

SLAC Microresonator RF (SMuRF) Electronics: A tone-tracking readout system for superconducting microwave resonator arrays

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We describe the newest generation of the SLAC Microresonator RF (SMuRF) electronics, a warm digital control and readout system for microwave-frequency resonator-based cryogenic detector and multiplexer systems such as microwave SQUID multiplexers (μ mux) or microwave kinetic inductance detectors (MKIDs). Ultra-sensitive measurements in particle physics and astronomy increasingly rely on large arrays of cryogenic sensors, which in turn necessitate highly multiplexed readout and accompanying room-temperature electronics. Microwave-frequency resonators are a popular tool for cryogenic multiplexing, with the potential to multiplex thousands of detector channels on one readout line. The SMuRF system provides the capability for reading out up to 3328 channels across a 4-8 GHz bandwidth. Notably, the SMuRF system is unique in its implementation of a closed-loop tone-tracking algorithm that minimizes RF power transmitted to the cold amplifier, substantially relaxing system linearity requirements and effective noise from intermodulation products. Here we present a description of the hardware, firmware, and software systems of the SMuRF electronics, comparing achieved performance with science-driven design requirements. We focus in particular on the case of large-channel-count, low-bandwidth applications, but the system has been easily reconfigured for high-bandwidth applications. The system described here has been successfully deployed in lab settings and field sites around the world and is baselined for use on upcoming large-scale observatories.

I. Introduction

Superconducting detector technologies have enabled highly sensitive measurements across a wide array of scientific applications, including particle physics, astronomy, cosmology, materials science, chemistry, biophysics, and quantum information science¹⁻⁷. As the instruments making these measurements move to ever-denser detector arrays to increase their sensitivity or throughput, there is a growing need for advancements in cryogenic multiplexing techniques to reduce thermal loading at sub-Kelvin stages, cost, and integration complexity. As an example, transition edge sensors (TESs) are a widely-used superconducting detector technology that have been fielded in arrays with over 10,000 detector channels.⁸ Cryogenic multiplexing of TESs on fielded experiments has been achieved with time-division multiplexing (TDM)⁹, frequency-division multiplexing (FDM)¹⁰, and Walsh code-division multiplexing (CDM) techniques^{11, 12}. These technologies operate at \sim MHz frequencies and achieve multiplexing factors of O(100) sensors read out per wire, but are limited by high cost, complexity, and/or noise penalties as they move towards higher multiplexing factors. To achieve future science goals which benefit from larger and/or more tightly packed detector

arrays, new strategies are being devised to push towards the much larger sensor counts required.

Microwave frequency-division multiplexing techniques offer one strategy for addressing these challenges by coupling signals to superconducting microresonators on a single RF feedline¹³. The high quality factors achieved by standard superconducting nanofabrication techniques along with the large readout bandwidth on the microwave feedline in the 0.1-10 GHz frequency range enable O(1000) multiplexing factors on a single coaxial line. In microwave SQUID multiplexing (μ mux), each detector is inductively coupled via an RF SQUID to a unique GHz-frequency resonator such that detector signals correspond to shifts in the resonator's frequency.^{14, 15} The resonator may also serve as both the readout and the detector simultaneously, as is the case for microwave kinetic inductance detectors (MKIDs), where incoming photons break Cooper pairs and induce a modulation of the resonance transmission³. In either scheme, many resonators with unique resonance frequencies can be fit in the available readout bandwidth, allowing large numbers of detectors to be read out on a single coaxial input/output pair.

Despite the promise of these microwave resonator based systems, challenges in both cold and warm components had until recently limited these systems from achieving their full multiplexing potential. Previous implementations of μ mux have traditionally been limited by resonator spacing and bandwidth achieved in fabrication. These challenges have since

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been largely resolved, and resonators with high quality factors, reproducible frequency spacing, and bandwidths suitable for O(2000) multiplexing factors are now being consistently produced.¹⁶ KID devices have similarly achieved sufficiently uniform fabrication and post-fabrication frequency adjustments to realize O(1000) channel multiplexing factor arrays.^{17,18} For several microwave resonator-based readout systems such as bolometric applications of μ mux, which operate at higher resonator drive powers, linearity of the cold amplifier and subsequent warm electronics have become important constraints in the achievable multiplexing factor.

The SLAC Microresonator RF (SMuRF) warm electronics system has been designed to meet these linearity challenges and read out up to 3328 channels in a 4 GHz band. The SMuRF provides both the RF and low-frequency signals required to enable full operation of a microresonator-based readout system. Here and throughout the text, we refer to RF as the frequency regime spanning about 1-10 GHz, while DC refers to both DC and low-frequency (up to about 1 MHz) signals. The SMuRF system is unique in its implementation of fast tone tracking via a closed-loop adaptive filter, which thereby reduces power on the cryogenic amplifiers and RF mixers. This in turn increases the RF dynamic range of the electronics, allowing for improved RF linearity and increasing the achievable multiplexing factor. An earlier version of SMuRF was described demonstrating multiplexed readout of nearly 500 μ mux channels.¹⁹ SMuRF has already been used to read out TES-coupled arrays in laboratory and field observing settings, achieving multiplexing factors of over 900 channels on a single coaxial pair.^{20,21} This system is being used for several upcoming Cosmic Microwave Background (CMB) experiments including Simons Observatory²² and BICEP Array²³.

While SMuRF was designed and optimized for the read out of TES-coupled μ mux resonators, it is flexibly reconfigureable for application to many other sensor and microwave readout technologies including KIDs, magnetic microcalorimeters (MMCs), and even potentially superconducting qubits^{3,24,25}. Many other warm RF systems have been developed for GHz resonator read out, particularly targeting the μ mux and MKID applications.²⁶⁻³³ Compared to those systems, SMuRF offers significant advantages including linear RF performance for high channel density over a 4 GHz bandwidth, integrated high-performance low-frequency signal generation, timing and data streaming integration with large-scale experiment infrastructures, and a unique tone-tracking capability.

In this paper, we present the design and performance of the current SMuRF electronics. We begin with a brief introduction to the superconducting resonator systems that SMuRF has been designed to read out in Section II. We pay special attention to μ mux for large-format arrays to motivate the design needs of the readout electronics. In Section III we outline the system requirements driven by the science goals and design constraints in anticipated applications. We follow with an overview of the system in Section IV and descriptions of its hardware and firmware in Sections V and VI respectively. We briefly describe the data streaming system, integration with instrument control systems, and the end user software envi-

ronment in Section VII. Finally, we present the system performance in Section VIII, in particular demonstrating that the SMuRF system meets or exceeds its design goals.

II. Superconducting Resonator Readout

There are many scientific applications that utilize superconducting microwave resonators as the sensing and/or readout element for ultra-sensitive circuits, including but not limited to superconducting cryogenic sensor arrays for both low-bandwidth signals and pulse detection, qubit control and readout, and quantum information storage and addressing.^{7,13,34,35} These applications have a large range of signal bandwidths ranging from < 1 Hz to 10s of kHz or faster. There is generically a tradeoff between signal bandwidth and channel count: to fit large numbers of sensors in the usable readout bandwidth the detectors must be bandwidth-limited, while larger-bandwidth detectors utilize a greater fraction of the total readout bandwidth and cannot be packed as tightly in frequency space. To maximize performance on the smaller number of channels being used in the higher-bandwidth cases, system linearity is more limited, meaning that the fewer channels may be run before performance begins to degrade.

Since the RF linearity requirements for the high channel density case drive the SMuRF hardware design in particular, we focus here on low bandwidth, high-channel-count application, such as transition-edge sensor (TES) bolometer or MKID arrays for astronomical sensing. We further focus on the case of microwave SQUID multiplexers (μ mux) coupled to low-bandwidth TESs, since the higher resonator drive powers and flux ramp demodulation present additional complications beyond KID systems. In this section we briefly review key terms pertaining to resonator readout and give a brief overview of the TES-coupled μ mux system and typical integration to orient the reader for later sections.

A. Microwave Resonators

The microwave resonators used for cryogenic detector readout are coupled to a common RF transmission line to enable multiplexing. The RF transmission line is excited with input microwave power to interrogate the state of each resonator simultaneously. This input power can be broadband noise, or as is more often the case for high-channel count systems with known resonance frequencies, probe tones tuned to each resonance frequency. We typically operate the RF line in transmission mode and adjust the power levels at the resonator inputs with attenuation on the input and amplifiers on the output. In the discussion that follows we assume that the resonators are coupled to the RF line such that they exhibit a dip in forward transmission (S_{21}) amplitude on resonance; other systems are easily mapped onto this case.³⁶

An example resonator S_{21} is plotted in Figure 1. In this idealized case, the resonance frequency f_{res} corresponds to the minimum of the transmission amplitude at the center of the resonance, though in general resonances can be asymmet-

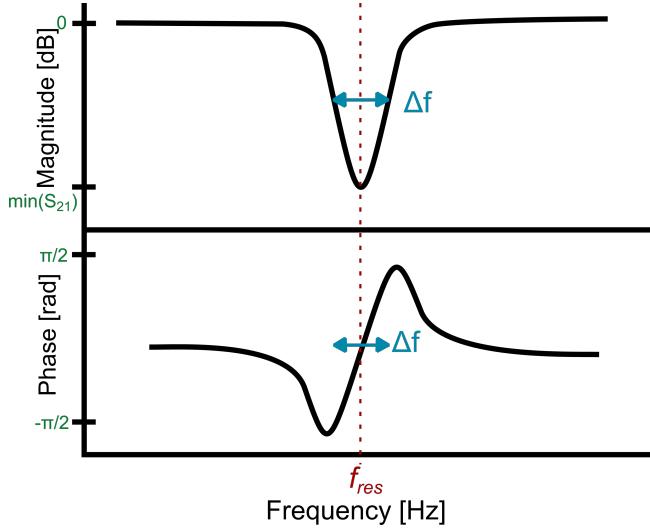


FIG. 1. Example S_{21} amplitude [Top] and phase [Bottom] versus frequency for an idealized resonator. The resonance frequency f_{res} corresponds here to the minimum of transmission. The width characterizes the sharpness of the resonance, which is inversely related to the quality factor Q .

ric. The complex phase undergoes a sign change through the resonance. The width of the resonance in frequency space is characterized by the resonator quality factor Q , which is related to the width of the resonance Δf at full width half maximum by $Q = f_{\text{res}}/\Delta f$. We further differentiate between the internal quality factor denoted Q_i where $Q_i \equiv Q/S_{21}^{\min}$ and the coupling quality factor Q_c , where the total Q is related to the coupling and internal quality factors by $1/Q = 1/Q_c + 1/Q_i$. The complex forward transmission through an idealized resonance may be parameterized as

$$S_{21}(f) = 1 - \frac{Q}{Q_c} \frac{1}{1 + 2jQx} \quad (1)$$

where $x = (f - f_{\text{res}})/f_{\text{res}}$ is the fractional detuning from the resonance frequency.³

We can equivalently plot the S_{21} from Figure 1 in the complex plane, where the resonance traces out a circle with the resonance frequency at the point of intersection with the real axis as seen in Figure 2. For an ideal resonance, this circle is positioned such that on resonance the transmission is entirely real and small changes in the resonance frequency result in changes in transmission along the imaginary axis, or phase quadrature of the resonator. The real axis corresponds to the amplitude quadrature. The effects of cabling and loss, however, may in general displace, scale, and/or rotate the resonance circle to an arbitrary position in the complex plane. We may equivalently refer to the amplitude and phase quadratures of the resonator as I and Q , respectively, which are related by the same rotation and scaling to the electronics I and Q quadratures.

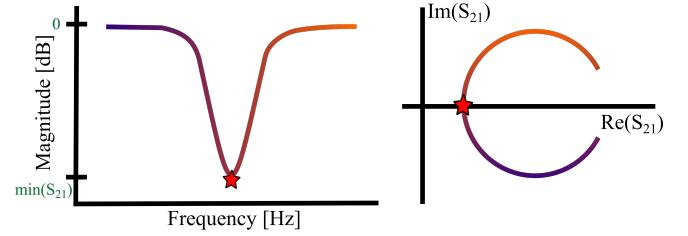


FIG. 2. [Right] Example resonance circle corresponding to an ideal resonance, with magnitude response [Left] for reference. The S_{21}^{\min} is denoted with a star in both cases. The color gradient denotes the same points in frequency in both panels. We see that the resonance generically traces out a circle in the complex plane, with the diameter of the circle corresponding to the depth of the resonance.

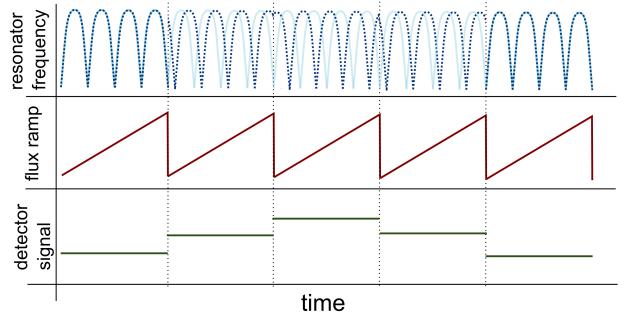


FIG. 3. An overview of the phase modulation scheme of μmux. The top panel depicts the resonance frequency, which shifts in response to flux in the SQUID loop. The middle panel gives the flux ramp, which is typically driven in a sawtooth pattern with amplitude sufficient to drive many flux quanta through the SQUID (in this case, 4). The bottom panel gives an example detector signal, which varies slowly compared with the flux ramp. The detector and flux ramp signals both contribute flux to the SQUID; thus, the resonator response to the detector signal is a shift in the phase of the flux ramp modulated frequency shift (dark dotted lines on top panel) from the detector signal-free case (light lines on top panel).

B. The Microwave SQUID Multiplexer

The microwave SQUID multiplexer (μmux) transduces a cryogenic detector signal, typically from a transition-edge sensor (TES) or a magnetic microcalorimeter (MMC), into flux in an RF SQUID loop.^{14,15} Flux in the SQUID loop changes the effective inductance of the microwave resonator, causing a shift in its resonance frequency. A separate and common flux ramp bias simultaneously applied to all channels linearizes the SQUID response without the need for individual feedback lines.³⁷ An incoming detector signal is therefore transduced into a phase shift in the SQUID response to flux ramp, which is periodic. A simplified picture of this phase modulation scheme is depicted in Figure 3.

The SQUID flux to resonance frequency relation is a quasi-sinusoidal relation parameterized at low drive power by the peak to peak swing of the resonance frequency Δf_{pp} and hysteresis parameter λ , where lower values of λ correspond to

more sinusoidal curves. It is commonly parameterized as³⁷

$$\Delta f(t) = B \left(\frac{\lambda \cos(\omega_c t + \theta(t))}{1 + \lambda \cos(\omega_c t + \theta(t))} \right) \quad (2)$$

where B is a constant related to Δf_{pp} . We note that Eq. 2 does not fully capture the complexity or dependencies of the SQUID curve, which is a subject of intense study.³⁸

The SMuRF system hardware was designed for high linearity use cases with large channel counts. The current firmware was optimized for reading out microwave SQUID devices for CMB TES bolometer readout fabricated by the Quantum Sensing Group at NIST Boulder.¹⁶ The design values for these resonators are resonator bandwidths of $\Delta f \sim 100$ kHz, peak to peak frequency swing of $\Delta f_{\text{pp}} \sim 100$ kHz, quality factor $Q_i > 100,000$, and $\lambda \sim 1/3$. They are optimally driven with input powers around $P_{\text{in}} \sim -75$ dBm. The devices are being developed for the eventual goal of achieving a multiplexing factor of 2000 channels from 4-8 GHz on a single RF line. Other resonator technologies such as MKIDs, TKIDs, and even room-temperature resonators have been successfully read out with SMuRF systems as well, but did not drive design choices.

The maximum multiplexing factor is set by the design value of the peak to peak swing Δf_{pp} and the requirement that the resonances are spaced apart by a factor of 10 times Δf_{pp} in frequency to avoid collisions and crosstalk. In general μ mux devices have roughly matched resonance bandwidth and peak to peak frequency swings to maximize SQUID sensitivity without causing hysteresis or pathological SQUID responses.¹⁵

Reconstruction of detector information is achieved by exciting the resonators with power and monitoring their state. The signal chain consists of a comb of microwave-frequency resonators capacitively coupled to a single RF transmission line. Each detector is uniquely identified by its resonance frequency through a 1:1 mapping between detectors and SQUID input coils. Since the initial resonance frequencies of the resonators can be measured in advance, the resonator excitation is often achieved with a comb of excitation probe tones, each tuned to a single resonator channel. In typical microwave resonator electronics systems, the resonator acts as a moving notch (narrow bandstop) filter that shifts off the probe tone, and each probe tone is monitored for changes in amplitude and phase. As will be discussed more in depth in § III B 2, the RF linearity requirements on a system that must accommodate the possibility of several thousand off-resonance probe tones are stringent, motivating SMuRF's unique tone-tracking capability.

C. Typical μ mux System Implementation

In this subsection we describe a typical implementation of a μ mux system to motivate the electronics and peripheral system designs. The superconducting detector arrays are mounted at the cold stages of a cryostat and connected to room-temperature electronics systems via cabling that is passed through the cryostat vacuum feedthrough. A sketch of a typical system integration is given in Fig. 4.

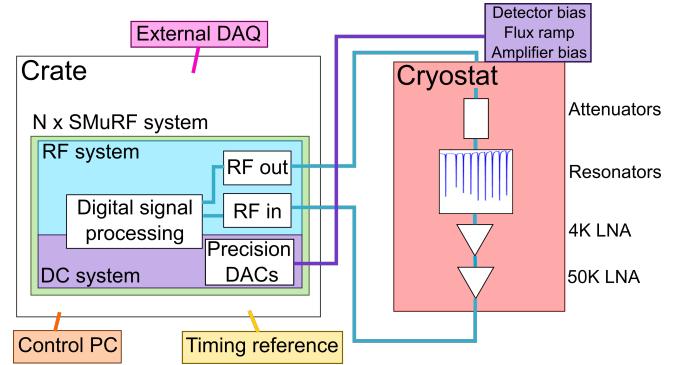


FIG. 4. A typical integration between the SMuRF system and cryostat. The SMuRF system consists of RF and DC components that interact with the cryostat. The number of RF pairs may change depending on the multiplexing factor achieved by the cold devices. The DC lines supply the biases for RF amplifiers and detectors as well as the flux ramp. The system additionally interacts with several peripherals, including an optional external timing reference and data acquisition system (DAQ). Throughout this paper, the color scheme is kept broadly consistent such that the main system computation and processing is given in green, RF components are depicted in light blue, DC components in purple, data output in pink, timing in yellow, the control in orange, and the cryostat and other externally derived parameters in red.

The SMuRF electronics interface with a cryostat through both RF and DC connections. The RF connections consist of two RF ports per transmission line, one each for input and output. The DC connection is made with a single multipin hermetic connector, which is carried to colder temperature stages by a wire loom that is broken out at the 50K and 4K temperature stages to provide amplifier biases and filtering of the flux ramp and detector bias lines.

The RF lines provide the signals that are carried via coaxial cables to the resonator transmission line. After interacting with the resonators, the transmitted power is amplified with one or more cryogenic RF amplifiers before being returned to the warm electronics for processing. In μ mux, the resonators are coupled to the TES detectors, which must be biased for operation. In addition to the TES bias lines and RF amplifier bias lines, the DC lines must also supply a flux ramp line to the SQUIDS.

III. Requirements

The SMuRF system and auxiliary components must satisfy many performance criteria in order to meet the design specifications of the experiments they are coupled to. Here, we outline briefly the system requirements driven by the expected functional requirements in § III A, the science goals in § III B, and the user needs in § III C. These design requirements are ultimately driven by the goal of reading out 2000 μ mux-coupled TES channels in a 4-8 GHz bandwidth, each with ~ 75 dBm input power, while maintaining total readout noise subdominant to the detector noise for a small-aperture telescope with no polarization modulation.

A. Functional Requirements

We first list the signals and connections that the SMuRF system is expected to provide, based on the typical system implementation discussed in § II C.

The RF system of SMuRF must provide one RF signal generation and readback pair per cryostat coaxial line. The multiplexing factors achieved by μ mux are set by the cryogenic devices, which are currently being designed for channel densities of about 500 resonators per 1 GHz of bandwidth across the 4-8 GHz range. For 1000x multiplexing factors we thus require two pairs of coaxial cables, while for 2000x multiplexing factors only a single pair is needed. Based on NIST CMB-style μ mux design specifications, we require that SMuRF can generate up to 2000 probe tones at about \sim 35 dBm each across 4 GHz of total bandwidth, allowing for cold attenuation prior to the resonator input.

For science operations, the TESs are usually operated with a constant low-noise voltage bias. In calibration mode and to drive the detectors out of their superconducting state, a higher bias level must be supplied, though this state is subject to looser requirements on the noise performance. Depending on the design and operating conditions of the TES, the bias line may be required to supply between \sim 10 μ A and \sim 5 mA of current.

The electronics should additionally provide low-noise DC biases for the cryogenic amplifiers. Previous measurements indicate that cascaded cryogenic amplifiers offer improved IP3 for minimal noise temperature penalty compared with typical noise in the cryogenic resonators.¹⁹ In this case, a lower gain, low noise 4K amplifier operates as the first-stage amplifier followed by a 50K amplifier, which may operate at higher powers and thus offers better linearity than a single 4K amplifier alone. For other applications, a single LNA may be sufficient. The LNAs are biased with a gate and optional drain DC line that supply no more than \pm 2 V and O(10) mA. We thus require the electronics supply bias lines for up to two cryogenic amplifiers per transmission line, which we refer to throughout the text as located at 4K and 50K, though they are not required to be fixed at these temperature stages.

Beyond these DC signals, the electronics must include a low-frequency signal generation system that supplies the flux ramp line common to all resonators on the same coaxial line. To modulate the TES signal above the $1/f$ knee of superconducting resonators³, we require that the flux ramp line achieve at least 10 kHz Φ_0 rate, defined as the sawtooth reset rate multiplied by the number of flux quanta swept per ramp. The required amplitude of the flux ramp depends on the mutual inductance to the SQUID, but must be able to sweep several Φ_0 . Assuming there is no current division in the cryostat to suppress noise, this places the minimum requisite current at about \sim 10 μ A for current device designs, which have couplings between the flux ramp coil and SQUID of about O(30) pH.^{16,37}

To mitigate against unwanted pickup from ambient radio frequency interference (RFI), cables carrying critical low frequency bias signals between the warm electronics and cryostat input connectors must be well-filtered and kept as short as feasible.

The electronics should interface with a control computer, allowing the user to command the system and receive data. The system should further provide the ability to interface with external timing systems such as provided from experiment or observatory control. While standalone laboratory test systems may be able to stream data directly to disk, we require that the SMuRF system be capable of interfacing with external data acquisition systems as required by large-scale experiments. We similarly require that a SMuRF system be capable of communicating with and operating alongside other SMuRF systems.

B. Science-driven Requirements

1. Noise

The ultimate figure of merit for the readout noise performance is the noise referred to the detector input. The noise performance has two components: the white noise and the low-frequency behavior. Achieving good noise performance places stringent constraints on all of SMuRF's subsystems, including but not limited to: RF tone generation and channelization, detector bias drives, flux ramp, and all digital signal processing algorithms. Since low-frequency noise on the detector bias line directly contributes low-frequency noise on the detectors, noise performance and stability of the low-frequency subsystems is particularly crucial.

White Noise

The white noise requirement of the readout, which includes both the cryogenic readout devices and warm electronics, is set by the requirement that it be subdominant to detector noise referred to the input. For uncorrelated noise contributions to the total detector noise the various noise components add in quadrature; thus, in order to have minimal impact we require the readout electronics induce no more than 10% penalty to the detector noise. CMB TES detectors are currently photon-noise limited, with noise-equivalent powers of O(40) aW/ $\sqrt{\text{Hz}}$ at 95 GHz.³⁹ We operate in the regime that total readout noise is subdominant to phonon and Johnson noise of the TES, which are in turn subdominant to photon noise. The amplitude of the TES photon noise-equivalent current is set by the operating resistance, which varies widely across experimental configurations from about 1 m Ω to 30 m Ω for DC-biased systems.^{21,40} For CMB detectors with antenna responses centered at 95 GHz, this translates roughly to O(100-160) pA/ $\sqrt{\text{Hz}}$ for a BICEP-like experiment with large operating resistance and Simons Observatory-like experiment with a low operating resistance, respectively.^{21,39}

To maintain low total readout noise, we require that the added white noise due to the warm electronics be subdominant to the intrinsic noise of the superconducting resonators and cryogenic amplifiers, whose combined noise in turn is required to be subdominant to the detected signals. For NIST CMB-style μ mux resonators, the dominant readout noise

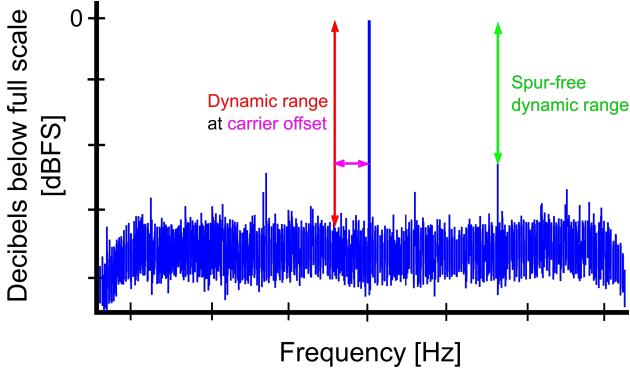


FIG. 5. A sketch of a carrier tone roughly centered within the displayed bandwidth to illustrate various terms relating to dynamic range. The dynamic range is measured at some offset frequency from the main carrier and represents the strength of the carrier signal relative to the noise floor. The noise floor may contain narrow-band spurs; thus, the spur-free dynamic range represents the strength of the carrier relative to the largest spur within the bandwidth.

sources include two-level systems (TLS) in the resonator dielectric, HEMT noise, and DAC noise.⁴¹ To incur no more than 10% penalty in the total current noise, this places an overall requirement on the readout at roughly $O(45\text{-}70)$ pA/ $\sqrt{\text{Hz}}$, which includes both the cryogenic devices and warm electronics.

For the purposes of setting a requirement on the warm electronics, we must relate the detector input-referred current noise to more practical units. Here we define several quantities relating to RF electronics systems that set the engineering specifications for the SMuRF systems.

In μ mux systems, detector signals are modulated into the sidebands of the drive tones at an *offset frequency* given by the product of the flux ramp sawtooth rate and the number of RF SQUID Φ_0 swept in each ramp, typically 4-6 kHz and 4-6 Φ_0 respectively to get above the strong $1/f$ -like component of the resonator noise. The difference between the power of the drive tone and the noise floor generated in the sidebands of the tone by the warm electronics at this offset frequency determines the maximum usable RF *dynamic range*. The dynamic range is often expressed in units of dBc/Hz, or decibels relative to the carrier integrated over the measurement bandwidth, where the carrier is taken as the resonator probe tone and the measurement bandwidth is typically 1 Hz. Alternatively, we can assign the maximum power of the carrier or ADC as “full scale” and express power levels relative to this maximum in terms of decibels relative to full scale (dBFS). While we are typically interested in the dynamic range above the effective noise floor, we may at times consider the strength of the carrier signal above the strongest spurs, or the *spur-free dynamic range* (SFDR). A sketch of these concepts for an arbitrary carrier and noise floor is given in Figure 5.

The maximum achievable dynamic range is limited by both noise and nonlinearity in the active devices used to amplify, up- and down-mix, and read out the tones. In general there are tradeoffs between noise and RF nonlinearity; while lowering drive power may improve the noise due to intermodula-

tion products produced by device nonlinearities, the dynamic range is eventually limited by the thermal noise floor of the electronics. Conversely, improving signal to noise over the thermal noise floor with higher drive powers eventually results in an increased effective noise floor due to intermodulation products of the probe tones. We operate in the regime that the electronics dBc/Hz is not changed by attenuation, as is generally the case for broadband RF electronics in this parameter space.

The referral from electronics noise in dBc/Hz to TES noise equivalent current involves converting noise in electronics phase to resonator frequency noise via resonator and SQUID device parameters and referring to detector input-referred noise via factors of SQUID couplings. The full derivation and measurement is deferred to a separate publication. To remain subdominant to the intrinsic noise of NIST CMB-style μ mux resonators, the requirement for the SMuRF electronics RF dynamic range per tone is -100 dBc/Hz at 30 kHz offset frequency when operating 2000 simultaneous channels. This channel density goal is driven by the maximum expected density of the cryogenic multiplexer.

The input-referred noise in the TES detector bias line is current divided between the shunt resistor and the TES. The ratio of the shunt resistance to the TES resistance under typical operating conditions is about 0.1. This sets the requirement on the noise on the bias line to 10 times the requirement on the current noise of the readout; for example, if the detector input-referred electronics noise requirement is 10 pA/ $\sqrt{\text{Hz}}$, then the noise on the TES bias line must be $\lesssim 100$ pA/ $\sqrt{\text{Hz}}$. The current noise requirement for a given application can vary widely depending on the TES operating resistance; but we require that the bias line current noise be < 50 pA/ $\sqrt{\text{Hz}}$, which is conservative for most DC voltage-biased TES arrays.

The flux ramp has a smaller mutual inductance on the SQUID input coil than the TES, typically by a factor of 10. In order to further reduce noise, it may be current divided inside the cryostat. We assume no current division in the worst case and the same conservative 5 pA/ $\sqrt{\text{Hz}}$ total electronics noise requirement, resulting in a flux ramp current noise requirement of < 50 pA/ $\sqrt{\text{Hz}}$.

Low-Frequency Noise

For mm-wave ground-based bolometric applications in particular, minimizing the $1/f$ noise of the detector time-ordered data is critical.⁴² The conversion between angular scale on the sky and time-ordered data frequency depends on the telescope scan strategy. CMB ground-based surveys targeting large-scale sky features, such as the BICEP/Keck program, require time-ordered data that is white down to ~ 0.1 Hz. Thus, in the absence of additional polarization modulation we target the $1/f$ knee of the electronics to be ~ 0.01 Hz. This low-frequency performance encompasses both the RF and DC subsystems, the latter presenting a particularly challenging constraint since any low frequency noise on the detector bias line directly translates to additional low frequency noise on the detector.

The primary components of the SMuRF system which drive

the low-frequency noise requirement are, in approximate order of impact, (1) the detector bias circuit, (2) the flux ramp circuit, (3) the RF signal circuit, and (4) the RF amplifier circuit. In particular, the detector bias circuit is critical since fractional drifts in detector bias current couple directly into fractional drifts in the mean detector signal and gain in the high loop gain limit of TES responsivity.

The low frequency drifts in the SMuRF-provided bias current are sourced by intrinsic $1/f$ noise of the DACs and opamps used in the bias circuit and by the coupling between bias circuit components (current-setting resistors and DAC voltage references) and the temperature of their environment. We choose DACs and op-amps such that their intrinsic $1/f$ contribution is subdominant to the temperature coupling effects. In this limit the dominant driver of $1/f$ noise is the temperature dependence of the passive circuit components. These components must maintain an $f_{\text{knee}} \sim 0.01$ Hz even when coupling to the extreme case of diurnal ambient temperature variations in the Atacama Desert where many CMB observatories are currently operating or being built, assuming no active temperature control. This sets the requirement for the combined temperature coefficient of all components in the TES bias circuit to be $\lesssim 20$ ppm/ $^{\circ}\text{C}$.

The flux ramp is typically operated at 10s of kHz, allowing for some relaxation of $1/f$ requirements on individual components. Since the flux ramp demodulation scheme relies on knowledge of the average flux ramp rate over long periods, the stability of this circuit is crucial. We find that the required component temperature stability and noise performance are comparable or subdominant to that of the detector bias circuit; thus, choosing similarly rated components for both circuits suffices to meet this requirement.

The RF signal circuit low-frequency noise performance, defined here as the dynamic range at small carrier offset, is set by the stability of the local oscillator (LO), discussed further in § V B. Due to the flux ramp scheme of μmux , the requirements on low-frequency noise in the RF circuit and low-frequency variation of the LO position itself are not as stringent, and are subdominant to other sources of $1/f$ noise.

The SMuRF provides circuitry for cryogenic RF amplifier biasing, as outlined in § V D. These amplifiers are typically sourced commercially and are robust to gain fluctuations due to noise in the bias circuit. Flux ramp modulation is expected to further suppress the majority of $1/f$ noise. The phase modulation scheme of μmux additionally offers some amount of immunity to overall amplitude fluctuations resulting from amplifier gain noise.

2. RF Linearity

Here, we set requirements on the system to minimize 3rd-order intermodulation products that degrade the RF dynamic range. The full octave of readout bandwidth used for this application places stringent requirements on the bandwidth of the RF components in the amplification and signal processing chain.

For *fixed-tone readout* where the RF probe tone is fixed

while the resonance is modulating, thousands of unattenuated tones are incident on the cryogenic amplifier as the resonances move in frequency. This is a particular challenge for μmux systems both due to the higher probe tone powers and since the resonance swings by $O(1)$ resonator bandwidth in operation. To accommodate this large change in power, the cryogenic amplifier must trade off noise temperature, thus degrading the overall readout noise. The interaction of the RF probe tones further generates third-order intermodulation products in-band; for 2000 RF tones in one octave of bandwidth there are over one billion third order intermodulation products whose total power grows as the cube of the fundamental tone power. These intermodulation products combine and form an irreducible pseudo-noise floor that degrades the dynamic range of the probe tones and may exceed the -100 dBc/Hz requirement discussed above, adding to the detector-referred readout noise. In addition to sourcing noise, intermodulation products induce crosstalk between channels that poses potential systematics for science analysis.⁴³

SMuRF addresses this linearity problem with a *tone-tracking algorithm*, which follows the modulating resonance with a closed-loop adaptive filter-based feedback. Previous results demonstrate that this tone tracking is sufficient to add negligible noise penalty when running $O(400)$ channels simultaneously versus a single channel at a time for NIST CMB-style μmux resonators.^{16,19} The impact of adding additional channels is dependent on resonator properties, most crucially the resonator dip depths. For the full system we design with the goal of sufficient linearity such that there is minimal penalty incurred by additional channels for up to 2000 NIST CMB-style μmux channels on an RF line. This sets a stringent requirement in particular on the RF components including the ADCs, DACs, and RF mixers, as discussed in § V B.

We emphasize that there is generically a tradeoff in RF systems between noise temperature and linearity; the noise may be improved at the expense of being able to operate fewer channels simultaneously. For the SMuRF system the large channel counts necessitated aggressive optimization for linearity at scale, but the noise performance of the SMuRF system for a small number of channels can be improved.

3. Bandwidth

The required sampling bandwidth for the readout electronics is driven by the bandwidth of the detectors and the bandwidth of the resonator. For CMB measurements, the science readout bandwidth is typically in the 0.1-300 Hz range.^{10,44} This gives us a minimum bandwidth for reading out TES bolometers of $O(1$ kHz) per detector channel across a minimum of 2000 channels. The detector noise itself extends higher in frequency; to keep the TESs stable they are designed to have bandwidth in the \sim kHz range.²

For applications that require the reconstruction of fast pulses such as calorimetry, the energy resolution is set by the rise time and height of the pulse.⁴⁵ In this case, the SMuRF must be reconfigured for lower channel count, higher band-

width channelization and signal processing. Further discussion of SMuRF optimization for pulse detection and processing is deferred to a future publication.

The requisite TES readout bandwidths stated above are for the flux ramp-demodulated detector data, whose sample rate is set by the flux ramp rate. While the tone-tracking ability of SMuRF allows for readout of resonator modulation faster than the resonator bandwidth, it is still useful to use the resonator bandwidth as a heuristic for required per-channel readout bandwidth. The resonator bandwidth for NIST CMB-style μ mux resonators is designed to be ~ 100 kHz, well above the TES bandwidth. To avoid aliasing penalties and get above the $1/f$ noise from two-level systems (TLS) in the resonator, we flux ramp-modulate the signal to $O(10)$ kHz. For the SMuRF to be capable of resolving $O(10)$ kHz flux ramp rates, the per-channel digitization bandwidth requirement is set at $O(1)$ MHz to ensure enough samples per flux ramp cycle to allow for demodulation.

Additional bandwidth and aliasing considerations of the μ mux devices are discussed in a separate publication.⁴⁶ For the purposes of checking performance for CMB-style resonators, we require that the system is capable of resolving and demodulating waveforms with frequencies up to ~ 1 kHz.

4. Crosstalk

Crosstalk between channels contributes to unwanted signal that biases scientific analyses. For CMB polarization analysis, the crosstalk between channels is held below 0.3%.⁴² The sources of signal crosstalk in the μ mux cryogenic subsystem have been investigated and largely mitigated through cryogenic device design.⁴³ Most pertinently for SMuRF, the noise floor induced in part by the higher order intermodulation products of tones contributes some amount to an “all-into-all” form of crosstalk. We expect that the RF linearity requirement outlined in § III B 2 is sufficient to address this form of crosstalk. In particular, the -100 dBc/Hz for 2000 tones requirement translates to a maximum -43 dBc leakage between arbitrary channels in 1 GHz bandwidth. To be conservative, this sets a minimum requirement of -60 dB isolation between channels in any given 1 GHz bandwidth.

Beyond the linearity requirement, there may remain some amount of crosstalk between channels in the electronics. The NIST CMB-style μ mux resonator is expected to dominate any crosstalk contribution from the SMuRF electronics.¹⁶ We thus require simply that the electronics-induced crosstalk be subdominant to the crosstalk induced by the cryogenic resonators.

C. Ancillary Design Requirements

We note here several additional design constraints guided by the unique operating environments encountered by SMuRF and similar microwave resonator-based systems.

- **Environment:** The SMuRF electronics are expected to operate in a wide range of ambient temperature conditions without noise penalty. In the worst case, the

electronics are air-cooled in ambient conditions. Additionally, many CMB experiments operate at high altitudes where oxygen levels are substantially lower than sea level. Extra care must be taken to ensure that the electronics do not overheat or otherwise fail in these conditions. The SMuRF was designed to operate at altitudes up to 5200 meters above sea level with ambient temperatures below 25°C.

- **Power:** Each SMuRF system’s maximum power consumption must fit within the power budget of the intended application. For high altitude astronomical observatories which often operate on a spare total power budget, this poses a significant constraint due to the high signal and switching speeds, the computationally intensive nature of the SMuRF processing algorithm, and the power demands of the electronics crate and peripherals. For reference, the total electrical power available for readout for current sites is about $O(1\text{-}3)$ kW for 20,000 channels.

- **Vibration Sensitivity:** The detector readout electronics on telescopes is often mounted as comoving with the receiver, which for CMB applications scans the sky at several degrees per second and turnaround accelerations of $O(1)^\circ/\text{s}^2$. The SMuRF should be able to withstand this movement without damage or noise penalty. In addition, multiple SMuRF systems are often run together in adjacent slots of one electronics crate. The presence of additional SMuRF systems in adjacent slots should not impact the performance of any given system.

- **Modularity:** Components of the system should be designed to allow for separate testing and integration, allowing for swapping and upgrading parts as necessary on deployed systems.

- **Ease of Use:** The user-facing interface and software should be accessible to end users who may desire to use SMuRF as a readout platform without expert knowledge of the electronics, firmware, or cryogenic multiplexer. This drives the SMuRF design towards modular components that are easily swappable, open source software, and compatibility with standard scientific laboratory computers and operating systems.

IV. SMuRF Electronics Overview

The SMuRF electronics are built on the SLAC Common Platform, a comprehensive FPGA-based firmware framework built on standardized hardware designed specifically for high-performance systems.⁴⁷ The hardware components are designed for modular use with commercial Advanced Telecommunications Computing Architecture (ATCA) crates.⁴⁸ The hardware architecture is based on a Xilinx FPGA carrier board, which supports up to two double-wide dual-height mezzanine cards for analog and RF processing and a rear transition module for high speed links and further customization.

A block diagram of the full SMuRF system is given in Fig. 6.

The system consists of an FPGA carrier card, which provides all real-time signal processing. Mounted to the carrier are one or two Advanced Mezzanine Cards (AMCs) which each consist of a base board and an RF daughter card. The base board contains the ADCs and DACs for the main signal path, clock generation, and local oscillator (LO) generation. RF daughter cards mounted to the AMC base boards contain the RF mixers, amplifiers, and bandpass filters. Opposite the AMCs on the carrier card is a Rear Transition Module (RTM) handling the low frequency signal generation, which is connected via a flexible multipin cable to a cryostat card that interfaces the low frequency signals with the cryostat. Additional peripherals such as the crate, timing, and networking are required to run the system, but may be shared between multiple systems.

Multiple parallel systems reside in a single commercial ATCA multislot crate, with each system occupying one slot. Timing information is distributed through the crate backplane. Typical crates contain $N + 1$ slots, with N carrier cards and one slot dedicated to a network switch (Vadatech ATC807) that allows for data streaming across multiple SMuRF systems. A 1 Gbit ethernet connection between the ATCA network switch and the control server, commonly a Dell R440, provides firmware register access for every SMuRF carrier in the 7-slot ATCA crate, allowing for software control and data acquisition. For large integrations, an additional PCIe card (Xilinx KCU1500) allows for streaming data from multiple carriers to the server. This specific hardware, including the PCIe card, Vadatech ATCA network switch, and Dell R440 computer server, have been qualified for operation in high-altitude environments as described in § VIII F.

Up to two AMCs fit into designated mezzanine slots on the FPGA, each with four up/downconverters corresponding to one of eight 500 MHz wide bands. The firmware framework is built on these 500 MHz-wide bands that can each accommodate processing and streaming of up to 416 simultaneous 2.4 MHz-wide channels. For each band, the tones are synthesized in the DACs and upconverted to the appropriate GHz frequency range with a local oscillator. Tones returning from the cryostat are split into bands and downmixed with the same LO before being sent to the corresponding ADC. The downmixed tones are digitized, channelized, and read into the FPGA, where they are processed. Each channel is typically associated with a unique resonance, thus creating a 1:1 mapping between readout channels and detectors. Thus, in total each SMuRF system contains the requisite hardware to read out 3328 resonator channels across 4 GHz of bandwidth, either in two parallel systems reading out resonators in 2 GHz of bandwidth or in two serial systems reading out resonators in across a 4 GHz range.

Each channelized tone is processed by the firmware to return data with desired levels of processing depending on the application. At the lowest level, the downconverted and channelized data can be streamed as the orthogonal I and Q components of the digital down conversion. Following calibration, these orthogonal components may be returned as the ampli-

tude and phase components of the resonator voltage fluctuations, or an approximation of the probe tone frequency and resonator frequency error Δf . With this frequency error approximation, a feedback loop may be applied to minimize Δf , thus following the resonator minimum. This “tone-tracking” feedback enables the applied probe tone to “follow” the resonance dip as it is modulated by the flux ramp and/or incoming detector power.

A more detailed discussion of the options for data processing is given in § VI. At the highest level of processing, the SMuRF returns a time-ordered data stream of flux ramp demodulated detector data for saving to disk or passing on to downstream data acquisition.

V. Hardware

A full picture of the hardware components is shown in Figure 7. The components displayed constitute the pieces of one SMuRF system, which is capable of reading out up to 3328 channels in one rack-unit of space, excluding peripherals that are shared between multiple systems. They are mounted in one slot of a multi-slot ATCA crate, and present SMA ports for RF connections from the front of the crate and a flexible multipin cable to a card mounted directly at the cryostat vacuum feedthrough at the back of the crate.

A. Carrier Card

The FPGA carrier card (Xilinx XCKU15P Ultrascale+) handles the real time signal processing, most notably the RF tone generation/channelization and tone tracking. The card provides interfaces for data, control, and timing. An Intelligent Platform Management Interface (IPMI) provides a serial control protocol for monitoring and power cycling. A single-lane 2.5 Gbit uplink provides on board timing and a frequency reference, while a 4-lane 10 Gbit/s ethernet interface to the ATCA crate switch via the backplane handles all other communication, including data transfer. The AMC (see § V B) interface with the carrier consists of eight 12.5 Gbit bi-directional JESD204b links to each AMC card for ADC/DAC data, digital clocks, and low-speed serial peripheral interfaces (SPI) for control and status readback. The carrier interfaces digitally with the RTM via SPI link and provides 51.2 MHz clock (see § V C). The RTM additionally routes two optional external 2.5 Gbit external inputs to the carrier, one for ethernet and the other for timing.

The firmware on the carrier card is programmed via VHDL and Xilinx System Generator for Digital Signal Processing and may be updated remotely. Further firmware details are addressed in § VI.

Due to the large resource consumption of the signal processing firmware and the need for many carriers to operate in close proximity in high altitude environments, the carrier is robustly temperature controlled with large heatsinks, delidded regulators, and high-velocity crate fans. The results of testing of these temperature control measures is described in § VIII F.

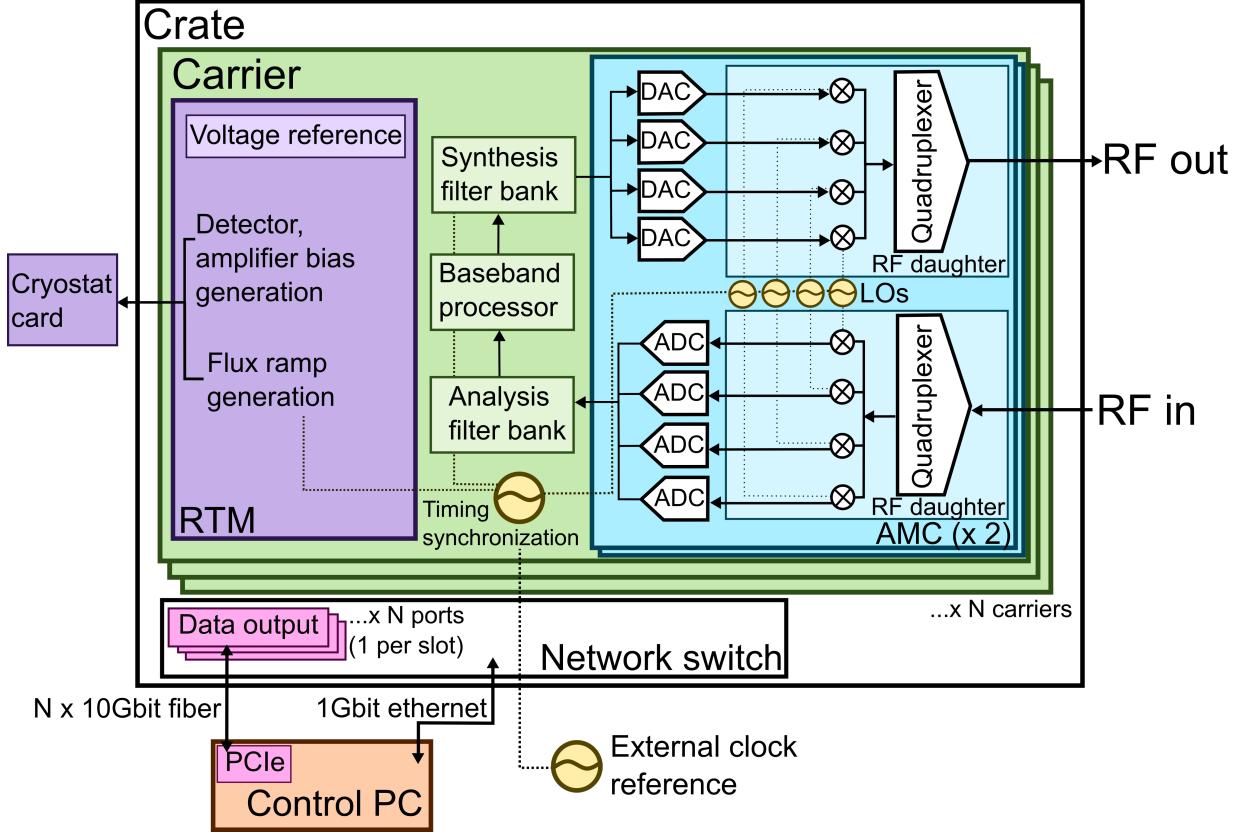


FIG. 6. A sketch of a full SMuRF system. Each system consists of an FPGA carrier board (green) that supports both the RF (blue) and DC (purple) systems. Multiple systems may be run in parallel in a single crate. The RF system (blue) consists of two Advanced Mezzanine Cards (AMCs), each with a pair of RF daughter cards that interface between the RF cabling to the cryostat and the ADC/DACs. The LOs used for up- and downmixing signals between the baseband and RF are derived from the main carrier but are generated on the AMCs. The DC system (purple) provides the amplifier biases, TES biases, and flux ramp generation on a Rear Transition Module (RTM). The voltage sources are further conditioned in a cryostat card that interfaces directly with the cryostat. The main signal processing functions take place on the carrier, and consist of a pair of polyphase filter banks for RF tone synthesis and readback and a baseband processor for per-channel computations. Several options exist for interfacing timing and data acquisition with the system; displayed here are streaming to a PCIe card installed in the control server and receiving an external clock reference. An ethernet connection between the the control server and network switch provides firmware register access to all slots simultaneously and may receive data streamed to disk for a small number of carriers. Not pictured is the ATCA crate backplane, which distributes power and timing information and connects the network switch ports to the carriers in each slot.

B. Advanced Mezzanine Cards and RF Daughter Cards

The Advanced Mezzanine Card (AMC) contains the ADCs and DACs for the main RF signal path, clock generation system, and LO generation for the RF daughter cards. Each AMC consists of a base board interfaced with an RF daughter card supporting a total of 2 GHz of bandwidth at either 4-6 GHz or 6-8 GHz. The RF daughter card provides the up- and down-mixing between the digital signal processing (DSP) band centered at 750 MHz and the RF frequencies of the microwave resonators. The current iteration of SMuRF has RF daughter cards for operation of either “low band” (4-6 GHz) or “high band” (6-8 GHz) resonators. The daughter cards may be arbitrarily swapped between AMC base boards, which require only minimal modification between low and high band types. Each carrier card accommodates two AMCs of any combination of low band and high band types. The LO frequency may be modified via firmware register. Combined with RF

hardware part swaps, this allows for for operation of SMuRF systems with cryogenic resonators at other frequency ranges and channel densities.

1. AMC Base Board

The AMC base board contains the hardware for interfacing the digital logic on the carrier with the analog electronics. A block diagram is shown in Fig. 8.

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

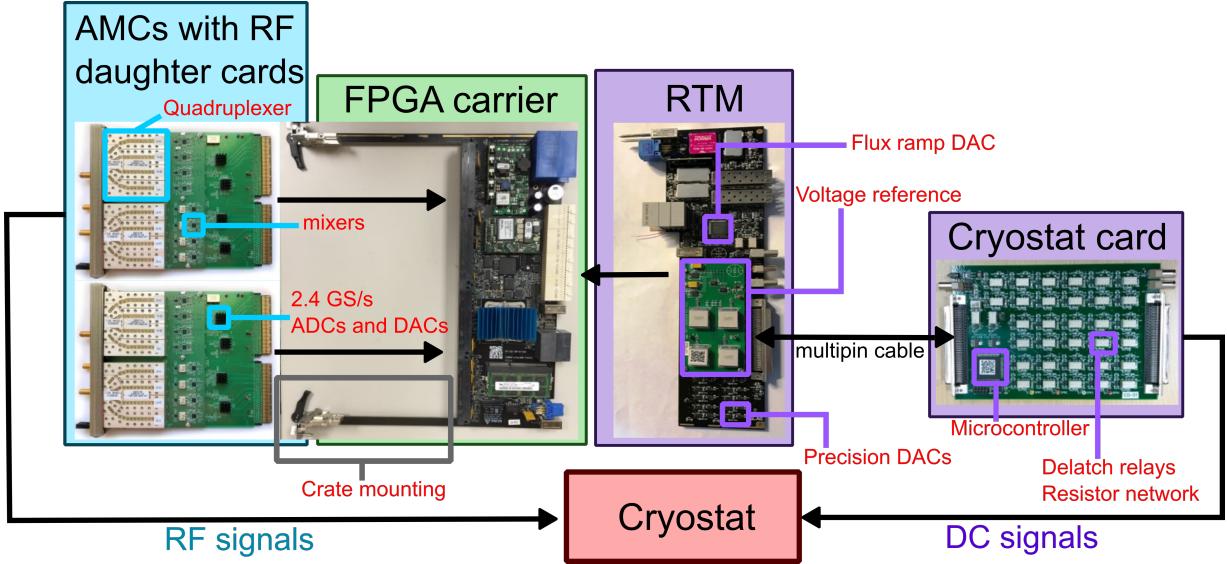


FIG. 7. A picture of the hardware for one SMuRF system accommodating up to 3328 resonator channels across one or two RF chains, excluding peripherals such as the server, data streaming, and timing. The pictured contents excepting the cryostat card occupy one slot in a multi-slot ATCA crate. The AMCs with RF daughter cards mount to the front of the carrier and present SMA connectors for RF connection to the cryostat, while the RTM mounts to the rear of the carrier and presents a multipin cable to a cryostat card, which mounts directly to the cryostat. The RTM and cryostat card combination supplies the low-frequency signal generation and conditioning.

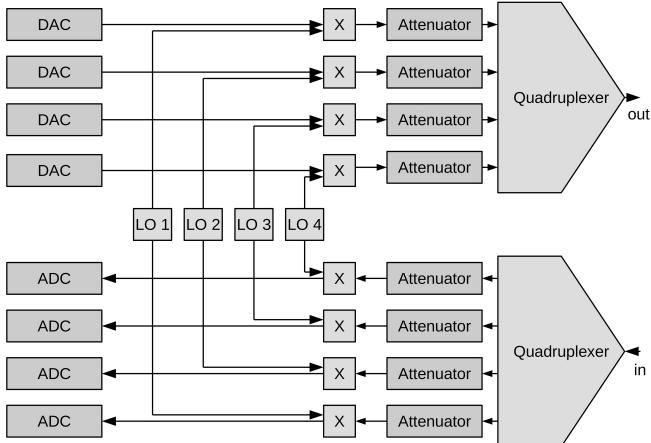


FIG. 8. Block diagram of a SMuRF Advanced Mezzanine Card (AMC), which is responsible for the RF signal input/output. RF signals enter the system between 4-8 GHz and are first split into 500MHz bands via a quadruplexer (bottom right). Each band is then processed in parallel through digitally controllable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

ultiplication with a directly synthesized intermediate frequency (IF), filtered, and downsampled to obtain the complex baseband signal. This relaxes the requirement on precisely match-

ing the gain and maintaining the phase offset between the I and Q components. However, DDC requires higher speed DACs and ADCs to do the I/Q combination and separation in firmware.

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

Each DAC/ADC pair is associated with a unique local signal generator to provide the LO that the RF daughter card uses to upmix or downmix between the 500 MHz-1 GHz band and the desired GHz-frequency band. Each 500 MHz band has its own LO, which is offset by the ADC/DAC numerically-controlled oscillator (NCO) frequency of 2.4 MHz from the centers of 3.5 GHz, 4.0 GHz, 6.0 GHz, and 6.5 GHz for the low band, and 5.5 GHz, 6.0 GHz, 8.0 GHz, and 8.5 GHz for the high band. We use the lower sidebands for the first two LOs and the upper sidebands for the second two such that the LOs fall outside the resonator bandwidth when operating AMCs of the same type (both low or both high band). These LOs are generated via a fractional-N locked phase-locked loop IC (Analog Devices ADF5355) with integrated low-noise voltage-controlled oscillators (VCOs).

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

Since the LO is used to generate all the tones in its respective 500 MHz band, its noise profile is critical to the overall noise performance. Other studies find that the noise profile of the LO can be seen in resonator channel noise spectra at high flux ramp rates, which does not impact typical science data but may constrain the maximum operable flux ramp rate.⁴⁶ The noise profile of the LO is set by the noise of the reference clock source, the charge pump used for loop gain adjustment, and the VCO. The phase noise of the clock reference source the LOs are locked to sets the close-in phase noise (below ~ 100 Hz). An external clock reference can be passed in and locked to the onboard clock either via a SMA input on the AMC front panel or through a timing distribution system on the ATCA crate backplane. If no external source is provided, the 122.88 MHz is derived from the FPGA.

Above 100 Hz the phase noise of the LO is set by the parameters of the charge pump phase-locked loop (CP-PLL) which is used to create a flat phase noise profile out to O(100 kHz). The noise level in this flat portion of the phase noise profile can be adjusted slightly by trading off bandwidth of the CP-PLL. Pushing to much higher bandwidth above ~ 100 s of kHz leads to instability in the CP-PLL. The phase noise after the CP-PLL rolloff is set by the intrinsic phase noise of the VCOs that generate the LO, which ultimately limit the phase noise level. For applications that require lower system phase noise, the VCOs can be exchanged with lower phase noise, fixed-frequency crystal oscillators. These were not used in the current design because flexibility of LO frequency while achieving the required white noise specification was favored over further improvements in phase noise.

2. AMC RF Daughter Cards

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

The RF daughter cards exist in both low band (4-6 GHz) and high band (6-8 GHz) variants. Quadruplexers (Lark Engineering SMCQ819-110 and SMCQ819-111) interface from the mixers on the four ADC/DAC pairs to a single input/output coaxial pair that connect with the cryostat or other test system via SMA connectors. Cavity filters within the quadruplexers define overlapping bandpass filters that split the RF frequency band into each 500 MHz band that is synthesized or channelized in firmware. Since the ADCs and DACs are clocked at 2.4 GS/s, the Nyquist frequency for the alias band

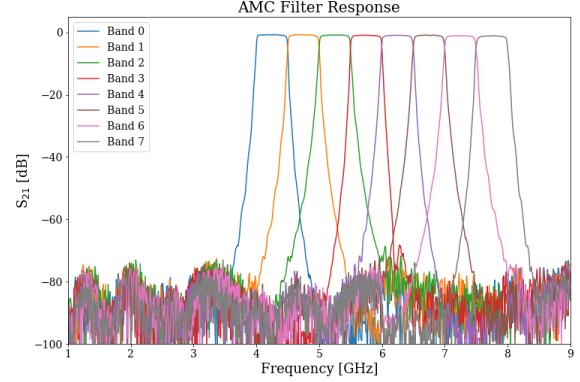


FIG. 9. The response of the AMC band-defining filters for each of the 500 MHz wide bands across the full 4-8 GHz SMuRF band. Band 0 corresponds to the 4-4.5 GHz band, and the bands count upwards in 500 MHz increments from there.

is at 1.2 GHz.

The AMC S_{21} response for each 500 MHz band is shown in Figure 9. We note in particular that the maximal passband of the filters is not quite 500 MHz wide to guarantee sufficient rejection out of band, set at 60 dB isolation between bands as described in § III B 4. The portion of the passband that is flat within -2 dB is about 460 MHz wide, while the edges of each band are about 3 dB down and isolation is > 65 dB between adjacent bands. The RF daughter cards additionally include programmable attenuators on both the RF input and output for each 500 MHz band to allow for flattening the RF signal levels over the full SMuRF RF bandwidth.

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

C. Rear Transition Module

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

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

The DC TES biases and cryogenic amplifier biases are provided by 33 low-noise 18-bit DACs (Analog Devices AD5780). To reduce low-frequency noise, the TES bias DACs are typically operated differentially, though they can be driven single-ended with no modifications to the SMuRF hardware or firmware to double the number of independent TES bias lines.

As discussed in §II C, for applications requiring high RF linearity it may be desirable to cascade two cryogenic amplifiers rather than use a single LNA. Thus, the RTM allocates eight DACs to biasing up to four total RF amplifiers, typically two amplifiers placed at 4K and 50K for two separate RF chains. Each amplifier is allocated two DACs for the gate and drain voltages, though some models require only a single bias line. Additional amplifier bias lines can be added at the expense of TES detector bias lines with no modification to the RTM hardware or firmware, if desired. These DACs, referred to as “slow”, can generate arbitrary waveforms up to about 1 kHz for calibration and testing purposes.

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

A daughter card on the RTM consists of a voltage reference providing buffered $\pm 10V$ reference signals to the precision DACs. The card uses two commercial voltage references (Analog Devices LTC6655) in parallel to reduce noise, and low-pass filters with low acoustic sensitivity capacitors (Rubycon 35MU106MD35750) to mitigate vibrational pickup. A daughter card implementation for the voltage reference provides the flexibility to easily and cheaply upgrade the performance of the voltage reference if required for a particular application.

The RTM presents a 100-pin SCSI connector containing the low-impedance flux ramp, TES bias, amplifier bias, DC power supplies, and digital communication to the cryostat card. The SPI communication link between the RTM and the cryostat card is default quiet to reduce noise during operation, with no active polling.

Two monitor ports with LEMO connectors are available for laboratory diagnostics. First, the flux ramp monitor port outputs a single-ended 50 Ω -impedance signal for monitoring the flux ramp output. This line is not guaranteed to have the requisite noise stability or bandwidth for properly driving the flux ramp. Second, the trigger output provides a low voltage TTL signal asserted at the start of each flux ramp period, typically at the resets for the sawtooth signal.

An SFP 1 Gbit ethernet port may be used for data streaming of a single SMuRF system without the need for interfacing with the rest of the ATCA crate. An additional SFP port receives timing data that may be used for synchronizing multiple SMuRF systems together, as discussed in § VI D.

D. Cryostat Card

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

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

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

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

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

The 4K and 50K amplifier gate bias voltages are each driven single ended via an RTM DAC with 20 μF bypass capacitors on the cryostat card for filtering. A resistor divider on the cryostat card allows for limiting the output gate voltages

to less than the maximum amplifier ratings if required. The drain currents may be read back to allow for bias adjustment using an inline resistor on the input to the regulator, which is amplified by an instrumentation amp and read back by ADCs on the microcontroller. The flux ramp and amplifier bias cryostat card circuits have jumpers that allow the user to manually select between sourcing from the RTM and inserting signals from an external source via a 2-pin differential LEMO connector.

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

The cryostat card provides an additional DC line at +12V with minimal filtering for powering an optional warm follow-on RF amplifier, which can be used to optimize the cryostat RF output signal power levels into the AMC RF ADCs. For general diagnostic purposes, it can report its temperature using an on-board TI TMP236 temperature sensor over the SPI bus.

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

E. PCI Express Card

For large integrations, a Xilinx KCU1500 data acquisition PCI express (PCIe) card using a custom firmware allows for multiple carriers' worth of data to be streamed to a server, with each slot outputting to a 10 Gbit fiber link on the network switch.⁴⁹ The PCIe firmware implementation is designed for a 7-slot ATCA crate configuration, with the PCIe card interfacing with up to six SMuRF carriers through the Vadatech switch via dedicated ethernet links to an eight lane PCIe 3.0 bus in a server computer (commonly a Dell R440). The dedicated ethernet link for each SMuRF carrier is implemented with 10 Gbit fibers which connect from the ATCA ethernet switch to two Quad Small Form-factor Pluggable (QSFP) transceivers in the KCU1500. The PCIe card presents two x8 interfaces bifurcated to a x16 edge connector to the server.

During data taking, each SMuRF carrier transfers data frames at the flux ramp rate to the PCIe card, where the frame stream is received and processed further in software. Common software processing steps include packaging the data with an application-specific header, which can include timing information, and optionally filtering and downsampling the data streams before writing the data to disk.

VI. Firmware

The FPGA carrier card firmware is responsible for SMuRF computation including RF tone generation, tone tracking, flux ramp demodulation, and data framing. A component block diagram is given in Figure 10. The SMuRF firmware components provide the following functionality:

- **Tone Generation and Channelization:** A pair of polyphase filter banks provide synthesis and analysis of the RF tones used to probe the resonators. The filter banks are shown as the blue boxes in Figure 10.
- **Frequency error estimation:** SMuRF uses an affine transform to translate the received complex response (S_{21}) into a resonator frequency error estimate. As this occurs on a per-channel basis, it is contained in the baseband processor (green box in Figure 10).
- **Tone tracking and flux ramp demodulation:** SMuRF uses a feedback loop to update the probe tone frequency, attempting to minimize its estimate of the frequency error. The feedback loop parameterizes the flux ramp modulation and also outputs the demodulated detector signal. This is also part of the baseband processor.
- **Miscellaneous:** The firmware handles miscellaneous timing and waveform synthesis tasks, such as flux ramp timing, arbitrary detector bias DAC waveform generation, data streaming, and synchronization between multiple SMuRF systems. These are shown heuristically with the timing (yellow), DC signal generation (purple), and DAQ (pink) boxes in Figure 10.

Intermediate data outputs with increasing levels of signal processing are available to examine for system characterization and debugging.

A. Tone Generation and Channelization

The SMuRF FPGA synthesizes and analyzes data from each of 500 MHz bands corresponding each to one RF ADC/DAC pair separately. The RF ADCs and DACs transform the IF band from 500 MHz-1 GHz to and from the complex baseband of ± 307.2 MHz, defined by the ADC sample rate of 2.4 GS/s and divide-by-4 complex decimation filter. From here, the data is downconverted from a 614.4 MHz-wide frequency band to individual channels to be processed. A conceptual sketch for the single channel downconversion chain is

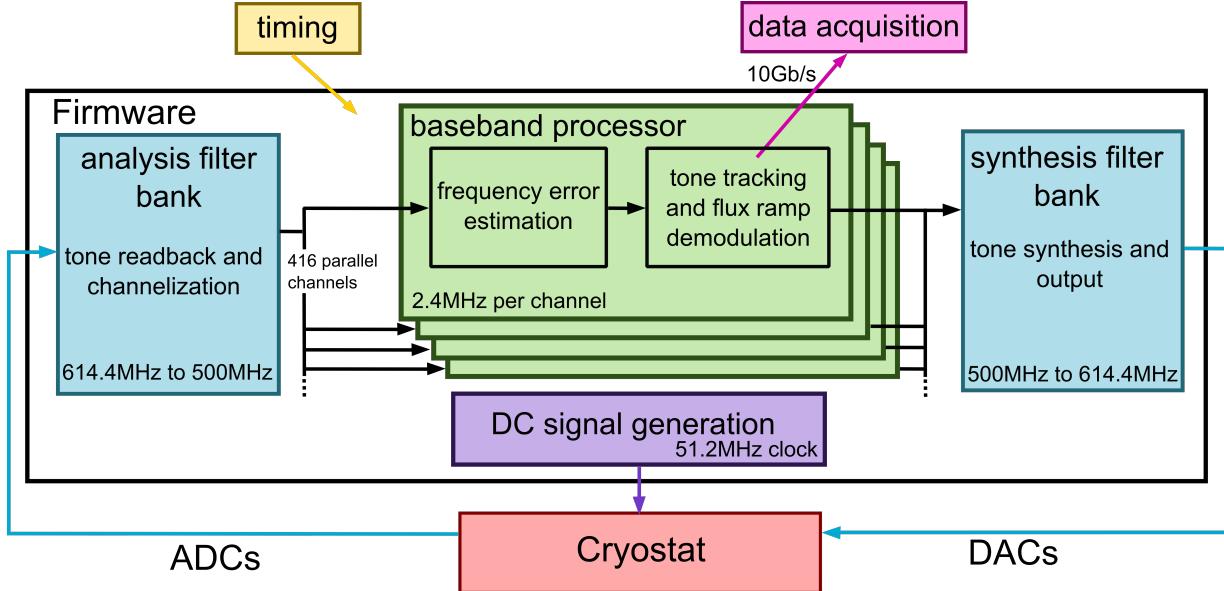


FIG. 10. Block diagram of the main components of the SMuRF firmware. Some miscellaneous tasks are not shown for simplicity. The blue boxes represent a pair of polyphase filterbanks that provide the synthesis and readback of the RF tones used to probe the resonators. The green box contains the baseband processor, which acts on each channel separately to perform the calibration and frequency error estimation, then the tone-tracking and flux ramp demodulation. The filter banks and baseband processor act on hardware-defined 500 MHz-wide bands. One SMuRF FPGA carrier card supports up to 8 bands, allowing for a total of 3328 channels per board. Auxiliary firmware functions such as timing synchronization, data framing, and low-frequency signal generation are represented by the yellow, pink, and purple boxes respectively.

given in Fig. 11. The upconversion works similarly, but in reverse.

A polyphase filter bank performs the channelization, converting from each 614.4 MHz I/Q stream to 512 interleaved 2.4 MHz-wide bins that are time multiplexed. The filter bank response is shown in Fig. 12. The filter bank oversamples by 2 such that each resonance falls into multiple bins; two resonances which fall in the same bin may be assigned to an overlapping bin such that both resonators are processed. Thus, a given 1.2 MHz frequency band may not contain more than 3 resonances such that every resonance can be assigned to its own bin. The firmware is capable of supporting processing multiple channels per bin if desired, but currently uses only one channel per bin. Placing a single channel per bin introduces additional latency due to the lack of parallelization, but reduces FPGA resource consumption. NIST CMB-style μ mux resonators aim for at least 1 MHz spacing between μ mux resonances for crosstalk considerations; thus the width of the filter bank channels does not significantly contribute to readout yield degradation.⁴³

The data from each bin is filtered with a 16 tap digital low-pass filter, resulting in a 4096 tap FIR filter (since there are 512 resulting channels, oversampled by 2). The filter bank achieves about 100 dB of rejection for out of band response and adds a group delay that is approximately constant across the 2.4 MHz-wide band.

This filter sets the majority of the processing delay due to SMuRF, on the order of 6 μ s out of a total signal processing delay of about 8.8 μ s. For comparison, the roundtrip cable delay through a typical CMB receiver cryostat may be about

100 ns, and the additional delay from the ADCs and DACs another $\sim 0.5 \mu$ s.

The FFT is performed with a streaming Radix 2² architecture to convert between the frequency and time domains.

The tone frequencies are defined with a 24-bit input to a *direct digital synthesis* (DDS) system. This results in a frequency resolution of 0.1431 Hz for probe tones, which is essentially continuous for practical purposes. A 10-bit look-up table and Taylor series correction generates the probe tone per channel, which is amplified to the desired output tone power and converted back to the full frequency-multiplexed bandwidth via the synthesis filter bank. The DDS system achieves greater than 100 dB spur-free dynamic range (SFDR) over the synthesis filter bank per-channel bandwidth of 2.4 MHz. The DDS continuous synthesis allows for fine frequency resolution and fast, phase-continuous frequency switching necessary for continuously updating the tone frequency with tone tracking, as discussed in § VI C.

The 512 time-multiplexed channels output from the polyphase filter bank are processed in two parallel streams by the baseband processor, which performs the per-channel calibration and feedback operations discussed in § VI B and VI C. While the filter bank channelizes the full 614.4 MHz bandwidth, we only use the center 500 MHz per AMC as defined by hardware filters. To reduce power consumption on the FPGA, the edge channels are dropped from processing, resulting in 499.2 MHz of bandwidth covering 208 time multiplexed channels per baseband processor stream for a total of 416 channels available per 500 MHz band. While dropping the edge channels adds the complication of converting

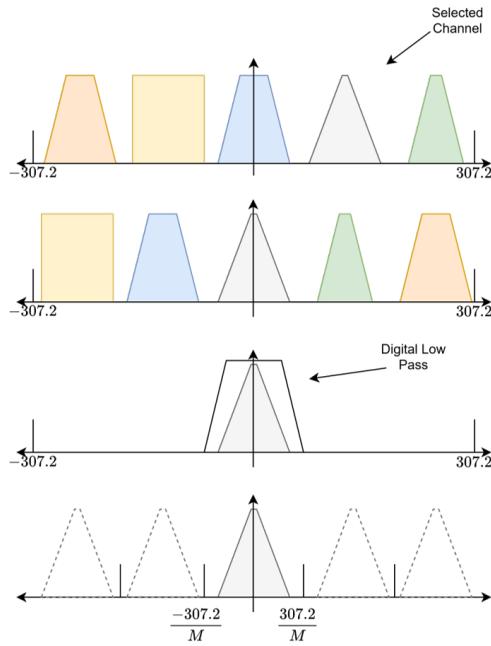


FIG. 11. Conceptual diagram of the downconversion process for a single SMuRF channel. Heuristically, the process works as follows: (1) The incoming data $x[n]$ consists of N channels separated in frequency space across the 614.4 MHz-wide output of the ADC, centered each at frequency θ_k for $k = 1, 2, \dots, N$.

- (2) Each channel is centered by multiplying with the inverse of the channel center frequency $e^{-j\theta_k n}$.
- (3) A digital low-pass prototype filter is applied.
- (4) We downsample by $M = 256$ to arrive at a 2.4 MHz-wide single-channel output.

This process is implemented in parallel for $N = 512$ channels via the analysis filter bank. The similar upconversion process occurs via the synthesis filter bank in the reverse order.

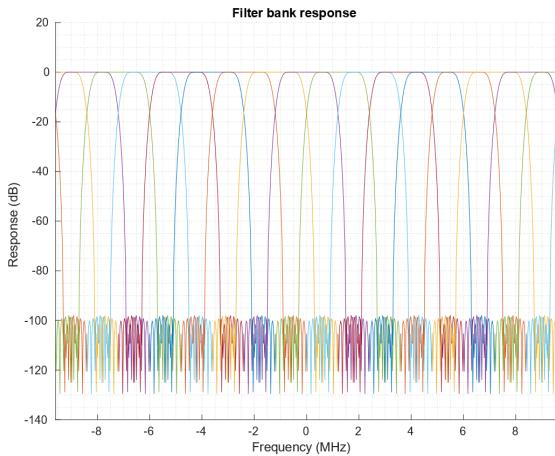


FIG. 12. A zoomed-in section of the filter bank response for 16 interleaved 2.4 MHz wide channelization bins. Note that the channels are oversampled by 2 such that the bins are overlapping with each other. The filter bank converts between 614.4 MHz of bandwidth and 512 time multiplexed bins that are each 2.4 MHz wide.

the 614.4 MHz stream from the ADC to a 499.2 MHz stream for processing, the slower clock speed is critical to fitting the baseband processor computations within the resource limits of the FPGA.

For each channel, the ADC output is digitally downconverted with the output from the DDS to center it about 0 Hz. At this point, the data may be optionally output as per-channel orthogonal I and Q streams for offline test and debugging purposes. Application of calibration (as discussed in § VI B) transforms this data into tone frequency and frequency error between the output tone and the resonator frequency. Finally, an adaptive loop filter configured with feedback parameters (as discussed in § VI C) outputs flux ramp demodulated data at the flux ramp frame rate, and feeds back to the DDS to update the output tone frequency during tracking.

After interacting with the baseband processor, the time-multiplexed processed channels are padded from 499.2 MHz back to the 614.4 MHz bandwidth and synthesized into a 614.4 MHz I/Q stream via the synthesis filter bank. A user-defined delay may be added to correct for any system delays due to cabling, processing, etc. before being passed to the DAC. The relevant frequency domains of the channelization and synthesis process is given in the main RF loop of Figure 10.

B. Frequency Error Estimation

Following the tone synthesis and analysis, the SMuRF electronics can be thought of as a system that is capable of generating tones and reading back the complex response as a channelized stream of orthogonal in-phase and quadrature components, typically denoted I and Q respectively. This I/Q stream may then be processed by the baseband processor, which acts on each channel serially. A conceptual diagram of the baseband processor is given in Fig. 13.

For μ mux systems, we are typically interested in modulation of the resonance frequency due to incoming power on the detector. We thus seek to transform our complex digital data to an estimate of resonator frequency shifts. We transform shifts of the resonance frequency into shifts entirely in the phase (Q) direction of the resonator response via a rotation and scaling of the resonator's complex response, allowing us to ignore changes in the amplitude direction (I). Equivalently, we rotate the resonator circle in the complex plane such that it is oriented as given in Figure 2, and approximate small movements near the resonance frequency as occurring along a single axis.

The resonance frequency is estimated as the point of minimum amplitude response, which is true for an ideal resonator and a good approximation for typical NIST CMB-style μ mux resonators. Resonance frequency estimation is performed in software; thus other metrics may be supported by the SMuRF system as various applications require.

The resonator has a response in the complex plane corresponding to the real and imaginary components of the S_{21} response. This response may be rotated and scaled by a complex number η such that for small changes in resonance frequency,

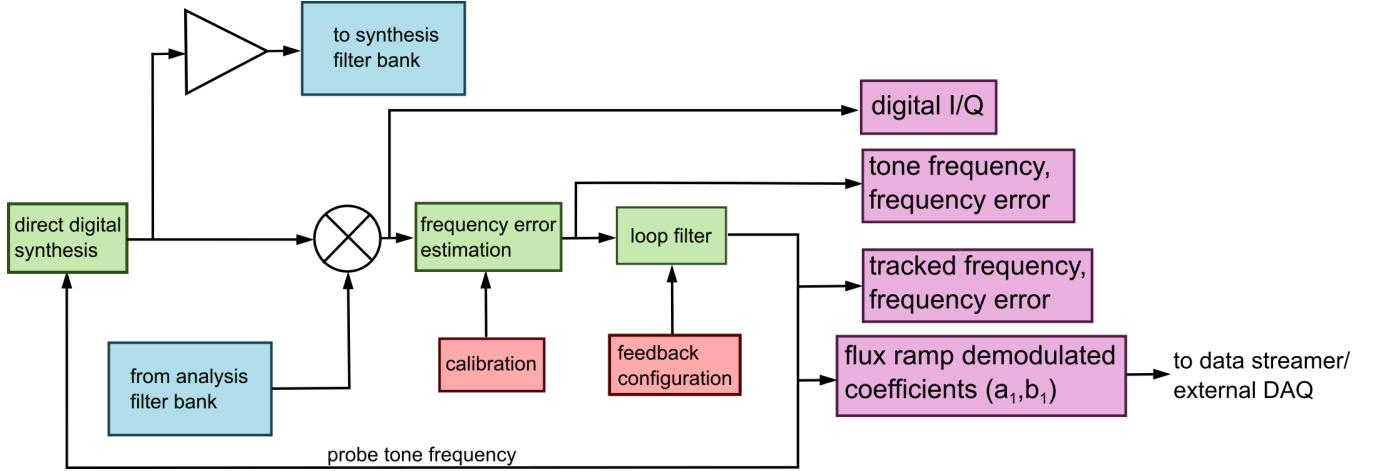


FIG. 13. The per-channel baseband processor. This acts on each of the channels serially, clocked at 499.2 MHz. Various intermediate points are available to examine for debugging and more detailed characterization, or the final demodulated data is passed to the data streamer for packaging and writing to disk or to external DAQ. External inputs are shown in red, FPGA computations are shown in green, interactions with the filter banks are given in blue, and options for data outputs are in pink.

the change in S_{21} is entirely in a single quadrature.

For (I_{\pm}, Q_{\pm}) the complex transmission measured at frequency offset $\pm\Delta f$ from the resonance frequency, we estimate η as

$$\eta = \frac{2\Delta f}{(I_+ - I_-) + i(Q_+ - Q_-)}. \quad (3)$$

We see that the denominator gives an angle in the complex plane by which to rotate the resonance circle, while the overall magnitude is set by the sharpness of the resonance. Thus, η has the effect of rotating and scaling the resonance circle such that the real and complex axes have physically meaningful units and with the S_{21}^{\min} rotated to be parallel to the imaginary axis.

With this calibration, the frequency shift may then be estimated as

$$\Delta f \sim \hat{\Delta}f \equiv \text{Im}[S_{21}(\Delta f) \times \eta] \quad (4)$$

where $S_{21}(\Delta f)$ is the complex transmission at the shifted resonance frequency. Throughout this text we denote estimates with a hat. This estimate is equivalent to projecting the rotated and scaled S_{21} response at the probe tone frequency onto a single axis. A sketch of this calibration scheme is given in Fig. 14.

The estimate holds for small changes in frequency and in the limit that shifting the resonance frequency away from the tone is the inverse of shifting the tone, ie

$$\frac{\partial S_{21}(f - f_{\text{res}}(\phi))}{\partial f} = -\frac{\partial S_{21}(f - f_{\text{res}}(\phi))}{\partial f_{\text{res}}(\phi)} \quad (5)$$

where f is the probe tone frequency and f_r is the resonance frequency, which depends on the flux ϕ . This assumption is valid for most resonator-based readout systems in the small signal limit, i.e. where the frequency shift Δf is small compared with the resonance bandwidth. Figure 15 shows the estimated frequency error $\hat{\Delta}f$ using the η calibration versus the

true frequency error Δf . In particular, we note that the frequency error estimate holds only for a small region within the resonance bandwidth.

The η estimate must be performed once per resonance, typically at the first cooldown for a set of resonators. The calibration is largely stable between data-taking sessions and fridge cycles for a given set of devices, and saved for reuse between device cooldowns. Subsequent to η calibration, the SMuRF is capable of channelizing tones and returning the output tone frequency and frequency error, interpreted as the resonator frequency offset from the probe tone, for each resonator simultaneously.

C. Tone Tracking and Flux Ramp Demodulation

We implement a feedback loop to minimize the SMuRF estimate of the frequency error magnitude and continuously update the probe tone to be positioned at the estimated resonance frequency.

When using the microwave SQUID multiplexer, the detector signal is reconstructed via phase shifts in the flux ramp response, which are linear in detector signal for detector signals much slower than the flux ramp rate. For recent generations of NIST CMB-style μ mux resonators with bandwidth of about 100 kHz, we typically modulate the SQUIDs with flux ramp sawtooth rates of several kHz and 3 to 6 flux quanta swept per sawtooth period.¹⁶

Rather than design a feedback loop to directly track arbitrary signals at 10s of kHz or faster, which is challenging due to latency from the filter bank, we rely on the periodicity of the flux ramp response. That is, we express the flux-ramp modulated frequency response of the resonator as in Equation 2, reproduced here for convenience:

$$\Delta f(t) = B \left(\frac{\lambda \cos(\omega_c t + \theta(t))}{1 + \lambda \cos(\omega_c t + \theta(t))} \right).$$

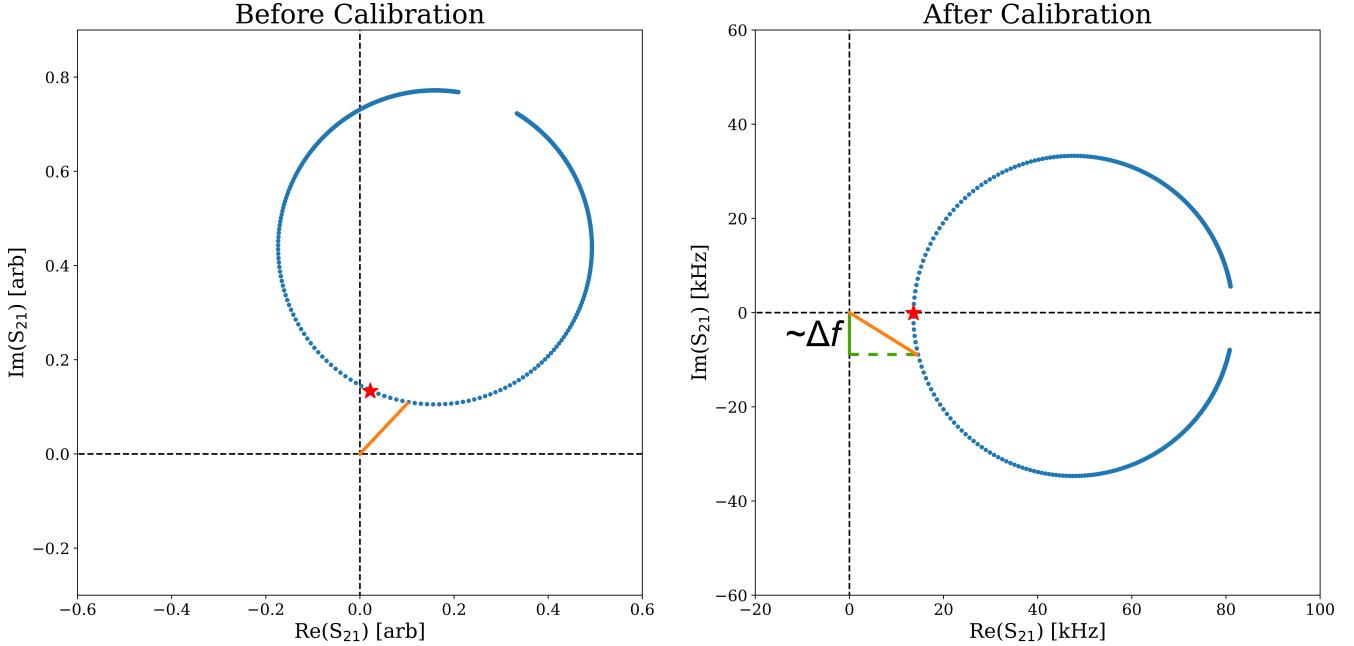


FIG. 14. A sketch of the calibration and frequency error estimation. [Left] An uncalibrated resonator generically has some response in the complex plane that may be arbitrarily rotated and scaled. A probe tone (orange) is initially tuned to the point of minimum transmission S_{21}^{\min} (red), but as the resonator is modulated by flux the S_{21}^{\min} will shift away from the probe tone. [Right] Following calibration by η as defined in Eq. 3, the resonance circle is scaled such that the axes are in physically meaningful units and the small changes in the resonance frequency correspond to movements in only one axis. The frequency error Δf between the probe tone and the new resonance frequency is estimated as the projection (green) of the probe tone complex response onto a single axis, corresponding to Eq. 4.

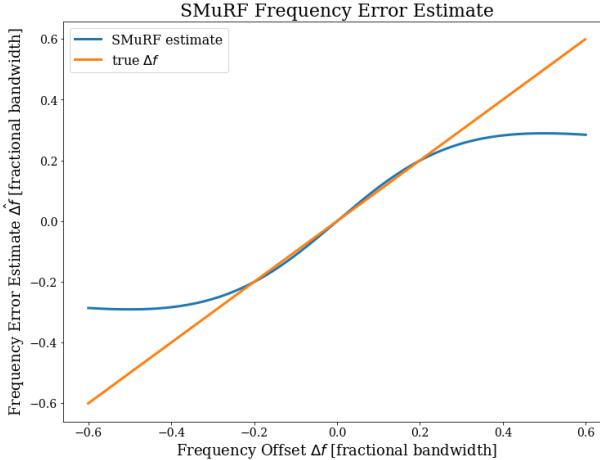


FIG. 15. Estimated frequency error with the η multiplication (blue) versus real frequency error for a simulated resonator (orange). The regime in which the estimated frequency error $\hat{\Delta}f$ (blue) matches the real frequency error is in a narrow band about the resonator frequency ($\Delta f = 0$). SMuRF tracks resonators with tones in this narrow bandwidth while flux ramping using an adaptive filter as described in Section VIC.

The carrier frequency $f_c = \omega_c/2\pi$ is the product of the flux ramp sawtooth reset rate and the number of flux quanta swept in a single flux ramp period. For NIST CMB-style μ mux resonators, the design values are a resonator peak to peak swing

of about 100 kHz and $\lambda \sim 1/3$, though this discussion applies without loss of generality to any sufficiently periodic signal.¹⁶ For detector signal $\theta(t)$ approximately constant relative to $\omega_c t$, we treat this as an estimation problem for multiple sinusoids in a truncated Fourier series, where we have M sinusoids of known frequencies. These frequencies are harmonics of the carrier with known frequency f_c but unknown in amplitude or phase.

Thus, the resonance frequency of a given channel as function of time is parameterized by

$$f[n] = \text{constant} + \sum_{i=1}^M A_i \sin(\omega_i T_s n + \theta_i) + w[n], \quad (6)$$

where we include a constant offset term and a Gaussian measurement noise $w[n]$. Here, n is the sample number and T_s is the time per discrete sample, which is $T_s = 1/(2.4 \text{ MHz})$ as defined by the filter bank per-channel bandwidth. Expanding the sine terms, we can rewrite the frequency as a sum of sines and cosines:

$$f[n] = \text{constant} + \left(\sum_{i=1}^M a_i \sin \omega_i T_s n + b_i \cos \omega_i T_s n \right) + w[n] \quad (7)$$

where

$$a_i = A_i \cos \theta_i, \quad b_i = A_i \sin \theta_i. \quad (8)$$

In the limit of slowly varying $\theta(t)$, the a_i and b_i are approximately constant, and the overall frequency estimate is linear in the a_i, b_i , and constant offset. The change in phase of each harmonic is given by $\Delta \arctan(b_i/a_i) = i \times \Delta\theta_i$. As the detector signal modulates the phase of the entire SQUID response curve, we seek an estimate of the phase of the fundamental harmonic $\theta_1 = \arctan(b_1/a_1)$. Due to FPGA resources, in the μ mux system we take the number of tracked harmonics M to be 3. In practice this provides a sufficient parameterization of well-behaved SQUID curves on NIST CMB-style μ mux devices.

For M harmonics of a fundamental frequency ω_1 , Eq. 7 may

$$\mathbf{H} = \begin{pmatrix} \sin_1[0] & \cos_1[0] & \sin_2[0] & \cos_2[0] & \cdots & 1 \\ \sin_1[1] & \cos_1[1] & \sin_2[1] & \cos_2[1] & \cdots & 1 \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ \sin_1[N-1] & \cos_1[N-1] & \sin_2[N-1] & \cos_2[N-1] & \cdots & 1 \end{pmatrix}, \quad (10)$$

where

$$\sin_m[n] \equiv \sin(\omega_m T_s n), \quad \omega_m = m \times \omega_1 \quad (11)$$

and similarly for $\cos_m[n]$. Our problem thus reduces to estimating the coefficients vector $\vec{\alpha}$.

The current SMuRF tracking and demodulation loop implements this estimation using a recursive algorithm. The fundamental tracking frequency $f_1 = \omega_1/2\pi$ is assumed known and must be measured using software tools (§ VII C) or input by the user. At each discrete timestep n , a probe tone centered on a resonance has frequency predicted as

$$\hat{f}[n] = a_1[n] \sin(\omega_1[n]) + b_1[n] \cos(\omega_1[n]) + \cdots + C \quad (12)$$

which may be thought of equivalently as the dot product of the n th row of \mathbf{H} with $\vec{\alpha}$. Simplifying notation, we can write the n th row of \mathbf{H} as $\vec{h}[n]$, given by

$$\vec{h}[n] = (\sin_1[n] \ \cos_1[n] \ \cdots \ \sin_N[n] \ \cos_N[n] \ 1). \quad (13)$$

Eq. 12 is then written more compactly as

$$\hat{f}[n] = \vec{h}[n] \cdot \vec{\alpha}[n]. \quad (14)$$

We measure the frequency error as described in § VI B following Eq. 4 as

$$\Delta f[n] \equiv f[n] - \hat{f}[n] \approx \text{Im}(S_{21}(\hat{f}[n] - f_{\text{res}}[n]) \times \eta) \quad (15)$$

Importantly, we measure only $\Delta f[n]$ and do not have access to the true $f_{\text{res}}[n]$.

Based on this readback, we update the prediction via a feedback loop such that

$$\vec{\alpha}[n+1] = \vec{\alpha}[n] + \mu \Delta f[n] (\vec{h}[n])^T \quad (16)$$

be rewritten as

$$\vec{f} = \mathbf{H} \vec{\alpha} + \vec{w} \quad (9)$$

where \vec{f} and \vec{w} are $M \times 1$ column vectors of resonance frequency values and measurement noise for N discrete samples per flux ramp frame. We define coefficient vector $\vec{\alpha}$ as a $(2M+1) \times 1$ column vector of sine and cosine coefficients $a_1, b_1, \dots, a_N, b_N$, constant for M harmonics and a constant offset term. The harmonic sample matrix \mathbf{H} is an $N \times (2M+1)$ matrix given by

where μ is a user-tuned gain parameter. This algorithm is equivalent to the stochastic gradient descent method, where we are attempting to minimize $\hat{\Delta}f[n]$.⁵⁰ Note that $\hat{f}[n]$ and $\hat{\Delta}f[n]$ are scalars corresponding to the probe tone frequency and estimated frequency error, respectively, of a single channel at discrete timestep n . Since ω_1 is known, the flux ramp rate is no longer limited by the bandwidth of the constant-only tracking loop, i.e. the $N=0$ case.

In this $N=0$ case of no flux ramp such as with KID systems or when slower tracking is desired, $\vec{\alpha}$ reduces to a single constant frequency error $\alpha = \Delta f$. We further see that $\vec{h}[n]$ is simply 1, and $\hat{\Delta}f[n]$ is the estimated frequency error, which is added to the previous constant offset term α with some gain μ . In this limit, a model with only the constant offset term is equivalent to having only proportional control in the tracking loop. The bandwidth of this operation alone is about several kHz, setting the limit for tracking bandwidth in non-flux ramped systems. Planned modifications for KID systems would use the resources freed up by no longer using the higher harmonic tracking to add a proportional gain term or a lead/lag compensator to increase the loop bandwidth, allowing for higher bandwidth applications such as calorimetry.

Since this tracking loop estimates coefficients of the harmonic content, the constant term encodes frequency shifts not directly sampled by the harmonics vector $\vec{h}[n]$. In the case of poorly selected tracking frequencies ω_i , this effectively aliases against the true modulation frequency and misconstrues the phase. The constant term C absorbs slow shifts in the resonator frequency or the amplitude of transmitted quadrature voltage through the resonator, potentially suppressing sources of noise including but not limited to amplifier gain fluctuations and microphonic pickup.

The user-defined parameters to the tracking algorithm are the fundamental harmonic frequency ω_1 and the gain μ . As is typical for generic tracking feedback systems, the gain must be large enough to guarantee convergence without causing sta-

bility issues. In practice, there exists a broad range of gain values for which the feedback loop is stable, with the gain determining the feedback loop bandwidth.

The fundamental tracking frequency f_1 is independent of the resonator bandwidth, and resonators centered in bins of the polyphase filter bank can in principle be tracked as fast as the per-channel Nyquist frequency, which for the SMuRF CMB firmware implementation is 1.2 MHz. Resonator tracking at rates faster than the resonator bandwidth may be desirable for several applications. The tracking loop updates for each channel at the per-channel digitization rate of 2.4 MHz; however, the stochastic gradient descent method is only expected to converge on the mean phase over the flux ramp period.⁵¹ Therefore, while the tracked frequency of each resonator is sampled at 2.4 MHz, detector information is only sampled at the flux ramp rate. Because the tracking feedback loop acts as a low pass filter, its bandwidth must be limited to at most half the flux ramp rate through the choice of tracking gain μ or pre-filtering to avoid aliasing detector noise into the signal band. Combined with the fact that increasing flux ramp rate modulates the detector signal band to a region of lower TLS contributions, higher flux ramp rates are generically preferred to reduce readout noise.⁴⁶

To avoid losing lock on the feedback loop during the flux ramp sawtooth resets, which cause a transient in the SQUID that are difficult to track, the tracking algorithm supports a user-tuned window within each flux ramp frame to be “blanked off”, or removed from the feedback loop. During this period, the coefficients α are held constant at the last updated value. The size of this window is typically tuned to cover the sawtooth reset period and subsequent transient response, and to maintain an integer number of flux quanta being tracked in the active loop. Blanking introduces an aliasing penalty due to the lower duty cycle, which is mitigated by keeping the blanking window size as short as possible.

D. Other Miscellaneous Firmware Functions

In addition to tone synthesis and readback, channelization, frequency error estimation, and tracking, the firmware handles miscellaneous functions such as the data streaming interface, timing synchronization between SMuRF and external systems, and control and configuration of the AMCs, RTM, and cryostat card.

In the carrier FPGA firmware, the streaming data framer sums the estimated coefficients of the first tracked harmonic a_1, b_1 as discussed in § VIC for each flux ramp frame and writes them to a dual port RAM at the start of each frame. This effectively averages over the flux ramp frame. From the dual-port RAM of each baseband processor, the per-channel coefficients and channel ID information are combined into a full 4096 channel RAM. The per-channel phase is computed serially as $\theta_1 = \arctan(b_1/a_1)$ with a coordinate rotation digital computer (CORDIC) algorithm, packaged with header information, and passed to the software streaming interface. The header includes information about the hardware configuration including the RTM DAC settings including detector bias set-

tings, flux ramp status, timing counters and synchronization bits, data rate, and optional user words. Additional coefficients from the higher harmonics of the tracking loop could in principle be added to the data stream for higher signal to noise performance, although in practice have not yet been found necessary and are not currently implemented.

The arctan computation is clocked at 156.25 MHz. Since all 3328 channels must pass through a single CORDIC for the phase calculation, this clock thus limits the per-channel streaming data rate to 156.25 MHz / 3328 channels ~ 50 kHz. Since the flux ramp reset rate in typical operation has not exceeded ~ 30 kHz, this is acceptable for current uses but could be upgraded to faster clock rates if necessary. Faster data rates are available for writing directly to disk for diagnostic and debugging purposes. These may be sampled up to the per-channel digitization rate of 2.4 MHz per channel, or the decimated 614.4 MHz ADC speed for the full band without channelization. Applications with fewer channels, such as high bandwidth pulse detection applications, would also have higher data streaming rates without the need for an upgraded clock.

The flux ramp sawtooth reset is triggered through one of three possible modes: (1) internally, (2) through a trigger from a common timing system, or (3) from an external TTL trigger on a LEMO port on the RTM. The internal timing mode uses a 32-bit programmable count limited counter in the FPGA, clocked at the complex baseband processing rate of 307.2 MHz. The trigger and pulse width are passed to the digital signal processing core for flux ramp demodulation and to the CPLD, which has a 32-bit counter to receive the external reset. The upper 20 bits of the CPLD counter are selected to output to the flux ramp DAC on the RTM, which allows for better flux ramp amplitude control. For large installations in which the timing information is important, the external timing option allows for higher precision and synchronization. The common timing system was originally developed for accelerator systems and designed for synchronizing large numbers of devices to sub-nanosecond precision.⁴⁷ In this case, a separate timing card distinct from the SMuRF carrier card generates a timing data stream at 2.4576 GS/s and broadcasts timing words over a serial fiber. The timing words are subsequently decoded by each individual SMuRF carrier card. These words may be derived from an external timing system to allow for synchronization between SMuRF and non-SMuRF timing systems within a given experiment.

While the bulk of the firmware is located on the carrier card, the PIC microcontroller on the cryostat card and the CPLD on the RTM additionally run firmware. The RTM CPLD is nominally clocked at 51.2MHz to accommodate the maximum frequency of the flux ramp DAC of 50MHz, though the flux ramp DAC is typically operated well below this frequency. An arbitrary waveform generation module containing a look-up table is capable of writing pairs of bias DAC values sequentially, which allows for up to ~ 1 kHz updates to the DAC output voltages. Due to firmware resource limitations, the CPLD interface is write-only, so to maintain a record of operations we cache written values in FPGA shadow memory (BRAM).

The large-scale PCIe streaming interface splits incoming

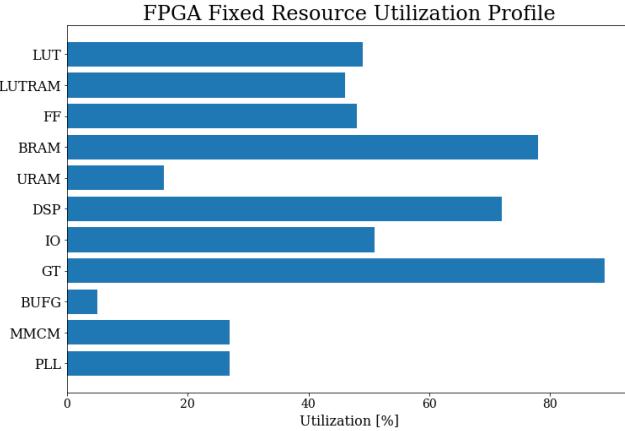


FIG. 16. Total consumption profile of fixed resources for the FPGA, a Xilinx Kintex UltraScale+ XCKU15P. The terms are as follows: LUT = look-up tables, LUTRAM = look-up table RAM, FF = flip-flops, BRAM = block memory, URAM = larger block memory, DSP = digital signal processing blocks, IO = input/output interfacing, GT = gigabit transceivers, BUFG = global clocking, MMCM = mixed-mode clocking manager, PLL = phase-locked loops for clocking.

data from the QSFP links to separate ports handling firmware register access and streaming data, respectively. Up to 6 carriers, typically run together in a single 7-slot ATCA crate with a network switch occupying the final slot. Each carrier's register access and data streaming ports are routed to the CPU via direct memory access (DMA) lanes. The register access port is low-latency for fast commanding and access. Every pair of carriers receives a dedicated 4 GB error correction code (ECC) DDR4 block, allowing for 2 GB of deterministic buffering per carrier before the CPU RAM to avoid dropping packets due to CPU back pressures on the DMA path. To avoid software data operations from back pressuring the register access, the DMA lanes for register access and data streaming are kept separate with independent PCIe DMA descriptors.

E. Firmware Resource Consumption

A challenge for SMuRF was fitting the firmware within the resource profile of the FPGA. The total FPGA resource consumption profile is given in Figure 16. We see that the most intensive elements for the SMuRF involve memory and signal processing. The high FPGA firmware resource usage and high 614.4 MHz clock rate is near the device limits and require substantial firmware optimization. We confirm that the requisite power draw and thermal environment can be maintained within device limits for this configuration even in high-altitude environments in § VIII F.

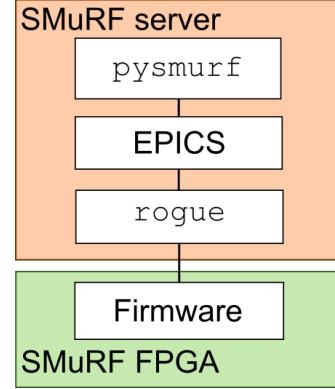


FIG. 17. A block diagram of the software components. A server (Dell R440) interfaces with the FPGA via Rogue via the EPICS framework. The `pysmurf` control software provides the end-user interface with the system.

VII. Software

A block diagram of the SMuRF software interface is given in Fig. 17. The FPGA is interfaced with a server via Rogue⁵², a Python-based hardware abstraction and data-streaming system. Rogue registers are accessed via the Experimental Physics and Industrial Control System (EPICS)⁵³, a control system architecture which acts as a communication protocol between the firmware registers and the end user's desired interface. For SMuRF end users, the EPICS layer is further wrapped in the open source `pysmurf` control software, which provides higher level processing and analysis tools.⁵⁴ Each of these layers is discussed in turn below.

To guarantee a consistent software environment across multiple deployed systems, the software suite is version controlled with `git`⁵⁵ and deployed to user institutions via Dockers⁵⁶. Furthermore, each hardware component has a serial number programmed in that may be remotely queried to facilitate hardware tracking across multiple deployments.

A. Rogue

The Rogue system serves as the communication layer between the FPGA firmware and the server software. It provides data streaming and hardware abstraction in a clean, modularized format that enables systems to be brought up from the prototyping phase to the system integration and deployment stage. Firmware registers are accessed via a hierarchical tree structure that encodes and optionally commands the configuration and status of each register. A memory API and data stream API separately handle FPGA memory access and data processing, respectively. The data stream may be passed directly to downstream DAQ programs such as telescope control suites, as described in § VII D. Metadata streams of any Rogue process variable can also be streamed alongside the data at the data rate, allowing for real time recording of important calibration and state information.

B. EPICS

The Rogue firmware registers are accessed via Experimental Physics and Industrial Control System (EPICS) process variables.⁵⁷ The EPICS interface provides a control system architecture with a distributed real-time database of register values and a client/server model with efficient communication protocol for passing data. Existing infrastructure for EPICS is built on decades of experience in designing and implementing large-scale control systems for scientific instruments such as accelerator systems, including pre-built software tools. It is currently actively developed and maintained by an international collaboration of users across several large scientific facilities. A channel access protocol connects the EPICS server, which wraps the Rogue registers, to the client-side process variables that may be queried, set, or monitored by any computer on the same client network. EPICS process variable interaction may be conducted in essentially any modern interface, including but not limited to C/C++, Java, perl, python, Tcl/Tk, sh, and Matlab.

C. `pysmurf` Control Software

Since python is an open source language used extensively for astronomical applications and laboratory experiment settings, our primary user software suite is built in Python3. The `pysmurf` control software serves as the interface for general users interacting with the SMuRF system. The software provides the following functionality:

- Wraps EPICS process variable control functions and provides data handling and low level commands, such as turning tones on and off, setting DAC voltages for detector and amplifier bias, and acquiring time-ordered data.
- Provides utility functions for finding and characterizing resonators, setting up tone tracking, and acquiring flux ramp demodulated data.
- Stores specific hardware configuration information such as in-line wiring resistance, band frequencies, SQUID couplings, and shunt resistances in a human-readable configuration table that may be edited or referenced in later analysis.
- Offers higher level control and analysis functions, including but not limited to noise analysis, TES detector characterization, and system optimization.

The repository is git controlled and actively managed at SLAC to ensure continuity and cross-system compatibility as the software advances.

D. Integration with Telescope Control Suites

The `pysmurf` software suite and Rogue streaming interface can be integrated with other experimental DAQ/control

suites, aligning detector data with information such as telescope pointing, housekeeping, and observatory state. It is possible to either encode syncwords in the frame headers or release frames based on sync signals for frame alignment. These syncwords, typically received from an experiment common timing synchronization system, allow for frames to be aligned offline, typically by a downstream frame packager or in post-processing. This integration has been successfully demonstrated with the generic control program (gcp) observatory control suite in a test deployment with the BICEP/Keck program, and has been built for the Observatory Control System (OCS) in anticipation of deployment for the Simons Observatory.^{20,58,59} Critical setup, calibration, and control functions may be received by SMuRF from a central telescope DAQ. Optionally downsampled data is digitally low-passed with a multi-pole Butterworth filter with user-defined rolloff frequency. Useful registers may additionally be encoded in frame headers, allowing for streaming of information such as detector bias levels and observatory timing synchronization.

E. Simulation Suite

To model the impact of the SMuRF firmware algorithms, particularly in frequency error estimation and flux ramp demodulation as described in § VI B and VI C, these algorithms are implemented in a publicly available Python simulation suite.⁶⁰ The simulation suite takes as input real or simulated resonator and SQUID characteristics, detector-referred input signals, and readout parameters such as flux ramp rate and tracking loop gain, and outputs the flux ramp demodulated detector phase as would be output by SMuRF. The details of this code and example studies of the impact of various user-selected parameters on final detector-referred demodulated data are presented elsewhere.⁶¹ Integration with the `pysmurf` user control software suite is planned for a future release.

VIII. Performance

SMuRF systems have been successfully integrated and used for both laboratory and field measurements of CMB detectors, demonstrating achievement of end-to-end system performance meeting the goals of CMB experiments.^{16,20,21} In this section we explicitly address the electronics requirements as outlined by § III.

A. RF Dynamic Range and Linearity

As outlined in § III B 1, the RF dynamic range of the electronics in dBc/Hz, or the level of the total RF noise floor, is critical to the readout white noise performance. The SMuRF was designed to achieve better than 100 dBc/Hz at 30kHz carrier offset for 2000 tones. At smaller carrier offsets, the dynamic range is expected to degrade due to the noise profile of the LO used to upmix the synthesized tones. The dynamic range of the SMuRF system is considered through

the full tone synthesis and analysis chain, which includes the ADCs, DACs, and the polyphase filter banks. In Figure 18 we show 2000 tones generated over 4–8 GHz by a SMuRF system and the dynamic range of one of those tones measured directly using a signal analyzer connected to the SMuRF RF output with no additional amplification or attenuation beyond a short length of coaxial cable. Note that measurements do not include the effects of the downmix, ADCs, or analysis polyphase filter bank which combined can both improve and degrade the achieved end-to-end RF dynamic range.

To check the impact of the entire digital processing chain, we acquire a 0.5 second timestream of the digitized I and Q orthogonal data components from a single channel with a low-band AMC connected in loopback, i.e. with a short coaxial cable connecting its RF input to its RF output. Since the noise performance of a single channel is limited by firmware rather than hardware choices, one AMC suffices to demonstrate performance without the need for a second AMC. The tone power is set to -33 dBm output from the DAC, with an additional 6 dB of programmable RF attenuation added to each the RF input and output on the AMC, as is typically used for NIST CMB-style resonators. The digital I/Q data are acquired without additional processing, corresponding to the state of the data at the output of the analysis filter bank as described in § VI A. We take the power spectral density of the time-ordered data and convert to units of dBc/Hz via normalizing by the magnitude of the time-ordered data stream. This is given in the left panel of Figure 19.

We note that the $1/f$ performance of a single channel is due to $1/f$ noise in the hardware components, particularly in the LO generation and RF mixers, which have an $f_{\text{knee}} \sim 100$ Hz. However, for μmux this performance is irrelevant due to the flux ramp modulation scheme: we focus instead on the noise at the carrier offset, where the sidebands due to flux ramp modulation are sampled. Thus, we repeat the measurement shown in Figure 19 while sweeping the generated tone across the entire frequency band and extract the noise level at 30 kHz offset, a typical flux ramp Φ_0 rate. This is given in the left panel of Figure 20. For applications requiring better $1/f$ stability in RF performance, upgraded hardware components have been identified and may be swapped in easily.

We see that the noise floor is well below the requirement of -100dBc/Hz across the entire frequency range. The noise rises slightly at the band edges, as is expected given the finite roll-off of the cavity filters on the AMC RF daughter cards.

A key design consideration for SMuRF is its ability to read out a large number of channels for cryogenic multiplexing applications. We trade off single channel noise performance for linearity across a large number of channels. We therefore require that the noise performance achieved in one channel is scalable to 2000 channels, corresponding to the expected multiplexing density for NIST CMB-style μmux resonators. In the right panel Figure 19, we show the noise spectral density of a single channel in loopback while 2000 channels at -33 dBm each are turned on from 4–8 GHz across two AMCs, which are chained in a configuration consistent with operating a full 4–8 GHz resonator set on a single coaxial line within the cryostat. This corresponds to the maximum expected chan-

nel density of a full operating SMuRF system, limited by the density of cryogenic channels. We see that the noise performance suffers minimal penalty as compared to the single channel case shown in the left side of the figure.

Similarly, with 2000 channels turned on from 4–8 GHz on two AMCs, we take the noise on each channel serially and extract the noise level at 30 kHz offset. This is shown in the right panel of Figure 20, with the single channel case for comparison given on the left panel. For more explicit comparison, in Figure 21 we plot the histogram of the counts in each side of Figure 20.

B. White and Low Frequency Noise Performance

Reading out newer, optimized (v3.2 or later) NIST CMB-style resonators not connected to TESs in dark cryostats with basic magnetic shielding (e.g. only a room-temperature mumetal can around the cold multiplexer volume), μmux -coupled SMuRF systems routinely achieve white noise levels of $O(40 \text{ pA}/\sqrt{\text{Hz}})$ TES current noise-equivalent with an approximately $1/f$ -like spectrum with $1/f$ knees of $O(10 \text{ mHz})$ while multiplexing $O(1000)$ channels.^{19,62} For ground-based CMB receivers instrumented with $O(10 \text{ m}\Omega)$ normal resistance TESs this level of readout noise is expected to increase the total per-detector noise-equivalent power, which is dominated by fluctuations in the observed photon intensity, by only an additional 1–2%. While a full analysis of the different noise contributions is deferred to a future publication, typical measured white noise levels measured at a $O(30 \text{ kHz}) \Phi_0$ modulation rate are dominated by resonator TLS noise, with a subdominant contribution from SMuRF’s finite RF dynamic range.

Confirming the TES and flux ramp drives achieve their low $50 \text{ pA}/\sqrt{\text{Hz}}$ output current noise specification (see Section III B 1) is difficult using conventional benchtop instrumentation due to the required sensitivity. From integrated system measurements, the estimated white noise contributions from the TES and flux ramp drives are consistent with exceeding their output current noise specification. At low frequencies, characterizing the noise performance of the TES and flux ramp drives is further complicated by additional noise sources including variations in passive component temperatures, microphonics, and RF phase drift. A method for monitoring the overall phase drift of the RF line for slow changes to enable feedback on tuning parameters or offline data cleaning has further been developed.⁶³ Further discussion of the low-frequency system performance, with particular focus on the SMuRF TES and flux ramp drives as well as mitigation techniques integrated into the software and hardware, will be addressed in a future publication.

C. Amplifier Linearity and Tone Tracking

As detailed in § III B 2, microwave resonator systems are often limited by the linearity of both warm and cryogenic RF components. We see above in § VIII A that the SMuRF RF

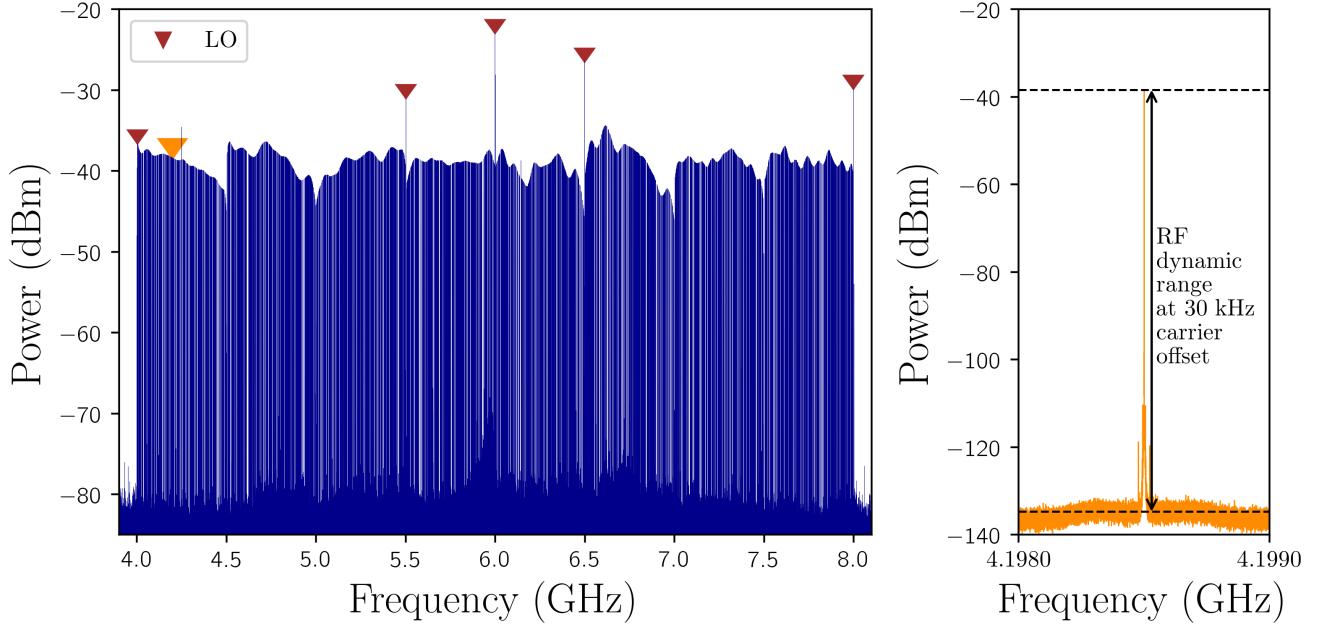


FIG. 18. [Left] 2000 fixed tones generated with SMuRF over 4–8 GHz (with 250 tones in each 500 MHz band) measured directly using a signal analyzer. LOs which fall within the frequency range of the plot are labeled with smaller brown markers. The average tone power in each 500 MHz band has been leveled using the programmable output attenuators. [Right] Zoom in of a single tone while playing 2000 fixed tones, indicated by a larger orange marker on the lefthand plot. The arrow indicates the measured RF dynamic range at a 30 kHz carrier offset. The signal analyzer trace has been median filtered (while preserving the maximum) to suppress narrow-band spurs. The dynamic range at 30 kHz carrier offset exceeds the SMuRF requirement of > 100 dBc/Hz after accounting for the signal analyzer noise floor (which is only ~ 6 dB lower) and noise which cancels when the RF signals are downmixed with their common LO.

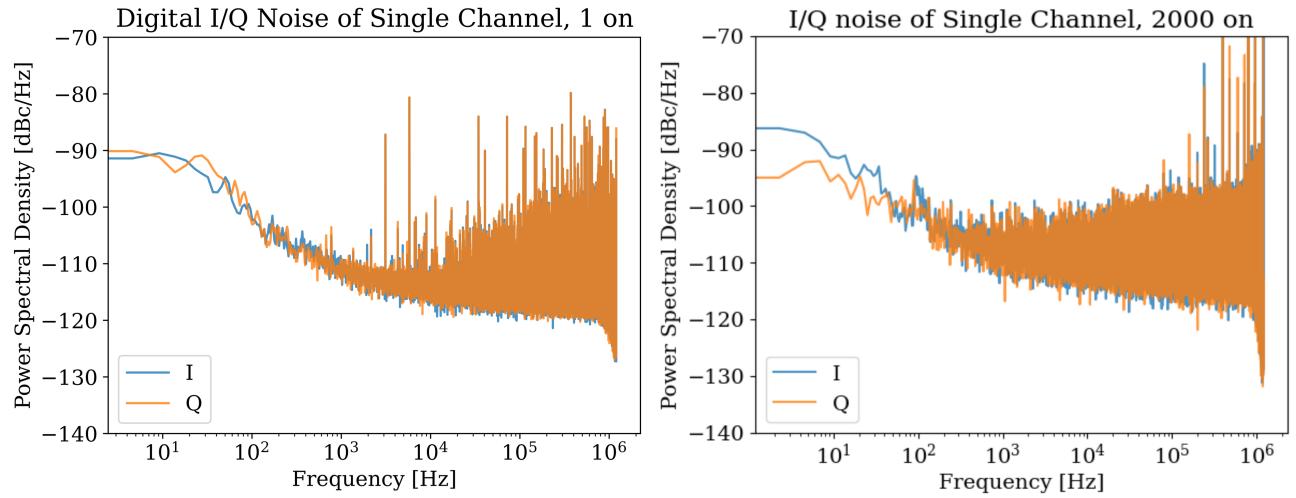


FIG. 19. [Left] The noise spectral density of a single channel looped back directly from RF out to RF in with only that channel on. The traces represent orthogonal data components I and Q, which should be identical given there is no resonator in the measurement. We expect the $1/f$ knee to be at several hundred Hz due to the LO generation and RF mixers. The white noise level is below -100 dBc/Hz, as expected, with a roll-off at 1.2 MHz defined by the Nyquist rate of the polyphase filter bank. Narrow-band spikes are due to a combination of hardware and firmware DSP effects, but do not contribute substantially to the total noise power. [Right] The same measurement as the left, but with 2000 tones turned on from 4–8 GHz. The minimal degradation in noise performance demonstrates the excellent linearity of the full SMuRF system.

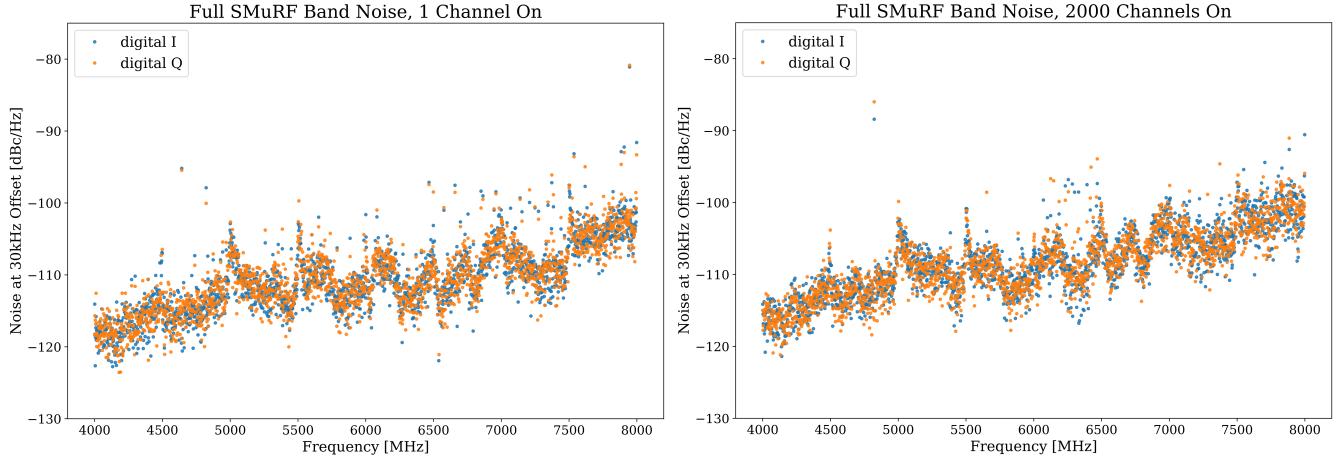


FIG. 20. [Left] The noise at 30 kHz carrier offset of a single channel in loopback across the entire 4-8 GHz frequency range. We see that the overall level remains below -100 dBc/Hz aside from a handful of outliers. The edges of each 500 MHz band are visible due to the slight roll-off in the edges of the passband of the cavity filters. A small overall tilt is expected due to the degradation of RF performance at higher frequencies for these broadband components. [Right] The same measurement taken while 2000 tones are turned on across 4-8 GHz, with 250 randomly chosen frequencies chosen in each 500 MHz band. We see that the overall level and shape are nearly identical to that of the single-channel case.

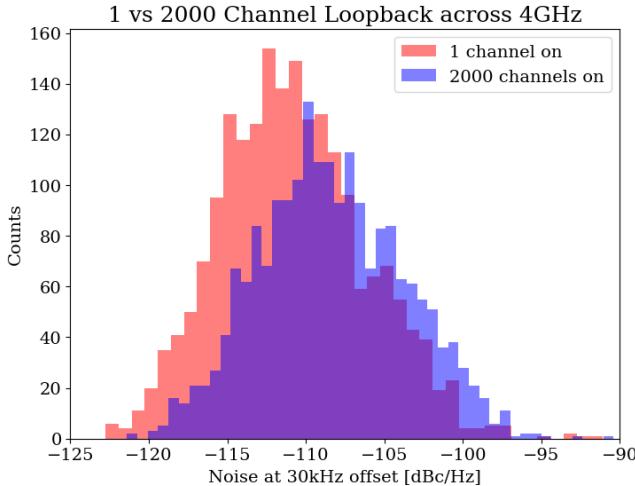


FIG. 21. Noise at 30 kHz offset of each channel in loopback with 1 or 2000 channels turned on using identical attenuation and amplification settings. The 2000 channel data was taken with 1000 channels each from a low band and high band AMC connected serially in loopback. We see that there is a modest penalty from turning on 2000 channels across the 4 GHz bandwidth, but the overall distribution falls within the -100 dBc/Hz specification from the initial requirement.

components achieve the requisite linearity: the warm system in loopback suffers minimal degradation when 2000 channels are operated simultaneously. The linearity of the full RF chain is therefore typically limited by the achievable linearity of the cryogenic amplifier and other cold RF components. As described in § VI C, SMuRF addresses this linearity problem by implementing a closed loop feedback algorithm to maintain the probe tone on the point of minimum amplitude transmis-

sion S_{21}^{\min} .

We show that this tone tracking system succeeds in reducing the power at the 4K cryogenic amplifier in a measured system in Figure 22. A resonator was calibrated using `pysmurf` software tools, with a directional coupler splitting the output of the cryostat between the SMuRF RF input on the strongly-coupled port and a spectrum analyzer on the weakly-coupled port. The power at the input of the cryogenic LNA chain was computed by taking the integrated power in a 100 Hz window around the probe tone frequency as measured by the spectrum analyzer and adjusting appropriately for the amplifier gains and coupler loss. When tracking is off, the probe tone is stationary and the total power at the 4K LNA input varies by over 10 dBm. After tracking is enabled, the probe tone moves with the resonance, and the power incident on the cryogenic amplifier is minimized and constant. Since the system linearity scales exponentially with the power incident on the RF amplifiers, this allows for substantial relaxation on the requirements for the linearity of the hardware components.

This minimization of tone power on the cryogenic amplifier is critical to reducing intermodulation products in band, which may contribute to crosstalk or noise. We demonstrate the improvement of the overall noise floor due to reduced intermodulation products in Fig 23. Here, a single NIST CMB-style μmux resonator chip with 65 SQUID-coupled resonators spanning 4.25-4.5 MHz is calibrated with SMuRF at a tone power roughly corresponding to -70 dBm per tone at the input of the resonators. After interacting with the resonators, the tones are amplified with a 4K HEMT (Low Noise Factory LNF-LNC4-8C-LG), a 50K follow-on amplifier (B&Z Technologies BZUCR-04000800-031012-102323), and room-temperature amplifier (MiniCircuits ZX60-83LN12+) and input to a spectrum analyzer. After feedback is disabled, no additional external flux is applied

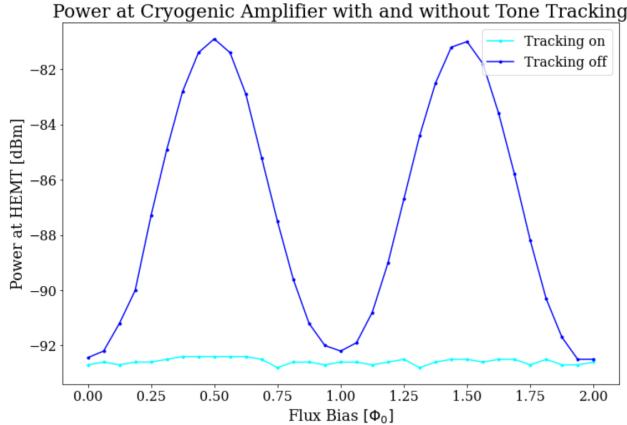


FIG. 22. The power incident on the first-stage cryogenic amplifier with and without tone tracking for a single tone tuned to a NIST CMB-style μ mux resonator. We see that the average power incident on the amplifier is reduced by about 5 dB for a single tone due to tone tracking, and by more than 10 dB for some bias points, significantly reducing the linearity requirement for the cryogenic amplifier performance.

to the SQUID. The measurement was limited by availability of cold multiplexer chips at the time it was performed; however, we see that even with a small subset of the full O(1000) tones used in typical operation, the overall noise floor due to the additional intermodulation products when feedback is disabled is elevated by ~ 3 dBm, thus degrading the RF dynamic range. This degradation is prohibitive for μ mux at full multiplexing density; thus, the tone-tracking capability of SMuRF is key to achieving the O(1000)x or higher multiplexing factors desired.

D. Bandwidth

The measurements in § VIII A and VIII C indicate that the SMuRF achieves the synthesis and analysis of tones across the full 4-8 GHz bandwidth while exceeding the requirement for RF dynamic range. We focus now on the achievable bandwidth of the detector readout firmware, including tone tracking and flux ramp demodulation. A discussion of the full bandwidth performance of the tone-tracking algorithm is deferred to a future publication. However, we see in Figure 24 that in the “DC tracking” state in which the flux ramp has not been activated, the power spectral density of tracked frequency and frequency error rolls off between 1 and 10 kHz depending on the gain settings used.

Using the adaptive feedback algorithm with tone tracking, previous measurements have indicated that the SMuRF is capable of tracking a flux ramp modulated resonance to flux ramp rates approaching 100 kHz on NIST CMB-style μ mux resonators, well in excess of requirements for CMB polarimetry measurements.⁴⁶ We show explicitly in Figure 25 that a 1 kHz sine wave injected on the detector bias line is resolved through the full tone tracking and flux ramp demodulation

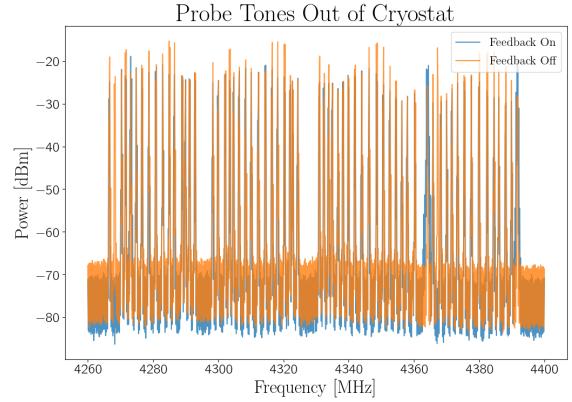


FIG. 23. Spectrum analyzer trace of the probe tone comb interacting with a single NIST CMB-style μ mux resonator chip containing 65 SQUID-coupled resonators. Probe tones approximately equivalent to -70 dBm per tone at the input of the resonators are calibrated to the resonators. The spectrum analyzer was placed at the equivalent of the RF input to SMuRF, after probe tones interact with the resonators and are amplified by 4K, 50K, and room temperature amplifiers. No additional flux is applied; thus, any change in resonance frequencies is due to noise only. We see that when feedback is disabled, the effective noise floor rises by about 3 dB.

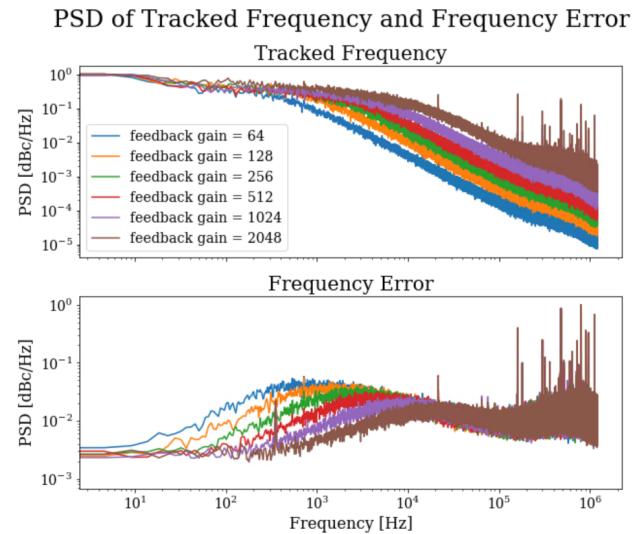


FIG. 24. Power spectral density of tracked frequency and frequency error for varying gains and no flux ramp. We see that the tracked frequency rolls off at a few kHz or above depending on the gain setting. The frequency error is suppressed at low frequencies, corresponding to the majority of the power being captured by the tracked frequency.

scheme. This measurement was limited by the ability of RTM slow DACs to play a pre-programmed waveform on the TES bias line, but is not near the capacity of the system.

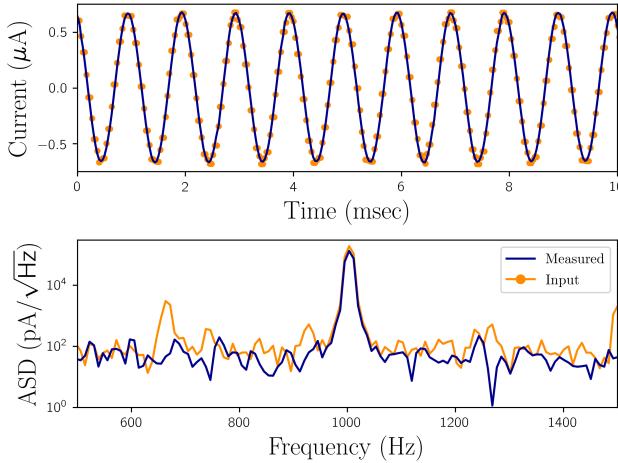


FIG. 25. [Top] Input signal and flux ramp demodulated signal played on the detector bias line while the detector was superconducting. The raw input signal is sparsely sampled and scaled and shifted to account for the transfer function of the superconducting detector's bias circuit and latency. [Bottom] Amplitude spectral density (ASD) of the input and measured signals. SMuRF easily generates and recovers a 1 kHz sine wave, here sampled at a flux ramp frame rate of 30 kHz with 1 Φ_0 per ramp. The white noise floor of $\sim 100 \text{ pA}/\sqrt{\text{Hz}}$ is due to Johnson noise generated by the superconducting detector's cold bias resistor.

E. Crosstalk

Since the polyphase filter bank achieves 100 dBc/Hz SFDR, it is expected that the crosstalk between channels is dominated by the hardware components. The linearity measurements from § VIII C indicate that the impact of additional probe tones being turned on in other channels does not contribute meaningfully to the measured characteristics of existing channels.

Previous measurements with SMuRF in a full field deployment setting including telescope scanning and streaming to external DAQ indicate that the crosstalk between non-pathological cold multiplexer channels integrated with TESs is at least 30 dB below the primary detector channel.⁶⁴

More recent laboratory measurements of the cold multiplexer only constrain the crosstalk of well-behaved channels to about 1 part in 10^4 , consistent with expectations for the crosstalk of μmux cold hardware.¹⁶ These suggest that the crosstalk of the electronics comfortably meets the existing requirements and dominated by the cryogenic elements.

F. Altitude

We demonstrate that the SMuRF system is operable in high-altitude environments despite its computationally intensive digital signal processing algorithm.

Due to high FPGA use and the 614.4 MHz maximum FPGA clock frequency, each SMuRF system (not including peripherals) consumes roughly 210 Watts. Out of this 210 W total,

a typical SMuRF system draws 110 W for the SMuRF carrier, 40 W for each AMC, and 20 W for the RTM. The power consumed by the SMuRF carriers is dominated by the main Xilinx Kintex UltraScale+ XCKU15P FPGA, which can draw as much as 80 W in steady state when reading out resonators over the full 4 GHz SMuRF bandwidth. The majority of the power drawn by the FPGA is consumed by the FPGA's 0.85V internal supplies, which draw more than 50A which is provided by an 80A Intel Empirion Power Solutions EM2280P01QI regulator on the SMuRF carrier. At these large current and power draws, the temperature limits of this regulator and the main FPGA are critical design considerations, particularly because the SMuRF systems are air cooled and intended for use at high altitude astronomical sites where the air pressure can be as much as 50% lower than sea level. For the regulator, this is compounded by the fact that the regulator efficiency degrades near its 80A limit, which can result in thermal runaway if not carefully controlled for in design. FPGA lifetime is an additional concern, since expected lifetime can degrade if the FPGA temperature exceeds the Xilinx temperature specification. The maximum regulator operating temperature is 120°C.⁶⁵ Xilinx specifies that the FPGAs must not exceed a temperature of 100°C for a 50A input current to ensure a > 10 year lifetime.

Early SMuRF prototypes exceeded these temperature limits in simulated high altitude conditions. This was addressed in the subsequent production systems by modifying the SMuRF FPGA firmware to significantly decrease resource usage and by making significant improvements to the FPGA and regulator heatsink designs and heatsink gluing materials and procedures. To verify performance, six full SMuRF systems, each consisting of a SMuRF carrier with two installed AMCs and an RTM, were tested in a 7-slot COMTEL ATCA crate along with all required peripherals, including a Dell R440 server and Xilinx KCU1500 data acquisition PCIe card, in a low pressure chamber.⁶⁶ With the chamber pressure pumped down to the equivalent air pressure at 17,000 ft (395 Torr) to simulate operating conditions at a high altitude astronomical site, a test routine was run simulating the full computational load expected for the read out of the maximum 3328 channels over the full 4 GHz SMuRF bandwidth for each of the 6 SMuRF systems simultaneously. The chamber temperature was ramped from -5°C to +25°C and then back down to -5°C on a 3 hour cadence 120 times over the course of 14 days, simulating the typical diurnal temperature swing expected at a typical high altitude astronomical site based on environmental data from the high Andes. No subsystem (including peripherals) failures were observed over this 14 day run, and the SMuRF subsystem temperatures and currents were logged while ramping the chamber temperature. At 20°C ambient temperature, the averaged logged temperature for the SMuRF carrier FPGAs and regulators during this test was 71°C and 67°C, respectively, and the average regulator current at 20°C was 57A. The measured regulator temperatures are well below their 120°C operating limit, and the FPGA currents and temperatures meet Xilinx's 10 year lifetime specification modified to account for these lower FPGA temperatures.⁶⁷ The system was also successfully cold booted at -5°C three times,

simulating remote recovery in winter conditions.

G. Vibration Sensitivity

The acoustic sensitivity of the SMuRF is determined primarily by the specific details of a given installation, such as the RF cabling used and the mounting between the crate and cryostat. While a comprehensive frequency sweep has not been exhaustively probed, prior field deployment on a scanning telescope indicated the SMuRF electronics suffered from no obvious additional noise due to the telescope motion, indicating that mounting schemes exist such that the SMuRF is sufficiently acoustically isolated.²⁰ Numerous existing lab measurement setups indicate no detectable acoustic sensitivity in the electronics from normal cryogenic laboratory settings, including the nearby operation of a pulse tube compressor and water chiller.

In order to sufficiently cool the boards within a crate, which may house arbitrarily many (typically up to 6) parallel SMuRF systems, we run the SMuRF crate fans at high speeds. This fan motion does not appear to cause additional noise in the electronics.

IX. Conclusion and Future Outlook

We have developed a digital readout and control system for use with large-scale microwave resonator-based arrays. The SMuRF electronics achieve > 100 dBc/Hz dynamic range at 30 kHz offset for 2000 simultaneous channels across its full 4–8 GHz bandwidth, with capacity to read out up to 3328 channels across its readout bandwidth. Each channel may return data with bandwidth up to 2.4 MHz and with a variety of processing options up to reporting flux ramp de-modulated timestreams. Compared to other microwave resonator readout systems, SMuRF is unique in its implementation of a tone-tracking algorithm, which allows for dramatically increasing the channel count without exceeding the linearity requirements of RF elements such as cryogenic amplifiers. The system has been designed for integration with other data acquisition and experimental control platforms, and has been successfully integrated with control suites for the BICEP/Keck and Simons Observatory CMB experiments. It has been fielded previously within the BICEP/Keck program and is baselined for use on the upcoming Simons Observatory and BICEP Array experiments.

Current efforts are underway to reconfigure the existing SMuRF firmware and hardware architectures for KID systems by removing hardware and firmware functionality required only for μ mux readout. It is estimated that for resonators fabricated with sufficient frequency accuracy, a single SMuRF card has the resources to support 20,000 KID-type resonators. This would allow for megapixel arrays to be read out with a single rack of equipment.

Future SMuRF development efforts include development of on-board pulse processing algorithms for high-bandwidth applications, improving the resonator tracking calibration

scheme for asymmetric resonances, and porting the existing software and firmware architecture to newly available “RF system on a chip” (RFSoC) systems. RFSoC systems allow for a large simplification in RF hardware design, compact profile, and lower power consumption, making them particularly attractive for balloon- and space-based missions. Preliminary results on a Xilinx ZCU111 evaluation board demonstrated comparable noise performance on 237 tracked channels, with the measurement limited by the availability of cold resonators at the time.⁶⁸ In addition to reproducing the existing SMuRF firmware and performance on the RFSoC, a number of improvements using the RFSoC platform are being pursued. We have developed new approaches to calibrating and improving the linearity of baseband I/Q modulation and demodulation schemes that provide the performance required for μ mux without requiring the large band-splitting cavity filters present on the current AMCs. This technique has been successfully demonstrated with test code on RFSoC hardware and currently is being fully built out and tested.

Acknowledgments

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Data Availability Statement

The data that support the findings of this study are available from the corresponding author upon reasonable request.

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