Multiplexed readout of kinetic inductance bolometer arrays

Hannu Sipola¹, Juho Luomahaara¹, Andrey Timofeev¹, Leif Grönberg, Anssi Rautiainen², Arttu Luukanen², and Juha Hassel¹

¹VTT Technical Research Centre of Finland LTD & QTF Centre of Excellence, PO Box 1000, 02044 VTT, Finland
²Asqella Oy, Kutomotie 18, 00380 Helsinki, Finland
E-mail: Juha.Hassel@vtt.fi

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Abstract. Kinetic inductance bolometer (KIB) technology is a candidate for passive sub-millimeter wave and terahertz imaging systems. Its benefits include scalability into large 2D arrays and operation with intermediate cryogenics in the temperature range of 5 - 10 K. We have previously demonstrated basic figures of merit compatible with typical requirements in security screening applications, and demonstrated the scalability into large arrays in terms of device fabrication, optics integration, and cryogenics. In this article, we address the last missing ingredient, the readout. The concept, serial addressed frequency excitation (SAFE), is an alternative to full frequency-division multiplexing at microwave frequencies conventionally used to read out kinetic inductance detectors. We introduce the concept, and analyze the criteria of the multiplexed readout avoiding the degradation of the signal-to-noise ratio in the presence of a thermal anti-alias filter inherent to thermal detectors. We also present radiometric detection experiments demonstrating the noise properties. Furthermore, we present a practical scalable realization of a readout system integrated into a prototype imager with 8712 detectors, and demonstrate working progress by detection with up to 456 detectors. We also present practical detection experiments demonstrating the capacity of SAFE in concealed object detection.
1. Introduction

Superconducting detectors have an established position in submillimeter wave, terahertz, and far infrared band radiometric imaging systems. The state of art in the detector integration level is today defined by superconducting transition edge sensors (TESs) [1] and kinetic inductance detectors (KIDs) [2, 3] used for the application of astronomical imaging. Detector arrays with up to 10000 sensing elements have been demonstrated [4, 5]. A further application of submillimeter-wave radiometry is security screening which relies on good penetration of radiation through dielectric materials, and a typical spatial resolution in the centimeter range as fundamentally limited by diffraction [6, 7, 8]. This enables the detection of concealed objects under the clothing. The integration level of such sub-millimeter wave security imagers has to date been limited to a maximum of about 200 sensing elements per system. Larger arrays would be beneficial in terms of system-level figures of merit such as spatial resolution, radiometric contrast, field of view, and mechanical simplicity. One limiting aspect in the scaling is the readout. This is emphasized in security screening applications in which the simplicity and cost-effectiveness are important factors. For TES-based systems, established solutions include rather involved SQUID-based readouts using either time-domain or frequency-domain multiplexing [10]. The readout of large KID arrays using RF or microwave frequency multiplexing typically requires state-of-art high-speed digital electronics operating at gigahertz sampling rates [9, 11].

We have previously introduced the detector technology based on kinetic inductance bolometers (KIBs) [12]. In brief, KIBs are thermal detectors based on sensing the change of the temperature-dependent kinetic inductance on a thermally confined membrane that is heated by the signal absorbed on the membrane. Compared to non-equilibrium KIDs and TESs relying in sub-Kelvin cryogenics, KIBs operate in the temperature range of 5 – 10 K. We have previously demonstrated the scalability into kilo-pixel arrays, and performance in line with the requirements of radiometric security imaging [13]. In this paper, we concentrate on a readout concept suitable for the KIB arrays.

Kinetic inductance detectors and sensors are inherently compatible with RF or microwave multiplexing, enabling separate resonance tuning of each sensor element with an individual readout frequency [2, 6, 14]. The readout is typically performed with frequency-division multiplexing addressing a number of detectors in parallel with a frequency comb, usually at microwave frequencies. In these approaches, the comb generation and demodulation require fast digital electronics. Here we introduce an alternative method, serial addressed frequency excitation (SAFE), which substantially simplifies the readout system. After introducing the concept, we discuss the criteria to avoid potential multiplexing-induced degradation of the signal-to-noise ratio (SNR), and present experimental results from the KIB-readout supporting the theory. We also describe an implementation of the electronics integrated into an imaging system, and show radiometric detection results from the simultaneous readout of >400 detectors.

2. The readout concept

2.1. Basic functionality

A readout-band schematic of a KIB array with N detectors corresponding to one readout channel is shown in Fig. 1(a). For KIBs, the signal at $i$:th bolometer induces a change in its temperature $T_i$, and causes a shift in the inductance $L(T_i)$. This in turn shifts the resonant frequency $f_i$ of the resonator formed by $L(T_i)$ and the shunt capacitance $C_i$. The coupling capacitors $C_{ci}$ are used to match the resonator into the readout line. In this configuration each detector within an array is probed by an individual readout tone at or near $f_i$. Figure 1(b) shows measured readout band transmission from one channel with $N=132$ detectors. The transmission minima represent the resonances. Figure 1(c) shows a simplified schematic of SAFE. During frame time $\tau_F$ all $N$ detectors are addressed, each within time slot $\tau_S = \tau_F/N$. After passing through the detector array, the modulated RF signal is amplified and demodulated down to the baseband with a mixer using the unmodulated readout tone as the local oscillator. The demodulation output contains the time-multiplexed detector signal, which is then analog-to-digital (A/D) converted at the sampling rate of $f_S = 1/\tau_S$. In this approach, a benefit is that the high-speed digital electronics is limited to a controllable RF source with a sinusoidal output. The digital signal processing requirements are essentially reduced down to the data rate of $f_S$ instead of the full readout band.

2.2. Noise dynamics

The risk in SAFE is a potential noise penalty due to the limited integration time per detector which is a factor $1/N$ of the non-multiplexed continuous readout, or a fully parallel readout. In case of wide-band white noise the penalty in the root mean square SNR for $N$ detectors would be $N^{1/2}$. If, however, the noise is correlated the situation is different. For example, band-limited white noise with low-pass cutoff at $f_c$ is correlated at time scales shorter than $\sim 1/f_c$. The penalty should be thus completely avoided for $\tau_F \ll 1/f_c$. This is quantified through simulations...
in Appendix A for limiting and intermediate cases of \( f_c/f_r \) (with frame rate \( f_r = 1/\tau_r \)), and for the different number of detectors in a readout channel.

For bolometers, the fundamental noise mechanism is the phonon noise \[15\]. It is characteristically bandwidth-limited white noise with the low-pass thermal cutoff at \( f_c = 1/(2\pi\tau_{th}) \). Here the thermal time constant is \( \tau_{th} = c/G \) with \( c \) the heat capacity of the bolometer thermal volume, and \( G \) the thermal conductivity from the volume to the thermal bath. In our practical devices \( f_c \) is in the range of tens of Hz up to over 100 Hz. Referring to Fig. A1, the noise penalty for a typical detector count \( N > 100 \) per readout channel with \( f_r/f_c = 2 \) is only about 30\% at maximum. Thus, frame rates in the order of \( >100 \) Hz suffice to avoid any significant noise penalty, and the data rates \( f_S \) corresponding to all detectors in a readout channel reside in the audio band. It can be noted that in the final data representation, such as image formation, the demultiplexed single detector signals, initially sampled at \( f_r \) are typically post-integrated into a lower rate.

A further criterion is that one needs to read out the resonator state within time slot \( \tau_S \). The prerequisite is that the resonator is electrically sufficiently fast, i.e., the electrical time constant \( \tau_{el} = Q_t/\omega f_i \) needs to be shorter than \( \tau_S \). Here \( Q_t \) is the total quality factor of the KIB resonator including the intrinsic losses and those from the coupling to the readout line \[12\]. For our KIBs, \( \tau_{el} \) is in the order of \( \mu s \), i.e. sufficiently faster than the audio-band sampling requires.

In a practical case there are often also other noise mechanisms of some significance. For KIBs, the electronics noise needs to be accounted for in some cases. It is typically wide-band white noise, thus, if dominant, giving rise to the worst-case multiplexing penalty. If the electronics noise is represented as the voltage noise spectral density \( S_V^{1/2} \) referred to the RF preamplifier input, it becomes dominant in multiplexing if \( S_V^{1/2} N^{1/2}/\Re \) exceeds the phonon noise \( \text{NEP}_{ph} \). Here \( \Re \) is the voltage responsivity of the detector. If this is the case it can be mitigated by increasing \( \Re \). For KIBs this can be in principle achieved by increasing the amplitude of the excitation tone. However, the maximum excitation is limited by the thermal nonlinearities due to the self-heating effects induced by the excitation \[12\]. For the presented multiplexing scheme the average heating power is, however, a factor \( 1/N \) of the non-multiplexed case if the excitation amplitude is held constant. If, further, the frame rate \( f_r \) sufficiently exceeds the thermal cutoff the self-heating is time averaged, and \( N \) times higher excitation power, or \( N^{1/2} \) times higher excitation...
voltage can be applied in the multiplexing. As the responsivity is proportional to the excitation voltage $v_{\text{ext}}$, this ideally compensates for the multiplexing penalty caused by the electronics noise.

3. Implementation

A block diagram of the electronics is shown in Fig. 2(a), along with the physical realization in Fig. 2(b). The system PC controls the microprocessor electronics, which is responsible for the real-time control of the readout sequence and analog-to-digital (A/D) conversion. The microprocessor electronics controls the channel electronics block through serial peripheral interface (SPI) bus. In this implementation, we have one microprocessor controlling up to 13 readout channels. The channel electronics hosts the main functionalities of the readout-sequence, i.e., RF excitation, RF preamplification, and demodulation. These are illustrated in Fig. 2(c). The RF excitation is performed by a direct digital synthesizer (DDS) chip enabling programming the excitation sequences by defining the frequency for each time slot $\tau_S$. The excitation signal first passes a digitally programmable attenuator, which defines its amplitude. After passing through the detector array the modulated RF signal is first amplified by a low-noise preamplifier. Thanks to the relatively high responsivity of the KIBs, the amplifier cascade can be entirely contained at room temperature, with the first-stage amplifier having a nominal noise temperature of 35 K. The modulated RF signal is then demodulated by an IQ mixer, generating the in-phase (I) and quadrature (Q) baseband components. The phase components contain the full information of the detector signal. However, for a given operating point, a single phase component with an optimal reference phase is sufficient for obtaining the detector signal. To halve the number of the required A/D channels, we have further implemented an analog summing circuit producing an output voltage proportional to the weighted sum of the voltages corresponding to the two phase components. As the relative weights are programmable, this process effectively corresponds to the post-selection of the phase reference. The output of the summing circuit contains the time-domain multiplexed (TDM) detector signals. To ensure the optimal integration of the signal with a negligible dead time, this signal is fed to a sample-and-hold (S/H) circuit [16] integrating the signal across each time slot $\tau_S$, and feeds the output to the A/D converter. The digitized TDM signal is then fed to the system PC responsible of demultiplexing and post-processing the data.

The above description corresponds to the full multi-channel readout electronics. It needs to be noted that the experiments considering proof-of-principle noise dynamics (Section 4.1) are performed with a more limited setup excluding the full microprocessor electronics and phase summing circuit. The RF excitation circuit is in this case based on a field-programmable gate array (FPGA) instead of the DDS. The preamplifier and demodulation circuits are, however, in this case essentially similar to the full system.

4. Experiments

4.1. Signal-to-noise ratio in multiplexing

We first study the noise dynamics with an experiment emulating this aspect of SAFE. The measurement platform is the setup reported in [13]. The measurement consists of recording the radiometric signal from a blackbody calibrator [17] ($T \approx 40^\circ\text{C}$) located at the image plane with 5 m standoff from the imager. Instead of performing full multiplexing, we observe here a single detector in a setting equivalent from the noise dynamics point of view: in analogy to Fig. 1(c) the signal is formed by running excitation pulses of length $\tau_S$ repeated at rate $f_F$. The IF output corresponding to each $\tau_S$ is integrated to form a sample, and the signal now sampled at $f_F$ is Fourier
noise properties in SAFE as a function of \(N = \tau_S/\tau_F\). The excitation power \(P_{\text{ext}}\) is increased linearly below \(N = 10\) while the maximum excitation of \(P_{\text{ext}} = 85 \text{ nW}\) is used for higher \(N\). The dashed line represents the ideal case with no multiplexing penalty while the solid line is the worst-case with \(N^{1/2}\) dependence expected for the fully non-correlated noise.

![Figure 3](image)

**Figure 3.** Signal-to-noise ratio in an experiment emulating noise properties in SAFE as function of \(N = \tau_S/\tau_F\). The excitation power \(P_{\text{ext}}\) is increased linearly below \(N = 10\) while the maximum excitation of \(P_{\text{ext}} = 85 \text{ nW}\) is used for higher \(N\). The dashed line represents the ideal case with no multiplexing penalty while the solid line is the worst-case with \(N^{1/2}\) dependence expected for the fully non-correlated noise.

The signal and noise are recorded at the frequency of the optical chopper, which is 25 Hz here. The result with varying slot time \(\tau_S\) and constant \(f_F \approx 120 \text{ Hz}\) is shown in Fig. 3 as a function of "multiplexing factor" \(N = \tau_F/\tau_S\). In accordance to the discussion in Section 2.2., the excitation power is increased linearly with \(N\) for \(N < 10\). For this system, however, the maximum available excitation power was 85 nW which was used for \(N > 10\). It is observed that SNR decreases by up to a factor of two for the maximum of \(N = 128\) in comparison to the formally non-multiplexed case with \(N = 1\). We can anticipate the reason to be that the electronics noise becomes dominant with high \(N\) as the excitation power is limited. However, even for the largest \(N\) the SNR still remains about a factor of 5 higher in comparison to the worst-case degradation scaling as \(N^{-1/2}\).

### 4.2. Calibration and Radiometric Detection

Next we consider detector calibration aspects and radiometric detection results in the context of two imaging systems utilizing SAFE readout concept. These are "CONSORTIS" radiometer whose basic features have been described elsewhere [18, 19, 20], and "LASTKID" system, which is a fully staring KIB-based imager with 8712 detector elements currently under development. From the readout point of view, the LASTKID detector array is divided into 66 readout channels with 132 detectors in each. The readout band response in Fig. 1(b) is recorded with the LASTKID system.

In a practical readout setting, a requirement is to find the correct RF excitation parameters, in particular the frequency and the phase reference, for each detector. We present here the calibration scheme and its execution in the LASTKID imager. The electronics output voltage \(V\) in a frequency sweep is proportional to the RF transmission in analogy to Fig. 1(b), albeit corresponding to just one phase component. To find the optimal frequency-dependent phase reference we first sweep the excitation frequency through the readout band. The transmission has a sinusoidal envelope due to the propagation delay of the excitation signal, which is depicted in Fig. 4(a). To fix the reference to a constant value at all frequencies, the phase reference post-selection circuit (see Fig. 2(c)) is programmed to compensate for the delay using the phase and frequency offset of the fitted envelope as the calibration. The transmission measurement is repeated with the corrected variable phase reference, and the result is plotted in Fig. 4(b). The responsivities of the detectors are proportional as \(R \propto dV/dL \propto f_i (dV/df)\). Here \(L = L(T_b)\) is the unmodulated thermometer inductance at the bath temperature \(T_b\). The derived responsivity spectrum is plotted in Fig. 4(c), and the peak values are used as frequency calibration points \(f_i\).

It can be noted that the responsivities of the individual detectors have some variation from detector to detector incident from the slight variation of the microwave matching conditions along the array. However, the responsivities as recorded here agree within a factor of \(\sim 2\) which can be considered tolerable.

Figures 5(a) and (b) illustrate a detection experiment with CONSORTIS imager. We note that the dataset is essentially similar to that reported previously in [18]. The test person is hiding an object made of dielectric materials (plastic, paper) under his jacket. The data has been recorded from a single readout channel addressing in this case 35 detectors. It is initially acquired with the frame rate of \(f_F \approx 440 \text{ Hz}\), and post-integrated into the rate of 10 Hz. The concealed object shows as a notch as expected due to its lower radiometric temperature as compared to the test person’s skin. While the demonstration is still limited in terms of scalability it already illustrates the capacity of a KIB-based system with SAFE readout to detect concealed objects. Figure 5(c)-(f) is an illustration of the working progress within LASTKID system to scale up the readout. Figure 5(c) represents the focal plane with the detector area of 100 mm x 200 mm. Currently, 2/3 of the detectors are mounted. Figure 5(d) shows a zoom-up of the detector plane, and Figs. 5(e),(f) illustrate detection results corresponding
to four readout channels with altogether 456 detector elements. In this case the initial frame rate is about 150 Hz, and the data is again post-integrated into the rate of 10 Hz. The full real-time detection data of Figs. 5 (b,e,f) is also presented in the Supplementary material.

5. Discussion and conclusions

In summary, we have presented a multiplexed electronics concept and an implementation designed to read out kinetic inductance bolometer matrices. We indicated that the SAFE readout scheme presented here has beneficial features in comparison to other concepts relying on full FDM. In particular, the method allows a significant simplification as compared to multiplexers relying on high-speed digital electronics operating in the full readout frequency band. The constraints in SAFE are that, in order to avoid the multiplexing penalty, the detector responsivity and noise properties with respect to the sampling rate need to fulfill certain conditions which we reviewed in the context of KIBs. Beyond this, SAFE may be useful for the readout of other types of RF coupled sensors and detectors but the detector requirements need to assessed for each case separately.

We discussed the criteria for avoiding the multiplexing penalty potentially degrading the SNR. This emphasizes the benefit of large detector arrays as an option to optomechanical scanning and a limited number of detectors: in the case of scanning the “scanning penalty” is always analogous to the worst-case multiplexing penalty, i.e. the RMS degradation of SNR is proportional to $N^{1/2}$. In this case $N$ represents the number of image pixels recorded with a single detector. We showed that this can be partially or completely avoided with large arrays and the SAFE multiplexing scheme. We also presented radiometric detection results from imaging systems indicating performance in practical real-time concealed object detection, and reported working progress towards a
full-scale readout system for ∼ 9000 detectors.

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Appendix A. Multiplexing penalty for band-limited white noise

To quantify the effect of noise penalty in case of band-limited white noise we numerically generate the noise and simulate the sampling process in our multiplexing scheme. The first step is to generate a time series of wide-band white noise as a series of Gaussian distributed random numbers. This is subjected to the first-order numerical low-pass filter with cutoff $f_c$ to impose the band limitation. The effect of multiplexing is simulated by dividing the time series in frames of length $\tau_F$, and integrating the signal within each frame over a slot time $\tau_S = \tau_F/N$. The integrated values form a sequence with sampling rate $1/\tau_F$, corresponding to the raw demultiplexed signal of a single detector. This is Fourier transformed, and the low-frequency limit $f \rightarrow 0$ of the RMS value of the power spectral density $S(f)^{1/2}$ represents the effective noise in the multiplexing scheme. The result is plotted in Fig. A1 as the function of the multiplexing ratio $N$ and frame rate $f_F = 1/\tau_F$.

![Figure A1. Simulated low-frequency limit of effective root-mean-square noise power spectral density $S(0)^{1/2}$ for $N$ detectors multiplexed with frame rate $f_c$. It is assumed here that the dominant noise mechanism is band-limited white noise with high-frequency cutoff at $f_c$. The results are normalized to the non-multiplexed case with $N = 1$. The dashed line represents the worst-case scenario valid for non-correlated noise ($f_c \rightarrow \infty$) proportional to $N^{-1/2}$.](image)

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