Author: Kumar, Manohar
Title: A study of electron scattering through noise spectroscopy
Issue Date: 2012-12-05

2

Measurement technique : Noise

In mesoscopic physics, the conductance measurement has clearly dominated the past two decades of its experimental and theoretical arena. The conductance measures the effective average transmission of electron near the Fermi surface. This measurement only focuses on the averaged properties of the system and clearly misses its phase and temporal behavior. The fluctuation in current is a fundamental quantity which in past few years attracted the attention of experimentalists and theoreticians. The fluctuations in current measures the charge statistics. The second order moment of the statistics is known as shot noise which measures the deviation from the mean number of the electron transmitted through the device under study. Measurement of this deviation gives detailed information about the nature of the charge of the particle and the characteristic of mesoscopic device. Measurement of shot noise is technically challenging. The design of the measurement setup is always a compromise between the different scientific requirements like low temperature, low noise, external noise filtering, broadband frequency domain measurements etc. Here I present briefly the different noise measurement techniques.

2

2.1 LOW FREQUENCY MEASUREMENT TECHNIQUE

A low frequency measurement set up is used to determine the electron counting statistics. It generally consists of sets of amplifiers and filters and either an analogue integrator or spectrum analyzer for measuring power spectrum. Analogue integrators were common in the past to measure the noise power, but with the advancement of the technology, they were replaced by FFT spectrum analyzers. They are fast and have better sensitivity. The biasing circuit is preferably battery operated to avoid spurious and 50Hz and its harmonics originating from power lines. The preamplifier amplifies the noise signal of the device above the noise floor of the spectrum analyzer but adds a technical constraint. The preamplifier generates voltage noise and current noise which add quadratically to the noise signal of the device. Hence, the noise measurement system should be calibrated with respect to the background noise and it should be subtracted from the measured noise power spectrum of the device. A typical low frequency noise setup is shown in the figure(2.1)[1]. Usually current noise is of more interest because of its closer physical institution to the current fluctuations. The current noise in the mesoscopic systems we will be considering is as of the order of a few pA$\sqrt{\text{Hz}}$ to fA/$\sqrt{\text{Hz}}$. Current noise is deduced from the voltage noise as
$\langle \Delta I_D^2 \rangle = \frac{\langle \Delta V_D^2 \rangle}{R_D^2}$, where R_D^2 is the resistance of the device.

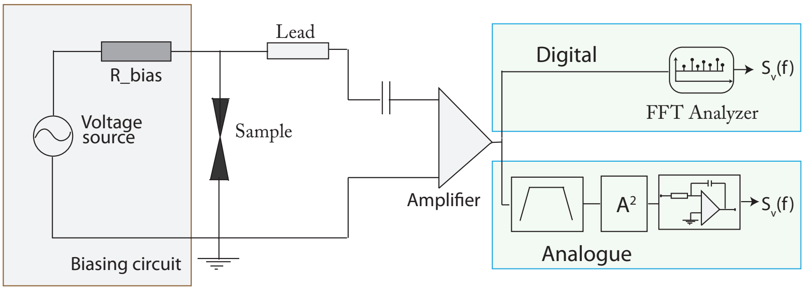

FIGURE 2.1: A typical low frequency noise measurement set up: Analogue and Digital.

2.1.1 SINGLE CHANNEL NOISE MEASUREMENT TECHNIQUE

A single channel noise measurement set up consists of a preamplifier coupled to the sample and a next stage amplifier which post amplifies the signal above the noise floor of the spectrum analyzer and background noise of the coupling environment. Such measurement setup is shown in figure (2.2 a). The main constraint of this design is the input noise of the preamplifier. This should ideally be at least an order of magnitude smaller than the noise signal of the sample. This is also the reason why for low impedance samples, voltage noise measurement is preferred over the current noise measurement as the thermal noise of the feedback resistor of transimpedance of current amplifiers limits the sensitivity of current noise measurements. The input noise of the second amplifier must be smaller than the effective noise seen at the input of the second amplifier due to the input noise of the preamplifier considering second amplifier completely noiseless. In this case noise due to second amplifier can be ignored and effective noise seen at the spectrum analyzer could be given by:

$$\langle V^2 \rangle = \langle \Delta V_D{}^2 \rangle + \langle R_D + R_L \rangle^2 \langle \Delta I_A{}^2 \rangle + \langle \Delta V_A{}^2 \rangle \tag{2.1}$$

Here $\langle \Delta V_D{}^2 \rangle$ is the voltage noise of the sample, $\langle \Delta I_A{}^2 \rangle, \langle \Delta V_A{}^2 \rangle$ are the current noise and voltage noise at the input of the preamplifier and R_D, R_L are sample resistance and lead resistance. Since R_L is very small in comparison toR_D, one can neglect the thermal noise of the leads. To determine $\langle \Delta V_D{}^2 \rangle$ accurately, $\langle \Delta I_A{}^2 \rangle, \langle \Delta V_A{}^2 \rangle$ should be measured separately and subtracted from total noise$\langle V^2 \rangle$ measured by the spectrum analyzer. Hence the sensitivity of the single channel noise measurement technique is limited by the input noise of the preamplifier. It suffers from $1/f$ noise which dominates at the lowest frequency. The effective noise resistance of the preamplifier is given as $R_{eq} = R_{eq_0}\left(1 + f_c/f\right)$ where f_c is the corner frequency of the $1/f$ noise. This limits the single channel noise measurement for the low frequency noise measurement. Especially, for the measurements below 100Hz, this technique is of no avail.

2.1.2 TWO CHANNEL CROSS CORRELATION TECHNIQUE

Generally, voltage noise of samples is in order of a few nV/$\sqrt{\text{Hz}}$ and designing a preamplifier with input voltage noise of nV/$\sqrt{\text{Hz}}$ is a technical feat. Moreover the $1/f$ noise makes the matter more difficult. One way to circumvent this issue is to employ a two channels cross correlation technique. A typical example for such setup is shown in figure (2.2b)[2]. Each channel measures the device noise and the input noise of the preamplifier (neglecting other sources of noise) *i.e.* $\langle V_{oi}{}^2 \rangle = \langle \Delta V_D{}^2 \rangle + \langle R_D + R_L \rangle^2 \langle \Delta I_i{}^2 \rangle + \langle \Delta V_i{}^2 \rangle (1 + f_c/f); i = A1, A2$. The uncorrelated noise of the amplifier of each channel averages out due to phase randomization while the correlated noise from the sample in both channels has same phase hence dominates over the non-correlated signal. Neglecting theR_L, the cross-spectral density is given by the cross conjugate product of the voltage noise measured by two channels. Here $*$ is cross conjugate product.

$$\langle V_{oA1} * V_{oA2} \rangle = \langle \Delta V_D{}^2 \rangle + \langle R_D \rangle^2 (\langle \Delta I_{A1}{}^2 \rangle + \langle \Delta I_{A2}{}^2 \rangle) \tag{2.2}$$

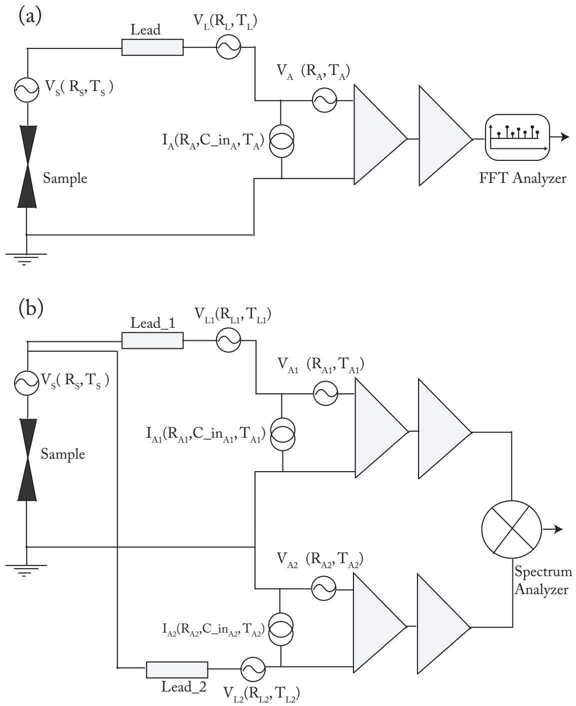

FIGURE 2.2: Low frequency noise setup. a) A Single channel noise measurement setup with a set of preamplifier and amplifier. b) A two channel noise measurement setup showing cross correaltion technique to measure the noise of the device below the noise floor of the preamplifier.

Over long measurement times, the uncorrelated noise reduces well below the other noise sources. The remaining noise is mainly due to amplifier's noise. The effective efficiency of cross correlation measurements is measured by a parameter called coherence. The coherence $R(f)$ is measure of the correlation between the two channels.It can be defined as :

$$|R(f)|^2 = \frac{S_{v12}^2(f)}{S_{v1}(f) * S_{v2}(f)}; \tag{2.3}$$

Where $S_{v1}(f)$, $S_{v2}(f)$ are the noise measured by channel 1 and 2 and $S_{v12}(f)$ is cross spectrum noise. The coherence value of the cross spectrum noise measurement lies between 1 and 0. Lower value indicates the less correlation between the two channel signal and value of 1 indicates the 100% correlated signal measured by both channel. Longer the integration time, higher the coherence in the measurement. Considering the $|R(f)| \sim 1$ being the ideal case for the cross spectrum analysis, N the number of averages, should satisfy the following condition[3]:

$$N \geq |\frac{1}{R(f)|^2} \tag{2.4}$$

The longer the averaging of data points, the lower will be the residual noise. The standard deviation of the residual noise can be defined as $\sigma_{V_{rcor}} = \langle {V_{rcor}}^2 \rangle / \sqrt{2N}$ where $\langle {V_{rcor}}^2 \rangle$ is residual correlated noise in cross spectrum measurement[4]. The sensitivity obtained by this method on $140k\Omega$ sample by Glattli *et al.*[5] is $S_v = 7.7 \times 10^{-20} \mathrm{V}^2/\mathrm{Hz}$ after 100s of averaging at 1kHz.

Limitation due to residual correlations between two channels :

The limitation to the cross spectrum analysis technique as seen by equation (2.2) is set by the residual correlation due to current noise of the preamplifier. Due to the ac coupling of the preamplifiers, its input voltage noise with the coupled becomes sample noise though the input capacitance *i.e.* C_{in} and stray capacitance*i.e.* C_{stray} of the coax connecting the sample to the preamplifiers. Considering both preamplifiers having the same noise behavior then residual correlation voltage noise could be written as[6][1]:

$$\langle {V_{rcor}}^2 \rangle = 2\langle {\Delta i_n}^2 \rangle {R_D}^2 + 2\langle {V_D}^2 \rangle \omega^2 {R_D}^2 \left(C_D + C_{stray} + C_{in}\right) C_{in} \tag{2.5}$$

If the current noise is small, the first term in equation (2.5) can be neglected, while the second term increases with frequency and becomes dominant with higher device impedance. This limits the use of this technique for high impedance at higher frequency. Generally, for higher impedances a current noise measurement scheme is employed where residual correlations are seen at the lower frequency end in contrast to the voltage noise measurement scheme. As per a rule of thumb for sample impedance ≥ 100kΩ a current noise measurement scheme is preferred and for sample impedance < 100kΩ a voltage noise measurement scheme with two channel cross correlation analysis is preferred. The exact cross over between the two measurement schemes is parameter dependent[2].

[1]Detail study has been done by Sampietro *et al.*[6].

[2]For our case, R_D, the sample impedance is in the $10k\Omega$ regime, hence we keep focus mainly on the voltage noise measurement technique in rest of the thesis.

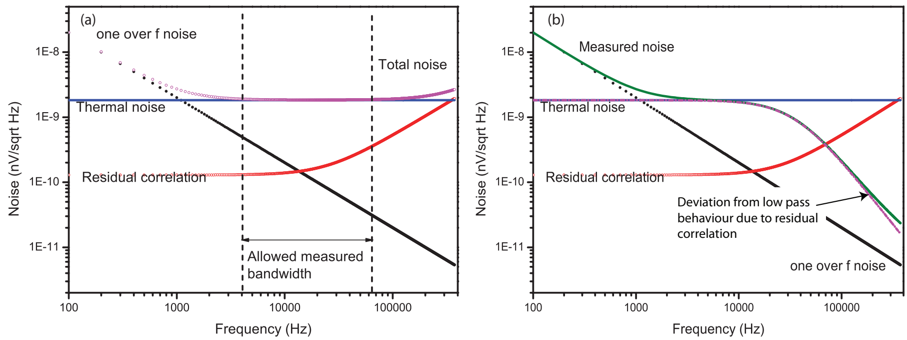

FIGURE 2.3: Residual correlation effect due to the current noise and voltage noise of the preamplifier: Using equation (2.5), residual correlated noise is calculated. Here $\langle \Delta i_n \rangle = 10\text{fA}/\sqrt{\text{Hz}}$, $\langle V_D \rangle = 2\text{nV}/\sqrt{\text{Hz}}$, $C_{in} = 50$pF, $C_{stray} = 300$pF, $R_D = 12.9\text{k}\Omega$ and $T = 4.7$K. a) Reduction in the bandwidth of noise spectrum due to residual correlation. (b)Residual correlation effect for the more realistic situation of a cryogenic set up, where the sample is situated far from preamplifier giving rise to the low pass behavior due to the stray capacitance of coaxes. A clear deviation from the low pass behavior is seen at 100kHz.

2.2 HIGH FREQUENCY MEASUREMENT TECHNIQUE

HIGH frequency measurement in MHz regime is useful for systems where $1/f$ noise limits the sensitivity of the low frequency noise measurement technique. In 2 dimensional electron gas systems and single-molecule systems, the corner frequency of the $1/f$ noise runs to a few hundred of kHz. The noise of the mesoscopic device is studied at sub cryogenic temperatures to avoid electron dephasing the high temperature phononic bath. The coaxes running from Kelvin sample bath temperature chamber to the room temperature low noise preamplifier have generally a capacitance of 100pF/m which gives an effective bandwidth of a few 10kHz. This can be easily seen from the figure (2.3b). The high sample impedance (12.9kΩ) combined with the cable capacitance (350pF) gives the low pass behavior with a cut off frequency of 35kHz which can be far below the $1/f$ noise corner frequency of a system like single molecule junction under high bias condition. To circumvent this technical issue, the stray capacitance needs to be reduced because the impedance of the sample is imposed by the experiment. This can be achieved by moving the low noise preamplifier very close to the sample. This puts a design constraint on the amplifier, which should be suited for the cryogenic environment. The Si based FET device have best noise performance but it suffers from the carrier freeze out at 77K. GaAs based FET suffers less from the carrier freeze out but their mobility decreases exponentially at the cryogenic temperature. Higher doping of the GaAs will not increase the mobility performance as the excess dopant atoms will act as sources of scattering and charge traps leading to high $1/f$ noise. One option is to mount them on the heat regulated finger but this will increase the heat load on the He bath. Alternative to this problem is to use

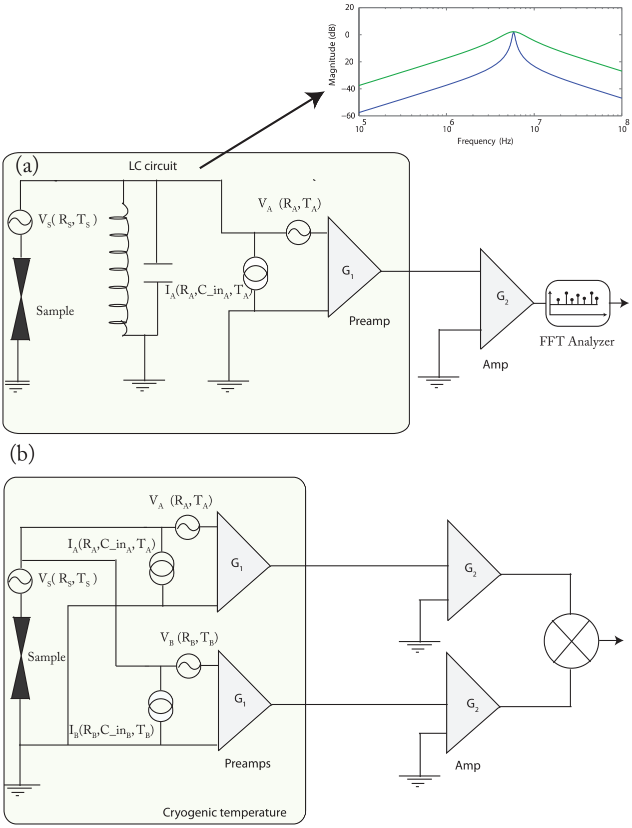

FIGURE 2.4: High frequency noise setup: a) A Single channel noise measurement setup with a chain of a low temperature preamplifier and a post-amplifier. A tank circuit is added to increase the central frequency of measurement band. Inset shows the typical measurement band of such system. Here parameter used for simulating this transfer functions are $R = 12.9\text{k}\Omega$, $C = 222\text{pF}$ and $L = 132.8\mu\text{H}$. This parameter gives sharp resonance peak at $f_0 = 5.8\text{MHz}$(Blue line). On reducing the C and L one could worsen the Q of tank circuit enabling the broadband measurement using tank circuit. The green line is obtained for $C = 22.2\text{pF}$, $L = 1328\mu\text{H}$ and $R = 12.9\text{k}\Omega$. b) A two channel noise measurement setup showing the cross correlation technique to measure the noise of the device below the noise floor of the preamplifier.

the high electron mobility transistor (HEMT). These are hetrojunction (GaAs-InAs) transistor. HEMT is suited best for the cryogenic amplifier due to its best noise and mobility performance[7]. The cryogenic amplifier are susceptible to the temperature drift and hence all the passive and specially active components needed to be thermalized properly. The proper thermalization ensures the stable operating point of the transistors. The HEMT amplifier are also preferred over Ge based FET due to its lower heat load. In order to reduce the heat load on the He bath, Adrian T. Lee *et al*[8] proposed the two stage amplifier. The cryogenic stage acts is cascode stage which acts as the impedance transformer while main gain comes from the room temperature part. The main advantage of the cascode stage is to reduces the Miller capacitance and hence it increases operational bandwidth. The main disadvantage of the two stage amplifier is high output impedance of the cryogenic amplifier. This give rise to reflection of the signal at the high frequency and hence giving rise to high frequency oscillation due to interference of reflected signals. Another alternative of the above design is to use the common-source gain stage with follower stage. This gives the better impedance matching at the cost of the bulkier design at the cryogenic end and higher heat load to He bath.

The design of rest of the high frequency noise set up is in analogous to the low frequency set up, apart from adding a cryogenic amplifier at the sample stage. Depending upon the sensitivity required by a single channel measurement or a two channels cross correlation method can be employed. A typical noise measurement setup is shown in figure (2.4) Achieving a broad spectrum, above few MHz can be achieved by just coupling the cryogenic amplifier very close to the sample. Even a few cm of wires coupling the sample to the preamplifier have a capacitance of a few pF. This limits the bandwidth of the noise measurement for high impedance samples to a low pass corner frequency of a few MHz. Adding a tank circuit (LCR) at the input of the amplifier increases the measurement frequency band to several of MHz. This technique can be employed with both two channel and single channel noise measurements. Noise is measured within the 3dB bandwidth around resonance peak. The sensitivity obtained by this technique is $1.1 \times 10^{-24} \mathrm{A}^2/\mathrm{Hz}$ [9] . It could be further improved to $6.7 \times 10^{-27} \mathrm{A}^2/\mathrm{Hz}$ by using the lockin technique[10].

2.3 VERY HIGH FREQUENCY NOISE MEASUREMENTS

VERY high frequency noise measurements are relevant for understanding the electron - phonon statistics in a quantum conductor. Detecting fluctuations of the order of a single electron detection is challenging. Even the cold amplifier having equivalent noise temperature of 40K with 50Ω input impedance will gives the amplifier input noise of $k_B T/50 \sim 10^{-23} \mathrm{A}^2/\mathrm{Hz}$. This brings the system design to the quantum limit. One of such design is shown in figure (2.5 a) which is based on M. Reznikov *et al.*[10]. In the frequency regime of GHz, all electronics component sizes are of the order of the wavelength of electromagnetic field.

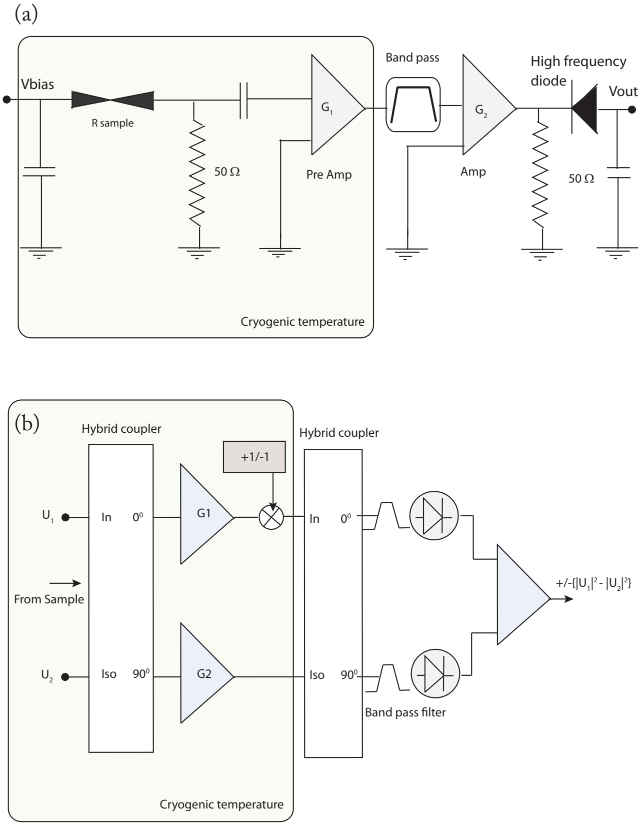

FIGURE 2.5: Very high frequency noise setup: a) A Single channel very high frequency noise measurement setup with a cryogenic low noise preamplifier and post-amplifier followed by a high frequency diode and capacitor load. b) A principle scheme showing the modulated double balanced amplifier set up.

Hence, all electronic circuit could be modeled as wave guides. A slight mismatch in the impedance leads to the reflection which further leads to instability of the electronics due to interference of reflected signals. In order to minimize such mismatch, all lines are 50Ω terminated. The high impedance sample is coupled to the 50Ω terminated input of a high frequency low noise cryogenic amplifier using a quarter wave impedance transformer. The very high frequency amplifier has a higher input noise temperature than the sample noise temperature which leads to measurements with very long integration times. Like in low and high frequency measurement techniques, the sensitivity of the noise measurement can be improved by using a lock in technique or cross-correlation technique. M. Reznikov *et.al.*[10] successfully used the lockin technique for noise measurement on quantum point contacts (QPC). The bias circuit is modulated with a low frequency signal and the amplified noise is synchronously measured with modulation signal. This improves the signal to noise ratio by the factor of $\sqrt{\Delta\tau \times \Delta\nu}$ where τ is integration time of the lock in amplifier and ν is measurement bandwidth. The sensitivity obtained in their experiment was $6.2 \times 10^{-27} \mathrm{A}^2/\mathrm{Hz}$. The sensitivity could be further improved by using a modulated balanced amplifier. Such technique was used by F.D. Parmentier *et al.*[11]. The key element in their setup is a 90° hybrid coupler. The sample is coupled to a cryogenic amplifier through a quadrature hybrid coupler.The coupler is 4 port device with two inputs in which the signal is equally splits in amplitude but 90° out of phase. Hence while measuring the difference across the output of coupler, amplifier noise is nullified. The sensitivity is further improved by modulating the biasing signal. The sensitivity obtained by F.D. Parmentier *et al.*[11] is an order improvement *i.e.*$2 \times 10^{-28} \mathrm{A}^2/\mathrm{Hz}$.

2.4 ON CHIP NOISE DETECTION

ON chip detection of noise is another approach in noise measurement techniques. All noise measurements mentioned above suffer from bandwidth problems, high frequency resonances and long integration times. If we follow the technological chronology of noise measurement setup: First amplifiers were used to increase the noise signal from the sample above the intrinsic noise of the spectrum analyzer or integrator, but this limits the measurement to the low frequency regime. Then amplifiers were put very close to the sample to reduce the stray capacitances and noise measurement in the MHz regime became possible. If we look here, all technological feats were done such that one could amplify the signal above the noise floor of the spectrum analyzer. If one could design a very low noise spectrum analyzer and coupled it directly to the sample this could solve above all technological problems. This is the basic motivation for the on chip detection. The idea of on chip detection was first suggested in D.E.Prober *et al.*[12] where the authors reported about the coupling of a quantum device directly to a quantum dot (QD) or a superconducting-insulator-superconductor(SIS) junction which acts as detector of the high frequency signal by converting it into a DC signal. This measurement provides a very large bandwidth, typically allowing to do frequency resolved measurements in GHz-THz regime. The current fluctuations in the quantum device induce fluctuations in capacitive environment coupling to SIS junction.This leads to photon assisted tunneling (PAT) of the quasi particle across the insulating barrier between the two superconductor[13]. The measures of this PAT gives the direct value of the noise in quantum device. In similar system, a quantum dot coupled to quantum device like quantum point contact can acts like on chip noise detector. The QPC current fluctuations induce the photo-ionization which drives the quantum dot out of the Coulomb blockade. This allows the sequential tunneling of the electron in the excited state. Hence the transient current through the quantum dot gives the information about noise in the quantum point contact[14]. The main advantages of on chip detection are fast and ultra high charge sensitivity *i.e.* $1.2\times10^{-5}\mathrm{e}/\sqrt{\mathrm{Hz}}$ as stated by R.J. Schoelkopf *et al.*[12].

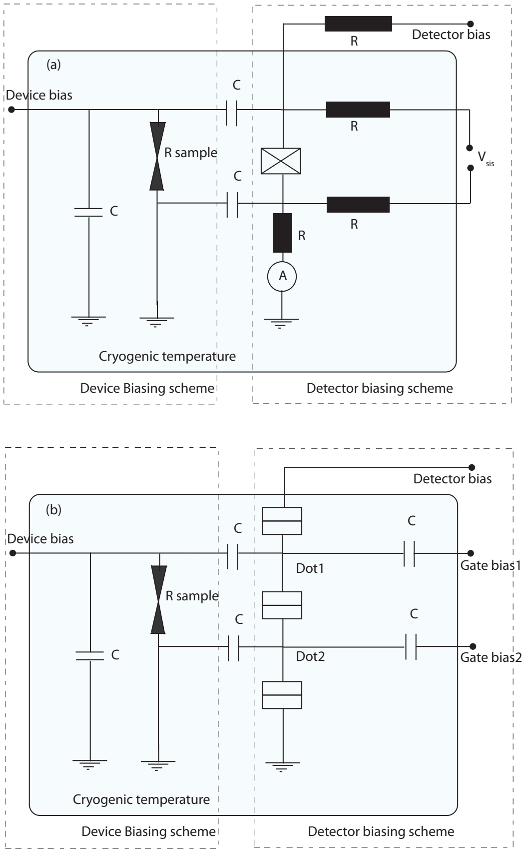

FIGURE 2.6: On chip noise detection noise setup: (a)A scheme showing electrical circuitry for the on chip noise detection using SIS junction capacitively coupled to quantum device. (b) On chip noise detection scheme using double quantum dot coupled to quantum point contact (Rsample).

Above both techniques are hampered by the low current levels. An improvement to above on chip noise detection schemes can be done by using double quantum dot capacitively coupled to the quantum point contact. This was first time proposed by Onac *et al.* [15]. The such measurement scheme for DQD coupled to quantum device is shown in figure (2.6). The current fluctuations in the quantum device induce fluctuations in the energy states of the DQD. This leads to the inelastic electron tunneling events. During inelastic tunneling events electrons exchange energy with the environment *i.e.* by absorption or emission of a photon. This inelastic tunneling current carries a measure of the fluctuations due to single tunneling of the electron with a photon. The modulation of the inter-dot barrier amplifies the inelastic current signal. The double dot system can be tuned to measure either emission or absorption spectra by tuning the energy difference between the dot in the bias window. The main advantage of this technique is wide detection window. The lower bound depends upon the inter-dot width of the resonant tunneling peak which is in the GHz regime and the upper end is given by the level spacing within the same dot which is in the THz regime. Both can be easily engineered to change the bandwidth. This concept was first realized by Gustavsson *et al.*[16] enabling them to do real time electron counting statistics. In another similar approach, Küng *et al.* used the above technique to do the cross correlation measurement[17]. This techniques enable them to measure the shot noise in μV regime. This techniques allows them to measure the third moment of full counting statistics on the QPC[18].

In this thesis we will use the low frequency noise measurement using the cross spectrum technique for noise measurement on the atomic contact. This is discussed in the chapter Shot noise measurement circuit . We also worked on the development of the high frequency noise measurement based on the Two channel cross correlation technique . This is discussed in the chapter The cryogenic amplifiers. Here we discussed the complete design of the noise measurement setup suited for the mechanical break junction. Criticality of both measurement scheme lies on the sensitiveness of our atomic and molecular junction on electro mechanical noise. The high frequency noise measurement setup is ongoing work and still needed to be tested on the atomic contacts.

REFERENCES

[1] L. K. J. Vandamme, in *Advanced Experimental Methods for Noise Research in Nanoscale Electronic Devices, Proceedings of the NATO Advanced Research Workshop, Nato Science Series II* (2003), vol. 151, pp. 1–370, URL http://rd.springer.com/book/10.1007/1-4020-2170-4/page/1.

[2] F. Crupi, G. Giusi, and C. Pace, in *Instrumentation and Measurement technology conference IMTC 2007* (Warsaw, 2007), pp. 1–4.

[3] J. Briaire and L. K. J. Vandamme, *Uncertainty in Gaussian noise generalized for cross-correlation spectra*, Journal of Applied Physics **84**, 4370 (1998).

[4] G. Ferrari and M. Sampietro, in *Advance Experimental methods for noise research in nanoscale electronics devices* (2003), 978-1-4020-2170-1, URL http://rd.springer.com/book/10.1007/1-4020-2170-4/page/1.

[5] D. C. Glattli, P. Jacques, A. Kumar, P. Pari, L. Saminadayar, and I. Introduction, *A noise detection scheme with 10 mK noise temperature resolution for semiconductor single electron tunneling devices*, Journal of Applied Physics **81** (1997).

[6] M. Sampietro, L. Fasoli, and G. Ferrari, *Spectrum analyzer with noise reduction by cross-correlation technique on two channels*, Review of Scientific Instruments **70**, 2520 (1999).

[7] A. T. Lee, *Broadband cryogenic preamplifiers incorporating GaAs MESFETs for use with low temperature particle detectors*, Review of Scientific Instruments **60**, 3315 (1989).

[8] A. T. Lee, *A low-power-dissipation MESFETs in parallel broadband cryogenic preamplifier utilizing GaAs*, Review of Scientific Instruments **64**, 2373 (1993).

[9] F. Wu, L. Roschier, T. Tsuneta, M. Paalanen, T. Wang, and P. Hakonen, *Setup for shot noise measurements in carbon nanotubes*, AIP Conference Proceedings **850**, 1482 (2006).

[10] M. Reznikov, M. Heiblum, H. Shtrikman, and D. Mahalu, *Temporal Correlation of Electrons: Suppression of Shot Noise in a Ballistic Quantum Point Contact*, Phys. Rev. Lett. **75**, 3340 (1995).

[11] F. D. Parmentier, A. Mahe, A. Denis, J.-M. Berroir, D. C. Glattli, B. Placais, and G. Feve, *A high sensitivity ultralow temperature RF conductance and noise measurement setup*, Review of Scientific Instruments **82**, 013904 (pages 8) (2011).

[12] R. Schoelkopf, P. Wahlgren, A. Kozhevnikov, P. Delsing, and D. Prober, *The radio-frequency single-electron transistor (RF-SET): A fast and ultrasensitive electrometer*, Science (New York, N.Y.) **280**, 1238 (1998).

[13] R. Deblock, E. Onac, L. Gurevich, and L. P. Kouwenhoven, *Detection of Quantum Noise from an Electrically Driven Two-Level System*, Science **301**, 203 (2003).

[14] E. Onac, F. Balestro, L. H. W. van Beveren, U. Hartmann, Y. V. Nazarov, and L. P. Kouwenhoven, *Using a Quantum Dot as a High-Frequency Shot Noise Detector*, Phys. Rev. Lett. **96**, 176601 (2006).

[15] R. Aguado and L. P. Kouwenhoven, *Double Quantum Dots as Detectors of High-Frequency Quantum Noise in Mesoscopic Conductors*, Phys. Rev. Lett. **84**, 1986 (2000).

[16] S. Gustavsson, M. Studer, R. Leturcq, T. Ihn, K. Ensslin, D. Driscoll, and A. Gossard, *Frequency-Selective Single-Photon Detection Using a Double Quantum Dot*, Physical Review Letters **99**, 2 (2007).

[17] B. Küng, O. Pfäffli, S. Gustavsson, T. Ihn, K. Ensslin, M. Reinwald, and W. Wegscheider, *Time-resolved charge detection with cross-correlation techniques*, Phys. Rev. B **79**, 035314 (2009).

[18] S. Gustavsson, R. Leturcq, B. Simovič, R. Schleser, T. Ihn, P. Studerus, K. Ensslin, D. C. Driscoll, and A. C. Gossard, *Counting Statistics of Single Electron Transport in a Quantum Dot*, Phys. Rev. Lett. **96**, 076605 (2006).