EFFECTS OF LOW-MOBILITY HOLES IN TCAD SIMULATED
GAN/ALGAN/GAN SUPER-HETEROJUNCTION HEMT FOR RF
AND HIGH-POWER APPLICATIONS

A Thesis in
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by
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ABSTRACT

Wide-bandgap materials, such as gallium nitride (GaN), are becoming the forefront in high power and high frequency, microelectronics because of their large breakdown fields and high carrier mobilities. GaN and AlGaN are a popular combination of materials because their heterojunction creates a polarization induced high density and high mobility two-dimensional electron gas (2DEG). The characteristics exhibited encourages an even further promising future for AlGaN/GaN heterostructure devices. High electron mobility transistors (HEMTs) are becoming common place device structures in applicable fields, such as, 5 GHz and 6+ GHz RF technology and radar systems, with the potential to replace travelling wave vacuum tubes (TWT) technology. Field Plates (FP) are implemented with these structures to improve breakdown voltage (BV) by reshaping the electric field. Superjunction (SJ) and super-heterojunction (SHJ) theory is growing rapidly due to its advancements in BV and advanced electric field redistribution.

A unique normally-on GaN/AlGaN/GaN lateral SHJ HEMT is proposed in this work. A p-type GaN SHJ structure with an ohmic Ni-Electrode electrically grounded to the source contact is introduced, along with a SJ charge balance between the P-GaN region and interfacial delta-doping at the upper U-GaN/AlGaN heterointerface. Silvaco TCAD simulation is used to examine the DC and RF characteristics under a Class A amplifier configuration. Data indicates SHJ region features improved uniform redistribution of electric fields over FPs. A decrease in switching performance with increase in frequency due to low mobility holes while switching is observed. Class-A operation of the device as it stands is limited to 1 GHz frequency. 10 GHz results show a PAE limitation of 25% due to low mobility holes. There appears to be an anomalous accumulation of holes in the i-GaN region, most significantly at 10 GHz. Future work must be completed to gain further insight on the source of the i-GaN holes. Current theories suggest possible simulation anomaly.
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Chapter 1

Introduction

The power and radio frequency (RF) semiconductor industry has boomed over the past few decades. Many industries are beginning to, or already, necessitate the use of microelectronic devices. One of the largest and arguably most essential industries are the electric power transmission and distribution utilities. The desire for high power, high temperature, and ultra-high frequency semiconductor microelectronics is most importantly driven by the demand to improve reliability and efficiency around the power grid. There is also great demand for high power and ultra-fast switching semiconductor research and development in the cellular and military communications industries, specifically involving 5 GHz, millimeter wave, and sub-6 GHz transmission and receiving technologies. These applications, alongside, automotive electric vehicles, aerospace vehicles, renewable energies, military and defense technologies, and consumer electronics are just the tip of the iceberg.

The driving mechanism behind these on-going and future innovations are III-V semiconductor materials, for example, silicon carbide (SiC), gallium arsenide (GaAs), and gallium nitride (GaN). They are chosen for the formerly described applications due to their characteristically large band-to-band energy gap and high breakdown voltages (BV), which unfortunately, the popular and affordable material silicon is incapable of providing because of its physical limitations. In line with the subject matter of the thesis, GaN is slowly becoming common place in high power and ultra-fast switching technologies. Aluminum gallium nitride (AlGaN), the alloy material of aluminum nitride (AlN) and GaN, is typically interfaced with GaN due to the substantial energy band offsets and polarization effects, inducing a two-dimensional electron gas (2DEG).

High electron mobility transistors (HEMTs) take advantage of the high density and high mobility 2DEG formation and are excellent choices for fast switching high power electronics. While GaN has a high BV, one caveat with traditional HEMT design is the
substantial peak electric field in the channel at the drain side of the gate when significant biases are applied, resulting in decreased device BV. An approach used to increase the BV is to create a Field Plate (FP) over the drain side channel, effectively reducing the peak electric field by redistributing it [1].

Superjunctions (SJ) are also an interesting semiconductor theory and have been applied to power devices to increase BV by redistributing the electric fields and reduce internal on-resistance ($R_{ON}$). The initial development began to improve the silicon ideal breakdown voltage and $R_{ON}$ limit, but their application naturally evolved to III-V wide-band gap semiconductors. It is a relatively new technology and has recently been utilized in GaN/AlGaN/GaN super-heterojunction (SHJ) Schottky diode works [2]. There are also reports of a vertical HEMT structure with a GaN SJ in the carrier drift region [3] and a lateral structure [4].

P-type magnesium doped wurtzite GaN is notorious for its extremely low hole mobilities. The implementation of SJs with GaN HEMT structures is quite untouched. Similarly, if not, less studied, is the effects of low mobility holes in GaN HEMT structures when in operation at high frequencies.

This thesis proposes a unique normally-on GaN/AlGaN/GaN SHJ HEMT structure for high voltage and high frequency applications, which can be viewed in Fig. 5-1. A P-GaN structure on top of the channel supplies the holes for the SHJ, and when balanced with the n-type delta-dopants at the U-GaN/AlGaN interface, would allow for a low $R_{ON}$ and a high BV. The P-GaN layer acts as a FP due to the nature and location of the SJ. The electric fields across the channel in the location of the SJ are redistributed. The study consists of the simulation and analysis of the device in Class A amplifier operation at a high drain voltage bias of 500 V and an AC frequency of 10 GHz. RF metrics such as power gain ($A_p$), output power ($P_{out}$), and power-added efficiency (PAE) are analyzed. A study on how the low mobility holes in the SJ affect the RF switching performance of the device is also presented. This is completed in two sets of assessments. The first experiment involves varying the low field mobility and observing for any effect in switching performance. The second test varies the input signal frequency from 1 MHz to 10 GHz to observe potential
changes in switching characteristics and hole concentrations. Lastly, there is a brief study on the important of the Ni-Electrode contacts and its requirement to be an ohmic contact.

1.1 Thesis Structure

The following chapters of the thesis are organized as such. Chapter 2 provides the fundamentals of Class A power amplifiers. Chapter 3 touches on the HEMT history and the FP approach. It further explores the theory behind the AlGaN/GaN heterojunction. Chapter 4 examines SJ history and theory, then introduces GaN-based SHJ HEMT devices, which integrate SJ technology into HEMT structures. Chapter 5 describe the research device in question and outlines the simulation structure, design, and simulation experimental procedures. Chapter 6 discusses the DC and AC large signal (RF) simulation results and the results of the P-GaN low mobility hole assessment. A discussion regarding the importance of the nickel electrode/P++-GaN structure work function is also done. Finally, Chapter 7 concludes the thesis presented and discusses future research pathways.
Chapter 2

Class A Power Amplifier Basics

RF power amplifiers are very common in industry, military, utility, and consumer products. The ability to operate at very high power and frequencies well into the GHz range (such as micro-wave and millimeter-wave wavelengths) make them highly desired. RF power amplifiers began as silicon devices, but over time, wide bandgap materials were introduced and now dominate the market and most industries. More background on this can be found in Section 3.1.

Amplifiers can be optimally engineered for precise power and frequency design specifications and can function in different operation classes. Many classes exist, such as A, AB, B, D, E, F, and J, and are chosen due to applications and desired characteristics, and can vary in features such as linearity/nonlinearity, power gain, efficiency, and required number of active and passive components. The SJ HEMT device of interest in this thesis will be designed to operate as a Class A amplifier, therefore, the discussion will be limited to Class A amplifiers.

In this chapter, RF power amplifier Class A operation will be introduced. In Section 2.1, Class A amplifier theory is established. Section 2.2 introduces a few important performance metrics.

2.1 Class A Amplifier Theory

As previously stated, there are many classifications of RF power amplifiers. The following will discuss the Class A amplifier for metal oxide semiconductor field effect transistors (MOSFET) or similar FET devices, and can be further extended to bipolar junction transistors (BJT).
An amplifier device in Class A operation is an ideally perfect linear amplifier. Minor nonlinearities do exist but can be reduced. Class A amplifiers benefit strongly from high power gain and simple circuit diagrams, due to requiring only a single active component, but have limited efficiencies to 50%. The simplest Class-A amplifier is illustrated in Fig. 2-1 and demonstrates a common source amplifier with a load resistance at the drain. Across the active device, or transistor, is a capacitor that represents the internal parasitic capacitances that may exist in the transistor.

**Figure 2.1** Class A Amplifier

When designing a Class A amplifier, an optimum load line matched with the DC characteristics of the transistor is desired. Ideally, one will assume the capacitance is zero, and therefore there is no imaginary reactance component as shown in Fig. 2-2. This means the load line is purely resistive by looking at the Smith Chart on the left-hand side.

The simple analytical equation in Fig. 2-2 shows how the load matching in the system can be represented as an impedance, where $Z_L$ is the total impedance of the load, $R_L$ is the applied load resistance, and $X_L$ is the load reactance associated with the system and occurs if $R_L$ is not the optimum value and if there are any capacitive or inductive elements in a non-ideal scenario. This is illustrated in the right-hand side Smith Chart, specifically for $R_L$ lower than the optimum value, where $R_{HI}$ and $R_{LO}$ are resistances higher or lower, than the optimum load resistance, respectively.
The Class A amplifier is an always on amplifier, signifying that it has a full AC drain current ($I_D$) swing from zero to the maximum drain current $I_{D,\text{MAX}}$ in a single RF period and the drain to source voltage ($V_{DS}$) swings from the knee voltage $V_{\text{KNEE}}$ to the maximum drain to source voltage, or the applied $V_{DD}$. The voltage $V_{\text{KNEE}}$ is illustrated in Fig. 2-3 and is the “turn-on” voltage [5]. The AC current and voltage are at a 180 degrees phase shift from each other, and consequently, $I_{D,\text{MAX}}$ occurs at $V_{\text{KNEE}}$ and the minimum drain current $I_{D,\text{MIN}}$ occurs at $V_{DD}$.

$$Z_L = R_L \pm jX_L = R_L \pm j0$$
To ensure a full voltage and current swing, the DC operating drain and gate voltages must be optimally set. Eqn. 2-1 is the optimum DC operating drain to source voltage (V_{DS,OP}). This is directly between the DC V_{KNEE} and V_{DD} voltages. Eqn. 2-2 is an equation for the DC operating drain current (I_{D,OP}), which is half I_{D,MAX}. It is desired to have an optimum DC operation gate to source voltage (V_{GS,OP}) at the center point of the linear regime of the I_{D} – V_{GS} characteristic plot of the FET to ensure the AC input can swing between zero and the saturation regime current. This is illustrated in Fig. 2-3(b). The RF AC voltage and current transient swings discussed are illustrated in Fig. 2-4.

\[
V_{DS,OP} = \frac{V_{DD} - V_{KNEE}}{2} \quad [2-1]
\]

\[
I_{D,OP} = \frac{I_{D,MAX}}{2} \quad [2-2]
\]

The optimal load resistance R_{L} of the purely resistive load line can be calculated using Eqn. 2-3 [5]. If a R_{L} is chosen that is lower than the optimal value, there would be an observed decrease in power.

\[
R_L = \frac{V_{DS,OP}}{\left(\frac{I_{D,MAX}}{2}\right)} = \frac{V_{DS,OP}}{I_{DS,OP}} \quad [2-3]
\]

**Figure 2-4** Class A Amplifier Always-On AC Voltage and Current Transient Plots [5]
2.2 RF Amplifier Performance Metrics

There are multiple key metrics used to analyze the performance of any RF power amplifier. The first is the amplifier’s power gain ($A_p$). Eqn. 2-4 shows how it is calculated. It is a ratio between the output power and input power in watts and is typically displayed in units of dB:

$$A_p = 10 \log_{10} \left( \frac{P_{out}}{P_{in}} \right) \quad [2-4]$$

where $P_{out}$ is the RF output power and $P_{in}$ is the RF input power when the load line is optimally matched with a load resistance. Eqn. 5-6 and Eqn. 5-7 calculate the RF $P_{in}$ and $P_{out}$, respectively. It is simply a multiplication of the sum of the maximum and minimum input and output currents and maximum and minimum input and output voltages.

To calculate the efficiency ($\eta$) of the device as a percentage, Eqn. 2-5 [6] is used:

$$\eta = 100 \left( \frac{P_{out}}{P_{sup}} \right) \quad [2-5]$$

where $P_{sup}$ is the total DC power supply of the amplifier and can be calculated simply using Eqn. 2-6:

$$P_{sup} = V_{DS,OP}I_{D,OP} \quad [2-6]$$

RF power amplifier performance is judged based on its ability to amplify power and efficiency when doing so. Power-added efficiency (PAE) is typically monitored over $\eta$ because it takes the power gain into consideration when calculating the device efficiency. This is done in Eqn. 2-7 [6] by subtracting $P_{in}$ from $P_{out}$:

$$PAE = 100 \frac{P_{out}-P_{in}}{P_{sup}} \quad [2-7]$$
Chapter 3

AlGaN/GaN HEMTs and Heterojunction Theory

Chapter 3 will introduce and examine the brilliance behind the high electron mobility transistor (HEMT), and more specifically, the heterojunction. Section 3.1 introduces the AlGaN/GaN HEMT and its short history. Section 3.2 introduces heterojunction basics, narrowing into the AlGaN/GaN heterojunction and its phenomenal polarization induced two-dimensional electron gas (2DEG) at the interface.

3.1 Introduction to GaN and AlGaN/GaN HEMTs

Over the last few decades, wide-bandgap semiconductors have become a very sought set of materials in the industry. Their large energy gap proves them useful in high-power and high frequency applications over the most popular material in the semiconductor industry, silicon. Gallium nitride (GaN), grown in a hexagonal-wurtzite lattice structure and typically used in conjunction with aluminum gallium nitride (AlGaN), is a direct wide-bandgap semiconductor material that is becoming widely present in today’s high power and high frequency industries. In less than two decades, the material’s market share has increased at an incredible rate.

3.1.1 The AlGaN/GaN HEMT and Brief History

The future of GaN lies in the development of a fundamental semiconductor device structure: the HEMT. HEMTs take advantage of the AlGaN/GaN heterojunction structure. Just like a MOSFET, there is a gate for electrostatic control of charge carriers with source and drain contacts to provide charge carriers to the channel. However, the main difference with an AlGaN/GaN HEMT is the polarization induced 2DEG at the AlGaN/GaN
heterojunction interface. The AlGaN/GaN heterojunction will be further discussed in Section 3.2. The 2DEG is well studied for its very high electron sheet density, easily greater than $2 \times 10^{13}$ cm$^{-2}$, impressive electron mobility up to 2000 cm$^2$/Vs at room temperature [7], and a predicted peak electron drift velocity of $2 \times 10^7$ cm/s at a field of $10^5$ V/cm [8]. A combination of the 2DEG, the high breakdown field of Wurtzite crystal GaN in the $10^6$ V/cm-range, low on-resistance, low thermal carrier generation rates, and high efficiency makes AlGaN/GaN HEMTs perfect for high speed, high power, and high temperature (operating at temperatures over 700 C) applications. Additional material electrical properties can be found on Table 3-1 in Section 3.3.

The 2DEG formation at equilibrium means that AlGaN/GaN HEMTs will naturally be always-on devices. This can be a problem when it comes to power savings and efficiency. There has been a lot of work done to create enhancement-mode HEMTs, such as using P-GaN gate structures [9].

The first AlGaN/GaN HEMT, illustrated in Fig. 3-1, was designed in 1993 and consisted of an AlGaN/GaN 2DEG channel, fabricated of a Sapphire substrate. It was an always-on device with its pinch-off voltage observed around a gate voltage of -6 V [10].

![First AlGaN/GaN HEMT Structure](image)

**Figure 3-1** First AlGaN/GaN HEMT Structure [10]

GaN HEMT technology research ramped up in the late 1990’s. Between 1996 and 1999, a number of HEMTs and metal-semiconductor field effect transistor (MESFET) device were fabricated and characterized. In 1996, an AlGaN/GaN HEMT was reported
with, at the time, a high breakdown voltage of 340 V with large transconductance, mobility of around 1500 cm²/Vs, and current drive of 150-400 mA/mm [11]. An extraordinary HEMT on silicon carbide exhibited a PAE of 29% at 10 GHz with a total output power of 2.6 W/mm and a power gain of 10 dB in 1999 [12]. A report of a HEMT at 10 GHz with a drain bias of 48 V achieved a PAE of 58% and a power density of 11.2 W/mm [13]. Through the early 2000’s, advancements continued to occur, with increasing PAE, power gain, and power density output. Yet, advancements do begin to gradually slow down due to fabrication costs and difficulties.

### 3.1.2 The Field Plate Approach

HEMT devices are notorious for their extremely large electric fields. In many devices, the electric fields reach a peak value at the gate, specifically on the drain side. To help reduce the peak electric field, which subsequently improves the breakdown voltage of the device with the trade-off of additional capacitance, is the addition of a field plate (FP) [9]. This is a metal plate deposited on top of a surface passivation layer across the AlGaN/GaN 2DEG channel from the gate towards the drain. There are implementations of the FG that connect to the source and are routed over the gate stretching across the 2DEG channel towards the drain. The optimal FP length is dependent on the geometry of the device. In the case of [14], the breakdown voltage increased, and the peak electric field decreased significantly with a FP as short as 1.7 microns. Research shows the use of multiple FPs are functional, as well as sloped FPs. Fig. (below) illustrates a FP stretching from the gate towards the drain over a passivation layer and the 2DEG channel. Simulation results in [14] demonstrate that the FP reduced the peak electric field at the drain side of the gate from $1.2 \times 10^6$ V/cm to $0.9 \times 10^6$ V/cm.
3.1.3 AlGaN/GaN HEMT Applications

Currently, GaN is used in many microwave and radio frequency (RF) applications due to its high-speed capabilities. High power and high frequency radar amplification and switching technology can be improved tremendously through GaN HEMTs. Improving efficiencies and reducing overall power dissipation in military and weather radar systems is desired at the moment. The material has recently been used in communication applications such as 4G and cutting-edge 5G, and may eventually be used in sub-6G bands and millimeter-wave devices [15]. Other applications include WiFi transmitters and receivers, and air traffic control communication systems.

Future adoptions of GaN HEMTs into the automotive industry for electric vehicles is promising since high efficiency power switching and conversion is desired. However, efficiency of GaN HEMTs must improve before then can be a viable option for long-range consumer vehicles.

Alongside radar systems, one of the most exciting potential future for GaN HEMTs is in the power electric utility industry. Improving power quality for customers and commercial/industrial entities by reducing outage time and eliminating power sags is an
on-going goal for electric utility companies. Devices for applications, such as, distribution system voltage regulators for power flow, solid-state high frequency transfer switches, solid-state high voltage circuit breakers, advanced transformer semiconductor components, frequency changers, and high current fault limiters are desired and a future likelihood for GaN HEMTs. Expectations of operating voltages up to 50 kV and 5 kA for GaN HEMTs or possibly other wide-bandgap semiconductors [8]. Fig. 3-3 illustrates the power flow from power generation to distribution and end-use by customers. The devices listed above generally exist in the transmission and distribution sectors, where the sources of outages and power quality issues occur. Overall, GaN HEMTs are predicted to provide much-improved stability, efficiency, and power dissipation characteristics.

Figure 3-3 Power Flow Diagram with Labelled Technology [8]

The leading technology today in high power and high frequency amplification is the traveling wave tube (TWT). Efforts in researching improved electric field management in HEMTs would make them a strong candidate to substitute TWTs in many applications, especially ones where power density scaling is desired.
3.2 AlGaN/GaN Heterojunction Theory

3.2.1 Heterojunction Basics

A homojunction is created when a p-type layer of a semiconductor interfaces with an n-type layer of the same material. When the two layers are interfaced and there is no externally applied bias, the fermi-levels of the two layers must align and be of equal energy. Since each side of the junction is doped of opposite carrier types, a built-in potential will occur. This internal electric field due to differences in the conduction band and valence band energies create equal and opposite diffusion currents. The charge carriers diffuse to opposing sides until a steady-state charge balance is created, also known as equilibrium. This can be identified by the bending of the conduction and valence bands.

In the case of a heterojunction, which occurs when two semiconductor materials of differing composition contact each other, there exists a discontinuity in the conduction band and valence band due to the differing bandgap, electron affinity, and work-function. The difference in conduction band energy and valence band energy between the two materials is expressed as $\Delta E_C$ and $\Delta E_V$, respectively. A heterojunction follows of the same charge carrier diffusion phenomenon and similar band bending discussed even if the materials are intrinsic. Although it may be minimal, this is because the intrinsic carrier concentrations differ between the two materials. Either way, the band discontinuity is present.

There are three types of heterojunctions illustrated in Fig. 3-4: Type I is straddling, Type II is staggered, and Type III is broken gap. Their naming convention comes from the orientation of the conduction and valence band energies. The AlGaN/GaN heterojunction in Fig. 3-6 is a perfect example of a type I heterojunction. The bandgap of AlGaN is quite large compared to the bandgap of GaN and the conduction band and valence band of GaN lie energetically between the AlGaN bands.
Figure 3-4 Three Types of Heterojunctions

A Type II heterojunction, labelled in Fig. 3-4, exists when one of the materials conduction band lies energetically within the bandgap of the second material, and the valence band lies below the valence band of the second material. Another situation can occur when the conduction band sits energetically above the second material’s conduction band, and the valence band sits in the bandgap of the second material. A Type III heterojunction, in Fig. 3-4, occurs when the bandgaps of the two materials do not intersect, i.e., the valence band of one material is energetically higher than both the conduction band and valence band of the second material.

There has been work in quantifying the $\Delta E_C$ and $\Delta E_V$ of the AlGaN/GaN heterojunction. Some common heterojunctions, for example with AlGaS/GaAs, have analytical equations dependent on the Al mol composition of the AlGaS material developed to calculate $\Delta E_C$ and $\Delta E_V$. Theoretical simulation shows a positive correlation between conduction band and valence band offsets with the Al mol composition of the AlGaN/GaN heterojunction. As the Al composition increases, $\Delta E_C$ for the $\Gamma$, $X$, and $L$ sub-bands and $\Delta E_V$ increase. Additionally, $\Delta E_V$ demonstrates significantly less change in energy as compared to the conduction sub-bands [16].

3.2.2 Polarization Theory

An interesting characteristic of the AlGaN/GaN heterostructure is the polarization effects it possesses due to mismatches of the lattices. Spontaneous polarization effect is fascinating since nitride materials are the only of the III-Vs to exhibit this. Polarization
effects are the source for the valuable 2DEG associated with this junction, which allows this structure to be implemented in high-speed switching and high-power devices due to the high mobility and very high sheet density of the 2DEG.

In the following theoretical analysis, one will assume the case of a AlGaN layer grown pseudomorphically on Ga-face wurtzite GaN and that the interface is abrupt. In this situation, as demonstrated on the left side of Fig. 3-5(b), there are two types of polarization in play that induce tensile stress: spontaneous and piezoelectric polarization. Spontaneous polarization occurs independent of an applied external stress [7]. In GaN and AlGaN, it is thought to be a result of the asymmetry in their independent lattices. GaN is a mixed ionic-covalent bond, meaning there is an electronic misbalance because the bond length along the c-axis differs from the a- and b-axis. Therefore, a net dipole electronic balance does not occur, and spontaneous polarization occurs, with its tensor in the negative direction.

When an AlGaN layer is grown on a GaN layer, there is tensile strain at the interface. In a similar fashion, the strain is a result of the change in $a_0$ and $c_0$ lattice constants with respect to the length of the bond parallel to the c-axis, or [0001] [17]. This is known as piezoelectric polarization, and as tensile strain increases, piezoelectric polarization increases because of further change in the lattice constants. It is also important to note that with Ga-face GaN, tensile strain induces a piezoelectric polarization with its tensor in the negative direction, while compressive strain induces a positive tensor.

With this, one can say that in the Ga-face wurtzite lattice, both polarizations are parallel and hence, can be summed as Eqn. 3-1:

$$P_{TOT} = P_{PE} + P_{SP}, \quad [3-1]$$

where $P_{TOT}$ is the total polarization, $P_{PE}$ is the piezoelectric polarization tensor, and $P_{SP}$ is the spontaneous polarization tensor. $P_{PE}$ can be expressed in terms of lattice axis strains and piezoelectric coefficients, shown in Eqn. 3-2:

$$P_{PE} = e_{33} \varepsilon_z + e_{31}(\varepsilon_x + \varepsilon_y), \quad [3-2]$$

$$\varepsilon_x = \varepsilon_y = \frac{a-a_0}{a_0}, \quad [3-3]$$

$$\varepsilon_z = \frac{c-c_0}{c_0}, \quad [3-4]$$
where $e_{33}$ and $e_{31}$ are piezoelectric coefficients that quantify the amount of polarization charge developed per applied unit of stress, $e_z$ is the c-axis strain in terms of the c-axis lattice constant $c_0$ (height of the lattice) and strained constant $c$, $\epsilon_x$ and $\epsilon_y$ are of the same value and represent strain in the a- and b-axis, respectively, in terms of the a-axis lattice constant $a_0$ (length of the edge of the lattice) and the strained lattice constant $a$.

Important values for GaN polarization theory can be found on Table 3-1.

![Figure 3-5](image)

**Figure 3-5 (a) Lattice Structure Orientation and (b) Heterojunction Structure Illustrations for Ga-face (left) and N-face (right) Lattice Orientations [17]**

In [18], equations developed from linearly interpolating datasets of elastic constants $C_{13}(x)$ and $C_{33}(x)$, piezoelectric constants $e_{31}(x)$ and $e_{33}(x)$, and lattice constant $a(x)$ dependent on the Al mol composition $x$ was completed and can be found in source. A similar linear interpolation of $P_{SP}$ data dependent on $x$ was also achieved, shown in Eqn. 3-5:

$$P_{SP}(x) = (-0.052x - 0.029) \frac{c}{m^2}.$$ [3-5]
Substituting the linear interpolations of $C_{13}(x)$, $C_{33}(x)$, $e_{31}(x)$, $e_{33}(x)$, and $a(x)$ into the $P_{PE}$ formula Eqn. 3-2, Eqn. 3-6 results:

$$P_{PE}(x) = 2 \left( \frac{a_0(x) - a(GaN)}{a_0(x)} \right) \left( e_{31}(x) - e_{33}(x) \left( \frac{C_{13}(x)}{C_{33}(x)} \right) \right). \quad [3-6]$$

Using the Eqn. 3-5 and Eqn. 3-6, a prediction of the expected total polarization in the Ga-face structure can be made by substituting them into Eqn. 3-1 with the corresponding linear interpolations. The work of [19] provides the solution for a Ga-face GaN lattice, shown in Eqn. 3-7:

$$P_{TOT}(x) = P_{PE}(x) + P_{SP}(x)$$

$$= -[(3.2x - 1.9x^2) \times 10^{-6} - 5.2x \times 10^{-6}] \text{C cm}^{-2} \quad [3-7]$$

For an N-face GaN lattice of an AlGaN/GaN heterojunction, the sum of the two polarization effects is demonstrated as a very similar equation in [18]. Additionally, this can be used to predict the amount of polarization induced charge at the lower AlGaN/GaN interface illustrated in Fig. 3-5(b).

### 3.2.2.1 Aluminum Composition and Relaxation

As a side note, there has been a great deal of work that demonstrates how an increase in Al mol composition in the AlGaN layer will increase both spontaneous and piezoelectric polarization [30]. One can go further and say that if the GaN layer is relaxed at the AlGaN/GaN interface (instead of stressed as assumed above), and furthermore, assuming the piezoelectric polarization is negligible, spontaneous polarization will nonetheless increase [17]. It is important to say that for added simplicity, the relaxation term in the above analysis is not included. A relaxation equation can be used in the case of partially relaxed, or fully relaxed, AlGaN/GaN interfaces, and is also a strong function of $x$. It would be inserted in Eqn. 3-6 [18].
3.2.3 Polarization Induced 2DEG

Grasping the idea of a polarization induced 2DEG is best understood through analysis via two routes: one, by analyzing the polarization of the structure, and two, by observing the structure’s band diagram. To begin, the polarization analysis from Section 3.2.1 is continued and an abrupt interface assumption is made.

Using Fig. 3-5(b), one can understand the charge balance of the system with respect to the polarization effects. The left side of the image presents a Ga-face GaN/AlGaN/GaN structure. When the layers contact each other, the spontaneous and piezoelectric polarization occur in the same direction due to a perpendicular direction tensile strain at the lower AlGaN/GaN interface. All the layers exhibit PSP due to their chemistry, however, only the AlGaN also exhibits PPE due to the tensile strain at the interface. The direction of the polarization towards the lower GaN layer produces a positive accumulation of charge on the AlGaN side of the lower AlGaN/GaN interface.

Furthermore, there must be a negative charge on the GaN side of the lower AlGaN/GaN interface to compensate the positive charge on the AlGaN side. It is predicted that not only will the lower GaN region become fundamentally negatively charged, but that a quantum-mechanically thin sheet of negative charge will exist within nanometers from the interface on the GaN side. This is also widely referred to in literature as a wide quantum well. The negative charge consists of a cloud of energetically free electrons in the conduction band of the GaN layer.

Further understanding of the 2DEG formation can be made by analyzing the energy band diagram. Fig. 3-6 illustrates an AlGaN/GaN abrupt interface band diagram in equilibrium. When the materials make contact, their Fermi levels must align in equilibrium. The $\Delta E_C$ and $\Delta E_V$, labelled in the diagram, are due to the discontinuity in bandgaps, and do not necessarily equal each other. Due to the polarization effects described above, on the AlGaN side of the interface, positive charge builds up. A corresponding negative charge must compensate the positive charge on the other side of the interface. At the same time, the GaN energy bands to bend downwards, and since the $\Delta E_C$ offset is sufficiently large enough, the conduction band becomes close to or lower in energy than the Fermi-level at
the interface. This induces a very large accumulation of free electrons in the conduction band, or a 2DEG.

![Figure 3-6 Band Diagram of AlGaN/GaN Heterojunction](image)

**Figure 3-6 Band Diagram of AlGaN/GaN Heterojunction**

To develop an equation for the polarization induced 2DEG sheet concentration, one can continue from the polarization equations discussed above. One can describe the charge density as the result of the derivative of the polarization, shown in Eqn. 3-8 [17]:

$$\rho_p = -\nabla P_{TOT}$$  \[3-8\]

where $\rho_p$ is the induced polarization charge density equal to the negated derivative, or gradient, of the total polarization in the enclosed system. Both spontaneous and piezoelectric polarization tensors can be presented as a magnitude of charge. Because the total polarization $P_{TOT}$ is a sum of the two polarization factors, it can be said that the polarization sheet charge density of the abrupt heterojunction is fixed and summed. It can be written elegantly as Eqn. 3-9 [18]:

$$|\sigma(x)| = |P_{PE}(AlGaN)| + |P_{SP}(AlGaN)| - |P_{SP}(GaN)|$$  \[3-9\]
Substituting in $P_{PE}$ into Eqn. 3-9 gives Eqn. 3-10:

$$|\sigma(x)| = 2 \left( \frac{a_0(x) - a(GaN)}{a_0(x)} \right) \left( e_{31}(x) - e_{33}(x) \left( \frac{C_{13}(x)}{C_{33}(x)} \right) \right) + |P_{SP}(x)| - |P_{SP}(0)| \tag{3-10}$$

where $\sigma(x)$ is the total polarization induced sheet charge density. Since the $P_{SP}$ and $P_{PE}$ tensors of the AlGaN are pointing in the same direction, they are added. The $P_{SP}$ of the lower GaN layer points in the same direction, so the charge of this polarization must subtract from the AlGaN’s.

Now suppose there is a Schottky metal gate contact on top of our Ga-face GaN/AlGaN/GaN structure, effectively creating the gate structure of a HEMT. In the quantum well on the GaN side of the lower AlGaN/GaN interface, the electron sheet concentration $n(x)$ can be calculated by subtracting the electrostatic barrier the electrons must overcome from the total fixed sheet charge density $\sigma(x)$.

$$n_s(x) = \frac{\sigma(x)}{q} - \left( \frac{\epsilon_0 \epsilon(x)}{q^2 d_{AlGaN}} \right) (q \phi_B(x) + E_F(x) - \Delta E_C(x)) \tag{3-11}$$

Eqn. 3-11 [17] is the electron sheet concentration, where $\epsilon(x)$, $E_F(x)$, $\phi_B(x)$, and $\Delta E_C(x)$ are the dielectric constant of AlGaN, the Fermi level, the combined barrier of the Schottky contact and the top GaN layer, and the conduction band offset at the lower AlGaN/GaN interface, respectively, all a function of Al composition $x$. Continuing, $q$ is the elemental charge, $d_{AlGaN}$ is the thickness of the AlGaN layer, and $\epsilon_0$ is the permittivity of free space. The linearly interpolated equations for $\epsilon(x)$, $E_F(x)$, $\phi_B(x)$, and $\Delta E_C(x)$ are defined in [17]. The Schottky barrier equation is dependent on the material of the contact.

The presence of a two-dimensional hole gas (2DHG) in a stacked P-GaN/U-GaN/AlGaN/GaN structure is also common, where U-GaN is unintentionally doped GaN due to the magnesium dopants in the P-GaN. The 2DEG on the bottom AlGaN/GaN interface is highly negatively charged, meaning the polarization charge on that side of the AlGaN is highly positive, and consequently, highly negative on the opposite side. This results in a 2DHG accumulation on the U-GaN side of the U-GaN/AlGaN interface.
3.3 GaN Material Properties of Interest

Table 3-1 lists various GaN material and electrical properties. Properties listed are important for the paper and is not comprehensive.

<table>
<thead>
<tr>
<th>Property @ 300 K</th>
<th>Symbol</th>
<th>GaN (Wurtzite)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandgap</td>
<td>$E_g$ (eV)</td>
<td>3.39 [21]</td>
</tr>
<tr>
<td>$e^-$ Mobility (Bulk)</td>
<td>$\mu$ (cm²/Vs)</td>
<td>&lt;1000</td>
</tr>
<tr>
<td>$e^-$ Mobility (2DEG)</td>
<td>$\mu$ (cm²/Vs)</td>
<td>&lt;2000</td>
</tr>
<tr>
<td>$h^+$ Mobility (Bulk)</td>
<td>$\mu$ (cm²/Vs)</td>
<td>&lt;200</td>
</tr>
<tr>
<td>$h^+$ Mobility (2DHG)</td>
<td>$\mu$ (cm²/Vs)</td>
<td>~10</td>
</tr>
<tr>
<td>Lattice Constant a-axis</td>
<td>$a_0$ (Å)</td>
<td>3.189 [17]</td>
</tr>
<tr>
<td>Lattice Constant c-axis</td>
<td>$c_0$ (Å)</td>
<td>5.185 [17]</td>
</tr>
<tr>
<td>Spont. polar. charge</td>
<td>$P_{SP}$ (C/m²)</td>
<td>-0.029 [22]</td>
</tr>
<tr>
<td>Piezoelectric Constant</td>
<td>$e_{31}$ (C/m²)</td>
<td>-0.49 [22], -0.32 [23]</td>
</tr>
<tr>
<td>Piezoelectric Constant</td>
<td>$e_{33}$ (C/m²)</td>
<td>-0.73 [22], -0.63 [23]</td>
</tr>
</tbody>
</table>

3.3.1 Electron and Hole Mobility

GaN is becoming widely used due to its relatively high carrier mobility. With a bulk electron mobility of around 800-1000 cm²/Vs at a temperature of 300 K, it is one of the highest mobility III-V semiconductors. In the 2DEG at the heterojunction interface, the electron mobility has been thought to reach up to 2000 cm²/Vs [7]. The mobility ceiling allows for high performance and fast switching devices.

However, GaN is also notorious for its significantly lesser hole mobility of below 200 cm²/Vs. Research has shown that in reasonably to significantly magnesium doped p-type GaN, hole mobility may reach a rock-bottom 10 cm²/Vs [24]. It is theorized that low enough hole mobilities in P-GaN may negatively impact switching speeds and power efficiency in potential GaN SHJ HEMTs.

The two plots in Fig. 3-7 illustrate electron and hole low-field mobility and its dependence on Wurtzite GaN with n- and p-type doping. The line is an analytical equation approximated from the experimental data points plotted. As doping concentration
increases, the mobility decreases. Various scattering mechanisms affect the electron and hole transport in GaN. These include polar, optical, piezoelectric, and acoustic mobility scattering mechanisms, and have scattering rates proportional with temperature [8].

Figure 3-7 Analytical Approximation of Electron (Left) and Hole (Right) Low-Field Mobility at Room Temperature with Experimental Data [25]
Chapter 4

Introduction to Superjunction Theory

In this chapter, the interesting story and theory of the semiconductor superjunction (SJ) is discussed. Incorporating its structure and theory in silicon power semiconductor devices has proven favorable in high voltage applications. Further development of the SJ using wide-bandgap III-V semiconductors, such as GaN, has been researched and is an obvious advantageous route. Section 4.1 will introduce the SJ by reviewing its history and importance in the continued effort to elevate power electronics to the next level. Section 4.2 establishes the theory behind the $R_{ON}$ and $BV$ relationship of the SJ. Section 4.3 combines the subject matter of Chapter 3 with this chapter, reviewing GaN-based SHJ HEMTs.

4.1 Introduction to the Superjunction

Power semiconductor devices are generally limited by their internal specific on-resistance ($R_{ON}$). The current pathways in semiconductor devices are resistive by nature and can only be reduced to an ideal limit characterized by the material and device geometry. This is called the material’s specific on-resistance theoretical limit, or ideal specific on-resistance ($R_{ON}*A$) relative to the cross-sectional area (A) of the current flow region. For high power devices, a high breakdown voltage (BV) is desired, and unfortunately, is inversely proportional to $R_{ON}$. Breakdown at high electric fields occurs due to an avalanche of impact ionization, which in most cases, renders the device irreparable.

Using wide-bandgap materials such as GaN and SiC improves performance due to their higher breakdown voltages at lower $R_{ON}$ values. The plot in Fig. 4-1 displays the $R_{ON}*A$ ideal limits for silicon, GaN, and SiC and labels a few researched devices that push the theoretical limits.
Nevertheless, improvement is always desired and clever ways can be used to attempt to surpass this theoretical limit. The SJ provides progress by means of a simple and intuitive structure. It is an arrangement of sufficiently thin, highly doped, p- and n-type alternating layers (or pillars). When the dopant charges are balanced equally in the structure, full depletion can occur even, forming a large carrier drift region at small applied biases. In this fully depleted region, electric field redistributes uniformly across the SJ. Ensuring full depletion of the device with a small bias is imperative to achieve the maximum possible breakdown voltage and the lowest $R_{ON}A$. It is also observed that increased number in SJ layers corresponds to a decrease in $R_{ON}A$. Fig. 4-1 shows the SJ $R_{ON}A$ with respect to BV for all three materials. It provides a decrease in $R_{ON}A$ up to, and in some cases more than, a factor of 10. GaN and SiC, along with other wide-bandgap materials, display very promising performance when compared to silicon because of their superior $R_{ON}A$ and BV relationship. The theory behind the improvements from the SJ are explained further in Section 4.2.

![Figure 4-1 Specific On-Resistance and Breakdown Voltage for Various materials and Devices][26]
4.1.1 Brief Superjunction History

Since silicon is now, and in the 1980’s and 1990’s, much more economical to fabricate microelectronic chips with than other materials, especially III-V materials, the SJ was first established to improve silicon power electronics. Moreover, the SJ was used over a decade and a half before the groundbreaking theory was established in [27]. It began with the multilayer varactor proposed in 1978 [28], which used multiple p- and n-type layers to increase the depletion region, consequently, the effective size of the conducting channel. Its structure can be observed in Fig. 4-2.

![Figure 4-2 Multilayer Varactor Diode Cross-Section](image)

The SJ was introduced into transistors using a similar structure as the varactor diode in Fig. 4-2 the V-Gate metal oxide semiconductor field effect transistor (MOSFET) in 1988 [29]. The insulated gate and drain contacts are deposited into the V-shaped wells for improved contact to the channel. The channel contains the lateral p- and n-type silicon stacked SJ layers which are sufficiently thin enough to deplete into each other.

The introduction of the SJ to silicon lateral MOSFETs followed the development of the double reduced surface (RESURF) trench-gate MOSFET [28]. Its typical cross-sectional structure is illustrated in Fig. 4-3(a) and demonstrates a lightly doped n-type drift region due to the depletion region with the upper p-type region. A SJ can be incorporated in the drift region of the double RESURF MOSFET. One application is illustrated in Fig. 4.3(b), where the SJ is multiple thin highly doped layers of p- and n-type silicon. The other possible orientation for SJ layers is alternating pillars through the width of the device.
(orthogonal to the cross-section). As expected, the addition of SJ layers decreases $R_{ON* A}$ and the number of SJ layers is inversely proportional to that value [29].

Vertical trench power MOSFETs can also have their lightly doped n-type drift regions converted into a SJ region for the same benefits. Vertical MOSFETS are arguably higher performance power transistors as compared to their lateral counterparts because of their much larger fully depleted drift region. However, the thicker the drift region is, the higher the resistance contribution is to the total $R_{ON*A}$. Planar transistors are quite limited by BV due to their narrower channel geometry. Fig. 4.3(c) is an example of a vertical trench power MOSFET with the SJ pillars formed vertically along the drift region in the x-axis direction. The SJ pillars can also be rotated 90 degrees and run along the y-axis, or orthogonal to the trench gates, however, it exhibits higher $R_{ON*A}$ [30].

![Figure 4-3](image)

**Figure 4-3 (a)** Conventional Double RESURF MOSFET, **(b)** SJ Double RESFURF MOSFET, **(c)** SJ Vertical Trench MOSFET

It has only been recent that power transistors have transitioned from silicon technology to wide-bandgap materials, such as GaN and SiC. It has been even much more recent that SJs have been incorporated into GaN-based HEMTs. Information about this is presented in Section 4.3.2.
4.2 Superjunction Theory

SJ theory was eventually published in the landmark paper by T. Fujihira in 1997 [27]. It laid the mathematical framework defining the physics behind the $R_{ON}\cdot A$ and BV relationship in the devices described in Section 4.1. Sections 4.2.3-4 details some of the theory, starting with the solution for the maximum electric field (or critical electric field) in the SJ, and finishing with the ideal specific on-resistance.

4.2.1 On-Resistance and Breakdown Explained

First, an understanding of the resistive path in a transistor must be grasped. Fig. 4-4 illustrates the resistive current path of a conventional silicon vertical power MOSFET from the source to the drain. When the device is biased into linear or saturation region, the electrons transport from the source contact, through the p-type well, and into the carrier accumulation region directly under the gate contact and dielectric. The carriers then drift through the relatively large n-type drift region and finally into the drain contact. The resistances in each region can be individualized, and added in series to make up the entire $R_{ON}$, as shown in Eqn. 4-1:

$$R_{ON} = R_s + R_n + R_{ch} + R_a + R_{JFET} + R_{drift} + R_D$$  \[4-1\]

In the equation, $R_s$ is the source contact resistance, $R_n$ is the resistance in the n$^+$ doped source region, $R_{ch}$ is the channel region resistance, $R_a$ is the accumulation region resistance, $R_{JFET}$ is the parasitic resistance that typically occurs in a junction field effect transistor (JFET), $R_{drift}$ is the drift region resistance, and $R_D$ is the resistance at the drain and n$^+$-type doped drain region.
Consequently, for vertical power MOSFETs with a large drain to source bias, the large drift region contributes almost all the resistance in the current path [26]. Replacing the single lightly doped n-type drift region with a SJ of multiple layers, or pillars, of highly n- and p-type doped material would allow for a significant decrease in resistance because of the large depletion region that would result. When the doping charges are properly balanced, the region will take on “intrinsic” characteristics. At the same time, when a significant drain to source biases is applied, the ionic positive and negative charges of the depletion region also screen, or “block”, the very large electric field in the drift region, causing a uniform redistribution of the electric field across the SJ. This permits the application of much higher drain to source biases, increasing the BV of the transistor.

Breakdown at high electric fields occurs due to an avalanche of impact ionization. When the electric fields in a crystalline material are high enough, it causes electrons entering the depletion region to become highly energized. They accelerate rapidly producing high-energy impact collisions with atomic lattice. This causes the atom to ionize by forming a new electron-hole pair. If the secondary electron has a high enough energy, it can then create another electron-hole pair through a secondary impact ionization event.
As the avalanche event occurs, the material can become damaged, often rendering the device severely irreparable.

### 4.2.2 Optimal Dopant/Charge Balance

For the SJ layers to deplete as perfectly into each other as possible, there must be an optimal balance of n-type and p-type doping with respect to the cross-sectional area of the SJ layers. Fig. 4.5 is an example of an ideal SJ FET, which is a simple SJ structure to be used to understand the SJ theory.

![Simple SJ FET Structure](image)

**Figure 4-5** Simple SJ FET Structure

The SJ is simply a set of pn-junctions. If it is assumed that the structure in Fig. 4-5 is an ideal FET, that the pn-junction regions have perfect geometry and equal depth along the x-axis orthogonal to the cross-sectional surface, and that there are equal number of p-type and n-type regions, simplifying the analysis to a single pn-junction. If one assumes the n-type doping (Nd) and the p-type doping (Na) are equal, then at a negligible bias, full
depletion of the two regions can be expected if their cross-sectional areas are equal. One can then state Eqn. 4-2:

\[ A_p N_A = A_n N_D \]  \[4-2\]

where \( A_p \) and \( A_n \) are the cross sectional areas of the p-type and n-type layers, respectively, and are equal to \( A_x \). This is scalable to any number of SJ layers or pillars, along with various cross-sectional areas.

In fabrication, there are two ways to go about balancing the dopant charges. One way is by controlling the doping of each material layer when deposited if each layer has the same geometry. A more economical way is by controlling the geometry of the layers if doping varies. For example, in [31], a GaN-based SHJ Schottky barrier diode was fabricated on a wafer with a predetermined p-GaN doped layer. Simply etching away at the p-GaN layer to a calculated thickness to reach charge balance with the AlGaN/GaN 2DEG was accomplished.

### 4.2.3 Critical Electric Field and Breakdown Voltage Derivation

To find the maximum breakdown voltage, the maximum electric fields must be found. Using the simple structure in Fig. 4-5 and assuming a dopant charge balance, i.e. p-type doping \( N_A \) equals n-type doping \( N_D \), as well as the junctions being abrupt, steps can be taken to understand this relationship beginning with a set of Poisson’s equations in Eqn. 4-3 and Eqn. 4-4 [27]:

\[ \frac{\partial E}{\partial z} = \frac{q N_A}{\varepsilon_s} \] \[4-3\]

\[ \frac{\partial E}{\partial z} = -\frac{q N_D}{\varepsilon_s} \] \[4-4\]

where \( E \) is the electric field, \( q \) is the elementary charge, and \( \varepsilon_s \) is the dielectric constant of the semiconductor multiplied by the permittivity of free space. Eqn. 4-3 is for a single p-type layer and Eqn. 4-4 is for a single n-type layer. Assuming the abrupt junction interface occurs at the origin of the z-axis and the electric field in the z-direction is zero, the integration of Eqn. 4-3 and Eqn. 4-4 with the bounds \( \{0 \leq z \leq z_p\} \) and \( \{z_n \leq z \leq 0\} \),
respectively, results in the electric field in the z-direction across the p-type layer (Eqn. 4-5 [27]) and n-type layer (Eqn. 4-6 [27]).

\[
E_z = \frac{qN_A(z-z_p)}{\varepsilon_s} \quad [4-5]
\]

\[
E_z = \frac{qN_D(z_n-z)}{\varepsilon_s} \quad [4-6]
\]

The maximum electric field in the z-direction is described at \( z = 0 \) where \( z_p \) and \( -z_n \) equal half the thickness, \( d \), of the p-type and n-type layer, respectively. The maximum electric field in the z-direction also must be less than the critical electric field \( (E_{crit}) \) of the semiconductor material to avoid impact ionization breakdown. This can be found in Eqn. 4-7 [27]:

\[
|E_z|_{max} = \frac{qN_Ad}{2\varepsilon_s} = \frac{qN_Dd}{2\varepsilon_s} < E_{crit} \quad [4-7]
\]

Expanding the analysis to the entire structure, once the layers are fully depleted, the y-direction electric field will now increase to an absolute maximum value once the gate and drain biases are applied. The max electric field in the y-direction occurs when the drain to gate voltage reaches the breakdown voltage [27], shown in Eqn. 4-8:

\[
|E_y|_{max} = |E_z|_{max} + \frac{V_B}{l_y} \quad [4-8]
\]

where \( V_B \) is the BV and \( l_y \) is the length of the SJ in the y-direction. The condition for breakdown of the simple FET is dependent on the maximum change in electric field in the y-direction. Lastly, the square of the maximum absolute electric field is a summation of the electric field in the x-, y-, and z-directions, and is equal to the critical electric field, as shown in Eqn. 4-9 [27]:

\[
|E|_{max}^2 = E_x^2 + |E_y|_{max}^2 + E_z^2 = E_{crit} \quad [4-9]
\]

where \( E_x \) is the x-direction electric field.

### 4.2.4 Ideal Specific On-Resistance Analytical Equations

As described above in Section 4.1 and Section 4.2.1, there is an inversely proportional relationship between \( R_{ON}*A \) and BV. In [27], the specific on-resistances with
respect to the BV for conventional lateral RESURF MOSFETs (Eqn. 4-10), the lateral
double RESURF SJ trench gate MOSFET in Fig. 4-3(b) (Eqn. 4-11), and the planar vertical
SJ trench power MOSFET in Fig. 4-3(c) (Eqn. 4-12) are:

\[
R_{ON} \cdot A_L = \left( \frac{27}{8} \right) \frac{v_B^2}{\mu \varepsilon \varepsilon_r E_{crit}^2},
\]

[4-10]

\[
R_{ON} \cdot A_z = \left( \frac{1}{\alpha (1-\alpha)^2} \right) \frac{dv_B^2}{l_z \mu \varepsilon \varepsilon_r E_{crit}^2},
\]

[4-11]

\[
R_{ON} \cdot A_y = \left( \frac{1}{\alpha (1-\alpha)} \right) \frac{dv_B}{\mu \varepsilon \varepsilon_r E_{crit}^2},
\]

[4-12]

respectively, where \( \mu \) is the majority carrier mobility, \( l_z \) is the z-direction length,
\( A_L, A_z, \) and \( A_y \) are their respective cross-sectional areas, and \( \alpha \) is the optimum coefficient.

The author of [27] sets \( \alpha \) to 1/3 and 1/2 for Eqn. 4-11 and Eqn. 4-12, respectively.
This means that the lowest possible (ideal) \( R_{ON} \cdot A \) value for the two SJ devices occurs
when the maximum z-direction electric field is (1/3)\( E_{crit} \) and (1/2)\( E_{crit} \). Therefore, the ideal
\( R_{ON} \cdot A \) for the lateral SJ device and vertical SJ device is Eqn. 4-13 and Eqn. 4-14,
respectively: [27]:

\[
R_{ON} \cdot A_z = \left( \frac{27}{4} \right) \frac{dv_B^2}{l_z \mu \varepsilon \varepsilon_r E_{crit}^3},
\]

[4-13]

\[
R_{ON} \cdot A_y = (4) \frac{dv_B}{\mu \varepsilon \varepsilon_r E_{crit}^2},
\]

[4-14]

Notice how the vertical SJ MOSFET has a linear relationship with BV while the
lateral SJ MOSFET has a square relationship with BV.

4.3 GaN Super-Heterojunction HEMTs

This section provides a brief introduction into the published work that brought both
SJ and GaN-based HEMTs together into one device. It introduces devices similar to the
structure of interest in this thesis, further explained in Chapter 5. The current research in
GaN SHJ HEMTs has been quite limited. A standout device is the vertical HEMT device
from 2013 imaged in Fig. 4-6 with a SJ region in the bulk region of the transistor. The n-
region pillar in the SJ structure allow current to flow vertically to the drain from the 2DEG
region parallel with the source. The P+-GaN region at the source ensures an ohmic contact
exists, and the P⁺-GaN under the gate gives the device its enhancement-mode characteristic. In simulation, it exhibited a breakdown voltage of 12.4 kV with an $R_{ON}$ of 4.2 mΩ-cm² [3]. A higher number of SJ pillars can be used, as well, if the doping charge is balanced and they can be formed in the given cross sectional surface area.

![Vertical GaN/AlGaN SJ HEMT](image)

**Figure 4-6** Vertical GaN/AlGaN SJ HEMT [3]

A lateral GaN SHJ HEMT demonstrated in 2011 is quite like the device proposed in this work. Fig. 4-7 shows the cross-section of the device. Notice the fourth electrode title base that is electrically connected to the gate. Lateral heterojunction devices do not perform as well as vertical devices do because of the geometric limitations of the channel, however, it still has a breakdown voltage of 1.1 kV with an $R_{ON}$ of 6.1 mΩ-cm² [4]. The SHJ used here can also be called a natural SJ due to the formation of the two-dimensional hole gas (2DHG) opposing the 2DEG layer on the U-GaN side of the top U-GaN/AlGaN interface.
The work of [31] describes the implementation of a GaN SHJ in a Schottky diode structure. Instead of allowing the formation of a 2DHG on the GaN-side of the upper GaN/AlGaN interface, delta-doping and a p-type doped GaN region is introduced, which are charge balanced as described in SJ theory. The delta-doping is an interfacial doping at the U-GaN/AlGaN interface, as illustrated in Fig. 4-8. It assists in increasing the 2DEG sheet density through thermionic excitement of electrons over the AlGaN barrier. The charge balance with the P-GaN region improves the reverse breakdown voltage of the diode by reducing the peak electric field and redistributing the field uniformly across the SHJ region. The HEMT device of interest in this thesis work includes both the delta-doping region and P-GaN region with the charge balance condition met for the purpose of similar electric field redistribution, allowing high operating voltages.
Chapter 5

Unique GaN Super-Heterojunction HEMT Structure and Simulation

This chapter introduces the unique high power and high frequency GaN/AlGaN/GaN SHJ HEMT device in question and details the simulation set up taken to achieve the discovered results, discussed in Chapter 6. Section 5.1 begins with an introduction to the device’s structure, Section 5.2 describes the Silvaco TCAD code displayed in Appendix B, and lastly, Section 5.3 explains the procedures used to extract the key RF performance metrics of the device and analyze the performance impacts due to low-mobility holes.

5.1 Super-Heterojunction HEMT Device Structure

The unique structure of the device can be found in Fig. 5-1. It consists of a 250 nm thick “intrinsic” GaN layer specified as a substrate in the material region definition. Typically, this would be a layer of grown undoped GaN on top of a substrate of differing material, possibly Sapphire. This detail is not included in the simulation.

Above is a 10 nm layer of undoped Al$_{0.25}$Ga$_{0.75}$N and a 10 nm layer of unintentionally doped U-GaN. Between these two layers is a 1 nm defined layer of delta doping. The purpose of delta doping is to increase the electron sheet density of the 2DEG, which is formed at the lower Al$_{0.25}$Ga$_{0.75}$N/i-GaN interface. The defined thickness should suffice since delta doping is defined as an interfacial atomically thin highly doped region. It is defined at the U-GaN/Al$_{0.25}$Ga$_{0.75}$N interface instead of the Al$_{0.25}$Ga$_{0.75}$N/i-GaN interface for two reasons: one, high doping concentrations introduce impurity scattering which would decrease the 2DEG mobility, and two, delta doping introduces higher-energy surface states making thermal excitation of carriers from the P-GaN and U-GaN over the Al$_{0.25}$Ga$_{0.75}$N barrier is more energetically favorable.
The purpose of the U-GaN region is to mitigate the downward leakage of Mg-dopants into the AlGaN and 2DEG region.

The ohmic source and drain contacts, with a work function equivalent to NiAu and 2 microns long, lie on opposite sides of the channel making bottom contact to i-GaN and sidewall contact to the U-GaN/delta-doping/Al0.25Ga0.75N structure. The Schottky gate, with a work function equivalent to Ni and 0.5 microns long, contacts the U-GaN for electrostatic control of the channel.

Most of the channel has a natural SHJ structure, with a P-GaN region 10 microns long by 200 nm thick, topped with P++-GaN and ohmic Ni-electrode (work function of Ni) layers each with dimensions 1 micron long by 10 nm thick. The purpose of the Ni-electrode is to allow holes to deplete from the structure when biased, completing the natural SHJ structure. It is electrically grounded with the source contact, unlike the device in Fig. 5-2 whose “base” electrode is connected to the gate. This makes the device quite unique. The P++-GaN regions purpose is ensure the Ni-electrode is an ohmic contact. The remaining space is filled with Si$_3$N$_4$. 

Figure 5-1 GaN-Based SHJ HEMT Cross-Section
The doping of each section, along with their dimensions can be found on Table 5-1. The doping for the P-GaN region is decided in regard to the doping balance SJ theory. Simulation simplifies this by allowing one to simply set a doping amount based on the geometry defined. In fabrication, it is a different story, and may require one to edit the geometry instead by means of selective dry etching. Appendix A discusses the dopant balancing procedure used.

<table>
<thead>
<tr>
<th>Region</th>
<th>Length (μm)</th>
<th>Thickness (μm)</th>
<th>Doping</th>
</tr>
</thead>
<tbody>
<tr>
<td>P**-GaN</td>
<td>1.000</td>
<td>0.010</td>
<td>$1 \times 10^{19}$ /cm$^3$</td>
</tr>
<tr>
<td>P-GaN</td>
<td>10.000</td>
<td>0.200</td>
<td>$3.75 \times 10^{17}$ /cm$^3$</td>
</tr>
<tr>
<td>U-GaN</td>
<td>12.000</td>
<td>0.010</td>
<td>~</td>
</tr>
<tr>
<td>Delta-Doping</td>
<td>12.000</td>
<td>0.001</td>
<td>$7.5 \times 10^{19}$ /cm$^3$</td>
</tr>
<tr>
<td>Al$<em>{0.25}$Ga$</em>{0.75}$N</td>
<td>12.000</td>
<td>0.010</td>
<td>~</td>
</tr>
<tr>
<td>i-GaN</td>
<td>16.000</td>
<td>0.250</td>
<td>~</td>
</tr>
<tr>
<td>Source/Drain</td>
<td>2.000</td>
<td>0.241</td>
<td>~</td>
</tr>
<tr>
<td>Gate</td>
<td>0.500</td>
<td>0.221</td>
<td>~</td>
</tr>
<tr>
<td>Nickel Electrode</td>
<td>1.000</td>
<td>0.010</td>
<td>~</td>
</tr>
</tbody>
</table>

**Table 5-1 Device Geometry and Doping**

5.2 Simulation Design Implementation

The philosophy behind creating the simulation is detailed briefly in Appendix C.

5.2.1 Variables, Mesh, Materials, Regions, and Electrodes

Appendix B is the Silvaco TCAD source code for the simulation of the device. It can be broken into multiple parts. The first portion of code sets variables used for specifying length and thickness geometric parameters. Underneath the variables is the mesh definition. In this section, the width of the mesh is specified to 1000 microns, or 1 millimeter thick, and the x-axis and y-axis mesh widths are defined. The mesh is defined to be finer in regions of higher carrier concentrations and changes of. The following section defines each region. The GaN and Al$_{0.25}$Ga$_{0.75}$N regions are defined with a parameter
“asub” and set it equal to 3.189. It sets the substrate lattice constant to ensure there is a net-polarization among the GaN and AlGaN regions.

The next section defines the electrodes. They include the source, drain, gate, and Ni-electrode. Uniform doping is then defined to simplify convergence. The doping values for each section can be found on Table 5-1. After detailing the regions and doping, the materials and their properties are defined. Table 5-2 defines the material property parameters used.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>edb</td>
<td>Donor energy level</td>
</tr>
<tr>
<td>eab</td>
<td>Acceptor energy level</td>
</tr>
<tr>
<td>taun0</td>
<td>SRH electron minority carrier lifetime</td>
</tr>
<tr>
<td>taup0</td>
<td>SRH hole minority carrier lifetime</td>
</tr>
<tr>
<td>copt</td>
<td>Radiative recombination rate</td>
</tr>
<tr>
<td>nsrcen</td>
<td>SRH electron concentration parameter</td>
</tr>
<tr>
<td>nsrchp</td>
<td>SRH hole concentration parameter</td>
</tr>
</tbody>
</table>

5.2.2 Physics and Mobility Models

The following section defines the physics models used. The “consrh” model enables the Shockley-Read-Hall (SRH) recombination model that is dependent on concentration. The following equation calculates SRH recombination rate [32]:

$$R_{SRH} = \frac{pn-n_{ie}^2}{\tau_p [n+n_{ie} e^{-\frac{ETRAP}{kT_L}}] + \tau_n [p+n_{ie} e^{-\frac{ETRAP}{kT_L}}]}$$  \[5-1\]

where $p$ and $n$ are the hole and electron concentrations, respectively, $n_{ie}$ is the intrinsic electron concentration, ETRAP is the trap state energy, $kT_L$ is the Boltzmann’s constant multiplied with the local lattice temperature, and $\tau_p$ and $\tau_n$ are the hole and electron minority lifetimes.

The “auger” model enables the Auger recombination model. Eqn. 5-2 is the Auger recombination rate in terms of auger coefficients AUGN and AUGP [32]:
\[ R_{\text{Auger}} = \text{AUGN}(pn^2 - nn_{ie}^2) + \text{AUGP}(np^2 - pn_{ie}^2) \]  

[5-2]

The “fermi” model enables fermi-dirac statistics. The model “incomplete” enables incomplete ionization of impurities and “optr” enables the optical generation/radiative recombination model. Eqn. 5-3 is the total rate of band to band generation/recombination [32]:

\[ R_{np}^{\text{OPT}} = \text{copt}(np - n_{ie}^2) \]  

[5-3]

where copt is the radiative recombination rate.

The model “POLARIZATION” enables the GaN polarization effects. It models spontaneous and piezoelectric polarization using equations similar to the ones in Section 3.2.2. The “polar.scale” statement allows one to modify the sign or magnitude of the polarization charge. The “calc.strain” statement enables the calculation of the strain induced by the lattice mismatch between the GaN and AlGaN materials.

The mobility models used are FMCT and GANSAT. It is found that the best selection of models for the device are as follows: “FMCT.n”, “GANSAT.n”, “GANSAT.p”, and “MUP0 = 10”. Roughly, a maximum low-field mobility (MUP0) of 10 cm²/Vs is used because work has been done that concludes one should expect this for p-type magnesium doped GaN [33]. The FMCT model is the Faramand Modified Caughey Thomas model for electrons and holes. It is composition and temperature dependent for low electric fields and is expressed as Eqn. 5-4 for electrons (and a similar equation for holes) [32]:

\[
\mu_n(T, N) = MU1N.FMCT \left( \frac{T}{300} \right)^{BETAN.FMCT} + \frac{\left( MU2N.FMCT - MU1N.FMCT \right) \left( \frac{T}{300} \right)^{DELTAN.FMCT}}{1 + \left( \frac{N}{NCRITN.FMCT} \right)^{GAMMAN.FMCT} \left( \frac{T}{300} \right)^{ALPHAN.FMCT} \left( \frac{T}{300} \right)^{EPSP.FMCT}} \]

[5-4]

where T is the local lattice temperature, MU1N and MU2N are mobility coefficients, N is the doping concentration, and NCRITN, ALPHAN, BETAN, GAMMAN, and DELTAN are other fitting parameters. It is important to note that the lattice temperature is set to be constant at 300 K and will not change.
The GANSAT model applies a high electric field mobility dependence for nitride materials and implements Eqn. 5-4 as the low field mobility parameter. Eqn. 5-5 is the mobility equation for holes (a similar equation is defined for electrons) [32]:

\[
\mu_p = \mu_p(T,N) + \text{VSAT} P \left( \frac{E}{E_{GANSAT}} \right)^{N_{GANSAT}} + \left( \frac{E}{E_{GANSAT}} \right)^{N_{GANSAT}} + \frac{E_{GANSAT}}{E_{GANSAT}} + \text{VSAT} \frac{E}{E_{GANSAT}} \left( \frac{E}{E_{GANSAT}} \right)^{N_{GANSAT}} \tag{5-5}
\]

where E is the electric field and the other variables are various fitting parameters based on the material.

When using “MUP0=10” with the current mobility model definitions, one can have a default electron low-field and high-field mobility dependence, while specifying the low-field mobility dependence for holes to 10 cm²/Vs. The statement “MUP0” allows one to set the \( \mu_h(T,N) \) value in Eqn. 5-5, which would have been calculated by Silvaco using the FMCT.h low field mobility model for holes. This statement is used to test the impact hole mobility has in the device simulation, discussed further is Section 5.3.

It is also noteworthy to say that impact ionization models are not used in this set of simulations. Breakdown voltage is not properly modelled in this device and may be a focus point in future papers.

### 5.2.3 Contacts, Solution parameters, and Solve Statements

The contact naming is then defined. Specifically, for AC large signal simulation, there is also a resistance applied to the drain to model an amplifier. The output parameters are then listed. They include conduction and valence band energies, charge, polarization induced charge, electron and hole mobilities, and surface charge. The method statement defines the solution runtime parameters, such as, number of iterations, maximum number of simulation convergence trap steps, maximum time step (RF large signal simulation only), total time (RF large signal simulation only), the convergence accuracy limit, and the current convergence criteria.

The solutions of the simulation depend on the desired output. Both DC characteristics and AC large signal simulations are executed. For both, an initial solver is
executed and an equilibrium structure file is saved. For DC characteristic simulations, initial solvers are used first. \( I_D - V_{GS} \) solutions used the following solve statements: an initial solver, followed by a drain voltage solver, and lastly a gate voltage solver in voltage steps that promoted simulation convergence. A similar setup is used for each curve in the \( I_D - V_{DS} \) characteristics: an initial solver, followed by a gate voltage solver, and lastly a drain voltage solver in voltage steps that promoted convergence. Every curve simulated resulted in a .log file, which contained raw data for plotting currents and voltage, and .str files, which contained two-dimensional device parameter data, such as, mobilities, carrier concentrations, and electric field.

AC large signal simulations required AC waveform specifications, and initial DC \( V_{GS} \) and \( V_{DS} \) biasing, and then a transient solver. The transient solver looped through all the AC waveforms, and then reported back maximum/minimum drain and gate voltages and currents, and their transient times. It then calculated input/output powers, power gain, and PAE. Finally, all data is exported to a .log file for plotting transient waveforms, a .dat file for chosen values of importance, and a .final file, that listed a neat output of the powers, power gain, and PAE of each AC waveform. More information is detailed in Section 5.3. All detailed simulation solvers, syntax, and calculated/extracted values can be found in Appendix B.

5.3 Simulation Experimental Procedures

5.3.1 Simulation of High Voltage 10 GHz Performance

The DC simulations are completed first with a specified low-field mobility \( M_{UP0} \) of 10 cm\(^2\)/Vs. This includes the simulation of a set of \( I_D - V_{DS} \) lines and then a corresponding \( I_D - V_{GS} \) plot of a calculated drain voltage. A drain voltage of 500 V is chosen because it is thought to be high enough to extract high operating voltage key performance metrics at 10 GHz and low enough to probably not be affected by breakdown. It is important to note that an absolute breakdown voltage is not tested for in this study, so
impact ionization physics models are not used. Gate voltages ranged from $V_{GS} = -5.0$ V to 1.0 V and the curves on the $I_D - V_{DS}$ graph are plotted in increments of 0.50 V. Using a Class-A amplifier load line model, the optimal drain voltage $V_{DS}$ is chosen with respect to the operating $V_{DD}$ and the knee voltage. The load resistance $R_D$ is also calculated from this data. The $I_D - V_{GS}$ plot is then simulated using the optimal $V_{DS}$ value calculated. From the plot data, the $V_{GS}$ value that correlates to half the maximum drain current $I_{D,MAX}$ is extracted.

After calculating the desired load resistance and optimal $V_{DS}$ and $V_{GS}$ DC values, the AC large signal simulation is completed. The AC frequency used is 10 GHz, and the input signal at the gate ranged from 0.75 V to 3.00 V in amplitude, in 0.25 V increments. All 12 waveforms are simulated for one period, and after each period, their input power $P_{in}$, output power $P_{out}$, power gain $A_P$, and power-added efficiency PAE (%) are calculated and plotted.

To calculate the metrics, various minimum/maximum currents and voltages are extracted from the gate and drains, along with their transient times. To calculate RF input power ($P_{in}$), RF output power ($P_{out}$), DC power supply ($P_{sup}$), power gain ($A_P$), and power-added efficiency (PAE), Eqn. 5-6 through Eqn. 5-10 are used, respectively:

$$P_{in} = \frac{|(V_{gs,max} - V_{gs,min})(I_{g,vgs,max} - I_{g,vgs,min})|}{8} \quad (W/mm),$$  \hspace{1cm} [5-6]

$$P_{out} = \frac{|(V_{ds,max} - V_{ds,min})(I_{d,vds,max} - I_{d,vds,min})|}{8} \quad (W/mm),$$  \hspace{1cm} [5-7]

$$P_{sup} = V_{DS,OP} \frac{I_{d,vdsmax} + I_{d,vdsmin}}{2} \quad (w/mm),$$  \hspace{1cm} [5-8]

$$A_P = 10 \log_{10} \left( \frac{P_{out}}{P_{in}} \right) \quad (dB),$$  \hspace{1cm} [5-9]

$$PAE = 100 \left( \frac{P_{out} - P_{in}}{P_{sup}} \right) \quad (\%).$$  \hspace{1cm} [5-10]

The simulation extracted variables used in these equations are explained in Table 5-3. The metrics are calculated for each AC amplitude at the end of each iteration’s period.
### Table 5-3 Simulation Extracted Variables

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{gs,max}$</td>
<td>Maximum gate voltage of the period</td>
</tr>
<tr>
<td>$V_{gs,min}$</td>
<td>Minimum gate voltage of the period</td>
</tr>
<tr>
<td>$V_{ds,max}$</td>
<td>Maximum drain voltage of the period</td>
</tr>
<tr>
<td>$V_{ds,min}$</td>
<td>Minimum drain voltage of the period</td>
</tr>
<tr>
<td>$I_{g,Vgs,max}$</td>
<td>Gate current measured at the time of the maximum gate voltage of the period</td>
</tr>
<tr>
<td>$I_{g,Vgs,min}$</td>
<td>Gate current measured at the time of the minimum gate voltage of the period</td>
</tr>
<tr>
<td>$I_{d,Vds,max}$</td>
<td>Drain current measured at the time of the maximum drain voltage of the period</td>
</tr>
<tr>
<td>$I_{d,Vds,min}$</td>
<td>Drain current measured at the time of the minimum drain voltage of the period</td>
</tr>
</tbody>
</table>

#### 5.3.2 Low Mobility Hole Performance Assessment Simulations

To assess if the low-mobility holes affect the performance of the device in high-frequency applications, an AC large signal simulation, as described in Section 5.4.1, is completed for multiple holes mobilities. The DC simulation does not change with varying mobility, and therefore, the same characteristics can be used to set up all the AC large signal simulations. The combination of the FMCT.n, GANSAT.n, and GANSAT.p mobility models are used. The statement “MUP0 = $\mu_{th}$” is used to define the low-field dependent hole mobility. Mobility values of 0.1, 1, 10, and 100 cm$^2$/Vs are used. Changes in $P_{out}$, $A_P$, and PAE values are analyzed for possible correlations. The transient waveform plotted from the .log file will also be visually observed.

The values 0.1, 1, 10, and 100 cm$^2$/Vs are chosen to see if a change in mobility in multiple factors of 10 could lead to positive or negative affects in the 10 GHz performance. A maximum low-field mobility of 10 cm$^2$/Vs is the expected value for p-type magnesium doped GaN [34]. Hole mobility lower than 10 cm$^2$/Vs in the P-GaN regions may be expected in real-world scenarios.

Setting a constant mobility of 1000 cm$^2$/Vs for the holes is also tested. The mobility model configuration for this is FMCT.n, FMCT.p, FMCT.MUP1=1000, and
FMCT.MUP2=1000. This is an extreme and constant mobility model to see if this also has any affect.

Lastly, varying the input frequency at the gate from 1 MHz, 10 MHz, 100 MHz, 1 GHz, and finally 10 GHz is another method for checking on device switching performance changes between frequencies. At higher frequencies, the holes may be too slow, i.e., when switching, the holes may not fully drift out or back into the device via the Ni-Electrode to respond with the changing potential and 2DEG electron concentration.
Chapter 6

Results and Discussion

Chapter 6 displays and discusses the simulation results found. Section 6.1 discusses the DC and 10 GHz simulation results. Section 6.2 presents data and discusses the effects of the low mobility holes on RF switching performance against what is optimally expected. It also provides a few potential theories on the origin of the hole accumulation in the i-GaN region. Section 6.3 concludes with a statement on the importance of an ohmic contact for the Ni-Electrode of the SHJ structure.

6.1 DC Characteristics and Initial RF 10 GHz Results

6.1.1 Simulation of High Voltage DC Characteristics

As described in Section 5.3, the DC characteristics are simulated first and include the $I_D - V_{DS}$ and $I_D - V_{GS}$ curves. The $I_D - V_{DS}$ curves are produced to choose a resistive load line and find the optimal DC operating drain voltage, gate voltage, and drain current. The $I_D - V_{GS}$ curve for the chosen DC operating drain voltage is simulated to find the DC operating gate voltage at the desired drain current. The AC 10 GHz transient characteristics are then run at the desired DC biases in a common source amplifier configuration.

Fig. 6-1 shows the $I_D - V_{DS}$ characteristics at various $V_{GS}$ values with a $V_{DD}$ sweep from 0 V to 500 V. The device is expected to operate with a negative threshold voltage due to the polarization-induced 2DEG and the metal/semiconductor gate structure. TCAD simulations do have issues with convergence at times, notably at simulation sweeps where large changes in current, electric fields, or charge carrier concentrations occur. Notice how convergence issues do occur at larger gate voltages ($V_{GS} > -1.0$ V) where the linear region begins to saturate. Another interesting effect that occurs at $V_{GS} > 0$ V is knee walkout,
which in physical HEMT devices, is typically caused by a combination of surface state charge trapping, 2DEG channel temperature increase, and material polarization changes. It also could be an effect of poor convergence at higher currents. This is an interesting outcome since polarization is included in the simulation, however, surface state physics is not, and temperature is set constant.

To avoid the poor convergence as illustrated in Fig. 6-1, the load line displayed in Fig. 6.2 is chosen. The knee voltage (V_{KNEE}) of 16 V is chosen by finding the intercept between the extrapolated triode region slope line and the lateral maximum simulated current (I_{MAX,SIM}) line, however, it could have been simply chosen arbitrarily. The maximum drain current (I_{D,MAX}) that occurs at V_{KNEE} is 0.520 A. The DC operating drain voltage (V_{DS,OP}) of 242 V with a V_{DD} = 500 V is found using Eqn. 2-1.

The V_{DS,OP} is chosen to achieve maximum AC voltage swing. The DC operating drain current (I_{D,OP}) of 0.260 A is chosen to be half I_{D,MAX} to ensure a full AC drain current swing from I_{D,MAX} to zero (at V_{DD}) occurs.
The load resistance \((R_L)\) at the drain of the common source amplifier configuration of 930.7 \(\Omega\) is calculated using Eqn. 2-3.

Finally, running an \(I_D - V_{GS}\) simulation with \(V_{DS} = V_{DS,OP}\) to acquire the \(V_{GS}\) value of -2.80 V at \(I_{D,OP}\) is completed. The plot can be found in Fig. 6.2.

![ID - VGS Plot at VDS = VDS,OP](image)

**Figure 6-2** \(I_D - V_{GS}\) Plot at \(V_{DS} = V_{DS,OP}\)

### 6.1.2 Super-Heterojunction and 2DEG Confirmation

Before discussing the RF simulation results, Fig. 6-3 displays the electron concentration and hole concentration contours at equilibrium (all biases equal zero). The electron concentration plot confirms the existence of the expected 2DEG formation in the i-GaN-side interface of the AlGaN/i-GaN junction. The hole accumulation is shown as a future reference point for the hole mobility discussion in Section 6.2.

Fig. 6-4 shows the electric field contours in the x-direction, y-direction, and total electric field. The biases are chosen arbitrarily, but with the goal to see the electric field of the device in an on-state. Notice how in the x-direction contours, much of the superheterojunction region to the left of the drain has a constant and relatively low electric field. A cutline is made through the U-GaN region across the entire device and the total electric
The electric field along the SJ region is uniform, and the significant peak electric field that would have been expected at the drain side of the gate is reduced significantly. There are now two smaller peaks in electric field. As described in Section 3.1.2, the electric field redistribution using a FP would not be as linear as illustrated in Fig. 6-5. This confirms the presence of the super-heterojunction theory in the device. This also confirms that the SJ is functionally better than the typical FP structure.

**Figure 6-3 DC Equilibrium Electron (Left) and Hole (Right) Concentration Contours (Log Scale, /cm$^3$)
Figure 6-4 Electric Field Contours (Log Scale, V/cm) with DC Applied $V_{DS} = 500$ V, $V_{GS} = -2.75$ V: X-Direction (Left), Y-Direction (Middle), Total (Right)

Figure 6-5 Electric Field (V/cm) Plot Along the Channel with DC Applied $V_{DS} = 500$ V, $V_{GS} = -2.75$ V
6.1.3 10 GHz RF Simulation Results

The AC Large-Signal 10 GHz simulation ran with $R_L = 930.7 \, \Omega$, $V_{DS,OP} = 242 \, V$, and $V_{GS} = -2.80 \, V$. The AC waveform voltage biases at the gate used simultaneously with the DC operating biases began at an amplitude of 0.75 V and increased in 0.25 V increments until 3.0 V. The transient waveforms for $I_D$, $V_{DS}$, and $V_{GS}$ are in Fig. 6-6. As expected, the $I_D$ and $V_{DS}$ waveforms are about 180 degrees out of phase. The $V_{GS}$ waveform is 90 degrees out of phase from the $I_D$ and $V_{DS}$ waveforms. The input power (dBm), output power (dBm), power gain (dB), and PAE (%) are plotted in Fig. 6-7.

![10 GHz RF Transient Characteristics](image)

**Figure 6-6 RF $V_{DS}$ (V), $I_D$ (A/mm), and $V_{GS}$ (V) Waveforms**

After observing Fig. 6.3, one can see that as the $V_{GS}$ amplitude increases, the $V_{DS}$ and $I_D$ amplitudes increase as well, until they become saturated at their minimum and maximum values. There are a few main takeaways from Fig. 6-6. The first is the positive shift in the $I_D$ and $V_{DS}$ waveforms. The waveforms are centered at values higher than the expected DC values of 0.260 A and 242 V, respectively. A second takeaway is that the minimum RF $I_D$ and $V_{DS}$ values are higher than expected, above zero amps and above the chosen knee voltage. This can be observed in Fig. 6-8 along with the specific values. Thirdly, at $V_{GS}$ RF amplitudes greater than 2 V, the minimum RF $I_D$ increases, and at the same time, the maximum RF $V_{DS}$ decreases. This is a result that correlates with the knee voltage characteristics shown in Fig. 6.1 because the maximum RF $V_{GS}$ values are now at
the point in Fig. 6.1 where convergence issues begin to appear (at the linear to saturation point of high gate bias curves) and the knee walk out is present. Overall, there is an RF swing compression at 10 GHz.

Fig. 6-7 shows a significant reduction in the expected PAE of 50%. The PAE observed reaches a maximum of about 25%, while $A_F$ is a reasonable ~32 dB and decreases to 21 dB as $P_{in}$ increases.

![10 GHz RF Simulation: Pout, AP, PAE vs Pin](image)

**Figure 6-7** 10 GHz: Pout (dBm), AP (dB), PAE (%) vs Pin (dBm)

6.2 Effects of Hole Mobility on HEMT Switching Performance

It is understood that the low-mobility holes in GaN/AlGaN/GaN SHJ devices may cause performance problems in high frequency switching applications. In the device structure of interest, illustrated in Fig. 5-1, the P-GaN region that creates the SJ provides holes to the system. After observing Fig. 6-8, one finds that the load line of the simulation is linear, and hence, is purely resistive when compared to the expected load line. There is no internal capacitance affecting the device’s switching performance. Additionally, the resistive load is exactly 930 Ω, the expected value.
Suppose the device is operating in the on-state (i.e. when the RF $I_D$ is at its maximum value). The 2DEG should be formed along the entire channel, as well as underneath the gate, and the P-GaN region should have an equally very large accumulation of holes due to its p-type doping. When the device is switched into the off-state (i.e. when the RF $I_D$ is at its minimum value), it is expected that the holes will drift out of the P-GaN region through the Ni-Electrode. At the same time, the electron concentration in the 2DEG channel will decrease equally. Due to the gate potential, there should be a minimal accumulation underneath the gate. When the device is then switched back to the on-state, the 2DEG will increase in sheet density and the 2DEG will form underneath the gate. At the same time, the holes will reenter the system and accumulate in the P-GaN region.

However, if the switching frequency is high enough and the hole mobility is sufficiently low enough, there is the possible outcome of a reduced reentry accumulation of holes when switching from the off-state to the on-state. In the same scenario, when switching from the on-state to the off-state, a decline in holes exiting the system occurs. This is the leading theory for the simulation in this work. After performing a variety of
simulation experiments, whose results which are discussed in this section, it has become clear that there is an impact on device performance due to low mobility holes in p-type doped GaN. It is also apparent that concurrently, there is an accumulation of holes in the i-GaN region with unknown origin.

6.2.1 Effects of Low Field Mobility at 10 GHz

For this simulation, the mobility model configuration is described in Section 5.2.2. It contains the following models: FMCT.n, GANSAT.n, GANSAT.p, and a MUP0 statement (for setting the low-field mobility of the holes through the GANSAT model). MUP0 is set to 10 cm²/Vs for the results shown in Fig. 6-5 and Fig. 6-6.

To find a direct correlation between the hole mobility and the poor PAE results, the MUP0 value is changed by multiple factors of ten: 0.1, 1, 10, and 100. To note, slightly different input powers are selected for this assessment, which explains the difference between Fig. 6-7 and Fig. 6-10. The slight change has no effect on the results. Fig. 6-9 displays the contour plots of the hole mobility, each at the largest 10 GHz RF V_{GS} waveform (3.0 V) in the off-state. The on-state showed very similar contours. Notice how the resulting high field mobility does not change very much with increasing MUP0 values. The maximum hole mobility in the center of the P-GaN region for all four test points is roughly 40.1 cm²/Vs. The MUP0 = 1000 simulation results show large mobilities up to 100 cm²/Vs deep in the i-GaN bulk. Fig 6-10 compares the RF P_{out}, A_{P}, and PAE values for the four MUP0 testing points. There is little to no difference.

Overall, two results come from this: One, that simply changing the low field mobility MUP0 for the given mobility model configuration does influence the resulting high field mobility. Two, changing the low field mobility with the given mobility model configuration has no effects on the RF switching performance metrics. Therefore, the conclusion can be made that setting low field mobility values is not a valid way to assess the performance of a switching mode device based on the mobility of its holes.
Fig 6-9 10 GHz Low Field Mobility Contours (Log Scale, cm$^2$/Vs) at Various MUP0 Values
6.2.2 Further Results with Large Constant Hole Mobility

As an extension of Section 6.2.1, to further test if directly changing mobility results in device RF performance improvements, the GANSAT.p mobility model statement is removed and MUP1.FMCT and MUP2.FMCT is used and set equal to 1000 cm²/Vs. Now, a constant mobility regardless of electric field, temperature, or dopant concentration is established. This final mobility configuration is simulated.

Fig. 6-11 shows the mobility contour plot and displays the RF performance metrics data. Further examination concludes that both PAE and $A_P$ are generally unchanged, and the conclusions of Section 6.2.1 holes here. Sections 6.2.1 and 6.2.2 illustrates how to not change the hole mobility in GaN HEMT TCAD simulation for RF testing. It is still unclear how the Silvaco algorithm exactly calculates the hole mobility in this model configuration. More specifically, further clarity on the order of precedence in calculation and execution
of each mobility model called is desired. All in all, poor PAE even with hole mobilities in the range of the electron mobility does raise concern in the HEMT’s ability to operate at 10 GHz.

Figure 6-11 Constant 1000 cm²/Vs Hole Mobility; Contour Plots (Left); $P_{\text{out}}$ (dBm), $A_P$ (dB), PAE (%) vs $P_{\text{in}}$ (dBm) (Right)

6.2.3 Frequency Assessment of Holes

Instead of changing the low field mobility, the frequency at which the device operates at can be changed. This is done by simply changing the frequency of the input RF waveform at the gate. To assess the switching performance and potential effects of slow holes, the RF input waveform amplitudes start at 0.75 V and end at 3.0 V, with 0.25 V increments. Each waveform is simulated for a single period, as before, and the operating frequencies vary by factors of 10, including: 1 MHz, 10 MHz, 100 MHz, 1 GHz, and 10 GHz. The Mobility model configuration is the same as Section 6.2.1 (FMCT.n, GANSAT.n, GANSAT.p, MUP0 = 10).

Fig. 6-13 displays the RF performance metrics between the five simulated frequencies. The color of the line distinguishes the RF performance metric plotted, and the
shape of the individual data points correlate to the simulated frequency. A few trends that can be extrapolated are: Firstly, as the frequency increases, $P_{in}$ increases, thereby correlating with the decrease in $A_P$. Secondly, PAE is quite similar between the 1, 10, and 100 MHz simulations, reaching the maximum 50% PAE expected by a Class A amplifier. The 1 GHz simulation exceeds the 50% expected maximum efficiency at higher input powers, reaching about 56%. This is due to $I_{D,MIN} < 0 \ A$, potentially caused by internal capacitive effects. The load line for the 1 GHz simulation displays negative current, therefore confirming an introduction of a negative reactance component to the load. The affect can be observed in Fig. 6-12, where the load lines for all five frequencies are plotted. They overlap with the same slope. Therefore, it can be stated they all have the same load resistance of 930 $\ \Omega$. The 1 MHz, 10 MHz, and 100 MHz load lines overlap perfectly with each other with end points at the calculated $V_{KNEE} = 16 \ V$ and a $V_{DS,MAX} = 500 \ V$. The corresponding $I_{D,MIN} \sim 0 \ V$ (nanoamp-range current). Continuing, at 10 GHz, the results for PAE, $A_P$, and $P_{out}$ all decrease significantly from the other four simulations, as previously seen in Section 6.1.3.

To dig deeper into the reasoning of the abrupt decline in performance, hole concentration contour plots for each simulation in the on-state and off-state can be observed in Fig. 6-14, with an RF amplitude of 2.50 $\ V$. This amplitude is chosen because it correlates with the presence of the knee walk out effect and the increase in minimum $I_D$ found in the 10 GHz transient waveform. In the on-state for all frequencies, the expectation is to observe a reentry of a significant amount of the holes that exited the Ni-Electrode during the previous off-state. While it is not realistic to expect all the holes to reenter the system at a frequency in the MHz or GHz range, charge balance between holes and electrons is required. For comparison, the total number of holes at equilibrium in the DC simulation can be observed in Fig. 6-3. In Fig. 6-14, the observation can be made that as the frequency increases, the number of holes reentering the P-GaN decreases, especially at 100 MHz, 1 GHz, and 10 GHz. Interestingly, as the number of holes in the P-GaN region decreases with frequency, there is an anomalous accumulation of holes in the i-GaN region that increases (albeit a relatively small concentration).
Furthermore, the off-state shows an increase in hole concentration with frequency with the 1 MHz, 10 MHz, and 100 MHz simulations. The 1 GHz and 10 GHz simulation display a decrease in hole concentration, and remarkably, contains an overall greater accumulation of holes than the on-state displays. Even more astonishing, it appears in the GHz-range of frequencies, there is a hole penetration effect from the i-GaN region through the 2DEG channel and into the P-GaN region in the off-state. The stream of holes moving vertically through the P-GaN region also occurs in the same location where there is a significant electric field, as seen in Fig. 6-4.

Fig 6-15 and 6-16 overlay the total current density as vectors with the hole concentration contours for the on-state and off-state 10 GHz simulation at the same RF amplitude of 2.50 V. Fig. 6-16 zooms into the channel where the gate and P-GaN meet the U-GaN layer. The off-state image on the right is quite interesting. It shows that the holes are responding to the off-state bias as expected, i.e., the holes are depleting the system through the Ni-Electrode at the top of the P-GaN region in the direction of the current density vectors. Confirmation of this is found in the right-side image (off-state) of Fig. 6-16. The on-state (left-side image) of Fig. 6-15 has no hole concentration of current density vectors penetrating the channel. Therefore, the conclusion can be made that the holes must have an origin that is not the P-GaN region, since holes do not appear to penetrate the channel and enter the i-GaN region from the P-GaN region in the on-state.

The decreases in hole accumulation with increasing frequency in the on-state of the device, and vice-versa in the off-state, at 10 GHz is conclusively due to the low mobility of the holes during reentry. Suppose that the 10 GHz device operation begins in the on-state and there are some holes present in the P-GaN to fulfill the charge neutrality requirement with electrons. When the device begins to turn off, holes will vacate the P-GaN region. In the off-state, the holes will have mostly left the system through the Ni-Electrode. When the biasing of the device once again reflects the on-state, the holes will begin to enter the P-GaN region again. However, if there is a combination of low mobility holes and/or a very high frequency, the holes may not exit or reenter the system in totality. Fig. 6-14 shows exactly this. Additionally, Fig. 6-13 reinforces this conclusion. The
increase in $P_{in}$ with frequency results in a decrease in $A_P$. At 10 GHz, $P_{out}$ and $A_P$ decrease with the increase in frequency, resulting in a decrease in PAE.

Nevertheless, the presence of holes in the i-GaN region is quite interesting. Their origin is unknown. One theory is that there is a generation of electron hole pairs in the i-GaN region due to the high electric fields. However, GaN has a relatively high bandgap of around 3.40 eV, making band to band generation statistically improbable. Fig. 6-17 plots the hole concentration, electron concentration, electric field, and recombination rate (from top left to bottom right, respectively). There is the presence of a recombination rate in the i-GaN region that correlates with the location of the hole concentration, along with the electric field. Unfortunately, the generation of electron hole pairs is not plottable, therefore, this theory is inconclusive as of now.

A second theory follows the line of the first theory. The impact ionization model is not included in any of the simulations. The reason for this is to improve convergence and simulation run-times. Another reason for the decision is because the study of interest is of the affects of low mobility holes in switching performance, not to evaluate the breakdown characteristics of the HEMT. The very large electric fields demonstrated in Fig. 6-17 maximize to $\sim 5.2 \times 10^6$ V/cm, which is well into the possible range for GaN critical breakdown field. Considering the i-GaN region is undoped, and critical breakdown field is proportional to doping concentration, it can be speculated that impact ionization breakdown events certainly would occur if the model were included. The Silvaco algorithm may expect generation or breakdown to occur in this situation regardless of if the model is included. Consequently, situational physics miscalculations or anomalies could be caused by from Silvaco’s self-consistent algorithm.

The possibility of simulation result anomalies from running an AC/RF simulation after a quasistatic DC simulation is highly possible. The quasistatic simulation is independent of time and may result in a fermi-level closer to the valence band in the left side of the i-GaN region, causing holes to accumulate when there is no physical origin for them to be arriving from. More work must be put into this.

A third theory is the possibility of physics-based miscalculations due to the lack of a doping statement for the i-GaN. The author’s personal experience recalls similar effects
in prior TCAD simulations. After comparing hole concentrations at 10 GHz in the on-state and off-state with and without i-GaN intentional doping in Fig. 6-18, it can be concluded that there is no effect. Similar plots can be made with the recombination rate and electric fields, resulting in near identical contours.

There also may be problems with the simulation mesh. Re-scaling the device to improve convergence and finding a more optimal mesh may be a good idea. Interesting physics can occur from poorly developed meshes.

A fifth theory is the possibility of hole injection from the ohmic source contact. Fig. 6-19 shows the band diagram of the Source contact with the i-GaN region. The barrier for holes to enter the i-GaN region is roughly equal to the bandgap energy of the i-GaN, making injection unlikely.

![Load Lines Plotted from RF V_DS and I_D Simulated Transient Values for 1 MHz, 10 MHz, 100 MHz, 1 GHz, and 10 GHz](image)

**Figure 6-12** Load Lines Plotted from RF V_DS and I_D Simulated Transient Values for 1 MHz, 10 MHz, 100 MHz, 1 GHz, and 10 GHz
Figure 6-13 $P_{\text{out}}$ (dBm), $A_{\text{p}}$ (dB), PAE (%) vs $P_{\text{in}}$ (dBm) for 1 MHz, 10 MHz, 100 MHz, 1 GHz, and 10 GHz
Figure 6-14 On-State Hole Concentration Contours (Log Scale, /cm$^3$) for Various Frequencies with RF Input Amplitude 2.50 V
Figure 6-15 P-GaN: 10 GHz Hole Concentration Contours (Log Scale, /cm$^3$) and Current Density Vectors (Log Scale, A/cm$^2$) Overlay with RF Amplitude 2.50 V: On-State (Left), Off-State (Right)
Figure 6-16 Channel: 10 GHz Hole Concentration Contours (Log Scale, /cm$^3$) and Current Density Vectors (Log Scale, A/cm$^2$) Overlay with RF Input Amplitude 2.50 V: On-State (Left), Off-State (Right)
Figure 6-17 10 GHz Off-State Contour Plots (Log Scale) with RF Input Amplitude 2.50 V; Hole Concentration (Top Left, /cm³), Electron Concentration (Top Right, /cm³), Electric Field (Bottom Left, V/cm), and Recombination Rate (Bottom Right, /cm³s)
Figure 6-18 10 GHz On- and Off-State Hole Concentration Contour Plots (Log Scale, /cm³) with RF Input Amplitude 2.50 V; Without i-GaN Doping Statement (Top), With i-GaN Doping Statement (Bottom)
When beginning the simulation, an original work function of 5.2 eV was used for the Ni-Electrode with the thought that using a highly doped P$^{++}$-GaN layer would assist in making the contact ohmic. The contact became a Schottky contact instead, therefore the holes are blocked from reentering the device originally. Removing the work function statement in the contact statement in the Silvaco code makes the electrode ideally ohmic. Fig. 6-20 shows the band diagrams for the Schottky Ni-Electrode and ohmic Ni-Electrode. Fig. 6-21 shows the hole concentration at difference between the Schottky contact and ohmic contact. The 1 MHz simulation is used to illustrate the serious impact the Ni-Electrode contact can have on the hole reentry from the off- to the on-state. Overall, ensuring that the Ni-Electrode is as ohmic as possible is important for optimal switching characteristics.
Figure 6-20 Ni-Electrode Simulated Band Diagram (eV): Schottky (Left), Ohmic (Right)

Figure 6-21 On-State 1 MHz Ni-Electrode Effect on Hole Concentration Contours (Log Scale, /cm$^3$) with RF Input Amplitude 2.50 V: Schottky (Left), Ohmic (Right)
Chapter 7
Conclusions and Future Work

The GaN/AlGaN/GaN SHJ HEMT presented in this thesis proves to be an excellent high power switching device at frequencies simulated up to 1 GHz, with PAE up to 50% from 1 to 100 MHz, and 56% at 1 GHz. At 10 GHz, PAE decreases to 25%. The polarization induced 2DEG channel due to the lower AlGaN/GaN heterojunction allows for a high DC saturated current density output simulated up to 0.8 A at a drain bias of 500 V. Substitution of a typical FP design for a P-GaN layer and delta-doping above the 2DEG channel creates a SHJ structure. Because of optimal charge balance as required in SJ theory, perfect uniform redistribution of the electric field, superior to the FP structure, with the objective to reduce peak electric fields at the drain-side of the gate is observed. The conclusion enforcing the necessity of an ohmic contact for the Ni-Electrode is produced. Optimal switching can be limited by the reentry of holes into the P-GaN region through the Ni-Electrode.

After studying the low mobility holes within the device, it can be concluded that they do negatively impact the switching performance of the HEMT, causing PAE and $A_P$ to decrease with increasing frequency. This is due to the limited movement of the holes when exiting and reentering the device through the Ni-Electrode. Concurrently, as frequency increases, an anomalous accumulation of holes increases in the i-GaN region. The cause of the holes is unknowns, however, the leading theory is they may be a physics-based miscalculation or simulation anomaly because of the lack of an impact ionization physics model.

There is a significant amount of possible future research work following this thesis. The beginning of the Author’s Ph.D. degree will consist of creating a detailed fabrication process and layout for the HEMT and successfully fabricating it. Possible fabrication roadblocks must be researched, such as how to dope GaN highly p-type for the ohmic Ni-Electrode, which may be difficult. A detailed comparison between the simulated HEMT
and the physical HEMT RF performance will be produced. Regarding the simulation, the first, and obvious, next step is to locate the source of the anomalous i-GaN hole accumulation. The study would begin with a re-simulation of the device with the addition of the impact ionization model. Other changes can be done, such as rescaling and remeshing the device to improve convergence and possibly reduce abnormalities in the simulation. Further research into the effects of holes in the device can be completed. A transient simulation would simplify the hole movement analysis and give further conclusive confirmation that the i-GaN hole accumulation is simply a quasistatic simulation anomaly. When there is a full understanding of the hole movement in the device, an analytical equation can be developed to model the movement and/or time it takes for a hole to exit the P-GaN region when switching into the off-state. Lastly, investigating a switching-mode Class-E configuration for the HEMT is an interesting future route to improve PAE and minimize power loss during switching.
Appendix A

Charge Balance Procedure

The charge balance for the device to fulfill the SJ theory required an equal number of P-GaN dopants and n-type delta dopants at the U-GaN/Al_{0.25}Ga_{0.75}N interface. The SHJ region exists where there is a P-GaN/U-GaN interface, or 10 microns in length along the channel. This is the region where charge balance must occur, and follows Eqn. A-1, which can be represented in terms of dopants and cross-sectional area of the regions:

\[ N_A A_p = N_D A_n \]  \[[A-1]\]

where \( N_A \) and \( N_D \) are the p-type dopant charge in the P-GaN region and the n-type delta dopant charge of the interface delta-doping, respectively, and \( A_p \) and \( A_n \) are the P-GaN area and simulation interface delta doping area, respectively. The area are defined as the product of the lengths and thickness of the regions, i.e. \( A_p = L_p t_p \) and \( A_n = L_n t_n \), where \( L_p, t_p, L_n, \) and \( t_n \) are the P-GaN length and thickness and delta-doping length and thickness, respectively.

Since the P-GaN region and delta-doping are of the same length at the SHJ, Eqn. A-1 can be reduced to Eqn. A-2:

\[ N_A t_p = N_D t_n \]  \[[A-2]\]

The interfacial delta doping is chosen to be \( 7.5 \times 10^{19} \) /cm\(^3\) for the simulation following the work in [2]. The region is also defined to be 1 nm thick, making the sheet dopant density of the interfacial delta doping equal to \( 7.5 \times 10^{12} \) /cm\(^2\). The thickness of the P-GaN region in 200 nm. Therefore, one can calculate \( N_A \) to be \( 3.75 \times 10^{17} \) /cm\(^3\) in the P-GaN region.
Appendix B

Silvaco TCAD Code

Simulation Code for all DC and AC Simulations:

```
go atlas
#******************************************************************************
# Title: AC Large Signal RF Simulation GaN/AlGaN/GaN Power HEMT w/
# doping balance.
# Date(Last): 03/07/2021
# Designer: Jesse Kemmerling
# Description: This code will simulate the Id-Vd, Id-Vg, and RF-AC char
# of a GaN/AlGaN/GaN power mosfet and produce Structure files
# for simulated biased HEMT
#******************************************************************************
#******************************************************************************
# Parameters
# Variables instantiated and defines
#******************************************************************************

# Geometric Parameters
### Thickness
set TDelta=0.001
set TPGaN=0.2
set TPPGaN=0.01
set TUGaN=0.01
set TALGaN=0.01
set TBulk=8.25
set TNielectrode=0.01
set TD=$TAlGaN+$TDelta+$TUGaN+$TPPGaN+$TPPPGaN+$TNielectrode #can be anything, really
set TS=$TD #can be anything, really
set TG=$TPPGaN+$TPPPGaN+$TNielectrode #can be anything, really
set TS1=$TG
set TS2=$TG
set TSphi=$TPPGaN
set TSphi2=$TPPGaN
set TSphi3=$TNielectrode+$TPPPGaN
set TTotal=$TBulk+$TD
### Length
set LNiElectrode=1.0
set LPPGaN=$LNielectrode
set LPGaN=10*$LNielectrode
set Lg=0.5
set LD=2.0
set LS=$LD
```
set Lsp1=0.5
set Lsp2=ILsp1
set Lsp3=ILsp1
set Lsp4=ILsp1
set Lsp5=ILsp4+(LPGaN+LNiElectrode)
set Lch=ILsp1+ILsp2+ILG+LPGaN+ILsp4 # (dependant on LGaN)
set Ldelta=Lch
set Lbulk=ILsp1+ILC+ILD # (dependant on LGaN)
set Ltotal=Lbulk
# Al mole fraction of AlGaN (25%)
set xcomp = 0.25
# Mesh Parameters
set Sdiff = 0.00001
set SBulk = 0.1

#******************************************************************************
# Define Mesh
# Creating the x and y-axis mesh
#******************************************************************************

# Mesh width (1 mm)
set width = 1000
mesh width = $width

# X-axis mesh definition
x.m l=0 s=0.11
x.m l=ILs s=0.1
x.m l=ILsp1 s=0.1
x.m l=ILsp1+ILG+ILsp2 s=0.1
x.m l=ILsp1+ILG+ILsp2+ILPGaN s=0.1
x.m l=ILsp1+ILG+ILsp2+ILPGaN s=0.1
x.m l=ILsp1+ILG+ILsp2+ILPGaN+ILsp4 s=0.1
x.m l=ILsp1+ILG+ILsp2+ILPGaN+ILsp4+ILD s=0.1

# Y-axis mesh definition (Bulk GaN Buffer)
y.m l=0 s=0.05
y.m l=(-0.9*SBulk) s=0.05
y.m l=-$SBulk s=0.01
#Y-axis AlGaN Mesh Y-region

#Y-axis Delta Doping Layer Mesh Y-region
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(0.0$*$T$Delta$) s=0.0005
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(1.0$*$T$Delta$) s=0.0005

#Y-axis Undoped GaN Mesh Y-Region

#Y-axis P-GaN Mesh
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(1.0$*$T$Delta$)-(1.0$*$TUGaN$)-(0.0$*$T$GaN$) s=0.001

#Y-axis top P++-GaN Mesh
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(1.0$*$T$Delta$)-(1.0$*$TUGaN$)-(1.0$*$T$GaN$)-(0.0$*$T$PP$GaN$) s=0.001

#Y-axis Ni-Electrode Mesh
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(1.0$*$T$Delta$)-(1.0$*$TUGaN$)-(1.0$*$T$GaN$)-(1.0$*$T$PP$GaN$)-(0.0$*$T$m$E$$l$$e$$c$$t$$r$$o$$d$) s=0.005
y.m.=$^n$T(hl$=1.0$*$TAlGaN$)-(1.0$*$T$Delta$)-(1.0$*$TUGaN$)-(1.0$*$T$GaN$)-(1.0$*$T$PP$GaN$)-(1.0$*$T$m$E$$l$$e$$c$$t$$r$$o$$d$) s=0.005

******************************************************************************

# Define Region
# Creating x/y geometric region of bulk GaN, top GaN, AlGaN,
# (Include delta doping thickness TDelta in other files)
******************************************************************************

###Semiconductor Regions

#Bulk GaN Region
region num=1 mat=GaN asub=3.189 \ 
  x.min=0 x.max=$^n$L$B$$u$$l$$k$ \ 
  y.min=0 y.max=$^n$T$B$$u$$l$$k$ substrate

#Dielectric Region 1
region num=6 mat=nitride \ 
  x.min=0 x.max=$^n$L$T$$o$$t$$a$$l$ \ 
  y.min=$^n$T$B$$u$$l$$k$ y.max=$^n$T$B$$u$$l$$k$ $^n$TAlGaN $^n$T$D$$e$$l$$t$$a$ $^n$TUGaN $^n$T$Sp$1

#AlGaN Region
region num=2 mat=AlGaN \


x.comp=xcom x.min=0 x.max=$LTotal \\
y.max=-$TBulk y.min=-$TBulk-$STA$Gan

#GaN Delta Doping Region
region num=7 mat=Gan asub=3.189 \\
x.min=0 x.max=$LTotal \\
y.max=-$TBulk-$STA$Gan y.min=-$TBulk-$STA$Gan-$STDelta

#Undoped-GaN Region
region num=3 mat=Gan asub=3.189 \\
x.min=0 x.max=$LTotal \\
y.max=-$TBulk-$STA$Gan-$STDelta y.min=-$TBulk-$STA$Gan-$STDelta-$TUGan

#P-GaN Region
region num=4 name=pgan mat=Gan asub=3.189 \\
x.min=$LSL-$LS$pl1+$LSL+$LS$Pl2 x.max=$LSL+$LS$pl1+$LSL+$LS$Pl2+$LPGan \\
y.max=-$TBulk-$STA$Gan-$STDelta $TUGan y.min=-$TBulk-$STA$Gan-$STDelta-$TUGan-$STPGan

#P++-GaN Region
region num=5 mat=Gan asub=3.189 \\
x.min=$LSL-$LS$pl1+$LSL+$LS$Pl2 x.max=$LSL+$LS$pl1+$LSL+$LS$Pl2+$LPPGan \\
y.max=-$TBulk-$STA$Gan-$STDelta $TUGan-$STPGan y.min=-$TBulk-$STA$Gan-$STDelta-$TUGan-$STPGan

*****************************************************************************
# Define Electrodes
#  Creating x/y geometric region of Source, Gate,
#  and Drain electrodes
#  (Include delta doping thickness TDelta in other files)
*****************************************************************************

#Source Definition
electrode name=source \\
x.min=0 x.max=$LSL \\
y.max=-$TBulk y.min=-$TBulk-$TS

#Drain Definition
electrode name=drain \\
x.min=$LSL+$LS$Ch x.max=$LTotal \\
y.max=-$TBulk y.min=-$TBulk-$TD

#Gate Definition
electrode name=gate \\
x.min=$LSL-$LS$pl1 x.max=$LSL+$LS$pl1+$LSL \\
y.max=-$TBulk-$STA$Gan-$STDelta $TUGan y.min=-$TBulk-$STA$Gan-$STDelta-$TUGan-$STG
Nickel Electrode Definition

```plaintext
electrode name=nickleelectrode \
  x.min=-$L5+$LSp1+$LG-$LSp2 x.max=-$L5+$LSp1+$LG-$LSp2+$LNielectrode \
  y.max=-$TBulk-$STAlGaN-$TDelta-$TUGaN-$TPGaN-$STPPGaN y.min=-$TBulk-$STAlGaN-$TDelta-$TUGaN-$STPPGaN-$STNielectrode
```

# Define Doping
# Defining the dopant concentrations

```plaintext
#n-Delta Doping interface/layer
doing uniform n.type conc=7.5e19 \
  x.min=0 x.max=$LTotal \
  y.max=-$TBulk-$STAlGaN y.min=-$TBulk-$STAlGaN-$TDelta
```

#P GaN Layer Doping
```plaintext
doing uniform p.type conc=3.75e17 \
  x.min=-$L5+$LSp1+$LG-$LSp2 x.max=-$L5+$LSp1+$LG-$LSp2-$LPGaN \
  y.max=-$TBulk-$STAlGaN-$TDelta-$TUGaN y.min=-$TBulk-$STAlGaN-$TDelta-$TUGaN-$TPGaN
```

#P++ GaN Layer Doping
```plaintext
doing uniform p.type conc=1e19 \
  x.min=-$L5+$LSp1+$LG-$LSp2 x.max=-$L5+$LSp1+$LG-$LSp2-$LPGaN \
  y.max=-$TBulk-$STAlGaN-$TDelta-$TUGaN-$TPGaN y.min=-$TBulk-$STAlGaN-$TDelta-$TUGaN-$TPGaN
```

# Define Model and Specific Material
# Specifying the GaN and AlGaN materials, along with their parameters

```plaintext
#define GaN material mat=GaN edb=0.015 eab=8.17 tau0=1e-10 tauup0=6.5e-9 copt=1.1e-8 nsrhn=4e18 nsrhp=4e18
#define AlGaN material mat=AlGaN edb=0.015 eab=0.17 tau0=1e-10 tauup0=6.5e-9 copt=1.1e-8 nsrhn=4e18 nsrhp=4e18
```

#Models
```plaintext
models constrh auger fermi incomplete print temp=300 optr model POLARIZATION polar.scale=0.8 calc.strain
```
AC Transient Simulation Code:

```plaintext
set Vd_DC = 242
system rm startVd1_smallamp_AC_LargeSignal.dat
solve init
save out
=startVd1_smallamp_AC_LargeSignal_PGaN_and_Delta_Doping_Equilibrium_GaN-AlGaN-GaN_HEMT_F.st
solve previous
solve name-drain vdrain=0 vfinal=500 vstep=5
solve name-gate vgale=5 vfinal=-2.88 vstep=0.1
save out=a_startVd1_smallamp_AC_LargeSignal_PGaN_and_Delta_Doping_GaN-AlGaN-GaN_HEMT_F.st
```

AC Transient Simulation Code:

```plaintext
# Define Mobilities
# Defines n and p-mobilities for the layers

mobility FMC.T n GaNsT.N GaNsT.P MU08-10

# Define Workfunction
# Creating contact work functions for Source, Drain, and Gate

#Source ohmic, drain(ohmic), gate(schottky), ninkle(ohmic) Workfunction
contact name-source workfunction=4.04
contact name-drain resistance=930.7+$\text{width}$ workfunction=4.04
contact name-gate workfunction=5.2
contact name-ninkle electrode
contact name-source common=nickle electrode

# Solution method, impact ionization, and output

set max_dt = 5e-12
method cklimit=1e-4 ir.tol=1e-22 tol.time=1e9 dt.max=$\text{max dt}$ itlimit=100 maxtraps=10
output con.band val.band charge polar.charge band.par e.mobility h.mobility qss

# Waveform Creation. 10 to loop through

waveform amplitude=0.75 elec.name-gate frequency=1e16 number=1 periods=10 sinusoid
waveform amplitude=1.00 elec.name-gate frequency=1e16 number=2 periods=10 sinusoid
waveform amplitude=1.25 elec.name-gate frequency=1e16 number=3 periods=10 sinusoid
waveform amplitude=1.50 elec.name-gate frequency=1e16 number=4 periods=10 sinusoid
waveform amplitude=1.75 elec.name-gate frequency=1e16 number=5 periods=10 sinusoid
waveform amplitude=2.00 elec.name-gate frequency=1e16 number=6 periods=10 sinusoid
waveform amplitude=2.25 elec.name-gate frequency=1e16 number=7 periods=10 sinusoid
waveform amplitude=2.50 elec.name-gate frequency=1e16 number=8 periods=10 sinusoid
waveform amplitude=2.75 elec.name-gate frequency=1e16 number=9 periods=10 sinusoid
waveform amplitude=3.00 elec.name-gate frequency=1e16 number=10 periods=10 sinusoid
```

AC Transient Simulation Code:
# Set up top of AC_LargeSignal.dat file
system echo "Large Signal Power Curves" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "10 4 4" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power In (Watts)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power In (dBm)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power Out (Watts)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power Out (dBm)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power Gain (dB)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power Gain (dBm)" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Power Output Efficiency (%)" >> startVd1_smallamp_AC_LargeSignal.dat

system echo "max_Vg" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "min_Vg" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "max_Vd" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "min_Vd" >> startVd1_smallamp_AC_LargeSignal.dat

system echo "max_Ig" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "min_Ig" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "max_Id" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "min_Id" >> startVd1_smallamp_AC_LargeSignal.dat

system echo "Vmax_t" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Vmin_t" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Vmmax_t" >> startVd1_smallamp_AC_LargeSignal.dat
system echo "Vmmin_t" >> startVd1_smallamp_AC_LargeSignal.dat

# Loop through 30 waveforms previously created above.
# Each waveform will be simulated one at a time, top to bottom
# Log voltages, currents, times, and powers
set wave=0
log outfile=startVd1_smallamp_AC_LargeSignal.log
loop steps=10
set wave=(1+\$wave)
# transient simulation of wave input into device. Data automatically placed into .log file
set tempwave=([wave]-1)/0.5*1e-10
solve wave=solve tfinal=`tempwave` dt=smax_dt
save outfile=wave_${wave}.maxVd.str
solve wave=solve tfinal=`(1-20)*solve` dt=smax_dt
save outfile=wave_${wave}.minVd.str

# Extract init ifile="startVd1_smallamp_AC_LargeSignal.log"
# Extract max and min drain and gate voltages
extract name="max_Vg" max(curve(time, v."gate"))
extract name="min_Vg" min(curve(time, v."gate"))
extract name="max_Vd" max(curve(time, v.int."drain"))
extract name="min_Vd" min(curve(time, v.int."drain"))

# Extract Transient Times (times when min and max drain and gate voltages occurred)
extract name="Vmax_t" x.val from curve(time, v."gate") where y.val=$\"max_Vg\"
extract name="Vmin_t" x.val from curve(time, v."gate") where y.val=$\"min_Vg\"
extract name="Vmmax_t" x.val from curve(time, v.int."drain") where y.val=$\"max_Vd\"
extract name="Vmmin_t" x.val from curve(time, v.int."drain") where y.val=$\"min_Vd\"

# Extract Currents at the same time of the min and max drain and gate voltages
extract name="max_Ig" y.val from curve(time, i."gate") where x.val=$\"Vmax_t\"
extract name="min_Ig" y.val from curve(time, i."gate") where x.val=$\"Vmin_t\"
extract name="max_Id" y.val from curve(time, i."drain") where x.val=$\"Vmmax_t\"
extract name="min_Id" y.val from curve(time, i."drain") where x.val=$\"Vmmin_t\"
# Extract Powers: Input, output, Gain, supply, and efficiency from previously extracted currents and voltages

# W
extract name="Pin" abs((($"max_Vg")-($"min_Vg"))*($"Vmax_Id")-($"Vmin_Id"))/8
extract name="Pout" abs((($"max_Vd")-($"min_Vd"))*($"Vmin_Id")-($"Vmax_Id"))/8
# dBm
extract name="Pin_dBm" 10*log10(($"Pin")/0.001)
extract name="Pout_dBm" 10*log10(($"Pout")/0.001)
# dBi
extract name="Power_Gain_dB" 10*log10(($"Pout")/($"Pin")
# dBm
extract name="Power_Gain_dBm" 10*log10(($"Pout")/($"Pin")/0.001)
# W
extract name="Power_Supply" $"Vd_DC"*abs(((($"Vmax_Id")+($"Vmin_Id"))/2)
# W/N
extract name="Efficiency" 100*(($"Pout"-"Pin")/($"Power_Supply")

# Unix system command to input four power values into .dat file in proper order
system echo "$Pin" "$Pin_dBm" "$Pout" "$Pout_dBm" "$Power_Gain_dBm" "$Power_Gain_dBm" "$Power_Gain_dBm" "$Power_Gain_dBm" "$Efficiency" "$max_Vg" "$min_Vg" "$max_Vd" "$min_Vd" "$Vmax_Id" "$Vmin_Id" "$Vmax_t" "$Vmin_t" "$Vd_max_t" "$Vd_min_t" >> startVd_smallamp_AC_LargeSignal.dat

l.end

# No more calculations

# Closing Statements
# Necessary for all programs

quit
ID – V_{DS} Example Plot Codes:

```plaintext
#Vg=-5, Vd=0...500
solve init
solve voltage=-5.0
log out=fix_Vg-5_Id-Vd_F.log
solve name=d电流 vcurrent=0.0 vfinal=150.0 vstep=2.50
solve name=d电流 vcurrent=160.0 vfinal=500.0 vstep=10.0
save out=Vg_-5_Id-Vd_F.str
log off
```

ID – V_{GS} Example Plot Code:

```plaintext
solve init
solve name=d电流 vcurrent=0 vfinal=240 vstep=5
solve name=d电流 vcurrent=240 vfinal=242 vstep=1
log out=step2vd242Vdd500_Id-Vg_F.log
solve name=gate voltage=-5.0 vfinal=-3.2 vstep=0.15
solve name=gate vcurrent=-3.1 vfinal=1 vstep=0.1
save out=step2vd242Vdd500S_Id-Vg_F.str
log off
```
Appendix C

Simulation Process Philosophy

When creating the simulation .in file, the goal is to create the device and simulate step-by-step to ensure the simulation is running correctly. First, the regions and dimensions of the device, and have it exported in a .str structure file. The physics and mobility models are then implemented. The device in Fig. 5-1 is designed without any material doping and with a “P”-GaN region only one micron long, making the SHJ structure that length. Doping is included in incremental steps for each region, starting with the delta-doping and finishing with the P-GaN and P++-GaN regions. Exported plots of the $I_D - V_{DS}$ and $I_D - V_{GS}$ characteristics from .log files are performed at each step to ensure the simulation is working as expected. An analysis of the .str files is performed at each stage, as well, to ensure that the dopant concentrations made realistic sense that the 2DEG is present.

The charge balancing procedure is then performed after realizing the P-GaN doping should be balanced with the delta-doping. This procedure is explained in Appendix A. Other parameters are edited at this stage, such as the electrode work functions and setting the width of the device.

At this point, the DC characteristics appeared promising for varying drain voltages between 10 V and 500 V, so the AC large signal simulation process could begin. After researching, the AC large signal components of the simulation are implemented, however, the output power after simulation execution is zero. This is because an external load resistance had to be included at the drain electrode. Computing the Class A amplifier load-line, the load resistance and operating voltages and currents are found and the simulations are executed. The DC and AC simulations are re-simulated multiple times as mobility models, material parameters, and physics models are edited to make the device operate as realistically as possible and to examine the effects of the low hole mobilities.

The .in files code discussed in this section can be seen in Appendix B.
BIBLIOGRAPHY


