Optimization of 1/f noise measurement setup

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A dissertation submitted for the partial fulfilment of BS-MS dual degree in Science

Indian Institute of Science Education and Research Mohali

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Certificate of Examination

This is to certify that the dissertation titled **Optimization of 1/f noise measurement setup** submitted by **Taseng Mancheykhun** (Reg. No. MS13086) for the partial fulfillment of BS-MS dual degree programme of the Institute, has been examined by the thesis committee duly appointed by the Institute. The committee finds the work done by the candidate satisfactory and recommends that the report be accepted.

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Dated: April 19, 2018

Declaration

The work presented in this dissertation has been carried out by me under the guidance of Dr. Ananth Venkatesan at the Indian Institute of Science Education and Research Mohali.

This work has not been submitted in part or in full for a degree, a diploma, or a fellowship to any other university or institute. Whenever contributions of others are involved, every effort is made to indicate this clearly, with due acknowledgement of collaborative research and discussions. This thesis is a bonafide record of original work done by me and all sources listed within have been detailed in the bibliography.

Taseng Mancheykhun
(Candidate)

Dated: April 19, 2018

In my capacity as the supervisor of the candidates project work, I certify that the above statements by the candidate are true to the best of my knowledge.

Dr. Ananth Venkatesan
(Supervisor)

Dated: April 19, 2018

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Taseng Mancheykhun

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Abstract

1/f noise has been observed in almost every material. It is the least understood type of noise. There is no general model and theory that explains 1/f noise in different materials, although there are models that describe certain observations or measurements in specific materials. In this project, I have tried to experimentally measure 1/f noise in resistors like metal film resistor and wire-wound resistor, using wheatstone bridge implementation and 0-90^o subtraction technique. I have shown the measurement improvement of the experimental setup with wheatstone bridge implementation for frequency as low as 15mHz as compared to measurement using four-probe and OFFSET feature of the lock-in amplifier.

Contents

Chapter 1

Introduction

1.1 Motivation

Electrical fluctuations are fluctuations in any electrical quantity (voltage, current, etc) that exist in an electrical transport measurement. The resistance of any resistive sample fluctuates about a mean value resulting in voltage fluctuations of the order of 0.5 nV-5 nV across the sample. These fluctuations in the resistance of the sample is called as noise. In a noise measurement, these fluctuations are recorded in time domain and a Fourier spectrum analysis gives the information of various frequency components in the fluctuations. Power Spectral Density is used to quantify the noise. There are primarily three types of noise classified on the basis of the Power Spectral Density: Johnson noise, 1/f noise and Shot noise.

1/f noise is low frequency noise for which the Power Spectral density is inversely proportional to the frequency. Its magnitude is high at low frequencies and decreases in higher frequency, hence the name 1/f. 1/f noise has been observed not only in electronics but also in music, biology, economics, etc. The sources of 1/f noise are still widely debated and much research is still being done in this area. The study of electrical noise is also important from the point of view of applications. This noise interferes with low frequency measurement and can be a limit to achieving the best performance in precision measurement applications. There are various techniques to reduce or eliminate this noise.

1/f Noise measurement is a highly sensitive and challenging measurement. Because every electronics has their own 1/f noise which shows up in the final result, the experimental circuit needs to be very carefully setup to eliminate all other unwanted noise and get the desired 1/f noise of the sample.

1.2 Power spectrum density (PSD)

Power spectral density is used to quantify noise[5]. The fluctuating quantity (like electrical voltage or current) is obtained in time domain. Using fourier transform, it is converted to frequency domain. The time average of the amplitude squared of the fourier transformed signal gives the Power spectral density (PSD).

$$PSD = \lim_{T\to\infty}\left(\frac{1}{2T}\right)\left|\int_{-T}^{T}\delta V e^{-2\pi i f t}dt\right|^2$$

Its SI unit is $\frac{V^2}{Hz}$.

1.3 Types of noise

Resistance fluctuations or noise can be classified primarily into three types on the basis on the nature of their Power spectral density and dependence on the type of biasing on the sample:[1]

1. **Johnson noise (or) Nyquist noise (or) Thermal noise** : This electrical noise is independent of frequency and hence, is a white noise. It is said to occur due to brownian motion of the electrons. Given by Nyquist Theorem[6], the Power Spectral density of this noise in a sample of resistance R at a temperature T is given by:

   $$PSD = 4K_bTR$$

   where K_b is Boltzmann Constant which has value $1.3806\times10^{-23}\ m^2\text{Kg}\ s^{-2}K^{-1}$. This noise depends only on the resistance R and the temperature T of the sample. It has no dependence on size and type of the sample. Since the PSD of Johnson noise can be calculated theoretically using the above formula, it is used to check the accuracy of the noise measurement system.

2. **Shot noise** : This noise is also white in nature. Its power spectral density is given by Schottky formula,

   $$PSD = 2e\langle I\rangle$$

This noise occurs due to the discrete nature of the motion of electrical charge carriers. It appears as fluctuating current. It is an indicator of the magnitude of charge carriers in the sample. It is the most challenging noise to measure because it is visible only at low temperatures when the Johnson noise is very low.

3. **1/f noise or Flicker noise** : Almost every resistive sample through which a current is passing through exhibits voltage fluctuations whose power spectral density goes as $1/f^{\alpha}$, the value of α lying between 0.8 - 1.2 . This type of noise is referred to generally as '1/f noise'.
   The power spectral density (PSD) of 1/f noise is found to follow an empirical formula called Hooge's empirical formula[3]:

$$PSD = \frac{\gamma V^2}{N_c f^{\alpha}}$$

   where, γ is an empirical constant;
   V is the applied bias across the sample;
   N_c is the number of charge carriers in the sample;
   α - the exponent of frequency (f) lies between 0.8-1.2 .

   1/f noise is the least understood noise. It is special because the spectral form of this noise is similar for a wide range of materials and that it is scale invariant over a very large range of frequency.

Chapter 2

1/f noise - Algebra and Theory

2.1 Algebra of 1/f noise

As defined in the last chapter, if the fluctuating quantity is voltage which is recorded in the time domain, the Power Spectral density is

$$PSD = \lim_{T\to\infty} \left(\frac{1}{2T}\right) \left| \int_{-T}^{T} \delta V(t) e^{-2\pi i f t} dt \right|^2$$

where δV is the fluctuation in the voltage.

According to Wiener-Khinchine theorem [6], power spectral density is the Fourier transform of the autocorrelation function C(τ) defined as[5]:

$$C(\tau) = \lim_{T\to\infty} \left(\frac{1}{2T}\right) \int_{-T}^{T} \delta V(t+\tau) \delta V(t) dt$$

PSD becomes frequency dependent because there is a finite relaxation time associated with the fluctuating variable. This relaxation can be quantified by the relaxation function $\phi(t)$.
The power spectral density therefore becomes,

$$PSD = C \int_{-\infty}^{\infty} \phi(t) e^{-i2\pi f t} dt$$

If the relaxation function is the Debye relaxation function with a single relaxation time τ so that $\phi(t) \propto e^{-t/\tau}$, the power spectral density becomes a lorentzian :

$$PSD \propto \frac{2\tau}{1 + (2\pi f \tau)^2}$$

If this function is plotted, PSD as a function of 'f' for a given τ, at the low frequency end i.e for $f << \tau^{-1}$, PSD is constant and at the high frequency end $f >> \tau^{-1}$, $PSD \propto 1/f^2$.

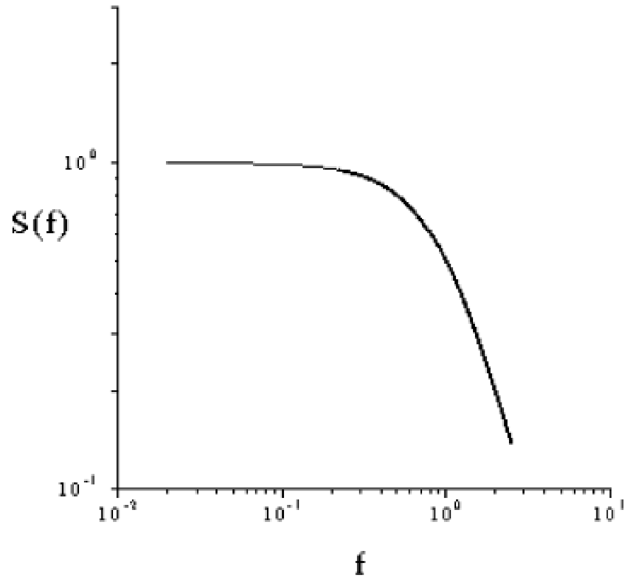


Figure 2.1: Lorentzian power spectra of a single fluctuator (1/τ=1 Hz)[5]

In presence of a distribution of relaxation time $F(\tau)$, the PSD is given as

$$PSD \propto \int_0^\infty F(\tau) \frac{2\tau}{1 + (2\pi f \tau)^2} d\tau$$

If the power spectral density is plotted as a function of frequency, in presence of many Lorentzians each with different relaxation time τ, the superposition of all the lorentzian gives a 1/f type power spectrum.

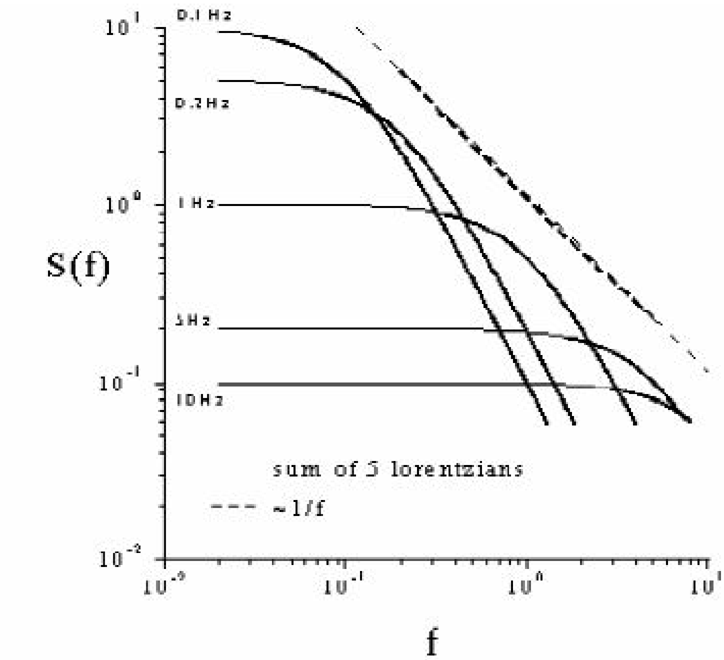


Figure 2.2: Superposition of 5 Lorentzians that generates a power spectrum of the type $1/f^{\alpha}$.[5]

2.2 Theoretical model for 1/f noise

There is no general theory of 1/f noise. But there are various theories/models that has been proposed to explain 1/f noise in specific materials that explain certain observations and measurements.

2.2.1 Hooge's model

This model[3] assumes that the noise arises from independent motion of N_c electrons from the whole bulk of the solid. According to this model, Hooge proposed an empirical formula where the power spectral density is given as

$$PSD = \frac{\gamma V^2}{N_c f^{\alpha}}$$

where V is the applied bias across the sample ; N_c is the number of charge carriers in the sample ; α - the exponent of frequency (f) that lies between 0.8-1.2 in many materials.

γ is an empirical constant which is dimensionless if PSD is expressed in the unit

V^2/Hz. It is not a universal constant and depends on the purity of the sample material and presence of point as well as extended defects.

The noise in a solid depends on its quality and purity. The characterization of materials on the basis of 1/f noise is based on two ruling paradigms[5]:

1. The 1/f noise is a manifestation of the resistance (R) or conductance (G) fluctuation so that the relative mean square voltage fluctuation,

   $$\langle(\delta v(t)^2)\rangle/\langle v\rangle^2 = \langle(\delta G(t)^2)\rangle/\langle G\rangle^2$$

   This has been experimentally established.[8]

2. The noise has a defect origin, which has been experimentally established, by experiments on metal films [2] as well as on semiconductors[4].

Most of the current theories/models that can be used for applications in materials science are based on the above two paradigms.

Chapter 3

1/f noise measurement techniques

The first thing to establish in 1/f noise measurement is to ensure that the 1/f noise originates from the sample. Achieving this is also the most difficult part of the measurement because many of the associated measuring electronics also have 1/f noise.

3.1 DC-biased four-probe method

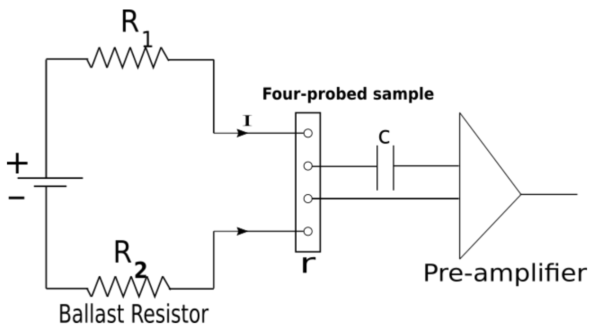


Figure 3.1: Schematic diagram showing four probe DC measurement setup

In this method, the sample is DC biased[7]. A four-probe configuration is used where the biasing current is applied through two separate leads and the steady voltage drop across the sample is blocked off by a capacitor bank forming a high pass filter. The fluctuating voltage across the sample is amplified using a low-noise preamplifier.

In order to separate out power spectrum due to background, the data are taken with zero bias current and then power spectrum is measured with a finite dc bias.The difference gives the noise due to the system under measurement.

3.1.1 Drawbacks

The drawbacks of doing four-probe dc noise measurement are several[7].

1. The use of capacitor in the circuit makes a high pass filter which limits the frequency to above $R_A C$ knee of the capacitor C and pre-amplifier input impedance R_A, hence making low frequency measurement not possible.

2. Because of use of dc power source (battery), the fluctuations in the battery voltage show up directly in the noise measurement.

3. The background noise is dominated by pre-amplifier's 1/f noise which greatly exceeds the intrinsic Johnson noise hence making low-frequency measurement very difficult.

To be able to measure low-frequency 1/f noise, improvement in the measurement technique is required. One very effective improvement is by implementing a Wheatstone bridge circuit and AC biasing.

3.2 DC biased - Wheatstone bridge circuit

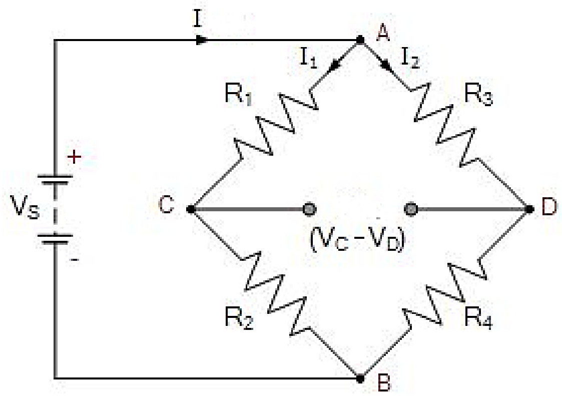


Figure 3.2: A wheatstone bridge circuit

A wheatstone bridge circuit consists of a combination of four resistors connected as shown in Figure 3-2. The bridge is said to be balanced when the voltage drop across the points C and D is 0. This balancing happens when,

$$\frac{R_1}{R_3} = \frac{R_2}{R_4}$$

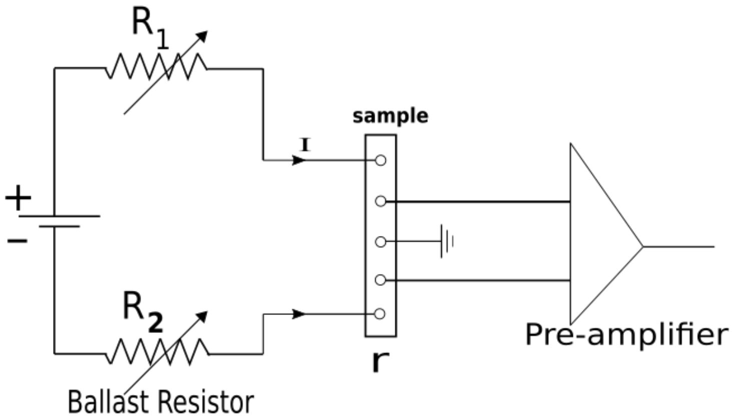


Figure 3.3: Wheatstone bridge circuit implementation for noise measurement[7]

Wheatstone bridge is implemented in the circuit by grounding the four probed sample in the centre, hence making it a five probed measurement. Grounding the centre divides the sample into two equal arms which forms the lower arms of the bridge[7].

So, how does this circuit helps in noise measurement? First, the bridge removes the average voltage drop and gives us only the fluctuation across the sample. Due to this, the limitation of lower frequency limit, $R_A C$ due to the capacitor is removed. Second, the noise contribution due to fluctuation in supply voltage of the battery is also removed. Third, due to bridge circuit, the noise measurement becomes insensitive to the bath temperature fluctuations.

The large ballast resistors[7] R_1 and R_2 reduce the effect of fluctuations in the current contact resistances. When $\frac{R}{r} >> 1 (j = 1, 2)$, the input impedance to the preamplifier is r.

3.3 Drawbacks even after putting bridge circuit

Note that we are still using DC as voltage source. Even after adding a bridge circuit, and eliminating various unwanted disturbances, the sensitivity of the noise measurement is still limited at low frequencies by preamplifier noise. The 1/f noise of the preamplifier at low frequency makes it impossible to measure 1/f noise of the sample.

To minimise the noise from the preamplifier as properly as possible, we need to implement the circuit in the "Eye of the Noise Figure" which is the region of minimum noise from the amplifier. For this, an impedance matching transformer can be used which makes the input impedance to match in the region of the eye of the noise figure of the pre-amplifier.

3.4 AC biased bridge circuit

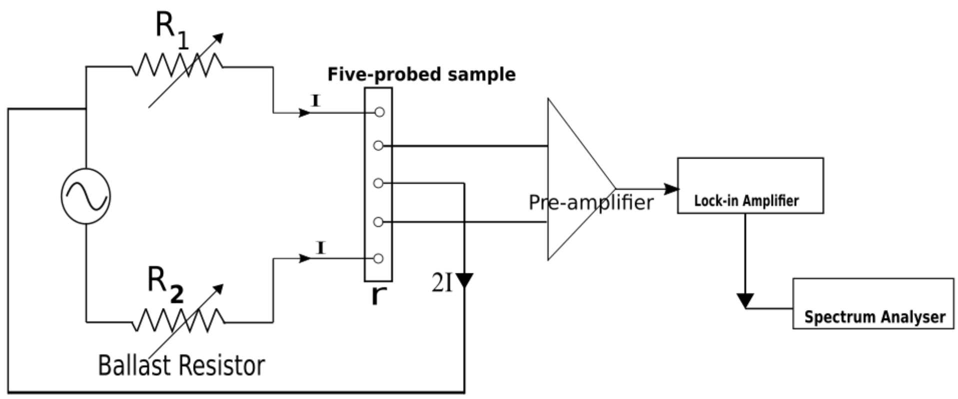


Figure 3.4: Five probe AC technique

Instead of DC, if now the sample is biased by an AC, we can use a carrier frequency f_o which is near the eye of the noise-figure of the preamplifier.

With an alternating current I(t) = I_o sin($2\pi f_o$t), the resistance fluctuations gets modulated with the carrier signal to produce noise sidebands. The signal is amplified by the pre-amplifier which will also add its noise in the signal but because the pre-amplifier is near its optimum frequency (i.e, near the eye of the noise figure), it will give relatively lesser noise. Then the carrier signal is demodulated to retrive the low frequency resistance fluctuations using a lock-in amplifier. The lock-in amplifier uses Phase Sensitive Detection technique to demodulate the fluctuations from the carrier signal.

Chapter 4

Measurement setup and apparatus

4.1 Basic noise measurement procedure

The basic idea to measure resistance fluctuation is by allowing current to flow through the sample and then record the fluctuations in voltage developed across it. To perform this task we use various instruments that will be discussed in this chapter.

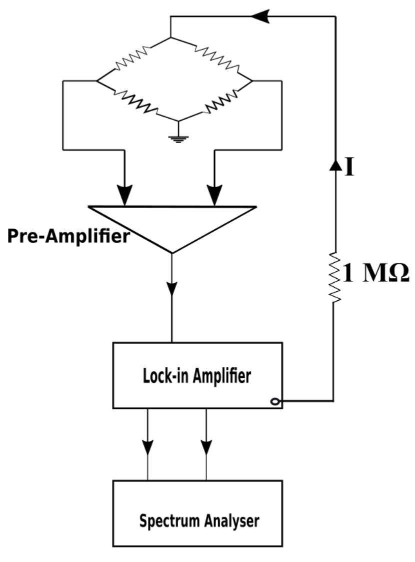


Figure 4.1: AC biased, Wheatstone bridge implemented noise measurement experimental setup

In the figure above, a schematic diagram for noise measurement setup is shown. A brief discussion on the measurement procedure is as follows:

A lock-in amplifier sends an AC voltage of a particular frequency, which passes through a resistor (whose resistance is much-much greater than the sample resistance) to make a current source. This current is now passed to the sample. The sample is connected in a wheatstone bridge arrangement. If the sample is a simple resistor, then two variable wirewound resistors can be used for the upper arms of the bridge and two sample resistors are used to form the lower arms of the bridge with the centre grounded. The bridge is balanced as properly as possible depending on the capability of the variable resistors. The error voltage signal developed across the bridge is then amplified using a low noise pre-amplifier. The output of the pre-amplifier is passed to the input of the lock-in amplifier. There are various parameters in the lock-in amplifier, like Time constant of the low pass filter, sensitivity, type of coupling, grounding type, etc to be adjusted depending on the experiment. These will be properly discussed in the subsequent sections. The X and Y output of the lock-in amplifier gives the in-phase and 90^o out-of-phase voltage across the sample respectively. The fluctuating voltage signal shown in the X and Y output of lock-in amplifier are in the time domain. This is passed to a Spectrum Analyser that gives the fourier transform of the signals. We expect the difference between fluctuating X and Y outputs to have 1/f feature which is the 1/f noise from the sample and fluctuating Y output to be white in nature which is the background noise of the sample and the associated electronics. These concepts will be discussed in the subsequent sections.

4.2 DSP lock-in Amplifier

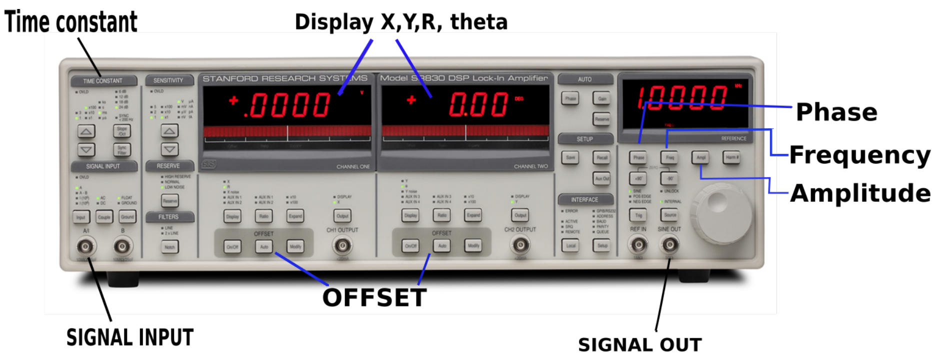


Figure 4.2: Stanford Research Systems Model-SR830 DSP Lock-in Amplifier.

A lock-in Amplifier is basically like an AC voltmeter/ammeter that can measure very small signals buried inside large unwanted signals. It does this by using a technique called Phase Sensitive Detection. A phase sensitive detector multiplies two ac voltage signals and gives out a resultant voltage signal only when the frequencies of the two signals are equal. The algebra of this is given below.

4.2.1 Phase Sensitive detection - Algebra

Suppose the input signal to the phase sensitive detector is,

$$V_{signal} = V_{sig} sin(\omega_{sig} t + \phi_{sig})$$

Now, if this signal is multiplied by another signal called the reference signal,
$V_{reference} = V_{ref} sin(\omega_{ref} t + \phi_{ref})$, we get,

$$V = V_{signal} V_{reference}$$

Using the trignometric identity, $\quad A sin\theta B sin\theta = \frac{AB}{2}[cos(\theta - \phi) - cos(\theta + \phi)]$
We get,
$V = V_{sig} sin(\omega_{sig} t + \phi_{sig}).V_{ref} sin(\omega_{ref} t + \phi_{ref})$

$V = \frac{V_{sig} V_{ref}}{2} cos([\omega_{sig} - \omega_{ref}]t + [\phi_{sig} - \phi_{ref}]) - cos([\omega_{sig} + \omega_{ref}]t + [\phi_{sig} + \phi_{ref}])$

If $\omega_{sig} = \omega_{ref}$, we get

$V = \frac{V_{sig} V_{ref}}{2} cos([\phi_{sig} - \phi_{ref}]) - cos([2\omega_{sig}]t + [\phi_{sig} + \phi_{ref}])$

When this signal is passed through a low pass filter, the second term which is high frequency term is filtered out and we get,

$V = \frac{V_{sig} V_{ref}}{2} cos(\phi_{sig} - \phi_{ref})$

$V = \frac{V_{sig} V_{ref}}{2} cos(\phi)$
where $\phi = \phi_{sig} - \phi_{ref}$, which is the phase difference between the input signal to the phase sensitive detector and the reference signal.

Further, a Dual Phase Lock-in Amplifier has a second Phase Sensitive Detector that multiplies the input signal V_{signal} by another reference signal,

$V^{*}_{reference} = V_{ref}cos(\omega_{ref}t + \phi_{ref})$

which is 90^{o} out of phase compared with $V_{reference}$ (because cosine function is 90^{o} out-of-phase of sine function).

By having this feature of dual phase sensitive detection, the lock-in can give outputs,

$$X = V_{sig}cos\phi$$

$$Y = V_{sig}sin\phi$$

X is called the in-phase output and Y is called the quadrature output.
Also,

$$R = \sqrt{X^2 + Y^2} = V_{sig}$$

$$\phi = tan^{-1}\frac{Y}{X}$$

Where R is the magnitude of the voltage signal we want to measure.

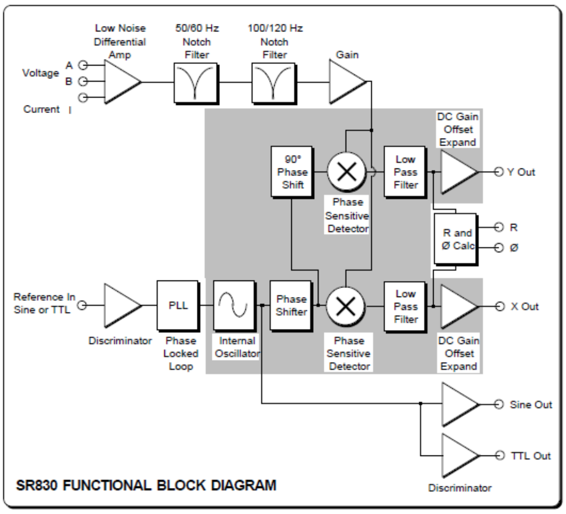


Figure 4.3: Schematic block diagram of a dual phase lock-in amplifier, SR830

A lock-in amplifier generates an AC sinusoidal signal using an internal Oscillator which is to be fed to the circuit. One can also use an external signal from an external Signal Synthesiser to be fed to the circuit depending on the experiment. When doing so, this external signal needs to be fed into the lock-in amplifier and is called the reference signal.

The signal from the circuit which is to be measured is fed into voltage input in single ended connection mode (A) or differential voltage connection mode (A-B) depending on the circuit connection setup. The input signal first passes to Notch Filters that filters out the 50/60/100/120 Hz powerline frequencies interfering the signal. Then the signal passes to the phase sensitive detector which extracts out the desired frequency signal using the phase sensitive detection method mentioned above. After that, the signal passes to the low pass filter whose time constant can be set by the user. This time constant sets the upper limit for the frequency of the output of the low pass filter. After passing through a DC amplifier, this signal is given out as the X signal which is in-phase with the signal from the internal oscillator or reference.

In a dual phase lock-in amplifier, the signal from the internal oscillator gets 90^o phase shifted and goes into another phase sensitive detector which multiplies this 90^o phase shifted signal with the signal coming from the circuit using phase sensitive detection algorithm. The output is again passed through the low pass filter already set up as discussed previously. This signal is given out as Y output which is 90^o out-of-phase with respect to the signal from the internal oscillator or reference.

The R value and ϕ value are also given out which the lock-in calculates digitally by using the appropriate formula discussed above.

4.3 Four probe resistance measurement

Whenever we do resistance measurement of a simple resistor using multimeter, we are doing 2 probe resistance measurement. The four probe resistance measurement is shown in the figure below :

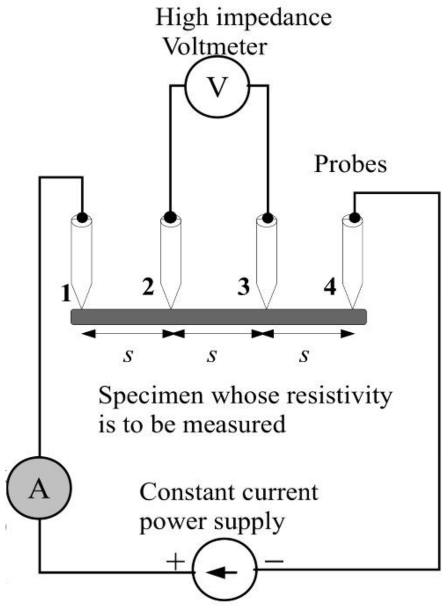


Figure 4.4: Schematic diagram showing four-probe resistivity measurement technique. [Source: Resistivity by Four Probe Method - Amrita University Virtual Lab]

In a four-probe measurement, four collinear equally spaced probes are in contact with the sample. Current is allowed to flow through 1 and 4 as shown in the figure, and voltage developed across 2 and 3 are measured. Using Ohm's law, resistance of the sample is calculated.

The advantage of four-probe resistance measurement over two-probe is that it eliminates the effects of contact resistance between the sample and electrical contacts and so, is more suitable for low resistance sample for accurate resistivity measurement.

4.4 Break-out box

A break-out box is a faraday box inside which the sample is soldered. It is used to reduce the unwanted noise from external electromagnetic radiations. In this experiment, a 5.5 mm iron walled box was used which is shown in the figure. It has BNC inputs for connections where the sample (resistors) were soldered appropriately and all other BNCS were grounded with grounding caps to minimise unwanted noises as effectively as possible.

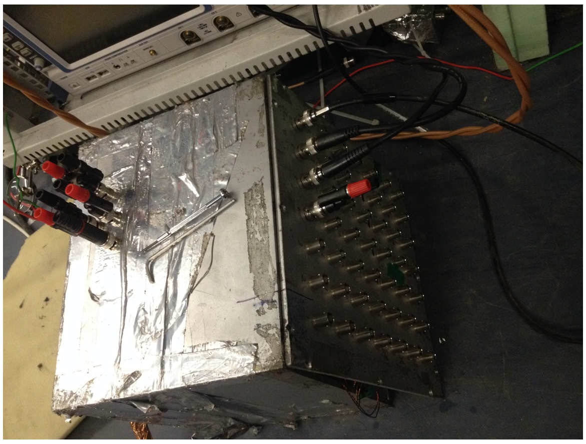


Figure 4.5: A break-out box with some grounding caps connected to it

The theory behind breakout box shielding is skin effect. It is the tendency of any AC signal to become distributed throughout the conductor (in our case, break-out box) such that the signal reduces with depth into the material depending on the signal frequency and conductor material. The formula for skin depth is:

$$\delta = \sqrt{\frac{2\ \rho}{2\ \pi\ f \mu_o\ \mu_r}} = 503\ \sqrt{\frac{\rho}{\mu_r\ f}}$$

where,
ρ = resistivity of the medium in Ωm.
f = frequency of the signal in Hz.
μ_r = relative permeability of the medium = $\frac{\mu}{\mu_o} = \frac{permeability\ of\ the\ medium}{permeability\ of\ free\ space\ (4\pi\ \times\ 10^{-7}\ H/m)}$

If a material is more magnetic, its skin depth is lower because of its higher μ_r. Higher the frequency of the signal, the lower is the skin depth. This means lower frequency signals can penetrate more inside the material.

In this experiment, with iron as the break-out box material with,
$\rho = 9.71 \times 10^{-8}\ \Omega$ m (at 20^{o}C)
$\mu = 6.3 \times 10^{-3}$ H/m $\Rightarrow \mu_r \approx 5000$ (for 99.8% pure iron)

$$\delta = 503\ \sqrt{\frac{9.71\ \times\ 10^{-8}}{5000\ f}} = \frac{2.215\ \times\ 10^{-3}}{\sqrt{f}}\ m$$

Hence, the 5.5 mm thick break-out box would shield frequencies above 162 mHz which is very effective.

4.5 Low noise Pre-amplifier

Figure 4.6: NF Electronics Model LI-75A low-noise pre-amplifier

A low-noise preamplifier is used to provide amplification gain to the voltage fluctuations occuring across the sample.

4.5.1 Noise Figure contour

Noise Figure of an amplifier is a measure of the amount of noise the amplifier adds over in the measurement in its operating temperature. Every amplifier comes with a Noise Figure contour plot from the manufacturer describing the amount of noise the amplifier produces at a particular frequency and at particular impedance at the input of the amplifier.

Noise Figure (in dB) of an amplifier for a particular carrier frequency 'f' and input impedance 'r' is given by :-

$$NF(f,r) = 20\ log_{10}\frac{S_V^o(f)}{4k_BTr} = 20\ log_{10}\frac{Output\ noise/Gain}{Nyquist\ noise}$$

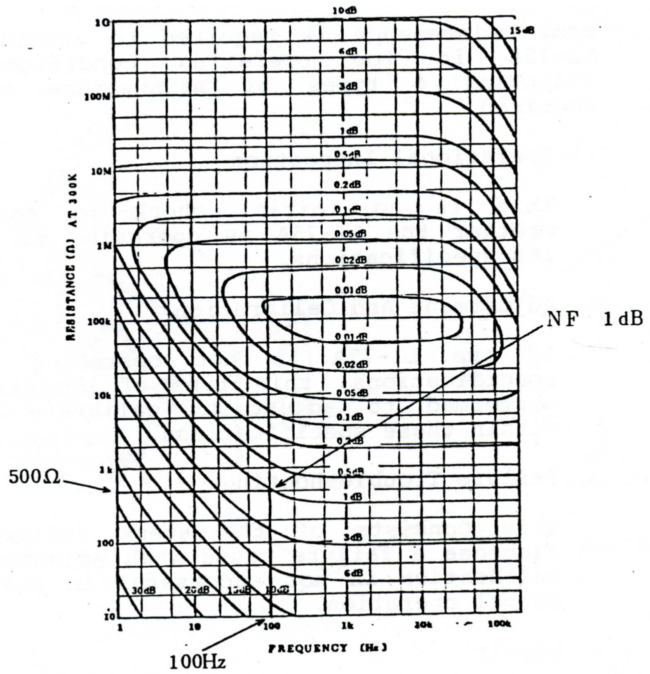


Figure 4.7: Noise Figure contour of NF-Electronics Model LI-75A low-noise preamplifier

From the Noise Figure contour, we can see that the LI-75A pre-amplifier produces the least noise at frequency 1 KHz with an input impedance 100 KΩ.

4.6 Low-pass filtering

The output of the phase sensitive detector passes through a low pass filter. The cutoff frequency of the low pass filter determines the upper cut of frequency of the output signal.

The time constant of the low pass filter is given by

$$\tau = \frac{1}{2\pi f}$$

where ' f ' is the cutoff frequency of the low pass filter.
We can set the time constant of the lock-in amplifier manually.

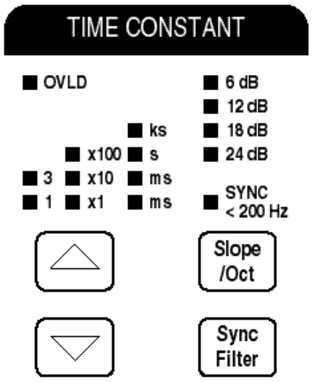


Figure 4.8: Time constant front panel in SR830 lock-in amplifier

The precaution to keep in mind when setting the time-constant in the lock-in amplifier is that the cutoff frequency of the low pass filter must be equal to or higher than the end frequency of the power spectrum plot. This needs to be ensured because the datas recorded above the cutoff frequency in the power soectrum is faulty.

4.7 FFT Spectrum Analyser

Figure 4.9: Stanford Research Systems, Model SR760 FFT Spectrum Analyser

A Spectrum Analyser is an instrument that takes a time varying input signal and computes its frequency spectrum i.e does a Fourier Transform. It first digitizes an input signal at a high sampling rate and mathematically transforms the signal into frequency spectrum using an algorithm called Fast Fourier Transform (FFT). Since in our noise measurement, we need the Power Spectral density as a function of frequency, a spectrum analyser is the right tool for it.

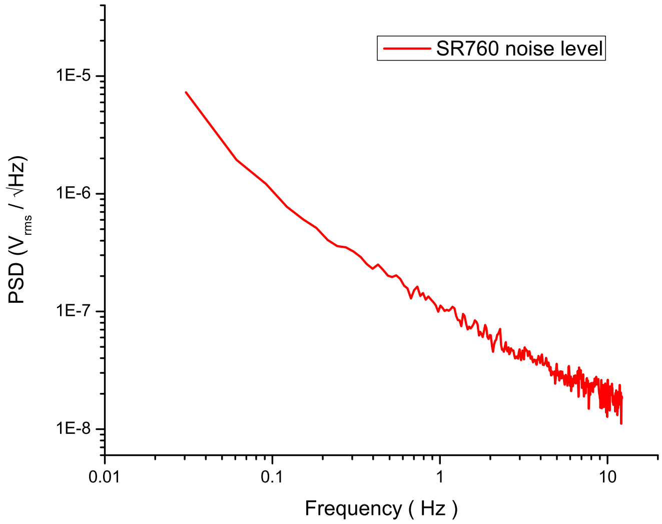


Figure 4.10: Power spectrum of SR760 Spectrum Anaylser taken by shorting its inputs showing 1/f spectral nature.

The Spectrum analyser itself has 1/f noise due to its electronics. A low noise Spectrum Analyser is essential for noise measurement. Because we have used a preamlifier, the multiplied gain has already increased the magnitude of 1/f noise from the sample to ensure that it is above the minimum noise level measured by the spectrum analyser. A low noise pre-amplifier is used to amplify the magnitude of the noise signal from the sample to keep it above the Spectrum Analyser's own noise.

The spectrum analyser used in the experiment, SR760 had many parameters to be set for the experiment. Windowing is used to reduce the amount of smearing in the spectrum from signals not exactly periodic with the time record. The different types of windows trade off selectivity, amplitude accuracy, and noise floor. Hanning window was used in the experiment. Averaging option in the spectrum analyser averages many spectra together and improves the accuracy and repeatability of measurements. Linear averaging of about 2000 or 4000 was used during the experiment. The input range on the SR760 varies from a maximum of 34 dBV full scale to a minimum of -60 dBV full scale. In the experiment, the input range was set to AUTO-RANGE mode and it was ensured that whenever a new measurement was taken, it was auto ranged again. This is because if the input signal level is very low and the input range is at high value, the real measurement signal gets overshadowed by unwanted signals and the measurement is not reproducible. The input coupling was set to DC because we want to measure low frequency as low as 15mHz. In AC coupling, low frequencies are attenuated by the high pass filter. The grounding was set to GROUND in which the shields are connected by 50 Ω to the chassis ground.

4.8 1/f noise measurement using OFFSET feature of lock-in amplifier

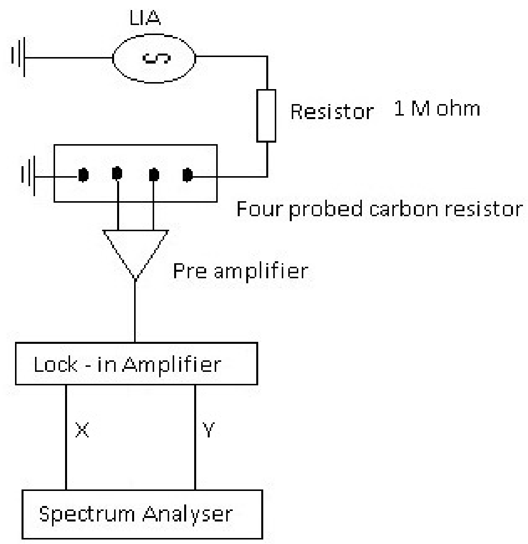


Figure 4.11: Measurment setup for Four-probe and use of OFFSET feature of LIA for 1/f noise measurement

A lock-in amplifier sends an AC voltage of a particular frequency, which passes through a resistor (whose resistance is much-much greater than the sample resistance) to make a current source. This current is now passed to the sample. The sample is connected in a four-probe arrangement as seen the figure. If the sample is a resistor, three resistors are connected linearly, hence, creating four ends. Current is passed to the first end and the last end is grounded using a grounding cap. The voltage across the resistor in the middle is passed to a low noise pre-amplifier for amplification gain in the voltage fluctuations developed across the resistor. The output of the pre-amplifier is passed to the input of the lock-in amplifier. There are various parameters in the lock-in amplifier, like Time constant of the low pass filter, sensitivity, type of coupling, grounding type, etc are set depending on the experiment.

AUTO GAIN and AUTO PHASE buttons are pressed for the lock-in to set the appropriate sensitivity and phase automatically. Then, the AUTO OFFSET is pressed. This feature substracts the average voltage developed across the sample from the instantaneous voltage and gives the instantaneous fluctuation. Basically, it gives $\delta V = V(t) - \langle V \rangle$. This values in X and Y output is expanded $\times 10$ or $\times 100$ times ensuring not to overload the lock-in.

The fluctuating voltage signal shown in the X and Y output of lock-in amplifier are in the time domain. This is passed to a Spectrum Analyser that can give the poswer spectral density of the signals. We expect the power spectral density of difference between X and Y outputs to have 1/f feature which is the 1/f noise from the sample and power spectral density of the Y output to be white in nature which is the background noise of the sample and also contributions from the associated electronics.

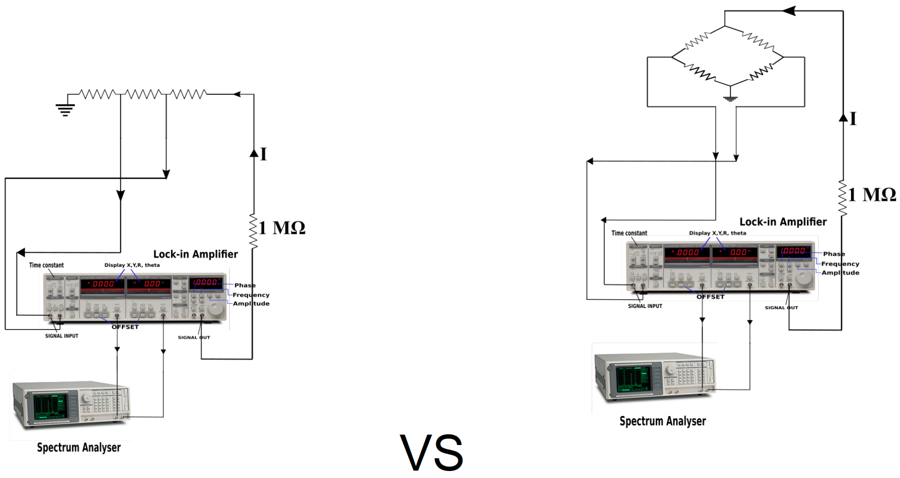
Chapter 5

Results and discussions

5.1 AC biased-wheatstone bridge implementation

1/f noise was measured using two different measurement setups :

- AC, four-probe and OFFSET feature of Lock-in Amplifier
- AC, wheatstone bridge implementation



(a) Four-probe and OFFSET implementation

(c) Wheatstone bridge implementation

The parameters kept in lock-in amplifier are:

Frequency (Internal)	1.1171 KHz
Amplitude	5V
Phase	Auto-phase
Time constant	30ms(corresponds to 5 Hz cutoff freq.)
Sensitivity	Auto-sensitivity
Coupling	DC
Gounding	Ground
Slope/Oct	24dB
Reserve	low noise
Filters	2 × line
Gounding	Ground

Table 5.1: Table showing various important parameters kept in LIA during the experiment.

The parameters kept in Spectrum Analyser SR760 are:

Frequency span	15.26mHz-3.067 Hz
Input range	Autorange
Auto-offset	Off
Input Coupling	DC
Grounding	Ground
Average	linear, 4000
Window	Hanning

Table 5.2: Table showing various important parameters kept in LIA during the experiment.

A-B input mode was used to get figure 5.1 with X and Y output of LIA in A and B inputs of Spectrum Analyser(SR760) respectively. B input mode was used to get figure 5.2 with Y output of LIA in A input of Spectrum Analyser SR760. Before every measurement, AUTORANGE was performed to bring the input range to appropriate value.

- 1/f noise measurement on sample : 1KΩ wirewound resistor(10W, 1%)
- Frequency span = 15.26 mHz to 3.067 Hz

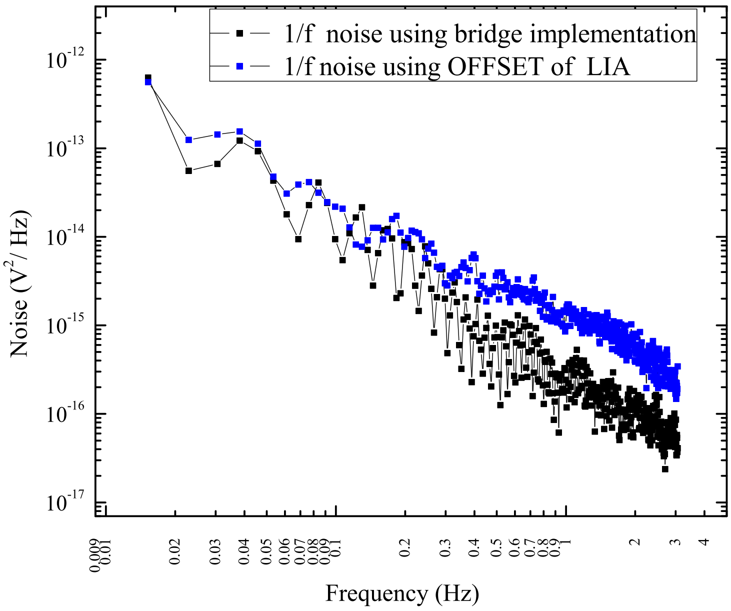


Figure 5.1: Comparison of 1/f noise measurement using wheatstone bridge implementation (Black) and four-probe OFFSET feature of Lock-in Amplifier (LIA) (Blue) in a 1KΩ wirewound resistor at 300K.

1/f noise is approximately same for both measurement setups.

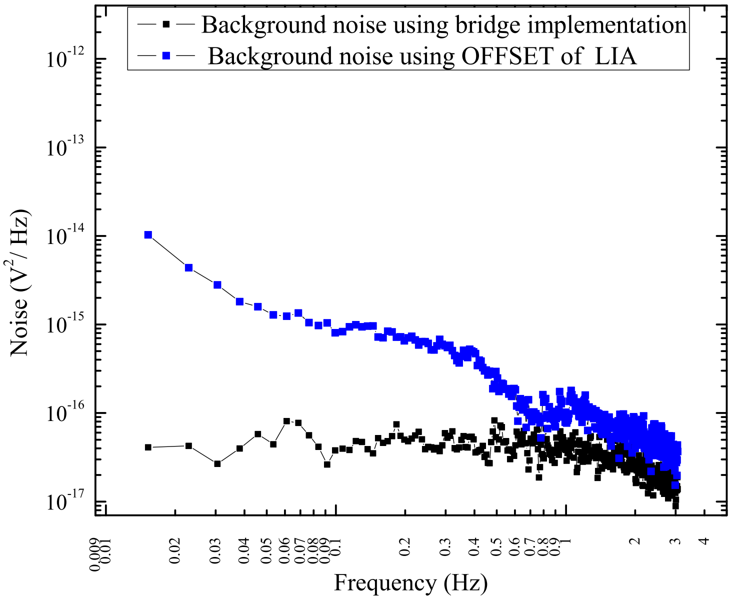


Figure 5.2: Comparison of background noise using wheatstone bridge implementation (black) and four-probe OFFSET feature of Lock-in Amplifier (LIA) (Blue) in a 1KΩ wirewound resistor at 300K.

The background noise obtained using wheatstone-bridge implementation has no 1/f feature at lower frequencies (measured as low as 15 mHz) compared to background noise obtained using four-probe and OFFSET feature of lock-in amplifier. This shows the improvement and superiority of the measurement using the wheatstone bridge implementation setup.

5.1.1 Voltage bias dependence

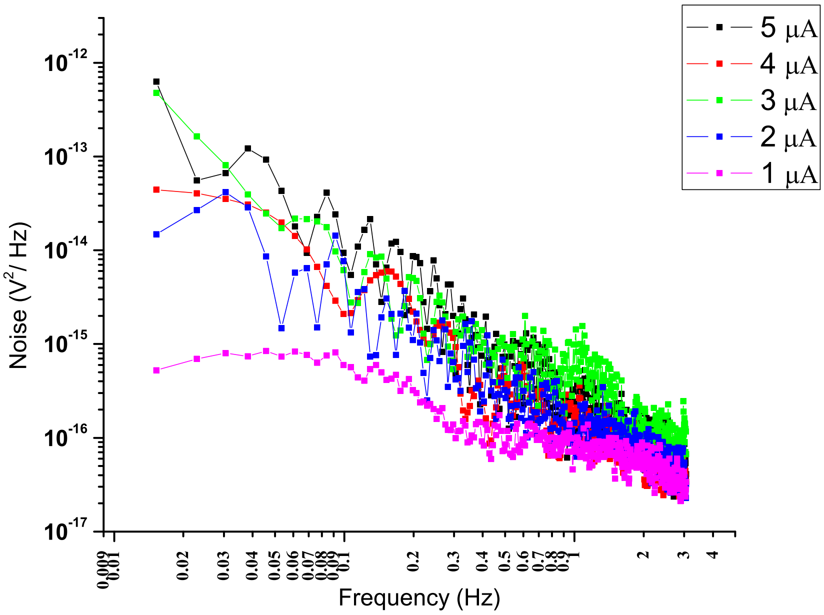


Figure 5.3: Current biasing dependance of 1/f noise in 1KΩ wirewound resistor in wheatstone bridge implementation.

The magnitude of 1/f noise seem to increase with increasing current biasing. The power exponent (α) of 'f' in Hooge empirical formula increases with biasing current shown in table 5.1 .

5.1.2 Calculations

From, Hooge empirical Formula,

$$Noise\ (\frac{V^2}{Hz}) = \frac{\gamma\ V^2}{N_c\ f^\alpha}$$

$$ln(Noise) =\ ln(\frac{\gamma\ V^2}{N_c}) - \alpha\ ln(f)$$

Plotting ln(Noise) vs ln(frequency) will give a graph with negative slope.

Hence, slope = power exponent (α) and y-intercept = $ln(\frac{\gamma\ V^2}{N_c})$.

Noise magnitude is calculated by doing integration over each plot in ORIGIN.

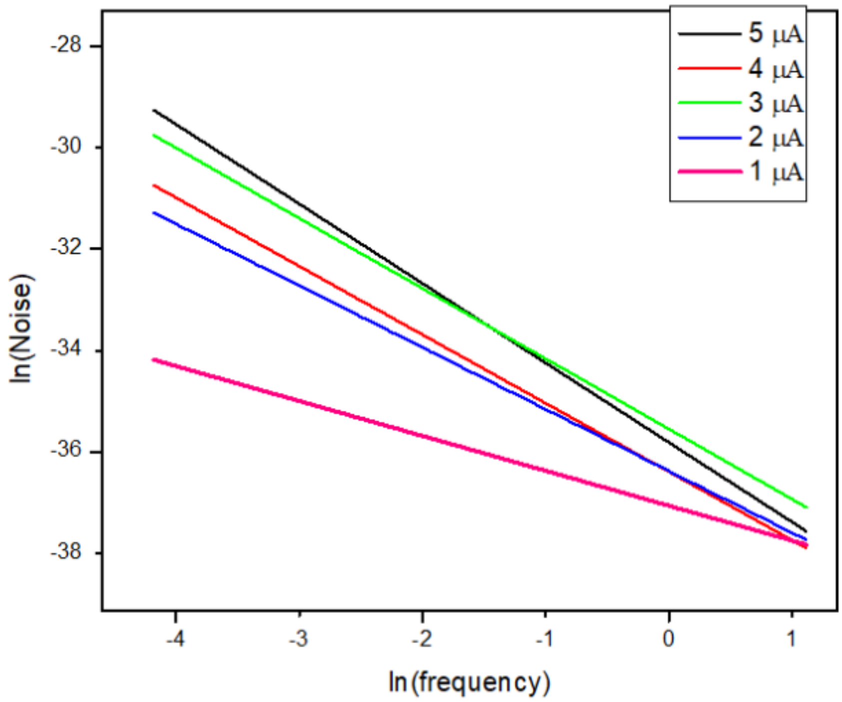


Figure 5.4: Voltage dependance of 1/f noise in 1KΩ wirewound resistor in bridge implementation.

Current(μA)	**Power exponent (α)**	**Noise (V^2)**
1	0.688	3.128×10^{-16}
2	1.217	1.890×10^{-15}
3	1.385	6.691×10^{-15}
4	1.351	2.523×10^{-15}
5	1.568	8.226×10^{-15}

Table 5.3: Table showing α and $Noise$ value of 1/f noise obtained in 1KΩ wirewound resistor for different current biasing.

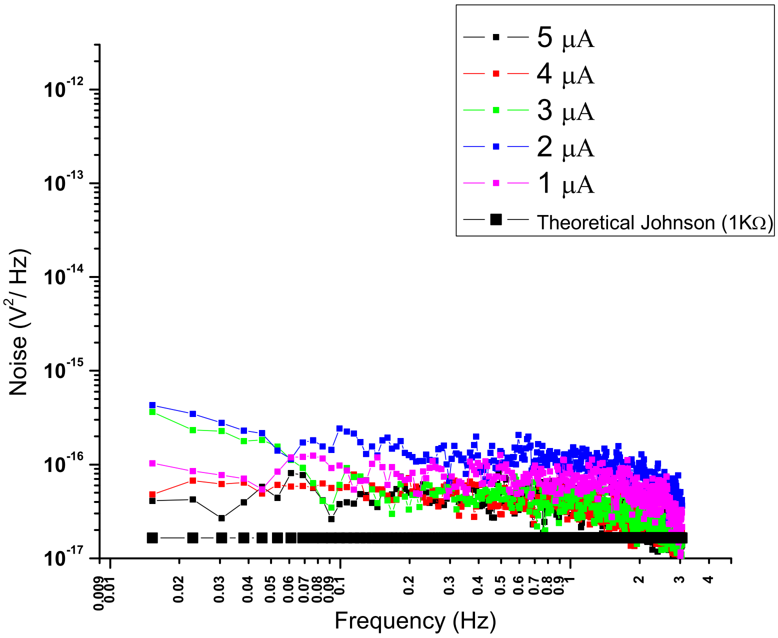


Figure 5.5: Background noise at different bias current across the sample 1KΩ wire-wound resistor

The background noise does not show any current bias dependance as expected.

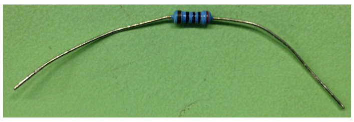
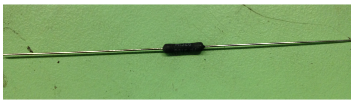
Theoretical Johnson (1KΩ) i.e, $[4K_BTR \times \text{frequency span}] = 5.041 \times 10^{-17}$

Current(μA)	**Noise (in unit** V^2)
1	1.783×10^{-16}
2	2.853×10^{-16}
3	1.098×10^{-16}
4	9.818×10^{-17}
5	1.004×10^{-16}

Table 5.4: Table showing α and Noise value of backgrund noise of 1KΩ wirewound resistor at different current bias.

5.2 Comparing 1/f noise in different types of resistors.

- 200Ω Metal film resistor (0.6W, 1%)
- 200Ω Wirewound Resistors - Through hole (1W, 1%)
- 1KΩ Wirewound Resistors - Through hole (10W, 1%)

(1)  (2) 

(3) 

(1) 200Ω Vishay Metal-film resistor - Through Hole, 0.6 W, 1%
(2) 200Ω Vishay Wirewound resistor - Through Hole, 1 W, 1%
(3) 1KΩ Vishay Wirewound resistor - Through Hole, 10 W, 1%

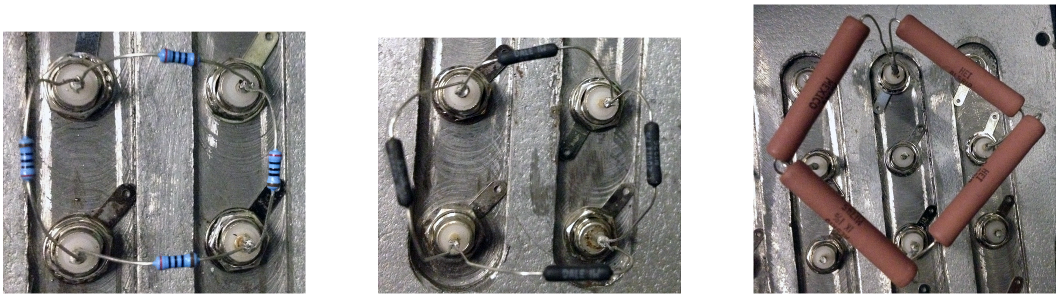


Figure 5.6: The three resistors as wheatstone bridge

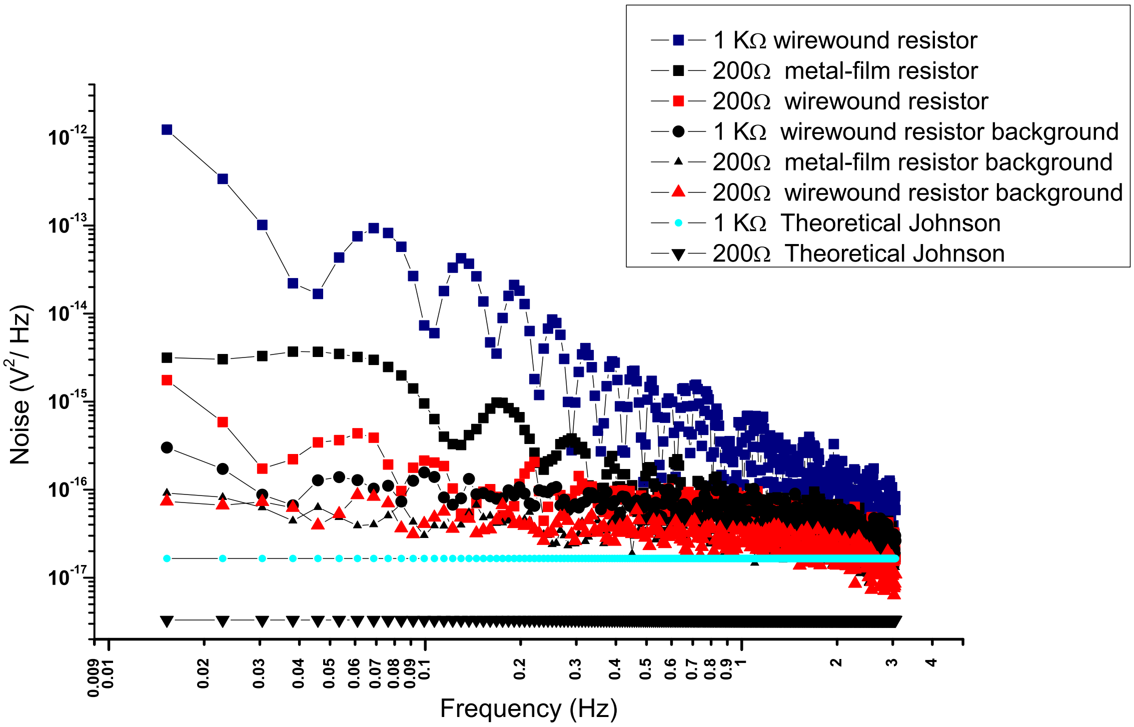


Figure 5.7: 1/f noise, background noise and Theoretical Johnson of the three resistors

1/f noise of three different types of resistors along with their background noise and theoretical Johnson noise $4K_BTR$ are shown in Figure 5.6. It compares 1/f feature of the noise from the resistors with their corresponding white background noise. The theoretical Johnson noise level for 200Ω and 1KΩ are also plotted to compare with the magnitude of the background noise obtained experimentally showing that the measurement still needs to be improved showing that the background noise contains noise due to other electronics associated with the circuit which has not been eliminated. Nevertheless, no 1/f feature in background noise shows that the wheatstone bridge implementation is working.

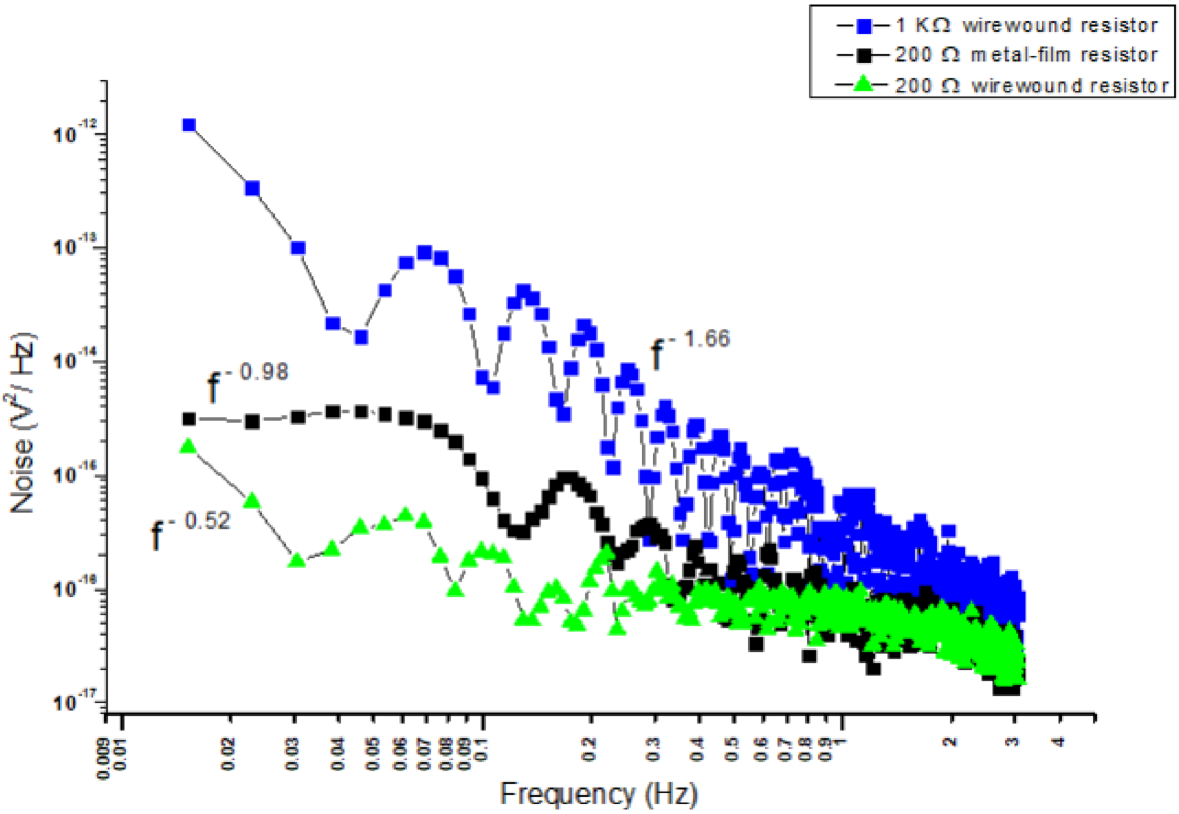


Figure 5.8: Comparing 1/f noise of the three resistors

1/f noise of metal-film resistor has more 1/f feature than wirewound resistor of the same resistance (200Ω) and tolerance value (1%). The power exponent (α)of frequency of the Hooge's empirical formula are also shown. Also, with increase in the resistance value, 1/f feature increases in a resistor as seen from comparison between 1KΩ (Blue) and 200Ω (green) wirewound resistor plots (both having the same 1% tolerance value).

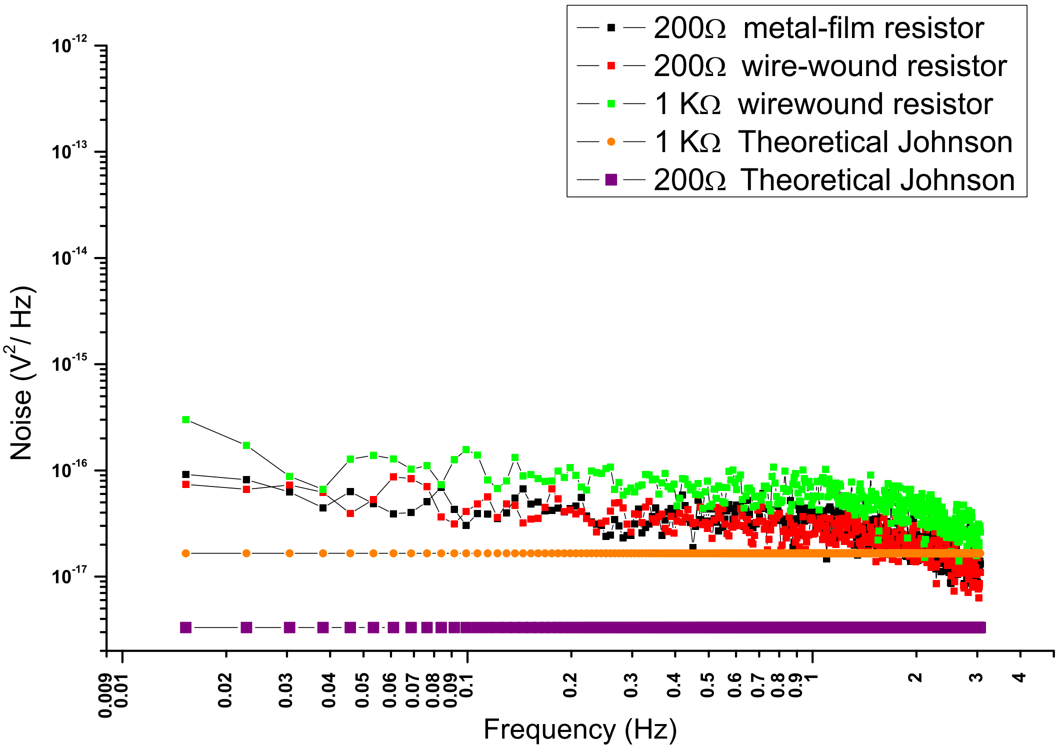


Figure 5.9: Comparing background noise of the three resistors

Irrespective of the type of resistors, the background noise remains the same for a particular resistance value as shown for 200Ω metal-film and wirewound resistor. But on increasing the resistance value to 1KΩ, the background noise increases as expected as seen from the green plot. The background noises does not coincide with the corresponding theoretical values showing that more critical optimization of the experimental setup is required to eliminate the unwanted background noise that maybe coming from various electronics associated in the circuit.

5.3 1/f noise in LAO-STO sample

5.3.1 Sample

LAO-STO sample was wirebounded in a PCB to perform a four-probe resistance measurement. The sample was kept in a dark environment and a resistance-vs-time measurement was performed.

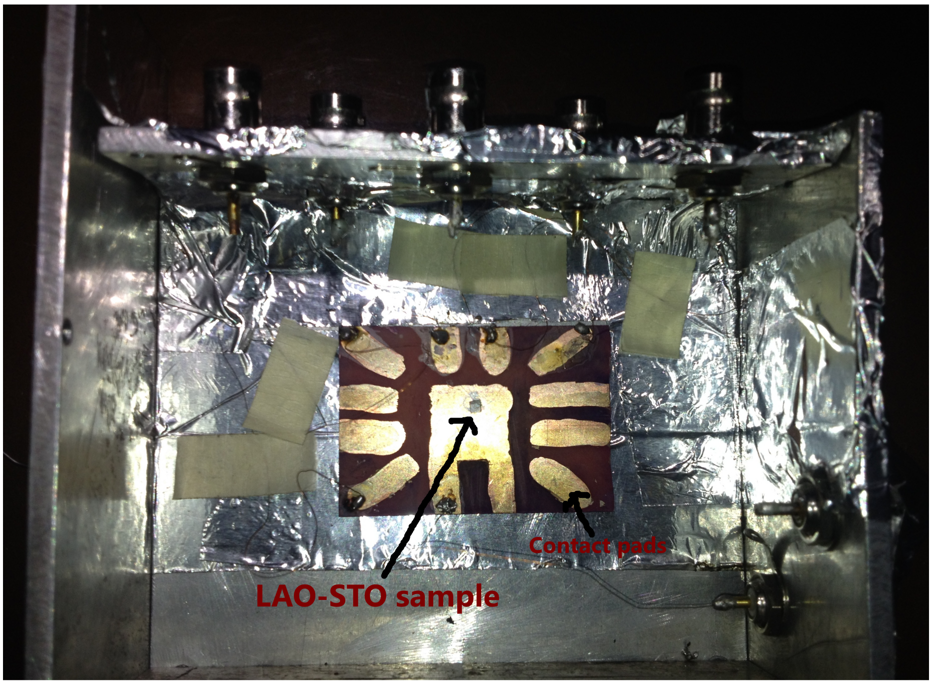


Figure 5.10: Sample soldered in a break-out box

5.4 Sample Preparation

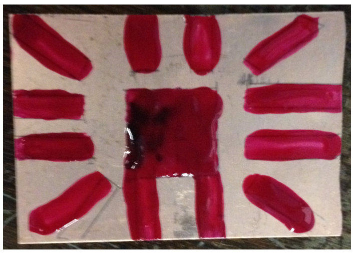


Figure 5.11: Nail-polish set into PCB in desired shapes for making contact pads

Nail-polish is set into the PCB in desired shapes for making contact pads. After the nail polish is dry, the PCB is dipped in $FeCl_3$ solution that will etch out the copper parts that does not have nail-polish covered over it. The chemical reaction happening is,

$FeCl_3$ + Cu $\rightarrow$ $CuCl_2$ (aq.) + Fe (aq.)

The nail-polish is washed off using acetone to give the desired copper pads. The region where copper has been etched out is painted with permanent marker. Then, Gold is evaporated over the copper pads. This is done because wirebounding happens only in gold surface.

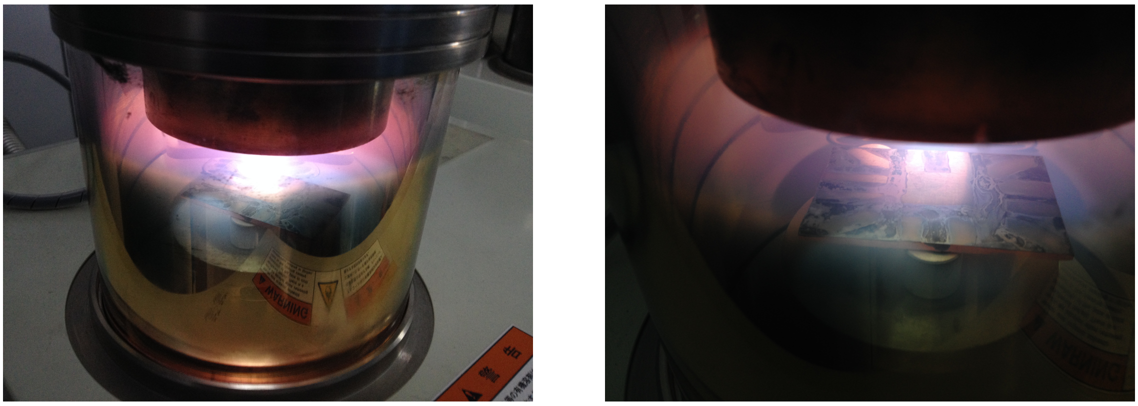


Figure 5.12: Gold plating over copper pads

LAO-STO sample is sticked in the appropriate area using silver paint and LAO-STO sample is wire-bounded with five connections on it.

(a) LAO-STO sample wirebounded in a PCB

(b) 5 wirebounded wires

5.4.1 Resistance-vs-time measurement of LAO-STO heterostucture.

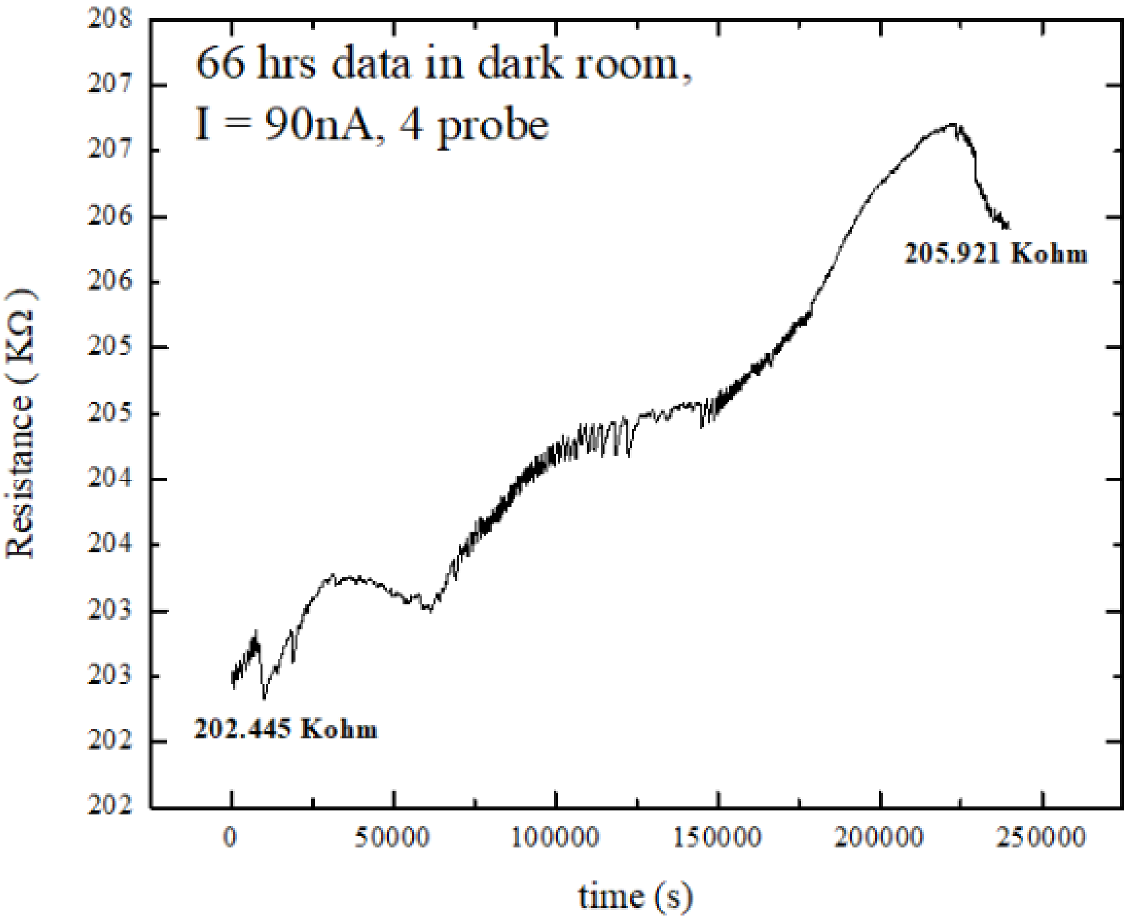


Figure 5.13: Resistance-vs-time plot measured for 66 hrs of LAO-STO sample of with current bias 90nA measured using four-probe method

The resistance seem to increase overall but only by approx 2% in 66hrs in dark environment.

5.4.2 Voltage bias dependance of 1/f noise

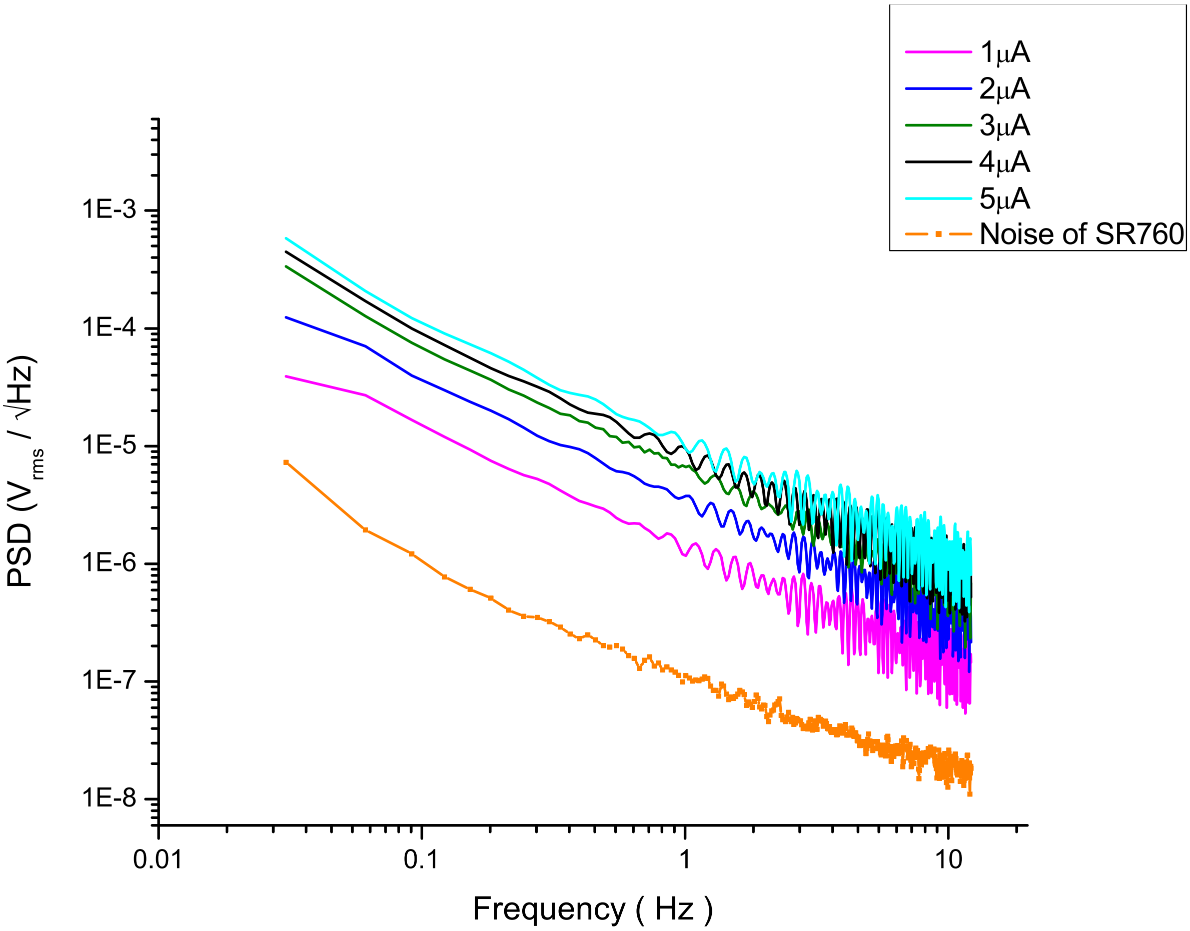


Figure 5.14: Power spectrum of LAO-STO sample with different current bias

The 1/f noise magnitude is observed to increase with increasing current bias across the LAO-STO sample and seems to follow Hooge's empirical law. This measurement was done using four-probe and OFFSET feature of lock-in amplifier.

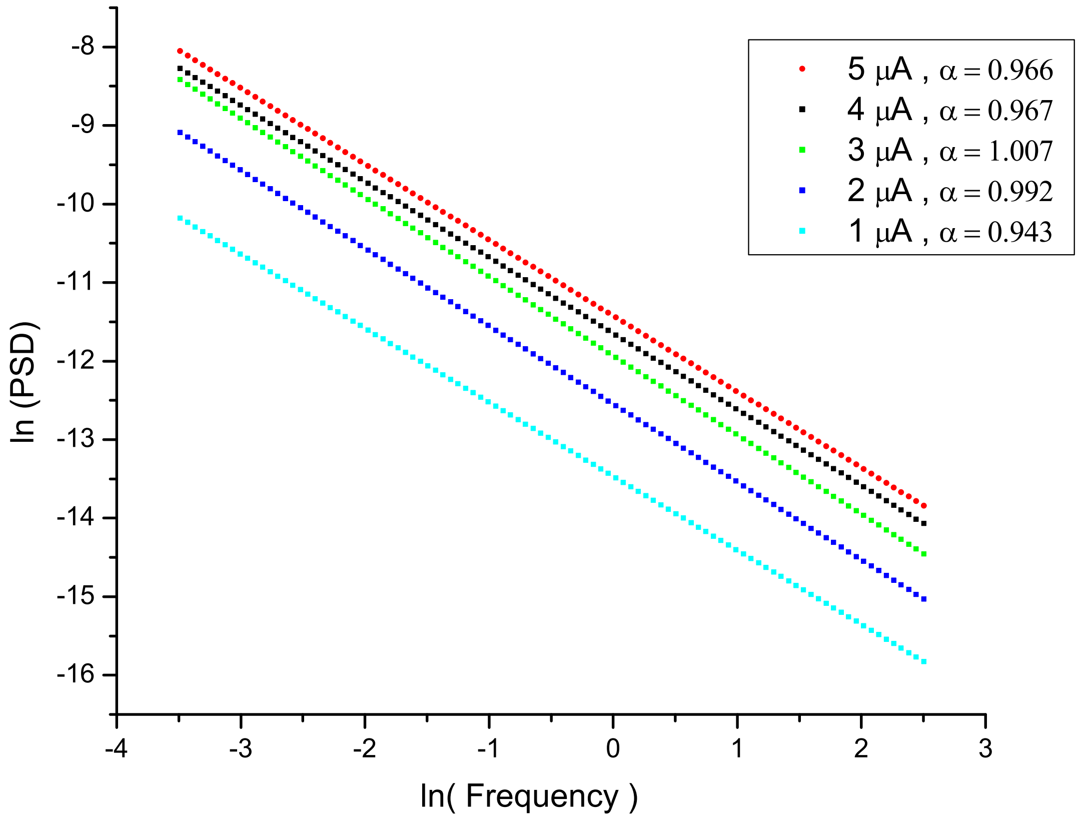


Figure 5.15: Linearly fitted plot of ln(PSD) vs ln(frequency) with corresponding power exponent value (α) for different current biasing.

5.5 Photoconductivity measurement

The sample was irradiated with a laser of 700nm (green laser) and the resistance vs time data was recorded for varying laser intensity.

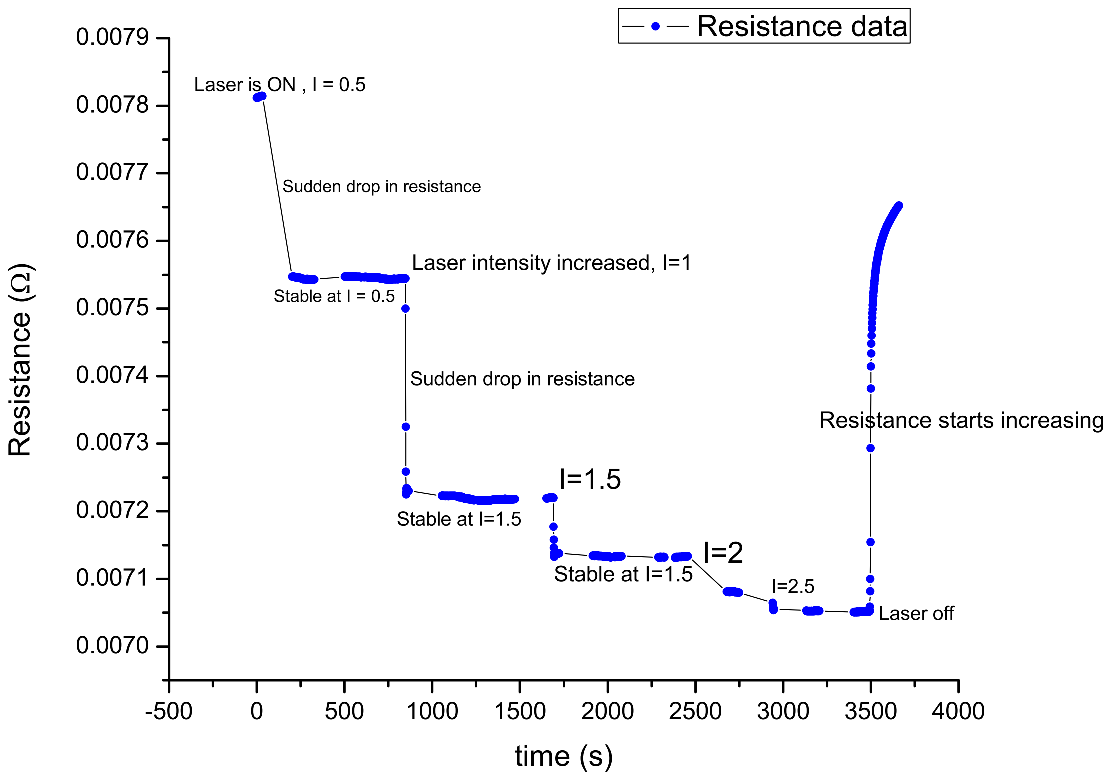


Figure 5.16: In the graph, "I" is the intensity of the laser beam, which was increased after every time the resistance of the sample became stable.

The plot shows that as soon as laser light is turned ON, the resistance of the LAO-STO sample decreased drastically and attained a minimum value and then became stable. On again increasing the intensity of the laser light, the resistance reduced further.

Chapter 6

Conclusions and Future plans

The wheatstone bridge implemented measurement setup is found to give good results for resistors upto frequencies as low as 15mHz. Depending on the resistance of the sample, good quality variable wirewound resistors needs to be used that can balance the wheatstone bridge down to 5nV-10nV.

The future plan is to try the measurement in samples like palladium thin films and palladium nanowires in hydrogen environment and vacuum and look at the behaviour of the 1/f noise in these samples.

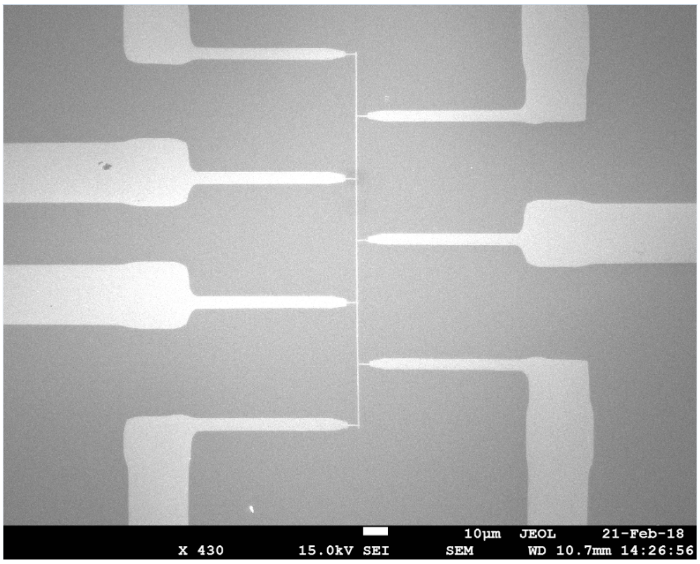


Figure 6.1: Aluminium nanowire with 7 contact pads fabricated using Electron beam lithography

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