An analytical two-dimensional model for AlGaN/GaN HEMT with polarization effects for high power applications
2.1 Introduction

As the industry faces increased pressures and a more competitive operating environment, staying on the edge of innovation is more important than ever before. To overcome it, we require new models for semiconductor manufacturing that redefine the value chain and demonstrate new levels of agility and efficiency in meeting the needs of their customers. In the recent years, the development of very large scale integration (VLSI) technology has been dramatic developed and has been led mainly by the miniaturization of semiconductor devices, availability of sophisticated equipments and very efficient CAD tools. In modern telecommunication application, Gallium Nitride has emerged as a competitor to existing silicon and other III-V based devices [30-34].

AlGaN/GaN HEMTs have surfaced as a strong candidate for high power application owing to their large band gap energy, high saturation velocity, large breakdown field and high thermal stability [35]. In addition, the presence of strong polarization (spontaneous and piezoelectric) fields leads to the enhanced performance of these devices. Furthermore, GaN HEMTs offer a very robust solid-state technology with operational voltages of 20V even for 60GHz operation. These devices use modulation doping, which reduces the ionized impurity scattering component of the charge carrier mobility. The large conduction band offset between GaN and AlN, and the large piezoelectric effect can be exploited to generate a large channel charge density, and along with the large breakdown field, high power can be handled [36].

In AlGaN/GaN HEMTs, the polarization charges, the conduction-band discontinuity and mole fraction are the important parameter that affects the sheet carrier density (2-DEG) at
the interface. The existence of spontaneous and piezoelectric polarization fields modifies the 2DEG density. An increase in aluminium composition (mole fraction) increases the density of the 2-dimensional electron gas, and electrons lie more closely to the interface [37].

In order to achieve further progress in the study of AlGaN/GaN, development of the 2-dimensional model must and detailed device simulation is necessary. Device structure optimizations plays a vital role for investigating the small signal parameters. Although significant progress has been made over the past few years, but these processes did not explain the complete behavior of the AlGaN/GaN HEMTs at the saturation level. No models are yet available to study the behavior of two dimensional electron gas in the saturation region and calculating the voltage at the interface of the saturation region, which is very important for determining the small signal parameters and for evaluating microwave performance of the device.

This chapter gives the details of the development of an accurate two-dimensional model for current–voltage characteristics and small-signal parameters, namely, drain conductance, transconductance, cutoff frequency, and transit time of AlGaN/GaN HEMTs, incorporating the effects of spontaneous and strain-dependent piezoelectric polarization fields. The effects of important technological parameters such as mole fraction, gate length, barrier thickness, and doping of the AlGaN layer on device characteristics have been analyzed in detail. Two-dimensional analysis of the device has been carried out in the saturation region and modified expressions of device transconductance and drain conductance has been given [38-39].
2.1.1 Types of models and analysis

Generally, there are three types of models for analysing current-voltage characteristics and small-signal parameters. They are empirical, physical and optimal model. The significant uses of these models are to predict or estimate performance information that is not available or easily obtainable by direct measurement. Analysis of these models gives very essential information to the Circuit-designers and Device-designers. The physically based models are appealing to device designers. Such models are also useful to circuit designers who have some control over the fabrication process because they allow simultaneous optimizations of both the devices and the circuits in which they are to be used [3].

A physical model is useful for predicting the effects of process variations on the electrical behavior of a device. Further, if the statistical distribution of the process parameters is known, yield prediction can also be obtained. Using such an approach, performance prediction information may be obtained purely from physical data describing the device (i.e., device geometry and semiconductor material properties). No electrical characterization of individual devices is required. The merits of this approach for device designer are obvious.
Empirical models are capable of predicting accuracies that approach measurement capability. The primary difficulty with this approach is that large amounts of tedious characterization data are often required to obtain such accuracy. In addition, minor changes in the device geometry or material require complete recharacterizations. Such models are also questionable when performing design centering or yield analyses because the empirical parameter distributions do not vary independently.

The optimal model for any given application depends on many factors. The model requirements for particular applications will vary depending on the type of circuitry required and the point in the design and fabrication process at which the model is to be used. The availability of the model within a circuit simulation routine is clearly a key factor. Computational efficiency of the model and accuracy of the model predictions are also important. No single model has emerged that optimally meets all current modeling needs. Instead, many different models are being used often within the same circuit simulation package.

These models have some merit and demerit this leads to a discussion of the merits of physically based models versus empirical models. Both physical and empirical modeling techniques are associated with certain advantages and disadvantages. The main disadvantage of the physical models is that it is not accurate as required for most circuit design applications. The inaccuracies arise from the assumptions and approximations required to perform the device analysis. A second problem with physically based models is that information concerning the physical design of the device can often be difficult or impossible to obtain.
The primary difficulty with empirical models is that large amounts of tedious characterization data are often required to obtain such accuracy. In addition, minor changes in the device geometry or material require the performance of complete recharacterizations. Such models are also of questionable value when performing design centering or yield analyses because the empirical parameter distributions do not vary independently [34].

2.1.2 High Electron Mobility Transistor’s Models

An exact modeling of the AlGaN/GaN HEMT entails a good understanding of the underlying material physics as well as an accurate knowledge of the material parameters used for the modeling. The AlGaN/GaN material system is relatively new and therefore, some of the parameters needed for the modeling are uncertain or not reported, so that needed values can only be obtained by reasonable approximations and interpolations. In this part, first I discuss about the history of the development of analytical models.

In short channel devices one-dimensional solution of Poisson’s equation is not sufficient [42], as the potential distribution under the gate is two-dimensional. Therefore, it is necessary to solve two-dimensional Poisson’s equation.

Hikaru Hida et al. [138] proposed a novel 2-DEGFET model based on the parabolic Velocity-field curve approximation. In this model, selectively doped heterostructure two-dimensional electron gas (2DEG) has been proposed. In order to take into account the strong field dependence of the 2DEG mobility, a parabolic approximation is employed for a velocity-field curve below a velocity saturation field. The nonlinear field dependence of
parasitic resistances has also been considered, which is important for a more accurate description of actual FET characteristics. The proposed FET model is useful for digital IC design.

In this context, Chang and Fettermen [122], proposed an analytical model for HEMT using new velocity-field dependency. This model explained the output current-voltage characteristics and microwave-signal parameters of the high electron mobility transistors. In this model, the two-dimensional Poisson’s equation has been used for explaining the behavior of the device. This model predicted the theoretical data that was compared with the experimental data.

In the same row, Tsung-Hsing Yu and Kevin F.Brennan [11] presented a theoretical model of an AlGaN/GaN high electron mobility transistor (HEMT) that includes a nonlinear model of the strain polarization field produced at the heterointerface. Experimental work has indicated that the macroscopic polarization in III-nitride alloy is a nonlinear function of the material composition. It is well known that the behavior of a GaN-AlGaN HEMT depends greatly upon the properties of the strain-induced polarization fields formed at the GaN-AlGaN heterointerface.

Rashmi et al. [21], have proposed model to investigate the small-signal microwave parameters of fully strained (FS) and partially relaxed (PR) Al$_{m}$Ga$_{1-m}$N/GaN high electron mobility transistors (HEMTs). It is observed that elastic strain relaxation of the Al$_{m}$Ga$_{1-m}$N layer imposes an upper limit on the maximum two-dimensional electron-gas sheet charge density and is, thus extremely critical in determining the microwave performance of high Al-content Al$_{m}$Ga$_{1-m}$N/GaN HEMTs.
In this way, these models are enhancing the performance information which is not directly obtainable. As the circuits merge towards the miniaturization (say nano gate length) analysis of these models give very important information to the Circuit-designer and Device-designer. So we can say physical based models are enhancing the technology, and that’s why these are very appealing to device designers and circuit designer. These models are able to explain every single parameter elaborately, which have an impact on device performance. After taking the knowledge of these model Circuit designers have some control over the fabrication process because these parameters allow simultaneous optimizations of both the devices and the circuits.

2.2 Basic AlGaN/GaN based HEMTs description

Wideband gap semiconductors are important competitors for the development of new technologies. These III-V nitrides are compatible with higher temperatures and caustic environments. Taking these advantages, the aerospace and power companies have continuously provided the impetus for the development of advanced high power technologies capable of operating at high temperatures and hostile environments. Thus, it was expected in late 90’s that a suitable high temperature nitride based technology would allow the bulky system to be replaced by heat tolerant on site control electronics.

The members of III-V nitride semiconductor family, aluminium nitride, gallium nitride and their alloys are all wide band gap semiconductors. They crystallize in both wurtzite, and zinc blend polytypes. Wurtzite GaN, AlN and InN have direct room temperature band gaps of 3.4, 6.2 and 1.9eV. In the cubic form, GaN and InN have direct band gaps while AlN is indirect. Thus when GaN is alloyed with AlN and InN, the III-V nitrides are
formed, and they may span a continuous range of direct band gap energies throughout much of the visible spectrum well into the ultraviolet wavelengths. Therefore, GaN, AlN, InN and their tertiary alloys are also useful for short wavelength optoelectronics device application [45-50].

GaN based field effect transistor is projected to be highly useful for power amplification and switching in high the temperature and high power environment [51]. Until now, the fabrications and electrical characterization of GaN FETs are at immature stage. Though some basic structures using GaN/AlGaN were reported, development in terms of modeling of small signal parameters is required for GaN based HEMTs to aid in technology development, device structure optimizations and advanced design.

Devices are usually fabricated on epi-films with high aluminum contents (in AlGaN/GaN devices) Higher current density also resulted with the higher mole fraction of Al. These HEMTs also showed high breakdown voltages in the range of 100-220V for different gate length structures. Improved carrier confinement, allowing a high mobility was also observed in AlGaN/GaN HEMTs [52-54].

High electron-mobility transistors (HEMTs) have also shown promising performance in high-speed integrated circuits owing to their very high switching speed and low power consumption. Over the past few years, AlGaN/GaN HEMTs have appeared as important devices for high-power high-temperature and high-speed applications at frequencies well into the microwave region [20, 55]. The presence of strong spontaneous and strain-induced piezoelectric polarization fields due to the lattice mismatch between AlGaN and GaN layers, in addition to the large band gap energy, high saturation velocity, large
breakdown field, large conduction-band discontinuity, and high thermal stability are some of the characteristic features of the AlGaN/GaN material system that lead to outstanding performance of GaN-based HEMTs [56-60].

The key to enhanced performance of HEMTs is also due to improved charge confinement and superior carrier transport that result from a very high two-dimensional electron gas (2-DEG) sheet charge density with concomitant high mobility. The conduction-band discontinuity between the AlGaN barrier and GaN channel layers and the doping density of AlGaN are the two most important parameters that can be tailored to attain high sheet charge density.

The upper limit on doping density is, however, imposed by the requirement of the nonleaky Schottky barrier and by electron traps in the barrier that reduce the conduction ability of the 2-DEG. Thus, increasing the conduction-band offset seems to be the best option to improve HEMT performance [61-62]. Furthermore, 2-DEG density in AlGaN/GaN HEMTs is predominantly determined by the piezoelectric and spontaneous polarization, which, in turn, is controlled by the alloy composition (Al content) of the AlGaN barrier. Piezoelectric polarization of AlGaN on GaN is also determined by the elastic strain in AlGaN, resulting from the lattice mismatch between AlGaN and GaN layers. Therefore, any factor that affects the strain in the AlGaN layer would be critical in determining the polarization charges.

Although significant progress has been made over the past few years, additional development work in terms of modeling of small-signal parameters is required for GaN-based HEMTs, especially for the high Al-content structures to aid in technology
development, device structure optimizations, and advanced design. Research efforts thus far in AlGaN/GaN HEMTs have focused on transport phenomena and charge control in fully strained (FS) devices. Investigations in high Al-content HEMTs have been directed toward analyzing the effect of relaxation on 2-DEG density and mobility to estimate the critical thickness for relaxation. No models are yet available to study the impact of strain relaxation characteristics due to spontaneous and piezoelectric polarization effects and calculate small-signal parameters of AlGaN/GaN HEMTs, which are extremely important for evaluating microwave performance of the device.

2.2.1 HEMT Structure (for analysis)

The high electron mobility transistor is a heterostructure field effect device. The term "high electron mobility transistor" is applied to the device because these structure takes advantage of the superior transport properties (high mobility and velocity) of electrons in a potential well of lightly doped semiconductor material.

Figure 2.01 presents a cross-sectional view of a conventional HEMT structure. Three metal electrode contacts (source, gate, and drain) are made to the surface of the semiconductor structure. Source and Drain are having metallic contacts while the gate contact is schottky.
The source and drain contacts are ohmic while the gate is a Schottky barrier. Further similarities between the HEMT and MESFET diminish rapidly as the description of an operation proceeds to the important phenomena-taking place within the semiconductor structure. This complexity is associated with fabrication difficulties, added costs, and lower yields[66].

This structure will serve as the focus of discussion of the first order operating principles of the HEMT. The most common materials used for this structure are AlGaN and GaN, respectively, as shown in the figure. The thickness and doping density of the AlGaN layer are designed so this layer is completely depleted of free electrons under normal operating conditions. The undoped GaN thickness is much less critical to the device design.
However, The GaN material quality (i.e., an absence of defects and deep levels) must be high [67].

The high free-electron concentration occurs over such a thin region that it is described as a two dimensional electron gas (2-DEG) and quantified in terms of sheet carrier density \( n_s \). Electrons traveling in this region do not encounter ionized donor atoms because the GaN is undoped. The electron mobility is highest for lightly doped material. Thus, transport properties in the 2-DEG are favorable to fast response times and high-frequency operation. Both depletion mode and enhancement mode devices can be fabricated.

Contact with the 2-DEG is made via the heavily doped, low resistance source and drain wells. For low values of drain-to source bias, a current flow from the drain to source through the electron gas. In depletion mode devices, current will flow in the absence of applied gate bias. This current is proportional to the applied drain-source voltage. As drain-source bias levels are increased, electron velocity and the current levels saturate. The saturated current level is determined primarily by the sheet carrier density of 2-DEG that forms in the structure.

The 2-DEG density is controlled by the gate bias. Increasing the negative bias applied to the gate decreases the depth (in electron energy) of the potential well at the AlGaN/GaN boundary. For depletion mode devices, no current conduction takes place until maximum reverse gate bias is applied. In the HEMT, gate bias controls the carrier density. Both of these effects, however, result in control of the maximum channel current. For large enough values of reverse gate bias, the sheet carrier concentration becomes negligibly
small and the channel current is pinched off. The gate voltage level that corresponds to
these phenomena is termed as pinch-off voltage.

2.2.2 Details of parameters that govern HEMT operation

In this section I discuss the importance of the various parameters in optimizing the
microwave performance of the AlGaN/GaN HEMT.

Most microwave FETs are depletion mode devices. This means that in the absence of
applied reverse gate bias, current can flow between the source and drain, contacts.
Enhancement mode devices do not conduct current between the drain and source, unless
forward gate bias is applied. In depletion mode devices reasonably low bias potential is
applied between the source and drain contacts. Enhancement mode devices do not
conduct current between the drain and source, unless forward gate bias is applied.

For depletion mode devices, when reasonably low bias potential is applied between the
source and drain contacts, a current flow through the channel. This current is linearly
related to the voltage across the terminals. For higher drain-source bias levels, the
electrons in the semiconductor material will attain their maximum carrier velocity
associated with the saturation of carrier velocity is the saturation of channel current.

The energy band bending produced by making schottky barrier contact with the
semiconductor creates a layer beneath the gate that is completely depleted of free charge
carriers. As no free carriers exist in this depletion layer, no current can flow through it.
The available cross-sectional area for current flow between the source and drain is
reduced by the existence of this depletion layer. As reverse bias is applied to the gate, the depletion layer penetrates deeper into the active channel. These further reductions in cross-sectional area result in further current reductions. The gate bias, then acts as a mechanism for limiting the maximum amount of source-drain current that can flow. When enough reverse bias is applied, the depletion region will extend across the entire active channel and allow essentially no current to flow. The potential required to accomplish this is termed the “pinch-off potential” or “pinch-off voltage.” In the same way there are many important parameters for which detailed analysis is required. The following parameters have a great impact on device operation.

(a) Aluminium Composition

The aluminium composition in the AlGaN charge supply layer is important because it determines the AlGaN bandgap and consequently, the conduction band discontinuity between the AlGaN and GaN at the heterojunction interface. In addition, the Al composition determines the conduction and valence band effective density of states (DOS), the effective mass of carriers and therefore their mobility in the AlGaN layer. It also determines the Schottky barrier height of the gate. Finally, and most important, due to the polarization effects present in these devices, the aluminium composition also strongly influences the magnitude of the charge in the 2DEG.

The dependence of each of these material and device properties on the aluminium composition is discussed in this chapter. The aluminium composition in the AlGaN charge supply layer along with the thickness of the AlGaN layer determines the charge due to polarization. The charge due to polarization has been modeled analytically. The
2DEG concentration and the conduction band discontinuity rise nearly linearly with the aluminium composition.

(b) AlGaN Thickness

The AlGaN layer thickness strongly influences the transconductance of the HEMT and hence, is an important parameter for optimizing the high frequency performance of the device. The AlGaN layer thickness determines the gate-to-channel distance in the HEMT and therefore, affects its transconductance, which is inversely proportional to the gate-to-channel distance. Hence, for maximizing the transconductance of the device, the gate-to-channel distance must be small as much as possible.

However, a decrease in the AlGaN layer thickness would lead to a reduction in charge in the channel at zero gate bias as the Schottky barrier’s influence extends into the channel, which negatively influences the HEMT’s transconductance. A forward bias can be applied to increase the charge in the channel, but the magnitude of bias is limited by the gate leakage current. Therefore, an optimum value of the AlGaN layer thickness has to be determined for maximizing the device’s transconductance and therefore its high frequency performance.

In the AlGaN/GaN HEMT, the thickness of the AlGaN layer also influences the charge induced due to polarization. The charge induced due to polarization increases with decreasing AlGaN thickness. The increase in charge would improve the transconductance of the device further.
(c) Field dependent mobility

The velocity-field characteristics give the important information about the field dependent mobility, which is essential for simulation of semiconductor devices. In AlGaN based devices like HEMTs, an additional complication arises from the effect of composition dependence, which influences electron transport and device performance.

To clarify optimum device parameters and to attain higher performance, a number of studies have been made on the modeling of HEMTs. An emphasis has been focused on the velocity-field relationship and less attention has been paid to the role of mobility at the low electric field. The electric field is not always high enough for carriers to drift at saturation velocity $v_s$ close to source under the gate, and pointed out the importance of low field mobility in the modeling of HEMT’s. The field dependent mobility can be modeled [21] using as,

$$\mu(x,m) = \frac{\mu_0(m)}{1 + \left(\frac{\mu_0(m)E_c-v_{sat}}{(E_c)v_{sat}}\right) \frac{dV_c(x)}{dx}}$$

$\mu_0(m)$ is the low field mobility, $v_{sat}$ is the saturation velocity and $E_c$ is the critical field. The low field mobility varies with Al mole fraction, temperature and sheet carrier concentration.

The band diagram for this structure under conditions of zero gate bias is illustrated in figure 2.02. A sharp dip in the conduction band edge occurs in the HEMT.
The high free-electron concentration occurs over such a thin region that it is described as a two-dimensional electron gas (2-DEG) and quantified in terms of a sheet carrier density, $n_s$. Electrons traveling in this region do not encounter ionized donor atoms because the GaN is undoped. Thus, transport properties in the 2-DEG are favorable to fast response times and high-frequency operation.

Contact with the 2-DEG is made via the heavily doped, low resistance source and drain wells. For low value of drain-to-source bias, a current flow from the drain to source through the electron gas. In depletion mode devices, current will flow in the absence of applied gate bias. This current is proportional to the applied drain-source voltage. As drain-source bias levels are increased, electron velocity and the current levels saturate. The saturated current is determined primarily by the sheet carrier density of 2-DEG that forms in the structure.

**Figure 2.02 : The band diagram of the HEMT structure**

(d) Modified threshold Voltage($V_{th}$)
The threshold voltage is defined theoretically as the applied gate voltage for which the channel is completely depleted of free carriers. The definition assumes that the expulsion of free carriers within the depletion region is total and that the substrate is a perfect insulator. Although both assumptions are good approximations over most of the operating range of the device, neither is absolutely true.

In an actual device, the boundary between the edge of the depletion region and the undepleted channel is gradual and occurs over a distance of several Debye lengths. A graded transition also occurs between the active channel and substrate. As the device is pinched off, these two graded regions are forced closer to each other.

For a conventional AlGaN/GaN HEMT with Si modulation doped layer, as shown in given figure 2.01, the polarization charges need to be taken into account in the calculation of HEMT’s threshold voltage. Modified threshold voltage, taking into account the effects of charge polarization, surface and buffer traps can be expressed as [76],

\[
V_{\text{th}}(m) = \frac{\Phi_{\text{B}}(m)}{q} - \frac{d \sigma(m)}{\varepsilon(m)} - \frac{\Delta E_{c}(m)}{q} + \frac{E_{\text{f0}}(m)}{q} \left( \frac{q}{\varepsilon(m)} \right) \int_{0}^{d} \left[ \int_{0}^{x} N_{\text{si}}(x,m) \, dx - \frac{q N_{\text{si}}(x,m)}{\varepsilon(m)} \right] - \frac{q N_{\text{st}} N_{\text{b}}}{C_{\text{b}}} \tag{2.01}
\]

Where \( \Phi_{\text{B}} \) is metal-semiconductor Schottky barrier height, \( q \) is the electronic charge, \( \sigma \) is overall net (both spontaneous and piezoelectric) polarization charge at the barrier—AlGaN/GaN interface. \( \Delta E_{c} \) is the conduction band discontinuity, \( d \) is AlGaN layer thickness, \( N_{\text{si}}(x) \) is Si-doping concentration, \( E_{\text{f0}} \) is the difference between the intrinsic Fermi level and the conduction band edge of the GaN, \( \varepsilon \) is the dielectric constant of AlGaN. \( N_{\text{st}} \) is the net-charged surface traps per unit area, \( N_{\text{b}} \) effective net-charged buffer
traps per unit area and \( C_b \) is the effective buffer-to-channel capacitance per unit area. The last two terms in eq. (2.01) describe the effects of the surface traps and buffer traps, respectively.

(e) Drain-current generalized equation

An analytical description of heterostructure field effect transistor is usually based on a charge control model. In a charge control model, the current in the channel is controlled by the charge induced into the channel due to the gate voltage. The device capacitances are found by the derivatives of the charge with respect to the terminal voltage. The most parameters of the charge control model are the device threshold voltage, the gate dielectric thickness \( W \), the effective mobility \( \mu(x, m) \), effective electron saturation velocity, source and drain series resistance, etc. The model developed in this chapter is based on the electron velocity saturation that happens at drain voltages higher than certain critical drain voltage.

In short channel compound semiconductor devices with gate length of 1\( \mu \)m or less, the extrinsic source and drain resistance are comparable to the channel resistance. Therefore, the voltage drop across these parasitic resistances can no longer be treated as a small perturbation to the external bias. Furthermore, useful device modeling and parameter extraction methods should be able to cover both the intrinsic characteristics and the extrinsic device characteristics. In this section I calculate the drain current with the inclusion of parasitic source and drain resistance.

The drain current (\( I_{ds} \)) can be calculated from the current density equation
\[ I_{d}(m,x) = W q \mu(x,m) \left( n_s(m,x) \frac{dV_c(x)}{dx} + \frac{K_B T}{q} \frac{dn_s(m,x)}{dx} \right) \]  

(2.02)

Where \( W \) is the gate width, \( n_s \) is the sheet carrier concentration, \( K_B \) is the Boltzmann constant, \( T \) is the absolute temperature and \( \mu(x) \) is the Al composition dependent mobility given[21] as

\[
\mu(x,m) = \frac{\mu_0(m)}{1 + \left( \frac{\mu_0(m)E_c-v_{sat}}{(E_c)v_{sat}} \right) \frac{dV_c(x)}{dx}}
\]

(2.03)

Where \( \mu_0(m) \) is low field mobility, \( E_c \) as the critical field and \( v_{sat} \) is the saturation velocity.

The sheet carrier density (2DEG) is given as

\[
n_s(m,x) = \frac{\varepsilon(m)}{q(d_d + d_i + \Delta d)} (V_{gs} - V_{th}(m) - V_c(x))
\]

(2.04)

Where \( V_{th}(m) \) is the threshold Voltage which includes the polarization effects [15]

### 2.2.3 Spontaneous and Piezoelectric polarization

Spontaneous and piezoelectric polarization effects play an important role in determining the charge in the 2DEG [15, 20]. The net effect of the polarization in these structures is that a thin sheet of positive charge in introduced at the AlGaN/GaN interface, which leads to an accumulation of electrons in the 2DEG even without intentional doping of the AlGaN charge supply layer. To model the charge analytically, Ambacher [15] has developed a set of equations which relate the charge in the 2DEG to the AlGaN layer
thickness and the Al composition in the AlGaN layer. This model assumes the layers to be dielectric, and though simplistic, gives an accurate determination of the charge in the channel.

Figure 2.03: Directions of piezoelectric (P_{PE}) and spontaneous (P_{SP}) polarization for different growth condition in AlGaN/GaN heterostructures [14].

A unique feature of the AlGaN/GaN material system is the high sheet carrier concentration, which can be achieved in the channel due to large band discontinuities at the interface. In fact, due to the piezoelectric and spontaneous polarization, high sheet carrier concentrations can be obtained even without intentionally doping the barrier. This
Fig. 2.03 (a). Variation of 2-DEG density with the Al mole fraction of the Al\textsubscript{m}Ga\textsubscript{1-m}N barrier layer for various values of doping density [21].

2-DEG density can significantly be increased by doping the AlGaN layer as shown in figure 2.03(a) [21]. Rashmi et al.[21] showed in fig. 2.03(a) that high charge density can be obtained at m = 0.2 in the absence of doping, which is attributed to the presence of strong spontaneous and piezo-electric polarization fields and increase in doping increases
2-DEG. In actual practice the 2-DEG density is limited by the presence of donor defects barrier traps and surface traps, which leads to increased scattering and reduced conduction. The 2-DEG density increases with an increase in the Al-mole fraction of the barrier owing to the increase in conduction band discontinuity and polarization induced charges. The 2-DEG density in a partially relaxed structure is much less than that expected to a fully strained device at the same Al mole fraction as shown in the experimental comparison of data in fig 2.03(a) with the results by Ambacher et al [31]. It is also seen from the figure 2.03(a) that the maximum 2-DEG that can be attained in an AlGaN/GaN depends more strongly on strained relaxation of the barrier than on the Al-content. So the inclusion of polarization effect in a calculation of 2-DEG density is more important to accurately calculate it for high Al content AlGaN/GaN HEMTs.

It has also been shown that from table 2.1 piezoelectric effects can have a fundamental influence on the electron distributions in strained nitride-based wurtzite heterostructures. In fact, in AlGaN-based transistors, the piezoelectric polarization of the top layer is more than five times larger than in analogous AlGaN structures. Furthermore, according to Bernardini et al. [14], a very large spontaneous polarization charge also exists which, must be taken in an account in the calculations of the sheet carrier concentration [10]. In short, the piezoelectric and spontaneous polarization charge is inextricable from the gate-induced and or charge-induced by doping in FETs, and must be carefully considered in device design and analysis [11].
If the electric field is absent then the total macroscopic polarization $P$ of a GaN or AlGaN layer is the sum of the spontaneous polarization $P_{sp}$ in the equilibrium lattice, and the strain-induced or piezoelectric polarization $P_{PE}$. It is due to sensitive dependence of the spontaneous polarization on the structural parameters, there are some quantitative differences in the polarization for GaN and AlN. The increasing no ideality of the crystal structure going from GaN to AlN corresponds to an increase in spontaneous polarization. Here I consider polarization along the [0001] axis, since this is the direction along which epitaxial films and AlGaN/GaN heterostructures are grown. The spontaneous polarization
along the c-axis of the wurtzite crystal is $P_{SP}$. The piezoelectric polarization can be calculated with the piezo-electric coefficients $e_{33}$ and $e_{13}$ as shown in the Table 2.1 given [15] as

$$P_{PE} = e_{33} \varepsilon_x + e_{31} (\varepsilon_x + \varepsilon_y)$$  \hspace{1cm} (2.04a)

Where $\varepsilon_x = \frac{(c - c_0)}{c_0}$ is the strain along the c-axis, and the in-plane strain $\varepsilon_x = \varepsilon_y = \frac{(a - a_0)}{a_0}$ is assumed to be isotropic. $a_0$ and $c_0$ are the equilibrium values of the lattice parameters.

The third independent component of the piezoelectric tensor, $e_{15}$, is related to the polarization induced by shear strain. The relation between the lattice constant of the hexagonal GaN is given [15] as,

$$\frac{(c - c_0)}{c_0} = -\frac{2}{C_{33}} \frac{C_{13}}{a_0} (a - a_0)$$  \hspace{1cm} (2.04b)

Where $C_{13}$ and $C_{33}$ are elastic constants.

Using above equation, the amount of the piezoelectric polarization in the direction of the c-axis can be determined by [15].

$$P_{PE} = 2 \frac{a - a_0}{a_0} (e_{31} - e_{33} \frac{C_{13}}{C_{33}})$$  \hspace{1cm} (2.04c)

Table 2.2. Measured and calculated elastic constants of wurtzite and cubic AlN, GaN, and InN.

<table>
<thead>
<tr>
<th>GPa</th>
<th>AlN</th>
<th>GaN</th>
<th>InN</th>
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<tbody>
<tr>
<td>Wurtzite</td>
<td>Exp. a</td>
<td>cal. b</td>
<td>exp. c</td>
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</tbody>
</table>
Since \([ e_{31} - e_{33} \left( \frac{C_{13}}{C_{33}} \right) ] < 0\) for AlGaN over the whole range of compositions, the piezoelectric polarization is negative for tensile and positive for compressive strained barriers, respectively. The spontaneous polarization for GaN and AlN was found to be negative, meaning that for Ga(Al)-face heterostructures the spontaneous polarization is pointing towards the substrate (Figure 2.03). As a consequence, the alignment of the piezoelectric and spontaneous polarization is parallel in the case of tensile strain, and antiparallel in the case of compressively strained top layers. If the polarity flips over from Ga-face to N-face material, the piezoelectric, as well as the spontaneous polarization changes its sign. In above figure, the directions of the spontaneous and piezoelectric

| \(c_{11}\) | 345 | 396 | 374 | 367 | 190 | 223 |
| \(c_{12}\) | 125 | 137 | 106 | 135 | 104 | 115 |
| \(c_{13}\) | 120 | 108 | 70  | 103 | 121 | 92  |
| \(c_{33}\) | 395 | 373 | 379 | 405 | 182 | 224 |
| \(c_{44}\) | 118 | 116 | 101 | 95  | 10  | 48  |
| B       | 201 | 207 | 180 | 202 | 139 | 141 |

| Zineblende | cal.\(^e\) | cal.\(^b\) | cal.\(^e\) | cal.\(^b\) | cal.\(^e\) | cal.\(^b\) |
| \(c_{11}\) | 304 | 304 | 296 | 293 | 184 | 187 |
| \(c_{12}\) | 152 | 160 | 154 | 159 | 116 | 125 |
| \(c_{44}\) | 199 | 193 | 206 | 155 | 177 | 86  |

\(^a\)Ref 70, \(^b\)Ref 71, \(^c\)Ref 72, \(^d\)Ref 73, \(^e\)Ref 74.
polarization are given for Ga-face, N-face, strained, unstrained AlGaN/GaN, and GaN/AlGaN heterostructures.

To calculate the amount of the polarization induced sheet charge density ($\sigma$) at the AlGaN/GaN and GaN/AlGaN interfaces in dependence of the Al-content $x$ of the Al$_x$Ga$_{1-x}$N barrier, I use the following set of linear interpolations between the physical properties of GaN and AlN:

$$a(x) = (-0.077x + 3.189) \times 10^{-10} \text{ m}$$  \hspace{1cm} (2.04d)

**Elastic constants:**

$$C_{13}(x) = (5x + 103) \text{ GPa},$$  \hspace{1cm} (2.04e)

$$C_{33}(x) = (-32x + 405) \text{ GPa},$$  \hspace{1cm} (2.04f)

**Piezoelectric constants:**

$$e_{31}(x) = (-0.11x - 0.49) \text{ C/m}^2,$$  \hspace{1cm} (2.04g)

$$e_{31}(x) = (-0.11x - 0.49) \text{ C/m}^2,$$  \hspace{1cm} (2.04h)

**Spontaneous polarization:**

$$P_{sp}(x) = (-0.052x - 0.029) \text{ C/m}^2$$  \hspace{1cm} (2.04i)

The amount of the polarization induced sheet charge density for the N-face GaN/Al$_x$Ga$_{1-x}$N/GaN is calculated. Both AlGaN and GaN layers have spontaneous and piezoelectric polarization effects, and the model assumes that the net charge in the layers can be modeled based on the difference in polarization between the GaN layer and the AlGaN layer. The increasing non-ideality of the crystal structure going from GaN to AlN causes an increase in spontaneous polarization. Furthermore, since the AlGaN layer doesn’t have inversion symmetry, there is a significant piezoelectric polarization present due to the
strain in these structures. The polarization sheet charge density at the interface is modeled as [15].

\[ \sigma(x) = P(\text{Top}) - P(\text{Bottom}) \]

\[ = [P_{\text{sp}}(\text{Top}) + P_{\text{pe}}(\text{Top})] - [P_{\text{sp}}(\text{Bottom}) + P_{\text{pe}}(\text{Bottom})] \]

\[ = [P_{\text{sp}}(\text{Al}_x\text{Ga}_{1-x}\text{N}) + P_{\text{pe}}(\text{Al}_x\text{Ga}_{1-x}\text{N})] - [P_{\text{sp}}(\text{GaN})] \]

(2.04j)

Where \( P_{\text{sp}} \) is the spontaneous polarization, \( P_{\text{pe}} \) is the piezoelectric polarization and \( x \) is the Aluminum composition in AlGaN.

### 2.3 Model Description

**Figure 2.04 AlGaN/GaN HEMT structure**

A basic AlGaN/GaN HEMT structure as shown in the figure 2.04. The structure is on a semi-insulating 4H-SiC substrate. The epilayer consists of a 100nm AlN buffer, 2µm undoped GaN, a 5nm undoped Al\(_{0.25}\)Ga\(_{0.75}\)N spacer a 10nm Si-doped Al\(_{0.25}\)Ga\(_{0.75}\)N charge supply layer and a 10nm undoped Al\(_{0.25}\)Ga\(_{0.75}\)N barrier layer. The device has a gate length of 120nm, a gate width of 100µm and a source drain spacing of 2µm. As shown in the
figure the region between the gate and the channel is rectangular. It is divided in two regions, i.e., the low field region with $0 < x < L_1$ and high field region with $L_1 < x < L_2$.

The current flowing through the two-dimensional electron gas channel as a function of the internal gate and drain voltage $V_{gs}$ and $V_{ds}$ is obtained by integrating the above equation, using the following boundary conditions.

![Diagram showing linear and saturation region](image)

Figure 2.05: Diagram for showing linear and saturation region

Substituting eqn. (2.04) and (2.03) in eqn. (2.02) and integrating it using the boundary conditions

$$V_c(x) \Big|_{x=0} = I_{ds}(m,x)R_s \quad \text{(Boundary condition for linear region)}$$

$$V_c(x) \Big|_{x=L} = V_{ds} - I_{ds}(m,x)(R_s+R_d) \quad \text{(Boundary condition for saturation region)}$$

the $I_{ds}-V_{ds}$ equation for the linear region is obtained as [20].

$$I_{ds}(m) = \frac{-\lambda_2 + \sqrt{\lambda_2^2 - 4\lambda_1\lambda_3}}{2\lambda_1} \quad (2.05)$$

Where
\[ \lambda_1 = \frac{(2R_s + R_d)}{E_1} - \frac{G_0}{2} \left( (R_d)^2 + 2R_sR_d \right) \]  

(2.05a)

\[ \lambda_2 = G_0 \left( V_{ds}(R_s + R_d) - V_{gs} (2R_s + R_d) \right) - L - \frac{V_{ds}}{E_1} \]  

(2.05b)

\[ \lambda_3 = G_0 \left( V'_{gs} V_{ds} - \frac{(V_{ds})^2}{2} \right) \]  

(2.05c)

\[ \frac{1}{E_1} = \left( \frac{\mu_0 E_c - V_{sat}}{E_c V_{sat}} \right) \]  

(2.05d)

\[ V'_{gs} = V_{gs} - V_{th} - \frac{K_BT}{q} \]  

(2.05e)

\[ G_0 = \frac{W \mu_0 \varepsilon(m)}{(d_i + d_i + \Delta d)} \]  

(2.05f)

Where L is the channel length, \( V_{ds} \) is the applied drain bias, \( V_{gs} \) is the gate bias, \( R_s \) and \( R_d \) are the parasitic source and drain resistance, respectively, with increase in drain bias the drain current saturates. The drain saturation current is obtained as

\[ I_{dsat} = \frac{G_0E_c}{(d_i + d_i + \Delta d)} (V'_{gs} - V_{dsat}) \]  

(2.06)
Where $V_{dsat}$ is the drain saturation voltage obtained by substituting $V_{ds} = V_{dsat}$ in eqn. (2.05) and equating $I_{dsat}$ (from eq. (2.06)) with $I_{ds}$ (m, x) (eq. (2.05)). Thus the drain saturation voltage is given as

$$V_{dsat} = \frac{-\beta_2 + \sqrt{\beta_2^2 - 4\beta_1\beta_3}}{2\beta_1}$$  \hspace{1cm} (2.07)$$

Where

$$\beta_1 = \frac{(G_0E_c)^2}{E_1} \left(2R_s+R_d\right) - (G_0) \left(\frac{E_c^2}{2}\right) \left(R^2_d+2R_s R_d\right) + G_0 \left(\frac{h_0}{v_{sat}} E_c \frac{1}{\frac{E_c^2}{2}}\right) - G_0 E_c (R_s+R_d)$$ \hspace{1cm} (2.07a)$$

$$\beta_2 = \frac{G_0}{E_1} \left(\frac{h_0 E_c V'_{gs}}{v_{sat} E_c} E_1 \right) - 2V'_{gs} \left((G_0 E_c)^2 \left(2R_s+R_d\right)- (G_0) \left(\frac{E_c^2}{2}\right) \left(R^2_d+2R_s R_d\right)\right) + G_0 E_c V'_{gs} (3R_s + 2R_d)$$ \hspace{1cm} (2.07b)$$

$$\beta_3 = \frac{(G_0 E_c)^2}{E_1} \left(2R_s+R_d\right) - (G_0) \left(\frac{E_c^2}{2}\right) \left(R^2_d+2R_s R_d\right) V'^2_{gs} - G_0 V'_{gs} E_c (G_0 V'_{gs})^2 E_c (2R_s+R_d)$$ \hspace{1cm} (2.07c)$$

By solving the above equations now I can easily calculate the voltage at the interface of the linear region and the saturation region as follows:

$$I_{ds} = \frac{G_0}{\left[1 + \left(\frac{1}{E_1} \frac{dV_c(x)}{dx}\right) (V'_{gs} - V_c(x)) \right]} \frac{dV_c(x)}{dx}$$ \hspace{1cm} (2.08)$$

Eq. (2.08) can be rewritten as

$$\frac{dV_c(x)}{dx} = \frac{I_{ds}}{G_0 \left(V'_{gs} - V_c(x)\right)} - \frac{I_{ds}}{E_1}$$ \hspace{1cm} (2.09)$$

Now boundary condition $x = L_1$, $V_c(x)|_{x=L_1} = V_{L1}$, $dV/dx = E_0$ and $I_{ds} = I_c$
Then eq. (16) becomes
\[
E_0 = \frac{I_c}{G_0 \{ V'_{gs} - V_{L1} \} - \frac{I_c}{E_1}} \tag{2.10}
\]

Rearranging the terms than I get the voltage at the point of saturation \( V_{L1} \) as
\[
V_{L1} = V'_{gs} - \left( \frac{1}{E_0 G_0} + \frac{1}{G_0 E_1} \right) I_c \tag{2.11}
\]

Substituting \( \rho = \left( \frac{1}{E_0 G_0} + \frac{1}{G_0 E_1} \right) \) in above equation, and get the following equation
\[
V_{L1} = V'_{gs} - \rho I_c \tag{2.12}
\]

By using above expression, I can easily calculate the interface voltage which is very significant for calculating the device transconductance \( (g_m) \) and drain conductance \( (g_d) \)

The length of the linear region \( L_1 \), which depends on the gate and drain voltages can be calculated by using the above equations and is obtained as Rajesh Tyagi et. al [176].
\[
L_1 = \left[ \left( \frac{G_0 V'_{gs}}{I_{ds}} - \frac{1}{E_1} \right) \left( V'_{gs} - (\rho + R_s) I_{ds} \right) \right] - \frac{G_0}{2} \left( \frac{V'_{gs}^2}{I_{ds}} + (\rho^2 - R_s^2) I_{ds} - 2 V'_{gs} \rho \right) \tag{2.13}
\]

For a given set of \( V_{gs} \) and \( V_{ds} \) I can determine the value of \( L_1 \) iteratively.

As it is evident from the experiments that in short channel devices the current in the saturation region is not constant but increases with the increase in drain voltage. To describe the characteristics of device in the saturation region accurately I have performed two dimensional analyses in this region. The length of the saturation region \( L_2 \), is determined by both the gate and drain biases.

**2.3.1 Solution of two-dimensional Poisson’s equation**
The potential distribution in the saturation region is obtained by solving two dimensional Poisson’s equation. It is assumed that the region is completely depleted under normal operating conditions and is rectangular in shape. The analytical solution is obtained by the method developed in and the boundary conditions similar to those of Chang and Fetterman [122]. For the calculation purpose it is assumed that there is continuity of potential, there is discontinuity of transverse electric field at the interface due to the presence of two dimensional electron gas and the discontinuity of the surface electric field at the gate edges.

The potential distribution in the saturation region is obtained as

\[
V(x',y) = \frac{2dE_0}{\pi} \sinh \frac{\pi x'}{2d} \sinh \frac{\pi y}{2d} + V_g + \frac{q}{\varepsilon(m)} \left( \int_0^d N_d(y) \, dy - \frac{L}{qW_{sat}} \right) y - \frac{q}{\varepsilon(m)} \int_0^y N_d(y) \, dy \]  

(2.14)

Where \( x' = x - L_1 \). Substituting \( y = d \) in above equation I can find the potential along the 2DEG channel. Now,

\[
V(x')_{x'=L} = V_L = V_d \quad \text{and} \quad V(x')_{x'=L_1} = V_{L_1}
\]

So in the saturation region I can write

\[
V_L - V_{L_1} = \frac{2dE_0}{\pi} \sinh \frac{\pi L_2}{2d}
\]

(2.15)

Where \( L_2 \) is the length of the saturation region and is obtained by substituting \( V_L = V_d \) and the value of \( V_{L_1} \) from above equation and is given by

\[
L_2 = \frac{2d}{\pi} \sinh^{-1} \left( \frac{\pi}{2dE_0} \left( V_d - V'_{gs} + (\rho - R_d) I_c \right) \right)
\]

(2.16)
Now \( L = L_1 + L_2 \). Substituting the values of \( L_1 \) from eq. (2.13) and \( L_2 \) from eq. (2.16) I have total length \( L \):

\[
L = \left( \frac{G_0 V'_{gs}}{I_{ds}} \right) (V'_{gs} - (\rho + R_s)I_{ds}) - \frac{G_0}{2} \left( \frac{V'_{gs}}{I_{ds}} - (\rho^2 - R_s^2)I_{ds} - 2 V'_{gs}(\rho) \right) + \frac{2d}{\pi} \sinh^{-1} \left( \frac{\pi}{2d E_0} \right) (V_d - V'_{gs} + (\rho - R_s) L_1)
\]

### 2.3.2 Small signal parameters

The small signal model is extremely important for active microwave circuit network. The intrinsic gain mechanism of the FET is provided by the transconductance. The transconductance \( g_m \) is a measure of the incremental change in the output current \( I_{ds} \) for a given change in input voltage \( V_{gs} \).

\[
L = \left( \frac{G_0 V'_{gs}}{I_{ds}} \right) (V'_{gs} - (\rho + R_s)I_{ds}) - \frac{G_0}{2} \left( \frac{V'_{gs}}{I_{ds}} - (\rho^2 - R_s^2)I_{ds} - 2 V'_{gs}(\rho) \right) + \frac{2d}{\pi} \sinh^{-1} \left( \frac{\pi}{2d E_0} \right) (V_d - V'_{gs} + (\rho - R_s) L_1)
\]

(2.17)

The current voltage characteristics in the saturation region can be obtained from eq (2.17) by rearranging the terms and replacing \( I_c \) by \( I_{ds} \). After obtaining the current voltage characteristics one can easily derive the small signal parameters by differentiating the drain source current with respect to the gate and drain biases.

### 2.3.3 Transconductance

The derivative of the drain-source current with respect to drain-source voltage while gate-source voltage is held constant is defined as the output conductance of the device. Note that the output characteristics of the device are often more conveniently expressed in
terms of output resistance. These quantities are related simply as the inverse of each other. The output conductance and resistance are given by

\[ g_{ds} = \frac{1}{r_{ds}} = \frac{dI_{ds}}{dV_{ds}} \text{ at } V_{gs} \text{ constant} \]

The output conductance of the device is an important characteristic in analog applications. It plays a significant role in determining the maximum voltage gain attainable from the device and is extremely important for determining optimum output matching properties. In general, for a device to have a low value of output conductance is desirable, or, equivalently, an extremely high output resistance.

The device transconductance is defined as the slope of the \( I_{ds}-V_{gs} \) characteristics with the drain-source voltage held constant. The mathematical statement of this is

\[ g_m = \frac{dI_{ds}}{dV_{gs}} \text{ at constant } V_{ds} \]

The transconductance of the device is one of the most important indicators of device quality for microwave and millimeter wave applications. When all other characteristics are equal, a device with high transconductance will provide greater gains and superior high-frequency performance.

Transconductance \( g_m \) is obtained mathematically by

\[ g_m = \frac{dI_{ds}}{dV_g} = \frac{dI_c}{dV_g} = \frac{\left( \frac{df}{dV_g} \right)}{\left( \frac{df}{dI_c} \right)} \]

(2.18)

Where
\[ f(I_c, V_{ds}, V_{gs}) = L_1 + L_2 - L = 0 \]

I have

\[
g_m = \frac{-V'_{gs} G_0 E_1 + 2 \left( \frac{\rho I_c}{G_0 E_1} + \frac{R_s I_c}{G_0 E_1} \right) - R_s V'_{gs} - \rho^2 I_c + R_s^2 I_c + \frac{2d}{\pi G_0} \sinh^{-1} \lambda_c + \frac{I_c (\rho - R_d)}{G_0 E_0 (\sqrt{1 + \lambda^2})} - \frac{L_c}{G_0}}{G_0 E_1} \]

Where \( \lambda \) is given as

\[
\lambda = \frac{\pi}{2d E_0} \left( V_{ds} - V'_{gs} + (\rho - R_d) I_c \right)
\]

The output conductance of the device is an important characteristic in analog applications. It plays a significant role in determining the maximum voltage gain attainable from the device and is extremely important for determining optimum output matching properties. In general, for a device to have a low value of output conductance is desirable, or, equivalently, an extremely high output resistance. Device dimensions and channel material properties both affect output resistance. In the developed model attempt has been made to calculate the small signal parameters accurately.

### 2.3.4 Output conductance

The output or drain conductance is obtained from

\[
g_d = \frac{dI_{ds}}{dV_{ds}} = \frac{dI_c}{dV_{ds}} = -\frac{dI_c}{dI_c} \frac{dV_{ds}}{dV_{ds}}\left( \frac{df}{dI_c} \right)
\]

And given as
\[
    g_d = \frac{E_1 I_c}{G_0 E_0 \sqrt{1 + \lambda^2}} \left\{ -V'_{gs} - \frac{\rho I_c}{G_0 E_1} - \frac{R_s I_c}{G_0 E_1} - \frac{I_c}{G_0 E_0} \left( 1 + \lambda^2 \right) \right\} - \frac{R_s V'_{gs} - \rho^2 I_c + R_s I_c + 2d \frac{d}{\pi G_0} \sinh^{-1} \lambda + \frac{I_c(\rho - R_d)}{G_0 E_0(\sqrt{1 + \lambda^2})} - \frac{L}{G_0}}{2 \pi LG_0} \right\}
\]

(2.21)

2.3.5 Cut-off Frequency

The unity gain cut off frequency \( f_T \) is an important figure of merit in determining the microwave application of the device. It is given by the ratio of the transconductance and the device capacitances. For simplicity I have taken the capacitance to be a constant quantity. The intrinsic frequency \( f_T \) is given as,

\[
    f_T = \frac{g_m \mu_0}{2 \pi L G_0}
\]

Substituting the value of \( g_m \) in above cut-off frequency relationship then I get

\[
    f_T = \frac{\left( V'_{gs} - R_s I_c - \frac{I_c}{G_0 E_1} - \frac{I_c}{G_0 E_0(\sqrt{1 + \lambda^2})} \right) - \frac{R_s V'_{gs} - \rho^2 I_c + R_s I_c + 2d \frac{d}{\pi G_0} \sinh^{-1} \lambda + \frac{I_c(\rho - R_d)}{G_0 E_0(\sqrt{1 + \lambda^2})} - \frac{L}{G_0}}{2 \pi L G_0} \mu_0}{2 \pi L G_0}
\]

(2.22)

The cut-off frequency, \( f_T \), or the gain-bandwidth determines the ultimate speed of a switching device; \( f_T \) becomes independent of the gate length in long channel devices.

2.4 Results and Discussions

A new two-dimensional analytical model is presented using a simplistic approach. To prove the validity of the model the results are compared with the experimental data and numerical results.
The current voltage characteristic proves that current increases with the increase in drain source voltage as shown in figure 2.06. Complete saturation is not observed due to the inclusion of two-dimensional analyses in the saturation region. The I-V characteristics for different gate voltages are compared with the experimental data. The device exhibited a maximum drain current density of 1.2 A/mm at a gate bias of 2V and a drain bias of 15V. This shows that AlGaN/GaN devices can be effectively used for high power applications. The calculations for drain current have been done for Al mole fraction (m) equal to 0.25. Since I have carried out mole fraction dependent analysis, an increase in mole fraction resulted in an increase in drain current since there is an increase in 2-DEG density at the interface.

![Figure 2.06: Variation of drain current (A/mm) with drain-source bias (V).](image)
The variation of drain current density with gate to source voltage is obtained and compared with the experimental data as shown in figure 2.07. Generally drain current density increases with the increase in gate to source voltage. With the increase in gate reverse bias, the depth of the potential well at the heterointerface decreases which result in a decrease in 2-DEG density and hence the drain current decreases. In addition, if an increase in aluminum mole fraction is made the curve shifts towards left, which indicates a decrease in the value of the threshold voltage ($V_{th}$). In the present case for $m = 0.25$ the value of threshold voltage is -5.2V.

![Graph showing variation of drain current density with gate to source voltage](image)

Figure 2.07: Variation of drain current density (A/mm) with gate to source voltage (V).

The device transconductance ($g_m$) is an important parameter showing device performance. The variation of $g_m$ increases with the increase in $V_{gs}$ and after obtaining a peak value it decreases. A maximum transconductance of 320mS/mm is observed at a drain bias of 8V. A sharp transconductance peak is observed in the present analysis ($L_g = 0.12\mu m$). For
larger gate lengths, the peak flattens and the decrease in $g_m$ for higher gate voltages is not prominent. However, as the gate length is reduced the $g_m$ peak tends to become sharper as the fall in transconductance is faster.

Figure 2.08: Variation of transconductance(S/mm) with gate to source voltage (V).

The peak $g_m$ occurs at the gate voltage that just begins to cause some noticeable occupations of ionized donors under the gate. For low gate voltages, $g_m$ decreases because of the resistive drop through the channel. At higher $V_{gs}$, the gate charge fills the AlGaN donors rather than the charge within the channel with electrons. Due to the strong polarization effects and high saturation velocities, a high value of transconductance is achieved.
Figure 2.09: Variation of Drain-conductance (S/mm) with drain to source voltage (V).

The variation of drain output conductance ($g_d$) is shown in figure 2.09. The drain conductance decreases with the increase in drain to source voltage and increases with the increase in gate to source voltages. Since $g_d$ strongly depends on Al mole fraction, in the present case choice of $m = 0.25$ resulted in $g_d = 0.65$ S/mm at $V_{gs} = -4$ V and $V_{ds} = 4$ V. The drain conductance increases with an increase in Al content and this increase is more prominent at lower gate voltages. This graph is not compared with the experimental data due to lack of availability.

The variation of unity gain cut off frequency ($f_T$) with gate to source voltage as shown in figure 2.10. At a drain bias of 8V and for mole fraction $m = 0.25$ I obtained $f_T = 124$ GHz. This value is slightly overestimated in my analysis since I have taken a constant capacitance value. In chapter 3rd I have explored the effects of various capacitances because in nanoscale transistors these capacitances have significant effects. For proper modeling it is very important to consider all effects whether these parameter have higher
or lower effects. This graph is not compared with the experimental data due to lack of availability.

![Graph showing variation of cut-off frequency (GHz) with gate to source voltage (V).]

Figure 2.10: Variation of cut-off frequency (GHz) with gate to source voltage (V).

A detailed analysis of gate source capacitance and gate drain capacitance can be done to obtain accurate value of $f_T$. Experimental results report $f_T = 121$ GHz for the present specifications. Maximum $f_T$ is attained at the gate voltage corresponding to the transconductance peak. As the gate voltage approaches the threshold frequency decreases sharply as the channel begins to pinch off and there is no current flow.

A large variation of $f_T$ was observed with the variation in drain current density as shown in figure 2.11. When drain current is low, transconductance is low; hence, I have low value of $f_T$. 

m = 0.25  
$L_g = 0.12\mu m$  
$W = 100\mu m$  
$N_A = 5 \times 10^{16} \text{cm}^{-2}$  
$v_A = 9 \times 10^6 \text{cm/s}$  
$V_{ds} = 8V$
Figure 2.11: Variation of cut-off frequency (GHz) with drain current (mA/mm).

Transconductance increases with the increase in drain current, which result in an increase in $f_T$. After peak $g_m$, a fall in $g_m$ results in a fall in $f_T$. Thus a high value of $f_T (>60$GHz) at higher drain currents exhibits the potential of AlGaN/GaN HEMT for high power at high frequency.

2.5 Conclusion

In conclusion, the present model uses a more realistic approach and gives a better insight into the modeling of AlGaN/GaN MODFET’s. All the small signal parameters and their characteristics are well explained in terms of device physics. The model is applicable to the device with gate length as small as 0.1µm.

The spontaneous and piezoelectric polarization effects have been included. Two dimensional analyses have been carried out in the high field region. The output characteristics, device transconductance and cut off frequency for 120nm gate length
device are obtained. Peak transconductance of 320 mS/mm and a cut off frequency of 124 GHz has been obtained.

In the model, the switching parameters are calculated simply from the charge variations in different regions, and the characteristics depend only on the basic device parameters and terminal voltages. The high value of $f_t$ proves the utility of the model in the design of low-power MMICs.