REVIEW OF NOISE IN SEMICONDUCTOR DEVICES AND MODELING OF NOISE IN SURROUNDING GATE MOSFET

END–QUARTER REPORT OF THE WORK DONE BY

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Contents

Introduction

The sensitivity of electrical systems is limited by noise. A very important source of noise in electronic systems is the electronic devices that form the heart of the signal processing and transmission components in these systems. These are irreducible sources of noise, and it is very important to realize their properties and characteristics.

This report is divided into three parts. The first part of this report comprises of a brief outline of various forms of electronic noise sources. This is based on a brief literature survey aimed at gaining a beginner's understanding of the various sources of noise, their origin and their characteristics. These are by no means detailed or mathematical, only a concise understanding gained by leafing through various textbooks and some journal publications are outlined in this report.

The second part consists of a brief overview of the structure of surrounding gate MOSFET and the basics of the noise modeling of the device.

The third part comprises of the simulation of noise in the surrounding gate MOSFET based on the experimental data and the results of MATLAB simulations.

Mathematical Background on Noise

Noise is a random process. The value of noise fluctuation cannot be predicted at any time even if the past values are known. However, in many practical systems, the average power of noise is predictable, and hence, statistical models can describe noise. The average power of a voltage signal *x(t)* is defined as

$$P_{av} = \lim_{T\to\infty} \frac{1}{T} \int_{-T/2}^{T/2} x^2(t)dt \qquad (1)$$

where P_{av} is expressed in V^2 [1].

The concept of noise power becomes more versatile if defined with regard to the frequency component of noise. The Power Spectral density, $S_x(f)$ (PSD) of a noise signal, *x(t)* is defined as the average power carried by *x(t)* in a one-hertz bandwidth around *f* . PSD is a very powerful tool in analyzing the effect of noise in systems. If a signal with PSD $S_x(f)$ is applied to a LTI system with transfer function *H(s)*, the PSD of the output signal is given by

$$S_y(f) = S_x(f)\left|H(f)\right|^2 \qquad (2)$$

It is also instructive to note that the autocorrelation function of a signal and its Power Spectral Density are Fourier Transform pairs.

$$S_X(w) \Leftrightarrow \overline{X(t)X(t+t)} \qquad (3)$$

Another important result used in Noise analysis is the Carson's theorem [2] which states that the spectral intensity of a stationary random variable *Y(t)* which is the superposition of a large number of independent events, $F_i(t)$ occurring at random at the average rate λ is twice the product of λ and the magnitude squared of the fourier transform of the independent event, *y(f)*.

i.e., if $Y(t) = \sum_i F(t - t_i)$ then $S_Y(f) = 2l\left|y(f)\right|^2$ (4)

Noise in Semiconductor Devices

A brief description of the fundamental types of noise observed in semiconductor devices is described in the following sections.

Thermal Noise

A conductor in thermal equilibrium with its surroundings shows, at its terminals, an open-circuit voltage or short-circuit current fluctuation. M. B. Johnson was the first to report careful measurements of thermal noise in 1927[3]. He discovered that the open circuit voltage noise power spectral density of the conductor is independent of the material of the conductor and the measurement frequency, and is determined only by the temperature and electrical resistance

$$\overline{S_v(w)} = 4KTR \qquad (5)$$

The corresponding short-circuit current noise spectral density is given by

$$\overline{S_I(w)} = 4KT / R \qquad (6)$$

This noise is referred to as thermal noise and is the most fundamental and important noise in electronic devices.

The physical origin of the thermal noise in a macroscopic conductor is a “random-walk” of thermally excited electrons [4]. An electron undergoes a Brownian motion via collisions with the lattices of a conductor. The fundamental properties of a Brownian particle were first studied by Albert Einstein and then formulated by M. P. Langevin twenty years before Johnson’s observation of thermal noise. The electrons in a conductor are thermally energetic via collisions with the lattice and travel randomly. The electron velocity fluctuation is a statistically stationary process. However, the mean-square displacement of an electron increases in proportion to the observation time. The electron position fluctuation is a statistically non-stationary process. Such a microscopic approach can indeed explain Johnson’s observation.

Nyquist employed a completely different approach to the problem. He introduced the concept of “electromagnetic field modes” as a degree-of-freedom (DOF) of the system by assuming a transmission line terminated by two conductors. He then applied the equi-partition theorem of thermodynamics to the transmission line modes. In this way he could explain Johnson’s observation without going into the details of a microscopic electron transport process.

Johnson-Nyquist thermal noise is the intrinsic property of a conductor at thermal equilibrium; that is, when there is no applied voltage and no net current (energy flow) in the system. However, the Johnson-Nyquist thermal noise formula is experimentally known to be valid even when there is a finite current flow across the conductor.

It also turns out that the spectral density of thermal noise actually increases with frequency, rather than remaining constant. This result follows from a more detailed analysis taking into account the actual distribution of carrier energies, modified by considerations related to Heisenberg’s Uncertainty Principle. Similar to the impossibility of simultaneous and accurate measurement of the velocity and position of an electron, it is impossible to measure accurately the current through and the voltage across a resistor simultaneously. Nyquist’s approach for determining therma noise is very general and is easily extended to include quantum noise. Callen and Welton established the microscopic theory of quantum noise in a conductor. The general expression for thermal noise voltage [4] based on these considerations is

$$\overline{S_v(w)} = 2\hbar wR \coth\left(\frac{\hbar w}{2kT}\right) \tag{7}$$

At room temperature, however, this expression has the same value given by equation (5) till around f = 2kT/h = 10THz.

Since FETs are essentially voltage-controlled resisters, they exhibit thermal noise. In the triode region of operation, detailed theoretical considerations lead to the expression for the drain current noise [5] of FETs as:

$$\overline{i_{nd}^2} = 4kTgg_{do}\Delta f \tag{8}$$

where g_{do} is the drain-source conductance at zero V_{ds}. The parameter γ has a value of unity at zero V_{ds}, and in long devices, decreases towards a value of 2/3 in saturation. Note that the drain current noise at zero V_{ds} is precisely that of an ordinary conductance of value g_{do}.

Measurements show that the short-channel MOSFET devices in saturation exhibit noise far in excess of values predicted by long-channel theory, sometimes by large factors (γ is typically 2-3, but can be larger). It has been established [6] that the origin of this excess noise is carrier heating by large electric fields present in short devices. The non-local transport behavior causes a small derivative of velocity with respect to the electric field. The resulting higher local ac resistance near the source junction increases the impedance field and is directly reflected in excess noise and strong gate length dependence.

In addition to drain-current noise, the thermal agitation of channel charge has another consequence: gate noise. The fluctuating channel potential couples capacitively into the gate terminal, leading to a noisy gate current. Although this noise is negligible at low frequencies, it dominates at radio frequencies. Van Der Ziel has shown that gate noise can be expressed [2] as

$$\overline{i_{ng}^2} = 4kTdg_g\Delta f \tag{9}$$

Where the parameter g_g is given by

$$g_g = \frac{w^2 {C_{gs}}^2}{5g_{do}} \tag{10}$$

Shot Noise

Another noise mechanism, known as shot Noise was first described and explained by W. Schottky in 1918. The term "shot" is not a corruption of "Schottky"; it is simply that if one hook up an audio system to a source of shot noise biased at low currents, the resulting sound is much that like of buck-shot (pellets) dropping into a hard surface [3].

The fundamental basis for shot noise is the granular nature of electronic charge. Two conditions must be satisfied for shot noise to occur. There must be a direct current flow and there must also be a potential barrier over which the charge carriers hop. The second condition implies that ordinary linear resisters do not generate shot noise, despite the quantized nature of the electronic charge.

The fact that charge comes in discrete bundles means that there are discontinuous pulses of current every time an electron hops an energy barrier. It is the randomness of the arrival times that give rise to the whiteness of shot noise. Shot noise is given by the formula

$$\overline{i_n^2} = 2qI_{dc}\Delta f \qquad (11)$$

where I_{dc} is the DC current in amperes

Shot noise also is ideally white, and has amplitude that possesses a gaussian distribution. The requirement for a potential barrier implies that shot noise will be only associated with non-linear devices, although not all non-linear devices exhibit shot noise. For example, whereas both the base and collector currents are sources of shot noise in a bipolar transistor because potential barriers are definitely involved there, only the DC gate leakage current of FETs contributes shot noise.

Traditionally, the gate oxide of a MOSFET has been considered as a perfect barrier for carriers allowing no current flow between the gate and silicon. In fact, there is tunneling of electrons from the vicinity of the electrode Fermi level through the forbidden energy gap into the conduction band of the oxide. Such a phenomenon is called Fowler- Nordheim Tunneling [7] and its current density can be expressed as

$$J = \frac{q^3E^2}{8phf}\exp\left[-\frac{4\sqrt{2mf^3}}{3\hbar qE}\right] \qquad (12)$$

where h is Planck's constant, q is the electronic charge, E is the electric field in the gate oxide, ϕ is the barrier, and m is the free electron mass. The perfect barrier assumption has been valid in most practical situations because the Fowler-Nordheim Tunneling current has been negligibly small. However, ultra-thin oxides below 4 nm exhibit drastic increase of leakage current, so called direct tunneling current. In this regime, the gate oxide capacitor would introduce an extra noise current source, possibly a shot noise current source, besides two classical noise sources: drain and gate current noise. Fortunately, the IR drop along the gate polysilicon due to the leakage current is negligible; also, the additional conductances (1/rgs and 1/rgd) associated with this tunneling across the gate oxide are small compared with ωC_{gs} and ωC_{gd} in the range above the 1/f corner frequency, which is usually few MHz in MOSFETs.

By contrast, the impact of the direct tunneling current on high frequency noise performance is becoming critical. The gate shot noise current generated in each segment of the MOSFET flows along the channel and subsequently creates drain shot noise current as well, because it is uncorrelated with the origins of the drain and gate current noise. Since the direct tunneling current can be substantial, the drain shot noise becomes comparable to the drain current noise in MOSFETs with oxides below 2 nm. While a rigorous modeling of the direct tunneling current is prerequisite to accounting for this effect, accurate modeling of tunneling in MOSFETs involves evaluation of the multi-dimensional Schrodinger equation – an unsolved problem to date.

Flicker Noise

Though no universal mechanism has been identified for flicker noise or 1/f noise, it is the most ubiquitous form of noise in nature [3]. Phenomena that have no obvious connection like heartbeat, cell membrane potential, financial data, DNA sequences and transistors exhibit fluctuations with a 1/f character.

As the term “1/f” suggests, a spectral density that increases without limit as frequency decreases, characterizes this kind of noise. The practical implication of this statement is that the total mean-squared noise depends on the logarithm of the ratio of the frequencies, rather than the difference between the frequencies as in thermal noise and shot noise. This implies that “huge” noise powers that might dominate signal power will occur only at frequencies low enough that corresponds to time scales much larger than the expected life times of practical electronic systems.

Flicker Noise shows up in Resistors, when it is called “excess noise”, since this noise is in addition to what is expected from thermal noise considerations. It is found that a resistor exhibits 1/f noise only when there is DC current flowing through it, with the noise increasing with the current. This is widely observed in carbon composition resistors, and the source of noise has been attributed to the formation and extinction of “micro-arcs among neighboring carbon granules.

In spite of more than 30 years of research, there exits no unique model for 1/f noise occurring in natural systems. Originally, it was thought that 1/f noise in semiconductor devices is a surface phenomenon, and is related to the Si-SiO_2 interface. Evidence has been advanced showing a good correlation with density of interface or of near oxide traps. The models based on this information fall under what is called the number fluctuations or ΔN theories, based on the McWhorter theory. The McWhorter model was proposed for 1/f noise in germanium and assumes that origin of fluctuations is the tunneling of charge carriers at the semiconductor surface to and from traps, which are located close to the interface. The expression for input referred noise spectral density is

$$\overline{S_{Vg}(f)} = \frac{S_{Id}}{{g_m}^2} = \frac{kTq^2}{8WL{C_{ox}}^2 a_t}\frac{N_{ot}(E_f)}{f} \qquad (13)$$

where α_t is the tunneling parameter and $N_{ot}(E_f)$ is the oxide trap density [8].

Opposite to the ΔN model is the so-called Δμ model, which considers mobility fluctuations at the origin of 1/f noise, and for a homogenous semiconductor, assumes a volume and not a surface origin for this noise. It is purely empirical in nature and has been proposed to explain 1/f noise in resistors provided with low-noise ideal Ohmic contacts. It has been proposed that the fluctuation of bulk mobility in MOSFETs is induced by fluctuations in phonon population through phonon scattering.

$$\overline{S_{Vg}(f)} = \frac{q}{WLC_{ox}}\frac{a_H(V_{gs} - V_t)}{f} \qquad (14)$$

where a_H is a material parameter given by

$$a_H = 2\times 10^{-3}\left(\frac{m}{m_{latt}}\right)^2 \qquad (15)$$

m being the carrier mobility and m_{latt} the mobility due to lattice scattering only.

Recently, Hung et al [9] have proposed a unified model for Flicker noise, based on the investigation of random telegraph noise (to be explained later) in sub-micron MOSFETs, which revealed that the charge fluctuations in the oxide traps generate noise by modulating the carrier mobility, in addition to the carrier number. The mobility fluctuation is attributed to Columbic scattering effect of the fluctuating oxide charge. This model has a functional form resembling that based on the conventional number fluctuation model, but at certain bias conditions, it can be reduced to the form compatible with bulk mobility fluctuation model.

E. Terzioglu et al [10] have extended the unified noise model to high lateral field conditions to explain the decrease in total device noise in short channel nMOSFETs with strong DIBL due to the suppression of noise induced by mobility perturbations at high lateral electric field.

Random Telegraph Signals

Another type of noise that plagues semiconductors is known as Random Telegraph Signals (also called burst noise, bistable noise and popcorn noise) [3]. It was first observed in point-contact diodes, but has also been seen in ordinary junction and tunnel diodes, some types of resistors and both discrete and integrated circuit junction transistors. This kind of noise is characterized by bi-modal, and hence non-gaussian amplitude distribution. That is, the noise switches between two or more discrete values at random times. The switching intervals tend to be in the audio range (~10µs) and the popping sound that is heard when a burst noise source is connected to an audio system is why this is also known as popcorn noise.

The so-called burst noise in reverse-biased p-n junctions and bipolar transistors is an example of discrete switching behavior in electronic devices. Although first observed nearly thirty years ago, the origins of burst noise still remain uncertain; dislocations, metal precipitates and the switching on and of surface conduction channels have all been implicated in its production. Generation-Recombination Noise observed in semiconductor devices is a typical example of RTS.

It has been found that that as the device area is scaled down, the total number of Si/SiO2 interface defects is correspondingly reduced. In small enough devices it is quite likely that only a handful of traps will have energy levels within *kT* or so of the surface Fermi level and thus will be fluctuating in occupancy. K. Kandiah and F. B. Whiting observed that resistance changes are consistent with a single electron being removed from the channel and captured in a localized defect state [4]. It was also noted that as the gate voltage changes the mark-space ratio changes as the separation of the trap energy level and surface Fermi level is altered. It is also quite clear that the switching rate is a sensitive function of temperature. In addition, one can see that at elevated temperatures, where several RTSs are active, the resistance fluctuations are beginning to resemble the trace one would observe for a l/f noise source.

The noise generated due to the trapping of electrons due to a single trap in the gate oxide of a MOSFET can be modeled as an RTS, and the power spectral density of a random telegraph signal X(t) is a Lorentzian given by

$$S_X(f) = \left\langle (\Lambda X)^2 \right\rangle \frac{4t}{1+(2pft)^2} \qquad (16)$$

where τ is the average rate of a trapping of an electron. It has been established that the superposition of Lorentzians corresponding to a log uniform distribution of $1/\tau$ leads to a 1/f noise spectrum [12].

Surrounding Gate MOSFET Device Structure

The surrounding-gate transistor is a variation on the traditional planar MOSFET. Like the planar MOSFET, a gate separated from the channel by an oxide modulates the current flowing through a channel between source and drain. Unlike the planar MOSFET, the structure is vertical rather than horizontal and the channel covers the three-dimensional surface area of the silicon pillar on which it is built. The surrounding-gate transistor is similar to SOI and dual gate MOSFETs in that the silicon film can be either partially or fully depleted. The silicon on which the device is built can be cylindrically symmetric giving rise to a cylindrical surrounding-gate transistor (Fig 1), it can be elongated in one horizontal direction such that the channel surrounds a fin of silicon (Fig 2), or it can take on any other shape such that the gate and channel surround the structure while drain and source remain at top and bottom.

Analytical and numerical simulations of the cylindrical surrounding-gate transistor show that the device has steeper sub-threshold slope, higher electron mobility in weak inversion, and increased sheet electron concentration in inversion when compared to an equivalent planar device when the diameter of the pillar is reduced below about 0.5µm. A scaling theory for fully depleted cylindrical surrounding-gate transistors has been derived showing that the natural length in Poisson's equation for these devices is 30 to 35% less than the natural length for dual gate devices [11]. This means that to maintain a given sub-threshold slope and DIBL characteristic at the same silicon and oxide thickness as in a dual gate device, the cylindrical device can have a shorter effective gate length. Finally, because of its vertical structure, the footprint of a surrounding-gate transistor is decoupled from the gate length, allowing for a smaller footprint than an equivalent planar device. Although a cylindrical surrounding-gate transistor has a limited drive current due to its narrow width, multiple pillars may be used in parallel to increase drive current and for use in high-speed applications.

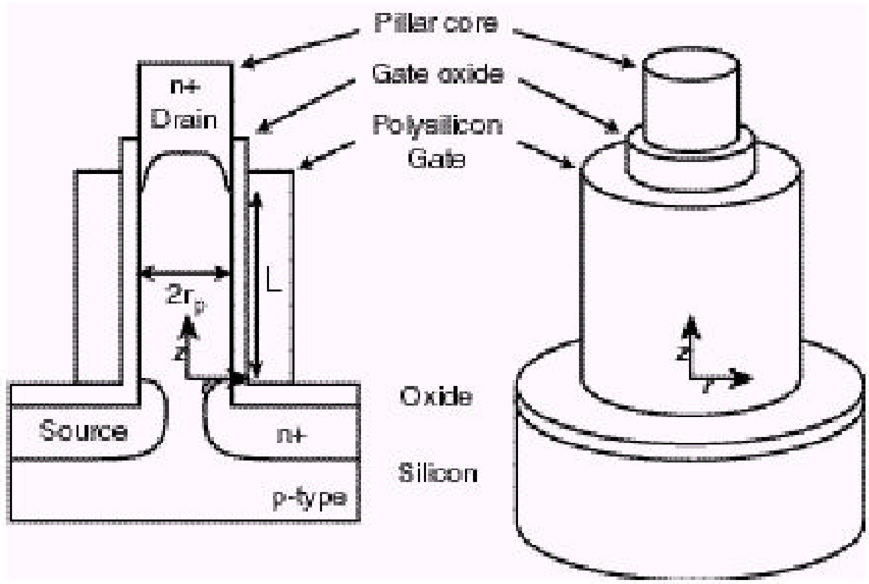

Fig 1: Schematic of the Cylindrical Surrounding Gate MOSFET

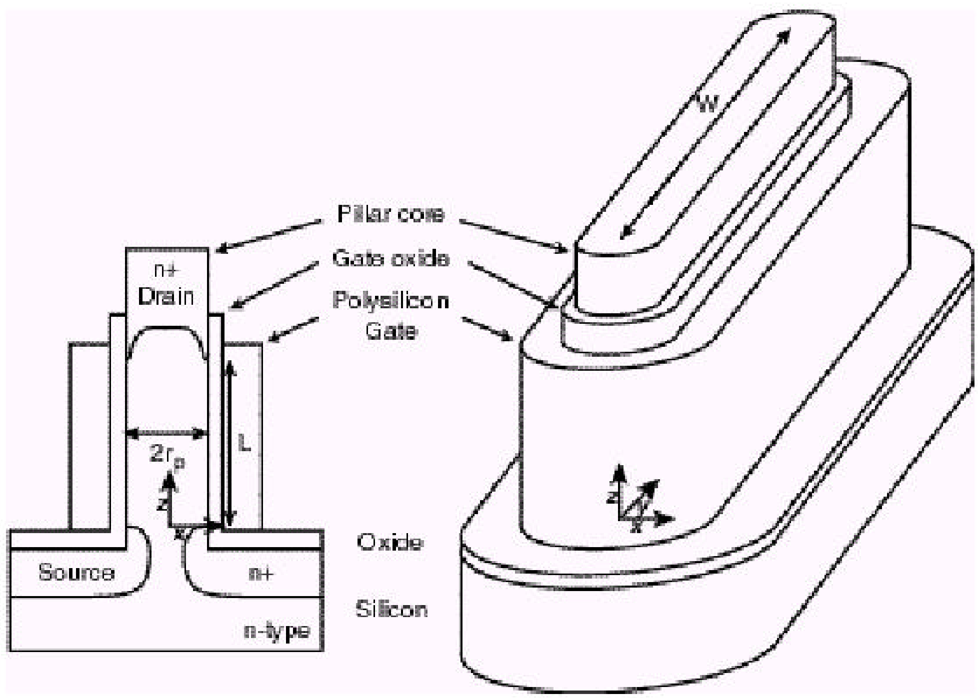

Fig 2: Schematic of the fin surrounding gate Transistor.

Noise Modeling of the SGT

Flicker Noise in surrounding gate MOSFETs is generated when carriers interact with the nearby traps in the gate oxide. The capture and emission of electrons lead to fluctuation in the number of carriers as well as in their mobility. The noise generated by a single active trap in time domain is a Random Telegraph Signal, which is equivalently represented by a Lorentzian in frequency domain. When there are multiple traps in the oxide, the net noise is the superposition of different Lorentizians leading to a 1/f noise spectrum. It is possible to build Surrounding Gate Transistors with area low enough such that the number of traps in the device can be less tan one (zero-trap device) [13].

The power spectral density of the drain current fluctuations in a cylindrical surrounding-gate transistor is derived following the unified model of Hung *et al*. The first step is calculating the fluctuations in drain current as a function of the fluctuations in oxide trap occupancy, taking into account the direct change in the number of channel carriers and the correlated change in mobility due to scattering from charged traps [14].

The spectrum of the trap occupancy fluctuations is calculated from the variance of the fluctuations by the Wiener-Kintchine theorem, which relates the spectrum of a fluctuating variable to its autocorrelation function. Calculation of this spectrum requires knowledge of the trap time constant τ and the variance of the fluctuations.

The time constant for electrons in oxide traps is calculated in a manner that is similar to the generation-recombination noise using Shockley-Read-Hall statistics. The capture and emission rates for electrons in the cylindrical n-channel surrounding-gate transistor are calculated in the case of either a distribution of traps through the oxide or a distribution of activation energies. Then the change in trap occupancy with time is found from the difference between the capture and emission rates. Finally, the differential equation for the fluctuations is solved to yield τ.

The variance of the trap occupancy fluctuations is calculated from the principles of Thermodynamics; the variance being equal to kT times the partial derivative of the number of electrons with respect to the chemical potential at constant temperature. The electrons are assumed to have a Fermi-Dirac distribution about the quasi-fermi level and the interaction of the channel electrons with traps of different time constants due to location and activation energy is taken into account by summing up the spectrum for each time constant's contribution.

Using the autocorrelation function of the fluctuations and integrating over all trap energies and locations, the final drain current spectrum is derived. Short channel effects are included, making the theory appropriate for 0.1µm cylindrical surrounding-gate transistors.

Simulations and Results

The total noise of a depletion mode fin is the sum of the contributions from the rectangular and cylindrical regions $S_{idtot} = S_{idr} + S_{Idc}$

A MATLAB Code has been written to simulate the noise of the rectangular region of the depletion mode fin of the SGT.

Neglecting the effect of trap activation energy, using the delta function nature of $f_t(1-f_t)$ and approximating the trap density to be independent of position makes the current noise spectrum for a rectangular depletion mode device

$$\frac{S_{I_{dr}}(f)}{I_{dr}^{2}} = \frac{W}{L^2} kTN_t \int_0^L \int_{r_p}^{r_p+t_{ox}} \left(\frac{g}{n_L(z)} + \frac{\frac{a}{W} m_{eff}}{1+(m_{eff}E_z/v_{sat})^2} \right)^2 \frac{4t(x,z)}{1+(2pft(x,z))^2} dxdz \qquad (17)$$

where

$$I_{dr} = qn_L v \qquad (18)$$

$$n_L(z) = W \int_{r_p - l_{\text{int}}}^{r_p} n(x,z) dx \qquad (19)$$

$l_{\text{int}} = m_{eff} m^* v_{sat} / q$ (mean free path from total mobility) (20)

$$N_{av} = \frac{n_L W}{l_{\text{int}}} \qquad (21)$$

$$m_{eff}(z) = v(z)/E_z(z) \qquad (22)$$

$$g = \frac{C_{nL}}{C_{Dl} + C_{itL} + C_{oxL} + C_{nL}} \qquad (23)$$

$$C_{nL} = -x \frac{q^2}{kT} n_L \qquad (24)$$

$$C_{oxL} = \frac{e_{ox} W}{t_{ox}} \qquad (25)$$

$$C_{itL} = qN_{it}W \qquad (26)$$

$C_{DL} = 0$ in full depletion (27)

$$t(x,z) = t_0(z)\exp(h(x - r_p)) \qquad (28)$$

$$t_0(z) = \frac{1}{c_n n(r_p, z)} \tag{29}$$

$$a = \frac{1.755\times10^{-16}\sqrt{\left(\frac{300}{T}\right)}(n_{av})^{-1/4}}{l_{perp}} \tag{30}$$

$$l_{perp} = l_{schrieffer} + 2\frac{1.73\times10^{-5}}{E_{eff}^{\ 1/3}} \tag{31}$$

$l_{schrieffer} = l_{noperp}(1-\exp(s^2)erfc(s))$ (Mean free path calculated from Schrieffer's model) (32)

$$s = \sqrt{2m^* kT}\frac{v_{sat}}{qE_z l_{noperp}} \tag{33}$$

$l_{noperp} = m_{noperp} m^* v_{sat} / q$ (Mean free path without the effect of the perpendicular field) (34)

$$m_{noperp}(z) = m_{eff}(z)\sqrt{1+\frac{E_{eff}(z)}{E_c}} \tag{35}$$

$$m_{nopar}(z) = m_{eff}(z)\frac{1}{\sqrt{1-\frac{v(z)}{v_{sat}}}} \tag{36}$$

Finally, after the integral was evaluated, the noise power was fit to an equation of the form

$$\frac{S_{I_{dr}}(f)}{I_{dr}^{\ 2}} = \frac{A}{f^n} \tag{37}$$

The noise power was evaluated at varying gate bias voltages. The results of the simulations are as follows.

The parameter n was found to lie between 1.0004 and 1.0009 for gate voltage variations from 0.4V to 1.2V

The parameter A was found to decrease when Gate Voltage is increased. This might be due to the fact that n_l was found to increase as gate voltage increased. The variation of A with V_g is shown in Fig 3.

The variation of the other parameters along the channel for different gate bias voltages is shown in Figures 4 to 14.

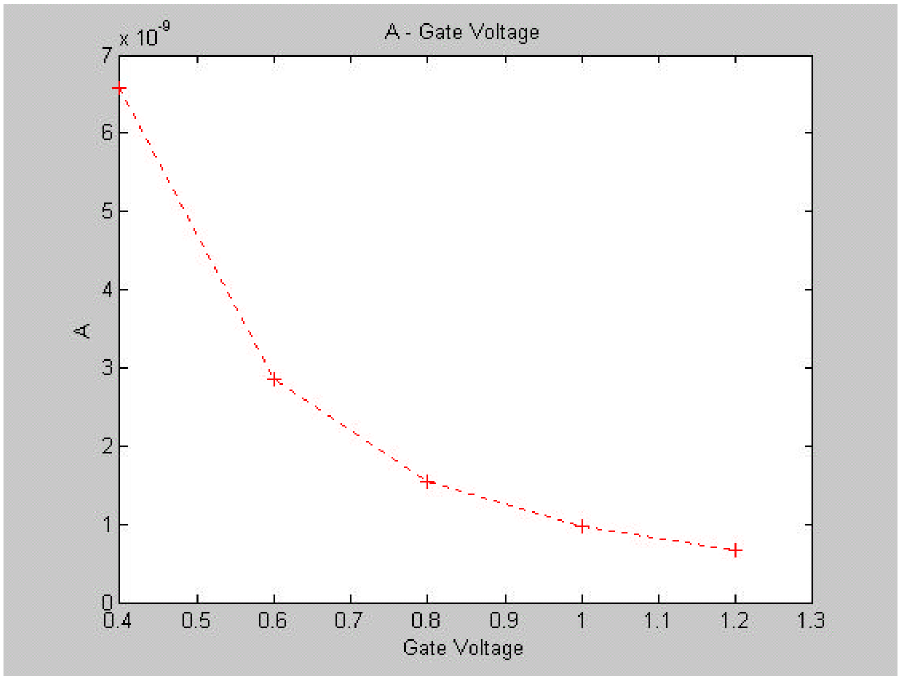

Figure 3. A - V_g

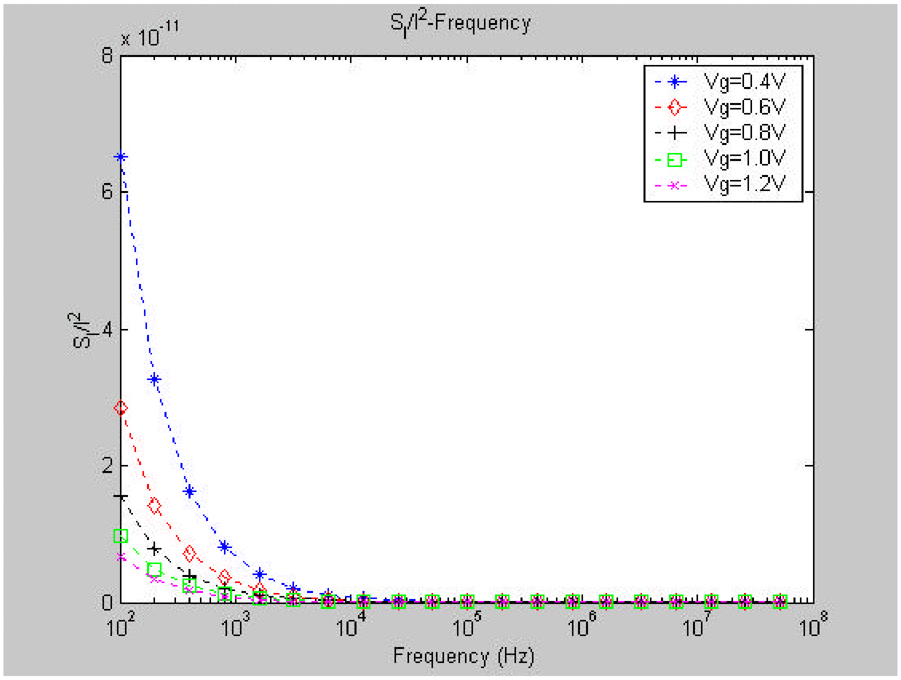

Fig 4. (S_I/I^2) - F for various V_g

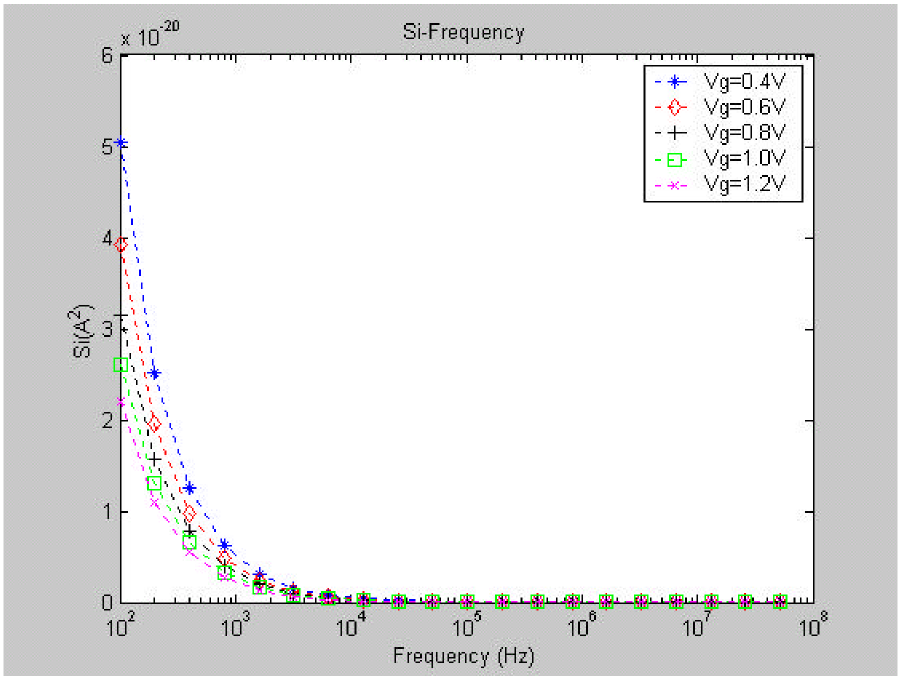

Fig 5. S_i-F

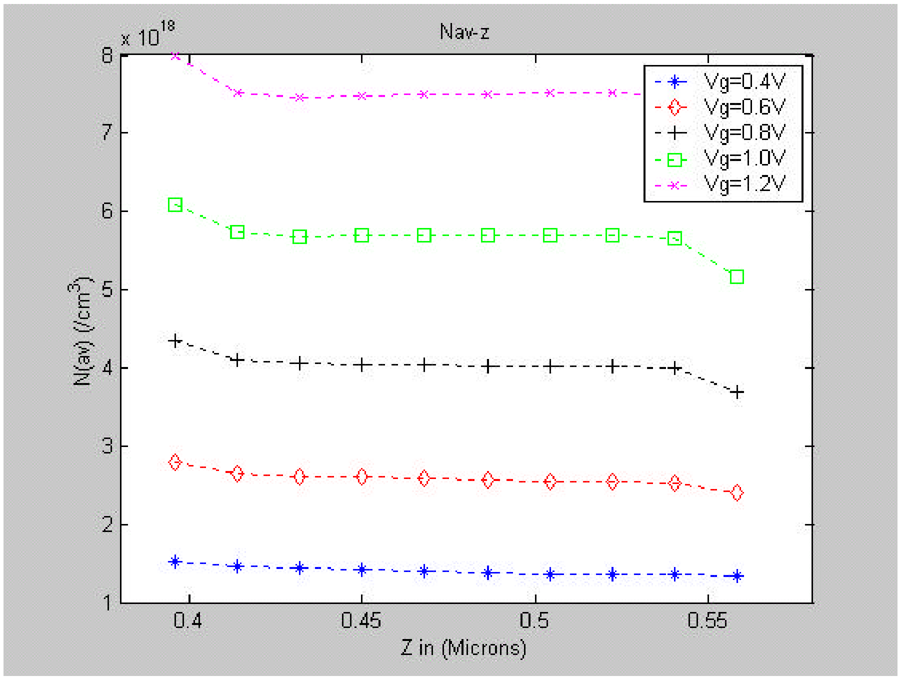

Fig 6. N_{av} -Z

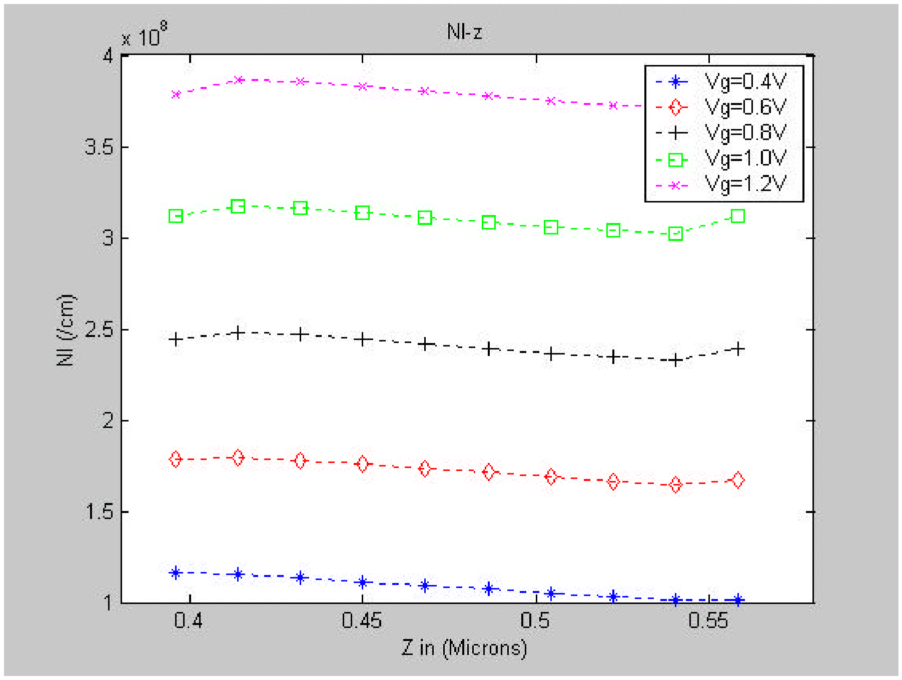

Fig 7. N_l –Z

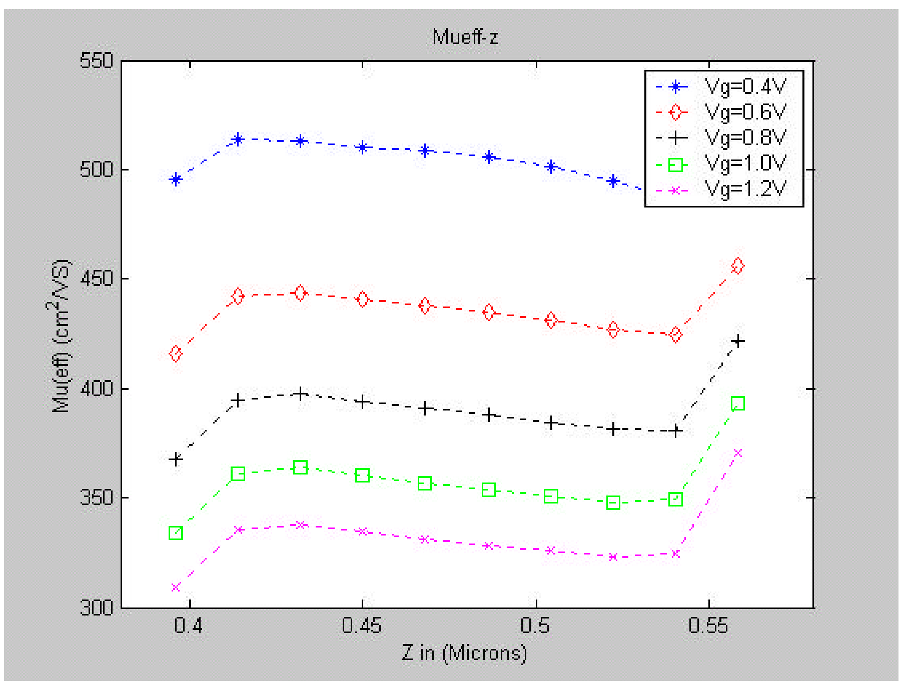

Fig 8. μ_{eff} - Z

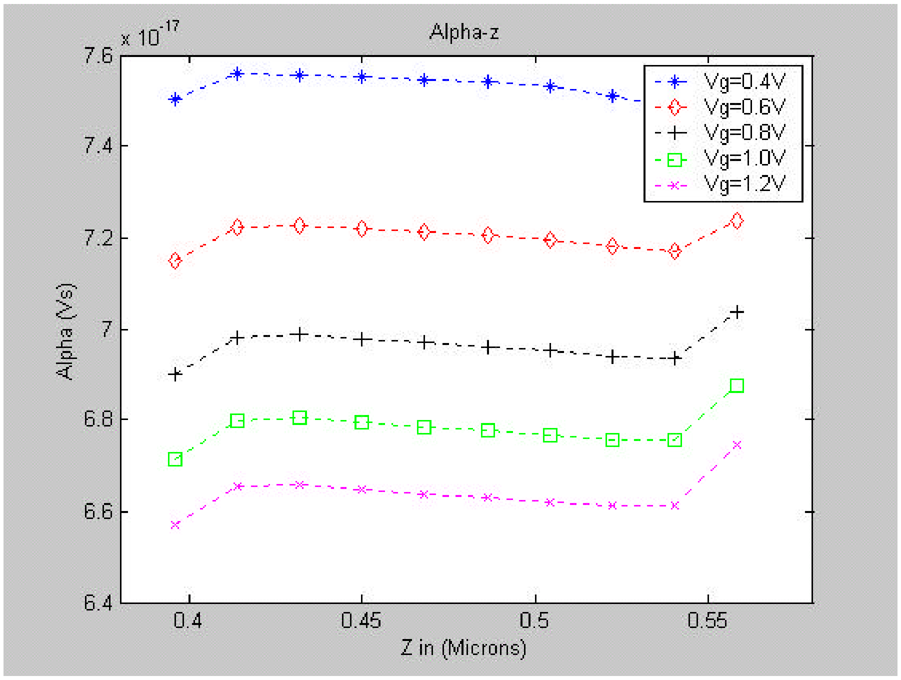

Fig 9. α -Z

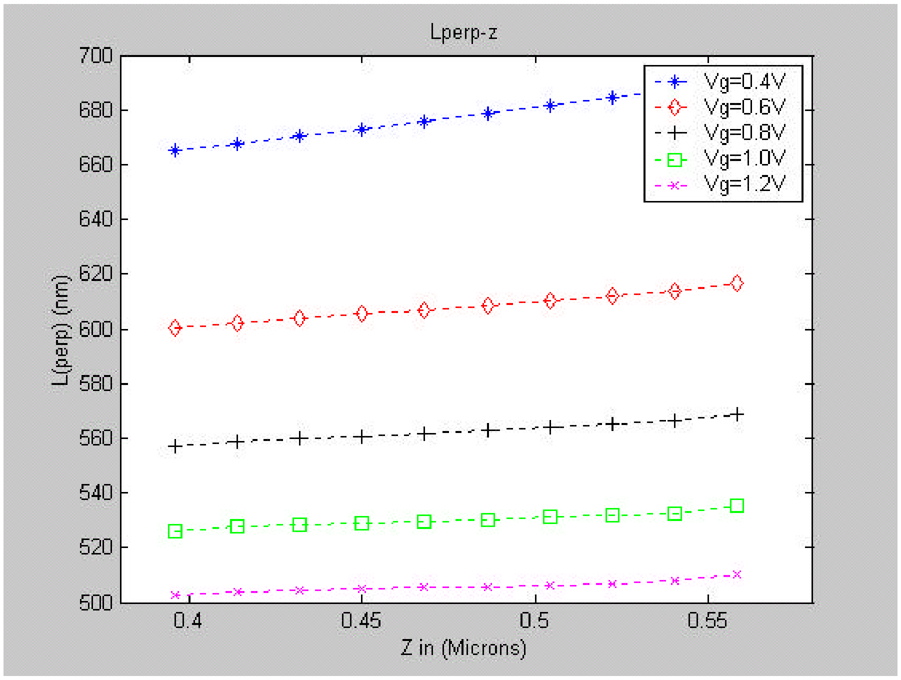

Fig 10. L_{perp} - Z

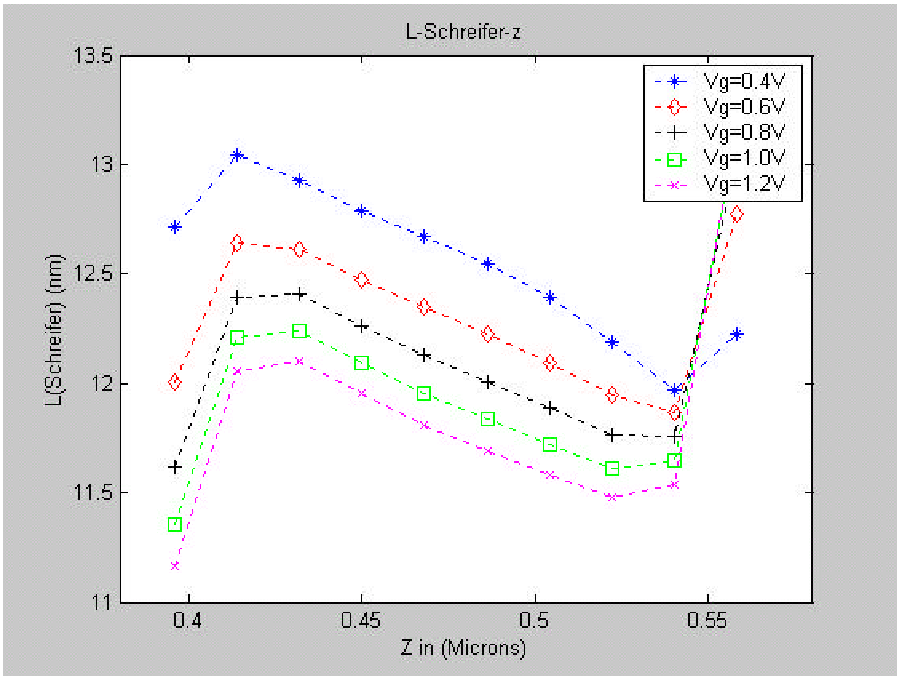

Fig 11. $L_{schriefer}$ - Z

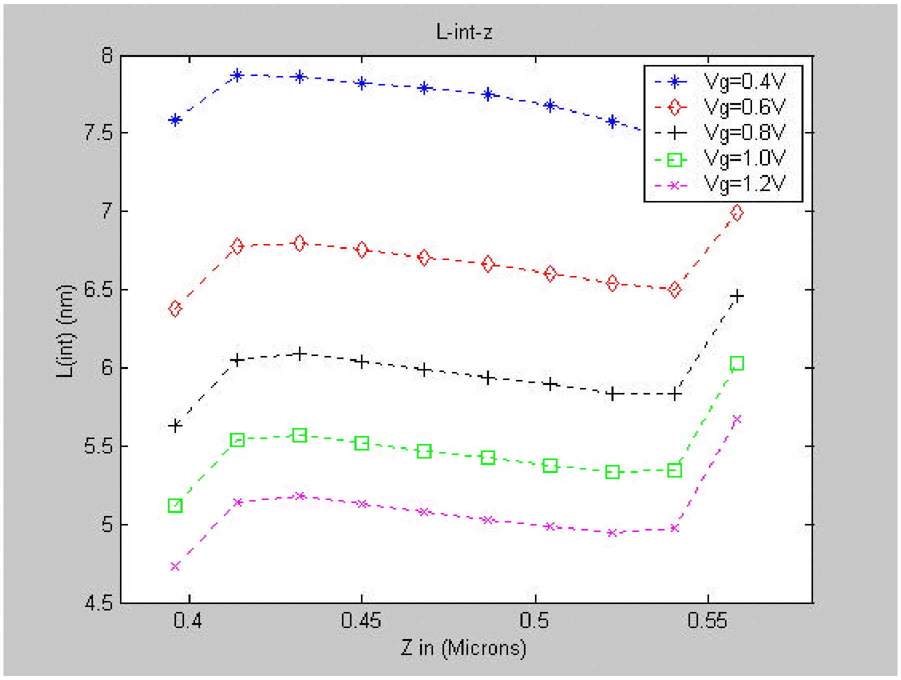

Fig 12. L_{int} - Z

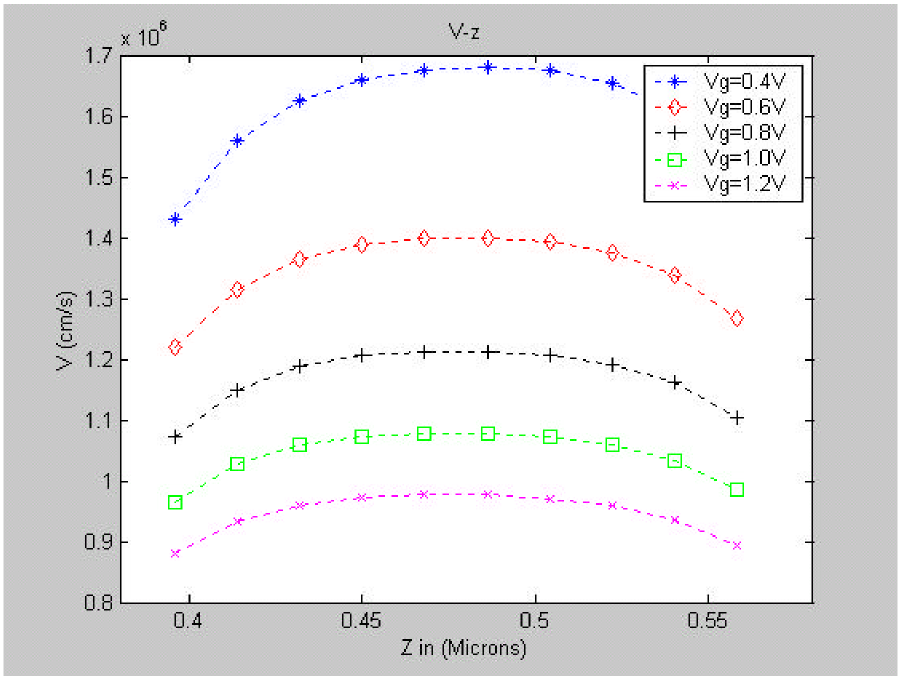

Fig 13. V - Z

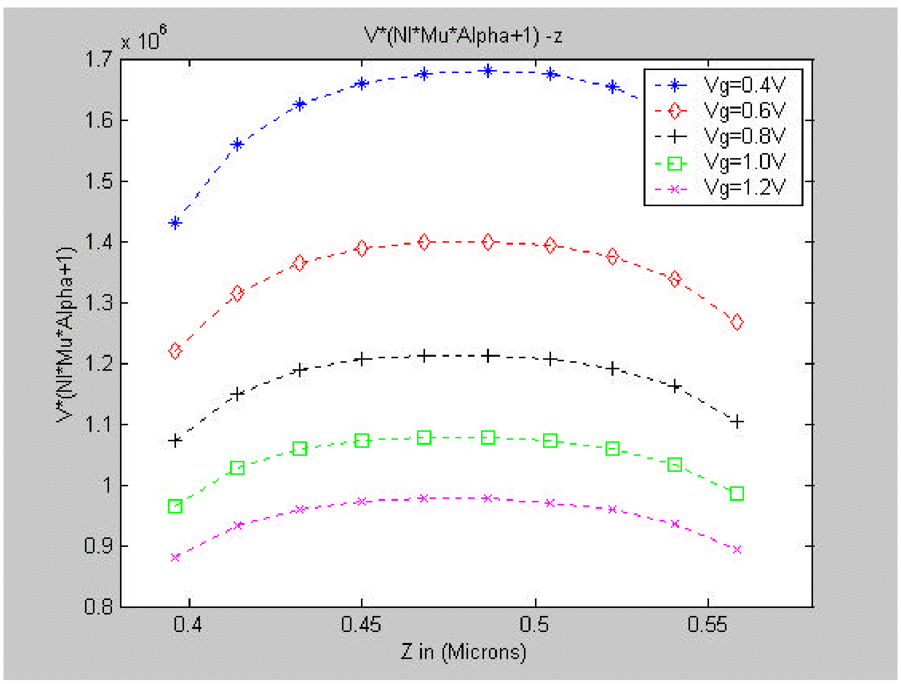

Fig 14. Determining Product – Z

Discussion

The results of simulations show that the noise power exhibits a $1/f^n$ spectrum, where n=1.000. Thus, the noise power exhibits a perfect flicker noise characteristic. This implies a uniform spatial distribution of traps.

It is also observed that as the gate voltage increases, the noise power decreases. Similar as well as contrasting behavior has been reported in literature before [9],[15].The observed nature of variation in our case can be explained on the basis of the variation of the other dependent variables with gate voltage.

As is obvious, the carrier density in the channel increases with Gate Voltage. It is also observed that the carrier velocity and hence, the effective mobility of the carriers in the channel decreases with the increase in gate voltage. The results of simulations also show that the scattering coefficient α decreases with increasing gate bias.

The parameters γ and $\mu_{eff}E_z/v_{sat}$ do not affect the noise much, as γ was close to 1 and $\mu_{eff}E_z/v_{sat}$ was much less than 1. Hence, to a first order approximation, the noise power varies as the product of velocity, v and the term $(1+\alpha\mu_{eff}E_z n_l)$ (from Eqn. 17). The variation of this "*determining product* " along the channel length for various gate biases is shown in Figure 14. Although, η increases with gate voltage, the decrease of α, μ_{eff} and E_z with V_g dominates the net product, and hence, noise power decrease with gate bias.

In this work, the simulation of noise in the rectangular portion of the fin has been completed. Minor modifications in the simulation code should be made to simulate the noise in the cylindrical portion of the SGT.

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