each other, crosstalk measurement between the CPWs becomes crucial. Other than the passive VNA measurement (Section 4.4), we have also done crosstalk measurements using both VNA and SMuRF. Figure 4.14 shows the crosstalk test setup of UMM-v3. UMM-v3 has one CPW chain (“thru-side”) populated with dummy CPW chips and connected to a VNA, while the other (“chip-side”) is populated with some real µmux chips and connected to a SMuRF. The SMuRF sends probing tones and keeps adjusting their frequencies to track the resonators, while the VNA simultaneously measures the $S_{21}$ frequency spectrum of the thru-side. The result shows two types of crosstalk features in the thru-side $S_{21}$ spectrum. The first type are ripple
features, and the second type are sharp peaks or troughs sitting at the tip of ripples. Each ripple and peak corresponds to a single resonator from the chip-side. The ripples are relatively shallow, with amplitude less than 0.1 dB, while the peak amplitude varies with resonators and can be as high as 1 dB.

![Figure 4.15: $|S_{21}|$ spectrum of the thru-side with different SMuRF operations to the chip-side resonators. The frequency range of the left plots is from 5.56 to 5.61 GHz with VNA sweep rate $\sim$250 kHz/s, while the right plots are the zoomed view of the frequency region associated with a single resonator from the chip-side with VNA sweep rate $\sim$10 kHz/s.](image)

To understand the crosstalk mechanism, we break the tuning process into individual steps according to SMuRF commands and functions, and take a VNA sweep of the thru-side after each step. Figure 4.15 shows the VNA measurement results in a frequency region that has the worst crosstalk effect, as well as the results related to
a single $\mu$mux channel. We outlined the SMuRF algorithm in Section 4.4. The five tuning steps of the chip-side by SMuRF shown in Figure 4.15 begins with a broadband signal to the chip-side from which SMuRF can roughly localize the resonators. The second step is to finely locate the resonators with a gradient descent algorithm and then send a single narrowband tone to each resonator. The third step is finding $\eta$ for each resonator as mentioned in Section 4.4. The system then automatically initiates the tone-tracking mechanism. The fourth step is turning off the feedback of tone tracking, and the fifth step is turning on both tone tracking and flux ramp current. No matter what tones the SMuRF sends or whether SMuRF sends any tones to the chip-side, the ripple feature persists. In contrast to the ripple feature, most of the peaks only show up when the tone-tracking is on (see Figure 4.15).

Figure 4.16: (Left): Zoomed view of the thru-side $|S_{21}|$ spectrum when the chip-side tone tracking was off, or on with and without the flux ramp current. One ripple feature and many peaks are visible. (Right): The amplitude and phase spectrum of the thru side $S_{21}$ when both the tone-tracking and flux ramp of the chip-side were turned on, showing individual peaks at different moments captured by the VNA sweep. The VNA sweep rate is around 10 kHz/s.

The two types of crosstalk features might be generated through different mechanisms. First and foremost, the bandwidth of the ripple feature is typically on the scale of 100 kHz. If the peaks are generated by the same mechanism as the ripples, applying the broadband tone (step one), which is composed of a pack of single tones that are densely distributed in frequency space, would be expected to amplify the ripple
in the same way that the peaks are amplified in the third row of plots in Figure 4.15. Instead, the $|S_{21}|$ spectrum stays the same except for becoming noisier, indicating that the peaks are only generated if the tone power is within a few kHz of the resonance frequency. This explains why the majority of peaks only show up when the feedback for tone-tracking is turned on, since then the tone frequency is constantly adjusted to be on resonance. Another piece of evidence comes from overlaying the $|S_{21}|$ spectra of step three and step five (Figure 4.16). When both the tone-tracking feedback and the flux ramp current are on (step 5), the flux ramp signal changes the resonance frequencies with a carrier frequency that is $\sim 16$ kHz, while the tone keeps following $f_0$, resulting in a rapidly moving peak in the $|S_{21}|$ spectrum of the thru-side. As the VNA sweeps from left to right, it captures in multiples moments the same resonator and therefore shows multiple peaks. If the peaks are generated by the same mechanism as the ripples, the flux ramp signal would also move the ripple along the frequency axis. On the contrary, the shape of the $|S_{21}|$ spectrum from step five is consistent with that from step three, which further confirms our assumption.
Figure 4.17: Complex plane plot showing the algorithm of tone-tracking. The blue curve is a resonator’s $S_{21}$ in the complex IQ plane, with three dots respectively at resonant frequency $f_0$ and $f_0 \pm 10$kHz. Here $I$ is the in-phase component and $Q$ is the quadrature component, which is $\pi/2$ out of phase. The red curve is the same resonator’s $S_{21}$ in the rotated $I'Q'$ frame, where $Q'$ component equals 0 at resonance. The complex tracking parameter $\eta$ defines $I'Q'$ relative to IQ. The green curve shows the resonator in $I'Q'$ when it experiences an extra phase shift of 5°, resulting an extra $Q'$ component that increases the frequency tracking error.

So far, little is known about how much effect the crosstalk will have on resonators because we have not had any arrays with two feedlines with overlapping resonator bands until very recently (The results from this and other more recent assemblies will be reported elsewhere). We therefore explore the crosstalk effect using simulations based on current data. We analyze the special case, in which two resonators from two different CPWs are close to each other in frequency space. We find out that due to the tone-tracking mechanism, resonators are affected by crosstalk through the $S_{21}$ phase shift, rather than the $S_{21}$ amplitude change. As mentioned previously, the system keeps tones on resonance by nulling the $Q'$ component of the complex transmission coefficient in the rotated $I'Q'$ frame. The $I'Q'$ frame is computed from a fine sweep of $S_{21}$ around $f_0$ and hence is fixed once tone-tracking is initiated. An extra phase shift of $S_{21}$ will then rotate the $S_{21}$ data in the complex plane so that the
$Q'$ component of the actual resonant point is no longer equal to zero, adding error to the $f_0$ tracking (Figure 4.17). We simulate the crosstalk effect by using the $S_{21}$ sweep of a resonator from the chip-side and treat it as if it were located on the thru-side. We use the tracking algorithm in [98], which first identifies $\eta$ and $f_0$ and then gets the transmission coefficients at $f_0 + 10$ kHz and $f_0 - 10$ kHz in the $I'Q'$ frame. The frequency tracking error is $\sim \frac{20kHz \times \text{Im}(S)}{\text{Im}(S_{+10kHz} - S_{-10kHz})}$, where $S$ is $S_{21}$ in the $I'Q'$ frame.

![Figure 4.18](image.png)

Figure 4.18: (Top left): VNA measurement of the ripple, showing the amplitude and phase spectra of the thru-side $S_{21}$. The linear phase shift is caused by the coaxial cable; we fit the phase spectrum from 5.5996 to 5.59975 GHz with a linear model and plot it in green. (Top right): The red data shows the extra phase shift due to the ripple feature, which is estimated as the total phase shift left after subtracting by the linear model. (Bottom): Simulated extra error of resonance tracking if a resonator is on this CPW and close to the ripple feature.

We start with the effect of the ripple features, which are relatively fixed in frequency space as described above. Figure 4.18 shows that apart from the linear phase shift due to the coaxial cable, the effect of which is largely canceled in the $f_{err}$ computation with $S_{+10kHz}$ and $S_{-10kHz}$, the ripple causes is an extra phase increase of $\sim 0.7^\circ$ with $\sim 100$ kHz width. We calculate the extra error in the resonance tracking
for a tone that is already right on resonance by:

\[
\Delta f_{err} = \frac{20kHz \times Im(S \times e^{i\alpha})}{Im(S_{+10kHz} - S_{-10kHz})},
\]

(4.5.1)

where \(\alpha\) is the extra phase. The result shows that 0.77° phase shift would give an extra 1 kHz tracking error, while the typical tracking error without any crosstalk is around 0.3 kHz. The typical amount of \(f_0\) movement with flux ramp current is on a scale of 100 kHz, depending on the resonator’s geometric design. If the original center frequency of the resonator is located right before the phase increase starts, half of each \(f_0\) shifting cycle will be boosted up with \(\Delta f_{err}\) from 0 to ~0.25 kHz.

Figure 4.19: (Left): VNA measurement of the peak feature, showing the amplitude and phase spectra of the thru-side \(S_{21}\). The measurement was taken when the tone-tracking feedback on the chip-side was turned on. Similar to the top right of Figure 4.18, the linear component of the phase shift has been removed here. (Right): Simulated impact on a resonator if it is located on the thru-side CPW, showing the extra tracking error versus the resonator’s \(f_0\).

As mentioned previously, the peak feature is dynamic since it is only generated when the tone is tracking the resonance and the feature shifts with flux ramp current. Although the peak itself is narrowband, lowering the chance that the resonator on the other CPW will run into it, there are“sidelobes,” whose amplitudes are small but whose phase shifts are not negligible. Figure 4.19 shows the phase shift of the thru-side when the flux ramp is off and tone tracking is on. There are three to four
sidelobes on both sides of the peak with maximal phase shift from $0.5^\circ$ to $1^\circ$. Similar to the case of ripple features, we simulate the $\Delta f_{err}$ caused by these phase shifts. Typical $f_{err}$ without the effect by extra phase shift is around 0.5 kHz to 2 kHz. Getting closer to the peak, $f_{err}$ can be as high as 4 kHz. The existence of sidelobes greatly extends the concept of two resonators from different CPWs colliding with each other. In this special case, if two resonators are less than 50 kHz from each other, their $f_0$ tracking will be boosted or deboosted by each other through the crosstalk effect, resulting in irregular SQUID response curves and associated degradation of sensitivity. In some sense, the crosstalk effect between peak features has negative feedback. If the performance of one resonator is greatly influenced by another, the tone frequency of the first will deviate the actual $f_0$, which immediately decreases the peak effect on the other CPW.

Figure 4.20: (Top): Flow chart of how the signal is modulated by the flux ramp and measured by the SMuRF system. The crosstalk effect mostly comes into play at the step of tracking $f_0$. (Bottom): Hierarchy of data processing from raw data to demodulated signal.

In summary (shown in Figure 4.20), the TES signal is combined with the flux ramp signal, which together turn into the SQUID response and subsequent $f_0$ shift.
of the corresponding resonator. The SMuRF system tracks the $f_0$ as raw data. The crosstalk effect mostly comes into the play at the level of tracking $f_0$, adding extra $f_0$ tracking error. Since the $f_0$ tracking error depends on the actual $f_0$, the frequency of the error shown on the raw data is related to the flux ramp carrier frequency.
Chapter 5

Conclusion

In Chapters 2 and 3, we have demonstrated that the last array of AdvACT, the LF array, with unprecedentedly low saturation power, has been successfully operating. The detector yield of the LF array is relatively low compared to the HF and MF arrays due to the increase of the packaging challenges and the incomplete site characterization to the date of writing. There are still much to explore about the operation of the LF array, such as the utilization of the dark detectors. The ACT receivers have been one of the most sensitive CMB instrument with high resolution, and will continue to provide rich CMB data for years.

The observation of the AdvACT instrument will end in October 2021, which is coincidentally around the same time when the next-generation CMB instruments, such as SO, start their observation. The detector techniques for AdvACT and SO do not change significantly. It is the cold readout system that is innovative. In Chapter 4, we demonstrated the design, packaging method and characterization of the prototype SQUID $\mu$mux system that is designed to readout $\sim$2000 TESes with only two pairs of coaxial cables. If succeeds, the SQUID $\mu$mux system will bring profound impact to the development of future generation CMB instruments.
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