An analytical model for electronic noise in a NTD Ge based bolometer readout circuit

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Abstract

This paper presents an analytical model to quantify noise in a bolometer readout circuit. A frequency domain analysis of the noise model is presented which includes the effect of noise from the bias resistor, sensor resistor, voltage and current noise of amplifier and cable capacitance. The analytical model is initially verified by using several standard SMD resistors as a sensor in the range of 0.1 – 100 MΩ and measuring the RMS noise of the bolometer readout circuit. Noise measurement on several indigenously developed neutron transmutation doped Ge temperature sensor has been carried out over a temperature range of 20 – 70 mK and the measured data is compared with the noise calculated using analytical model. The effect of different sensor resistances on the noise of bolometer readout circuit, in line with the analytical model and measured data, is presented in this paper.

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I. INTRODUCTION

A cryogenic bolometer detector with neutron transmutation doped (NTD) Ge thermistor offers an excellent energy resolution and high sensitivity, which makes it an ideal choice for several rare decay experiments such as neutrinoless double beta decay (NDBD) and dark matter searches\textsuperscript{1,2}. Ideally the performance of a bolometer detector is expected to be superior due to its good intrinsic resolution\textsuperscript{3}. However, in practice, resolution is limited by the noise contributed from the external sources. One of the major external factor is the noise contributed due to the electrical and electronic components in the bolometer readout circuit. The readout circuit consists of an NTD Ge sensor at mK temperature, a constant current source for sensor biasing (usually generated by applying a constant voltage through a high bias resistor), a front-end preamplifier and higher stage amplification, and the capacitance associated with the connecting wires and cables. In order to understand the effect of noise from each of these sources and to predict the noise for a given readout circuit, a thorough noise analysis of the bolometer readout circuit is required. The analytical noise model will also provide key information for the design and optimization of low noise readout circuitry for bolometer.

A tin cryogenic bolometer detector (\textit{TIN.TIN}–The INdia-based TIN detector)\textsuperscript{4} is being developed to search for NDBD process in \textsuperscript{124}Sn in the upcoming India-based Neutrino Observatory (INO). At Tata Institute of Fundamental Research, Mumbai, a prototype bolometer test setup with a sapphire absorber and indigenously fabricated NTD Ge sensor\textsuperscript{5} is developed for initial testing in a cryogen-free dilution refrigerator CFDR-1200 (Leiden Cryogenics)\textsuperscript{6}. The detailed noise measurement of an NTD Ge sensor and the effect of external noise pickups on the performance of the bolometer is already reported in Ref. 7. In this article, we present an analytical model to predict the RMS noise in a bolometer readout circuit. The proposed model is initially verified by taking noise measurements at 300 K on a bolometer readout circuit consisting of standard SMD resistors as a sensor and load capacitor. A test setup with four indigenously fabricated NTD Ge sensors of size 1 mm × 1 mm is developed and noise measurements were taken in a temperature range of 20 – 70 mK. The effect of cable capacitance and different sensor resistances on the measured noise as well as the shape of baseline noise is studied in line with the proposed noise model.
II. ANALYTICAL MODEL FOR BOLOMETER NOISE

FIG. 1. (a) Bolometer readout circuit (b) its equivalent noise model. Here, $R_L$: Bias resistor, $R_S$: NTD Ge resistor, $C_L$: Cable capacitance, $V_S$: Sensor bias voltage, $e_{n,RL}$: Thermal noise of $R_L$, $e_{n,RS}$: Thermal noise of $R_S$, $i_{n,a}$: Input current noise density of amplifier, $e_{n,a}$: Input voltage noise density of amplifier.

The bolometer readout circuit and its equivalent noise model is shown in Fig. 1. A pseudoconstant current source, consisting of a voltage source $V_S$ in series with a high bias resistor $R_L$, is used for sensor biasing. Here, we have not considered the noise from the sensor bias voltage $V_S$ as its thermal noise voltage (~ 60 nV/√Hz) is negligible as compared to the thermal noise of bias resistor $R_L$ (~ 20 GΩ, equivalent to a thermal noise voltage of 18.2 µV/√Hz) at room temperature. The thermal noise for bias resistor $R_L$ and sensor resistor $R_S$ is defined as

$$e_{n,RL} = \sqrt{4kT_L BR_L}$$  \hspace{1cm} (1)  

$$e_{n,RS} = \sqrt{4kT_S BR_S}$$  \hspace{1cm} (2)  

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where, \( T_L \) and \( T_S \) are temperatures (in Kelvin) of bias resistor and sensor resistor, respectively, \( B \) is bandwidth in Hz and \( k \) is Boltzmann constant. The contribution of noise at amplifier’s input from each of the sources can be calculated separately by switching off the remaining noise sources.

(a) Noise contribution (\( e_L \)) at amplifier’s input due to bias resistor \( R_L \):

\[
\begin{align*}
|e_L| &= \sqrt{4kT_L BR_L} \frac{R_S}{\sqrt{(R_S + R_L)^2 + (\omega R_S R_L C_L)^2}} \\
R_L &= \left[ R_S / (R_S + R_L) \right] + \frac{1}{j \omega C_L} \\
e_L &= e_{n,RL} \frac{R_L / (1 / j \omega C_L)}{R_L + \left[ R_S / (1 / j \omega C_L) \right]} = \sqrt{4kT_L BR_L} \frac{R_S}{R_S + R_L + j \omega R_S R_L C_L} \\
\end{align*}
\]

Eq. 3 depicts the behavior of a low pass filter (LPF). It can be seen from Eq. 3 that only a small fraction \([R_S / (R_S + R_L)]\) of thermal noise from \( R_L \) (\( e_{n,RL} \)) reaches the amplifier’s input at \( \omega = 0 \). For \( R_L = 20 \, \text{GΩ} \) at 300 K \( (e_{n,RL} = 18.2 \, \mu\text{V/√Hz}) \) and \( R_S = 500 \, \text{MΩ} \) (typical resistance of NTD Ge sensor, which cools to around \( T_S = 35 \, \text{mK} \) in our case), the contribution of thermal noise from \( R_L \) at the amplifier’s input is \( e_L = 0.44 \, \mu\text{V/√Hz} \), which is still a large value. The noise contribution \( e_L \) further reduces as the frequency is increased for a finite cable capacitance \( C_L \).

(b) Noise contribution (\( e_S \)) at amplifier’s input due to sensor resistor \( R_S \):

\[
\begin{align*}
|e_S| &= \sqrt{4kT_S BR_S} \frac{R_L}{\sqrt{(R_S + R_L)^2 + (\omega R_S R_L C_L)^2}} \\
R_S &= \left[ R_L / (R_S + R_L) \right] + \frac{1}{j \omega C_L} \\
e_S &= e_{n,RS} \frac{R_L / (1 / j \omega C_L)}{R_L + \left[ R_S / (1 / j \omega C_L) \right]} = \sqrt{4kT_S BR_S} \frac{R_L}{R_S + R_L + j \omega R_S R_L C_L} \\
\end{align*}
\]

Eq. 4 also depicts the behavior of an LPF. It can be seen from Eq. 4 that almost unity fraction \([R_L / (R_S + R_L)]\) of thermal noise from \( R_S \) (\( e_{n,RS} \)) reaches the amplifier’s input at \( \omega = 0 \). For \( R_L = 20 \, \text{GΩ} \) at 300 K and \( R_S = 500 \, \text{MΩ} \) at \( T_S = 35 \, \text{mK} \) \( (e_{n,RS} = 33.23 \, \text{nV/√Hz}) \), the contribution of thermal noise from \( R_S \) at the amplifier’s input is \( e_S = 30.32 \, \text{nV/√Hz} \). The noise contribution \( e_S \) further reduces as the frequency is increased for a finite cable capacitance \( C_L \).

(c) Noise contribution (\( e_{n,i} \)) at amplifier’s input due to input current noise density \( i_{n,a} \):
\[ e_{n,i} = i_{n,a} \left[ R_{eq} \parallel (1/j\omega C_L) \right] = i_{n,a} \frac{R_{eq}}{1 + j\omega R_{eq} C_L} \]

\[ |e_{n,i}| = i_{n,a} \frac{R_{eq}}{\sqrt{1 + (\omega R_{eq} C_L)^2}} \]  \hspace{1cm} (5)

where, \( R_{eq} = R_s \parallel R_L \). Eq. 5 also shows a behavior of an LPF. At \( \omega = 0 \), \( e_{n,1} = 0.78 \, \mu V/\sqrt{Hz} \) for \( i_{n,a} = 1.6 \, fA/\sqrt{Hz} \).

(d) Noise contribution (\( e_{n,a} \)) at amplifier’s input due to input voltage noise density:

The input voltage noise density of an amplifier (\( e_{n,a} \)) is due to two factors: the first one is a white noise (a frequency independent noise) whereas the other contribution is flicker noise (inversely proportional to frequency). The input voltage noise density (\( e_{n,a} \)) can then be written as,

\[ e_{n,a} = \sqrt{e_{white}^2 + e_{flicker}^2} = e_{white} \sqrt{1 + \frac{f_c}{f_n}} \quad 1 \leq n \leq 2 \]  \hspace{1cm} (6)

where, \( e_{white} = \sqrt{4kT_aBR_A} \), \( T_a \) is temperature of amplifier in Kelvin, \( R_A \) is equivalent noise resistor of amplifier and \( f_c \) is flicker corner frequency. Eq. 6 is a frequency dependent term due to flicker noise. There is strong frequency dependence for \( f \leq f_c \). For higher frequency \( (f > f_c) \), \( e_{n,a} \) coincides with the thermal noise floor (\( e_{white} \)).

Therefore, the total input voltage noise density, \( e_{n,in} \), (referred to input of the amplifier) and the total output voltage noise density, \( e_{n,o} \), (referred to output of the amplifier) can be written as,

\[ e_{n,in} = \sqrt{e_L^2 + e_S^2 + e_{n,i}^2 + e_{n,a}^2} \]  \hspace{1cm} (7)

\[ e_{n,o} = A \sqrt{e_L^2 + e_S^2 + e_{n,i}^2 + e_{n,a}^2} \]  \hspace{1cm} (8)

where \( A \) is voltage gain of amplifier in V/V.

The RMS noise at the output of the amplifier can be extracted from the frequency response of the output voltage noise density \( e_{n,o} \) by integrating Eq. (8) over a given acquisition bandwidth as follows:
where \( f_h - f_i \) is the acquisition bandwidth of noise and \( df \) is the resolution bandwidth. The peak to peak noise can be calculated from the RMS noise as follows:

\[
e_{n,o}(pp) = 6 e_{n,o}(rms) \tag{10}
\]

For peak to peak noise calculation, \( \pm 3 \) standard deviation is taken so that the probability of noise lying between the estimated peak to peak noise is 99.7%. To check the magnitude and nature of noise for different sensor resistances ranging from 10 M\( \Omega \) to 500 M\( \Omega \), peak to peak noise with different acquisition bandwidth is calculated using Eqs. (8), (9) and (10). For noise calculation using the analytical model, we are using the technical specifications of a differential amplifier (Femto DLPVA-100-F) which has a typical noise characteristic of \( e_{\text{white}} = 6.9 \text{nV/} \sqrt{\text{Hz}} \), \( i_{n,a} = 1.6 \text{fA/} \sqrt{\text{Hz}} \), \( f_c = 80 \text{Hz} \) and it is set to a voltage gain of 60 dB (\( A = 1000 \)). Typical value for sensor temperature (\( T_S \)) is taken as 35 mK for the calculation. The acquisition bandwidth and resolution bandwidth for noise is taken as 1 kHz and 0.1 Hz, respectively, assuming the rise time of bolometer pulse is greater than 10 msec. Table I shows the magnitude of noise voltage for different sensor resistances with load capacitance of 50 pF and 1 nF. For a very small load capacitance of 50 pF, the peak to peak output noise voltage, \( e_{n,o}(pp) \), over an acquisition bandwidth of 0.1 – 1000 Hz increases rapidly with the increase in sensor resistance. For higher sensor resistance (\( R_S = 500 \text{ M\( \Omega \)} \)), contribution of low frequency noise below 10 Hz is almost comparable to that in the high frequency region above 10 Hz. However, with a larger load capacitance of 1 nF, low frequency noise in the range of 1 – 100 Hz is heavily suppressed due to high filtering action which results in lower \( e_{n,o}(pp) \) as compared to small load capacitance. We can see from Table I that there is no significant variation in output noise above 10 Hz for a larger load capacitance of 1 nF. For much higher sensor resistance (\( R_S = 500 \text{ M\( \Omega \)} \)), low frequency noise below 10 Hz is the dominant contribution in the output noise of the bolometer. Table II shows the noise contribution from different components of the readout circuit for a sensor resistance of 500 M\( \Omega \) and load capacitance of 1 nF. In the low frequency region below 10 Hz, output voltage noise density is mainly contributed by the noise from the bias resistor \( R_L \) and input current noise.
Table I: Magnitude of noise voltages for different sensor resistances and acquisition bandwidth

| Acquisition bandwidth (Hz) | $e_{n,o}$ (pp) (mV) | $e_{n,i,o}$ (pp) (mV) | $e_{n,a,o}$ (pp) (mV) |
|---------------------------|---------------------|----------------------|----------------------|
|                           | $R_S = 10 \, \text{M}\Omega$ | $R_S = 50 \, \text{M}\Omega$ | $R_S = 100 \, \text{M}\Omega$ | $R_S = 500 \, \text{M}\Omega$ |
|                           | $C_L$ (pF) | $C_L$ (pF) | $C_L$ (pF) | $C_L$ (pF) |
| 0.1-1                     | 0.94       | 1.08       | 1.44       | 5.45       |
|                           | 0.56       | 1.06       | 3.28       | 12.59      |
|                           | 1.15       | 0.62       | 6.12       | 9.86       |
|                           | 2.34       | 3.38       | 3.51       | 3.56       |
|                           | 2.83       | 5.69       | 7.91       | 17.27      |

Table II: Noise contribution from different components for $R_S = 500 \, \text{M}\Omega$ and $C_L = 1 \, \text{nF}$

| Acquisition bandwidth (Hz) | $e_{L,o}$ (pp) (mV) | $e_{S,o}$ (pp) (mV) | $e_{n,i,o}$ (pp) (mV) | $e_{n,a,o}$ (pp) (mV) |
|---------------------------|---------------------|---------------------|----------------------|----------------------|
| 0.1-1                     | 1.61                | 0.11                | 2.83                 | 0.93                 |
| 1.1-10                    | 0.79                | 0.05                | 1.39                 | 0.44                 |
| 10.1-100                  | 0.26                | 0.02                | 0.46                 | 0.46                 |
| 100.1-1000                | 0.08                | 0.01                | 0.14                 | 1.25                 |
| 0.1-1000                  | 1.81                | 0.12                | 3.18                 | 1.68                 |

density of the amplifier. Whereas, in the high frequency region above 100 Hz, input voltage noise density of the amplifier is the dominant noise contribution.

### III. NOISE MEASUREMENT FOR STANDARD RESISTORS

Measurement of RMS noise, $e_{n,o}$(rms), and peak to peak noise, $e_{n,o}$(pp), is taken with a bias resistor, $R_L \sim 20 \, \text{G}\Omega$ (See Fig. 1). Different values of standard SMD (surface mount device)
resistors in the range of 0.1 – 100 MΩ is connected between the terminal ‘a’ and ‘b’ as shown in Fig. 1. Noise measurements are taken at 300 K with default wire capacitance ~ 50 pF and a standard ceramic capacitor of 1 nF. Time domain measurements of output noise, $e_{n,o}$ is carried out using a data acquisition card (DAQ) from National Instruments. A total of 100k samples are acquired in a time window of 10 sec, with a sampling rate of 10 kHz. During noise measurement of each resistor, data in total of six time-windows are acquired, which requires a measurement time of 60 sec. A typical 4 sec noise data for two SMD resistors (0.1 MΩ and 100 MΩ) with two different load capacitance of 50 pF and 1 nF are shown in Fig. 2. The bandwidth of the amplifier is set to 1 kHz during measurements. It is observed from Fig. 2 (d) that the magnitude of low frequency noise below 10 Hz is larger as compared to the noise in the high frequency region for the case of $R_S = 100$ MΩ, $C_L = 1$ nF, which is also expected from the analytical model. Measured peak to peak noise ($e_{pp-exp}$) along with the noise obtained from analytical model ($e_{pp-model}$) for different SMD resistors and load capacitors are given in Table III. The measured noise is found to be in good agreement (within 20%) with the analytical model prediction.

FIG. 2. Typical noise data for standard SMD resistors
Table III: \( e_{\text{pp-expt}} \) and \( e_{\text{pp-model}} \) for standard SMD resistors

| SMD resistor (MΩ) | Default wire capacitance ~ 50 pF | Ceramic capacitance ~ 1 nF |
|-------------------|----------------------------------|---------------------------|
|                   | \( e_{\text{pp-expt}} \) (mV)   | \( e_{\text{pp-model}} \) (mV) | \( e_{\text{pp-expt}} \) (mV) | \( e_{\text{pp-model}} \) (mV) |
| 0.1               | 10.3                             | 7.9                       | 8.2                         | 7.5                         |
| 1                 | 26.4                             | 24.1                      | 11.4                        | 11.7                        |
| 10                | 42.3                             | 49.0                      | 13.1                        | 12.3                        |
| 100               | 53.2                             | 54.5                      | 13.6                        | 12.3                        |

IV. MEASUREMENTS WITH NTD Ge SENSORS

Four indigenously fabricated NTD Ge sensors (M713, M715, M716 and M720) were directly coupled to the copper holder and the test setup is mounted on the mixing chamber stage of the CFDR. These sensors are basically 1 mm × 1 mm NTD Ge sensors with a face-type contact\(^8\). The sensors mounted on the copper holder are shown in Fig. 3. More detailed description of sensor fabrication and its mounting are given in Refs. 5, 8.

FIG. 3. NTD Ge sensors mounted on a copper test setup
A. Resistance measurement of NTD Ge sensor

Resistance measurements for NTD Ge sensors were taken in CFDR-1200 over a temperature range of 20 – 400 mK. The resistance, $R_S$, is measured by applying an excitation current signal and measuring the voltage signal across the sensor as described in Ref. 7. Twisted pair cables are used to bring the sensor signal from mixing chamber (MC) stage to 300 K. For resistance measurement in NI-DAQ card, a square wave excitation $V_S$ is applied to the sensor through a high bias resistor ($R_L = 20 \, \Omega$) and sensor resistance $R_S$ is evaluated from the measured output waveform $V_o$ (See Figure 1a). For measurement of resistance below 2 MΩ, AVS-47B resistance bridge is used. The measured resistance as a function of MC temperature ($T_{MC}$) is plotted in Fig. 4. Resistance measurement for all the sensors, except M716, was done for a temperature range of 20 – 400 mK. Due to 2-wire configuration for M716, its resistance was not measured above 100 mK using AVS resistance bridge. It can be seen from Fig. 4 that all the four sensors follow Mott-behavior, as given by Eq. (11), over the temperature range of 30 – 400 mK. Slight saturation of sensor resistance is observed below a temperature of 30 mK.

\[ R = R_0 \exp \left( \sqrt{\frac{T_0}{T}} \right) \]  

\[ (11) \]

FIG. 4. Measured resistance of NTD Ge sensors as a function of $T_{MC}$
B. Noise measurement of NTD Ge sensor

Noise measurements in the temperature range of 20 – 70 mK were taken for all the four sensors using NI-DAQ card, as described in Section III. For noise measurement of sensors, 12 time-windows, each of 10 sec durations, are recorded. A typical 5 sec noise data for all the sensors at the two extreme temperatures, 20 mK and 70 mK, are shown in Fig. 5 and Fig. 6. It can be seen from Fig. 5 that the low frequency fluctuations override the baseline noise for M713 and M715. These fluctuations are mainly contributed by the noise in the low frequency region below 10 Hz due to high sensor resistance (~ 1 GΩ) and large cable capacitance in the readout circuit, as expected from the analytical model. The low frequency noise component is even higher for other two sensors (M716 and M720) due to their higher resistance (~ 4 GΩ) as compared to M713 and M715. The $e_{pp-expt}$ is also higher, 5.7 mVpp (M716) and 6.5 mVpp (M720), as compared to 4.4 mVpp (M713) and 4.8 mVpp (M715). As the sensor is warmed-up to 70 mK, the low frequency fluctuation disappears below the baseline (See Fig. 6). Due to the lower sensor resistance (< 5 MΩ) at higher temperature of 70 mK, the low frequency noise gets obscured by much larger magnitude of higher frequency noise component. $e_{pp-expt}$ of 2.6 mVpp, 2.7 mVpp, 3.1 mVpp and 2.8 mVpp are obtained for M713, M715, M716 and M720, respectively.

![Typical noise data for NTD Ge sensors at 20 mK](image)

FIG. 5. Typical noise data for NTD Ge sensors at 20 mK
$e_{pp\text{-model}}$ is also derived from the analytical model and it is plotted in Fig. 7 along with the $e_{pp\text{-expt}}$ for all the four sensors. The calculation of $e_{pp\text{-model}}$ requires the knowledge of load capacitance $C_L$, which is estimated by measuring the RC time constant of the output square waveform ($V_o$) for all the sensors at 20 mK. Since the sensors are strongly coupled with the Copper holder, the thermal time constant has been neglected in the present case. For M713, $C_L$ comes out to be in the range of 1.67 – 1.72 nF and a mean load of 1.7 nF has been taken for further calculation.
Similarly, the mean cable capacitance is derived from the RC time constant for other sensors, which comes out to be in the range of 0.95 – 1.15 nF. It can be observed from Fig. 7a and Fig. 7b that \( e_{pp-expt} \) and \( e_{pp-model} \) for sensors M713 and M715 almost saturates and becomes independent of sensor temperatures above 40 mK, where resistance value falls below 10 M\( \Omega \). This is due to the fact that the noise contributions from \( e_L \) and \( e_{n,i} \) are significantly less as compared to that from \( e_{n,a} \), which is sensor independent for the lower sensor resistance. Except at the lowest temperature of 20 mK, where the noise voltage is high (~ 4.5 – 5 mVpp), there is a residual excess of ~ 1 – 1.5 mVpp between \( e_{pp-expt} \) and \( e_{pp-model} \). This may be due to the fact that external noises from various system components such as vacuum pumps and gauges in the system, diagnostic and control electronics of the cryostat etc. get induced on the sensor readout circuit and increases the output voltage noise. For sensors M716 and M720 (Fig. 7c and Fig. 7d), \( e_{pp-expt} \) of ~ 6 mVpp is obtained at 20 mK due to very high sensor resistances. The output voltage noise is sensitive to temperature variations up to 50 mK, as compared to other two sensors (M713 and M715) with relatively lower resistances. In the case of sensor M716 (See Fig. 7c), \( e_{pp-model} \) is larger as compared to \( e_{pp-expt} \) below 25 mK. It has been observed from the analytical model that the peak to peak voltage noise is very sensitive to the load capacitance for higher values of sensor resistance. The range of \( C_L \) for M716, estimated from the RC time constant of different pulses, was between 0.86 – 1.04 nF. The \( e_{pp-model} \), derived from the analytical model, will be between 6.6 – 7.8 mVpp for the given range of \( C_L \). Therefore, the reason for over prediction of output noise in case of M716 can be attributed to the 10 – 20 % error in estimating the load capacitance from the RC time constant.

V. CONCLUSION

In this paper, noise in a bolometer readout circuit has been modelled and expressed analytically to understand the effect of noise from different circuit components. The analytical model is initially verified by measuring the noise on standard SMD resistors in the range of 0.1 – 100 M\( \Omega \). It is found that the measured noise agrees well with the noise derived from the analytical model within 20%. Noise measurement is also carried out on four indigenously fabricated NTD Ge sensors which are mounted on a copper holder. The sensor setup is cooled down to 20 mK in CFDR-1200 and noise is measured in the temperature range of 20 – 70 mK. It is observed that for a sensor temperature of 30 mK or above, the measured noise for all the
sensors, except M716, is higher as compared to the noise derived from the analytical model. This may be due to the additional contributions arising from the various sources like microphonics, vibrational, external EMI pickups etc., which have not been incorporated in the model. It is shown that the present analytical model is able to predict the nature of peak to peak voltage noise with respect to sensor temperature and its resistance, cable capacitance and various components in a bolometer readout circuit. The analytical model is helpful in highlighting the major sources of noise in the biasing and measurement circuitry and will be critical during optimization of the noise performance.

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