AlGaN/GaN HEMT TOPOLOGY INVESTIGATION
USING MEASURED DATA AND DEVICE MODELING

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of the requirements for the degree of
Master of Science in Engineering

By

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ABSTRACT


Investigation has been done on procedure, development, and verification of transistor topology for Aluminum-Gallium Nitride/Gallium Nitride (AlGaN/GaN) High-Electron Mobility Transistor (HEMTs). To date various models have been published that address modeling issues dealing with AlGaN/GaN HEMTs. Many models rely on analytic parameter solutions to help define model behavior. In order to find an optimum transistor at X-band frequency, 8-12 Gigahertz (GHz), layout, testing, modeling and simulation was conducted on various transistor topologies. Theoretical and experimental analysis has been completed for device operating point selection in measurement and modeling to account for self-heating and radio frequency (RF) dispersion effects. The model extraction allows for observations in parameter extraction changes that occur from growth imperfections to heating effects on the different topologies. The model extraction techniques allow for precise parameter extraction, resulting in predicted direct current (DC), scattering parameter (s-parameter), and large signal power performance verification at X-band.
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I dedicate this thesis work to my loving, and understanding motivator

My wife, Cheri
1. Introduction

The development and integration of modern communication and sensor systems, within commercial and military applications, are continuously increasing capacity requirements. These requirements drive the need for improvements in the electronic circuitry’s radio frequency (RF) front-end performance. This holds true for microwave power amplifiers where bandwidth, power added efficiency (PAE), and linearity are essential. As device processing continues to mature for wideband gap materials, aluminum gallium nitride/gallium nitride (AlGaN/GaN) high-electron-mobility transistors (HEMTs) are slowly becoming the transistor of choice for power applications.

Device modeling improvements for AlGaN/GaN HEMTs have helped with optimization of these transistors. Using a bottom-up approach to modeling the HEMTs has helped to understand performance, reliability and overall device behavior. Modeling allows researchers to look at how changes in material growth, metal contact and passivation affect the performance of HEMTs.

Choosing various devices and modeling techniques, my research involves narrowing the devices down to find an optimum device at X-Band operation. Analysis of the measured data along with constructing modeling simulations will direct my research to find the optimum device. As an outcome of this research, I will demonstrate a reliable AlGaN/GaN HEMT device modeling procedure for production of high power and improved efficiency device characterization through class AB operation.
2. AlGaN/GaN HEMT Device

This chapter describes the fundamentals of the physics and operation of a AlGaN/GaN HEMT. The basics of the material system, the technology, the structure, and the performance of AlGaN/GaN HEMT are described. The information presented is very important for accurate device modeling.

2.1 Basic HEMT Operation

The high electron mobility transistor (HEMT) is a heterostructure field effect transistor. The structure of the device takes advantage of the superior transport properties of electrons in a potential well of lightly doped semiconductor material. The following diagrams, Figures 2.1 and 2.2, show a simplified AlGaN/GaN HEMT structure.

![Figure 2.1 AlGaN/GaN HEMT structure](image)
Figure 2.2. Band diagram for a Schottky contact made to a heterostructure typical of fabricated HEMT device

The figures show a wide bandgap semiconductor material (AlGaN) lying on a narrow band material (undoped GaN). The band diagram shown in Figure 2.2 demonstrates the sharp dip in the conduction band edge at the AlGaN/GaN interface. A high carrier concentration results at the dip within a narrow region (quantum well) in the source-drain direction. The distribution of electronic states in the quantum well is continuous in the two dimensions parallel to the interface and discrete perpendicular to the interface due to the very small thickness of the quantum well in comparison to the width and length of the channel. The charge density is defined as a two dimensional electron gas (2DEG) and quantified in terms of sheet carrier density $n_s$. An observation about the 2DEG for the AlGaN/GaN interface is how it can form even when there is no intentional doping of the AlGaN layer.
In AlGaN/GaN HEMT, the formation of 2DEG at the heterointerface is different from other HEMTs like AlGaAs/GaAs. Due to the presence of a strong polarization field across the AlGaN/GaN heterojunction, a 2DEG with the sheet carrier density up to $10^{13}$ cm$^{-2}$ can be achieved without intentional doping. [1] Ibbetson et al. found that surface states act as a source of electrons in 2DEG. [2] The built-in static electric field in the AlGaN layer induced by spontaneous and piezoelectric polarization greatly alters the band diagram and electron distribution of the AlGaN/GaN heterostructure. The electrons in the surface states transfer to the AlGaN/GaN heterointerface, this leads to a 2DEG with high density as observed in the band diagram.

AlGaN/GaN heterostructure have three important characteristics 1) high sheet carrier density ($n_s \approx 1 \times 10^{13}$ cm$^2$), which produces high $I_{\text{max}}$; 2) high electron mobility ($\mu = 1200 - 1500$ cm$^2$/Vs), which is responsible for low on-resistance (low knee voltage) since the channel resistance is related to $1/(q n_s \mu E)$ at low electric field; 3) high breakdown voltage; 4) high current density; and 5) high channel operating temperature. [1] These characteristics help to achieve high operating frequency ($f_T$) and high drain power added efficiency (PAE).

Figures of merit are used to describe the impact of material characteristics on the performance for AlGaN/GaN HEMT and other semiconductor devices. The Johnson figure of merit, defines the power-frequency product of the device. Another figure of merit is Baliga figure of merit that defines material parameters to minimize the
conduction loss in the device. The material properties of GaN compared to the competing materials is presented in Table 2.1. From the table, it is evident that GaN has greater advantage over conventional semiconductors.

<table>
<thead>
<tr>
<th>Material</th>
<th>$\mu$</th>
<th>$\varepsilon$</th>
<th>$E_g$ (eV)</th>
<th>$v_s$ (cm/s)</th>
<th>$E_{br}$ (V/cm)</th>
<th>JFOM</th>
<th>BFOM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Si</td>
<td>1500</td>
<td>11.9</td>
<td>1.12</td>
<td>1x10$^7$</td>
<td>0.3x10$^6$</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>GaAs</td>
<td>8500</td>
<td>13.1</td>
<td>1.43</td>
<td>1x10$^7$</td>
<td>0.4x10$^6$</td>
<td>1.8</td>
<td>14.8</td>
</tr>
<tr>
<td>GaN</td>
<td>1250</td>
<td>9.0</td>
<td>3.45</td>
<td>2.2x10$^7$</td>
<td>2x10$^6$</td>
<td>215.1</td>
<td>186.7</td>
</tr>
</tbody>
</table>

Table 2.1 Properties of Si, GaAs, and GaN

2.2 Basic HEMT Structure

In [1], Anwar Hasan Jarndal explains the relation between the polarization effects and the formation of the 2DEG channel. He explains how the AlGaN/GaN HEMT exhibits large polarization effects. The effects originates in the high polarity of the GaN material and the large lattice constant difference between GaN and AlGaN. Research has been conducted on the polarization effect and formation of the 2DEG channel. [1]

2.2.1 Polarization Effects in AlGaN/GaN HEMT

AlGaN/GaN HEMT polarization effects are both spontaneous and piezoelectric. The spontaneous polarization refers to the built-in polarization field present in an unstrained crystal. The electric field exists because the crystal lacks inversion symmetry.
and the bond between the two atoms is not purely covalent. This results in a displacement of the electron charge cloud towards one atom in the bond. In the direction along which the crystal lacks inversion symmetry, the asymmetric electron cloud results in a net positive charge located at one face of the crystal and a net negative charge at the other face. The electric field and sheet charges present in a Ga-face crystal of GaN and AlGaN grown on c-plane is illustrated in Figure 2.3. [1]

![Diagram showing electric field and sheet charges](image)

**Figure 2.3 Electric field and sheet charges present (a) due to only spontaneous polarization in GaN and AlGaN crystals; and (b) due to only piezoelectric polarization in a AlGaN layer.**

The piezoelectric polarization is the presence of a polarization field resulting from the distortion of the crystal lattice. Due to the large difference in lattice constant between AlGaN and GaN materials, the AlGaN layer which is grown on the GaN buffer layer is strained. Due to the large value of the piezoelectric coefficients of these materials, this strain results in a sheet charge at the two faces of AlGaN layer as illustrated in the previous diagram. The total polarization field in the AlGaN layer depends on the orientation of the GaN crystal. Epitaxial layers grown with Metal Organic Chemical Vapor Deposition (MOCVD) produces GaN crystal orientation that makes the sheet charges caused by spontaneous and piezoelectric polarizations add constructively, as in Figure 2.3 above. As a result of the MOCVD epitaxial growth, the polarization field in the AlGaN layer will be higher than that in the buffer layer. Due to the discontinuity of
the polarization field, a high positive sheet charge is present at the interface. Figure 2.4 illustrates the combined piezoelectric and spontaneous polarization field in AlGaN/GaN structure.

As the thickness of the AlGaN layer increases during the growth process, the crystal energy increases. After a certain thickness, the internal electric field becomes high enough to ionize donor states at the surface and cause electrons to drift toward the AlGaN/GaN interface. As the electrons move from the surface to the interface, the magnitude of the electric field is reduced, thereby acting as a feedback mechanism to diminish the electron transfer process. Under equilibrium condition, a 2DEG charge at the interface will be generated due to the transferred electrons and a positive charge on the surface will be formed from the ionized donors. The effect is illustrated in Figure 2.5.

Figure 2.4. Combined piezoelectric and spontaneous polarization field in AlGan/GaN structure

As the thickness of the AlGaN layer increases during the growth process, the crystal energy increases. After a certain thickness, the internal electric field becomes high enough to ionize donor states at the surface and cause electrons to drift toward the AlGaN/GaN interface. As the electrons move from the surface to the interface, the magnitude of the electric field is reduced, thereby acting as a feedback mechanism to diminish the electron transfer process. Under equilibrium condition, a 2DEG charge at the interface will be generated due to the transferred electrons and a positive charge on the surface will be formed from the ionized donors. The effect is illustrated in Figure 2.5.
2.2.2 Surface States (Traps)

Ibbetson et al investigated the mechanism of formation of charged surface states and the importance of these states for generation of the 2DEG channel in AlGaN/GaN HEMT. A conclusion from the investigation was that for non-ideal surface with available donor-like states, the energy of these states will increase with increasing AlGaN thickness. At certain thickness the states energy reaches the Fermi level and electrons are then able to transfer from occupied surface states to empty conduction band states at the interface, creating 2DEG and leaving behind positive surface sheet charge. For ideal surface with no surface states, the only available occupied states are in the valence band. The 2DEG exists as long as the AlGaN layer is thick enough to allow the valence band to reach the Fermi level at the surface or interface. It is necessary that a significant part of the GaN conduction band be pulled below the Fermi surface. The electrons in the valence
band will be excited into these states at most temperatures. This allows electrons to transfer from AlGaN valence band to the GaN conduction band, leaving behind surface holes. These accumulated holes produce a surface positive sheet charge. With either case, a positive charge must exist at the surface in order for the 2DEG to be present in the AlGaN/GaN interface. [1]

The surface states act as electron traps located in the access regions between the metal contacts. Proper surface passivation prevents the surface states from being neutralized by trapped electrons and therefore maintains the positive surface charge. If the passivation process is imperfect, then electrons leaking from the gate metal under the influence of a large electric field present during high power operation can get trapped. Reduction in the surface charge due to the trapped electrons will produce a corresponding reduction in the 2DEG charge, and therefore reduce the channel current. The amount of trapped electrons and current reduction depends on the applied bias voltages and the extent to which the device is overdriven beyond the linear gain. [1] The trapped electrons are modulated with the low frequency stimulating voltages, and therefore can contribute to the 2DEG channel current. However, trapped electrons cannot follow the high frequency stimulating voltages, and therefore produce a channel current reduction. This reduction in the current under RF operation is called current dispersion, or more precisely, surface traps induced current dispersion. [1]
3. Description of Large HEMTs

Layout is important when characterizing HEMTs. Large HEMTs can be described as parallel or series. [3] The overall width is set by the width of the gate over the active mesa area, $W_g$, times the number of gate fingers. Large HEMTs are set by the gate width, $W_g$, as shown in Figure 3.1.

\[ W_g = W \times \text{(number of fingers)} \]  

(3.1)

![Figure 3.1. Image showing gate width](image)

The parallel layout minimizes area by sharing drain and source connections between HEMTs. Two benefits are achieved with parallel layout: 1) smaller layout size and 2) reduction of source and drain depletion capacitances. The reduction in depletion capacitance is important in analog design, and latch-up for inverter circuits. Widths of HEMT’s laid out in parallel add to form an equivalent width HEMT. The parallel concept is useful for layout design. It also helps to find the optimal extraction frequency range. As
the gate width of the HEMT increases, the frequency range for reliable model parameter extraction decreases. The physical origin of this behavior is the small signal modeling limitation with respect to signal frequency. [4] Additionally, it is useful when defining width-to-length ratio for HEMTs.

\[
\text{Width-to-length ratio} = \frac{W_g}{L_g} \tag{3.2}
\]

The following figures show the HEMT layouts used for this research. Each HEMT has four fingers. The gate width and pitch between gates are modified on each transistor. The width and pitch changes are investigated to help identify variation in the s-parameter response for each transistor. Air bridges connect the sources together. Air Force Research Laboratory (AFRL) and REMEC, Inc provided the transistor layout for the HEMTs. Using Design Workshop 2000, the masks were laid out and inspected before constructing photolithography masks. Incorporating AFRL’s transistor designs into my thesis project allows me to investigate changes in the small signal equivalent circuit and large signal equivalent circuit based on modifications to each transistor layout. Through my research, one of the transistor designs should have performance that stands out based on modeling and simulation, IV characteristics, and large signal gain.
Figure 3.2. Four Finger HEMT, \( W_g = 100 \mu m \) with 15, 25, and 35\( \mu m \) pitch

Figure 3.3. Four Finger HEMT, \( W_g = 85 \mu m \) with 15, 25, and 35\( \mu m \) pitch

Figure 3.4. Four Finger HEMT, \( W_g = 65 \mu m \) with 15, 25, and 35\( \mu m \) pitch
Figure 3.5. Four Finger HEMT, Wg = 50µm with 15, 25, and 35µm pitch
4. Testing HEMTs

Testing HEMTs requires DC, RF, pulsed IV and load-pull measurements. The following sections explain the testing methods required for HEMT testing.

4.1 DC Testing

The 12 HEMTs used for my research were tested at DC for a) IV characteristics, b) pinch-off voltage, c) max current, d) peak current, e) breakdown voltage, f) knee voltage, g) threshold voltage and h) leakage current.

Equipment used to test epitaxial layer, metal contact layers at DC Bias is the Keithley Yieldmax 450 with Electroglas prober. The Keithley is capable of completing automated measurements for the following parameters: resistance, current isolation measurements, breakdown voltage, and sheet resistance. The Keithley assists with process control testing for on-wafer characterization monitoring throughout HEMT processing. Figure 4.1 has diagrams showing the variation of the parameters across the wafer.
Figure 4.1. DC testing parameters

The semiconductor parameter analyzer is used to test HEMTs for DC characteristics. The semiconductor parameter analyzer consists of HP 8510C, HP4142, Cascade prober and computer. DC measurements obtained from the parameter analyzer are max current, peak current, breakdown voltage, knee voltage, threshold voltage, and leakage current. The following diagrams show the variation of the parameters for the 4x50 p15 and 4x100 p 35 HEMTs across the wafer.
Figure 4.2. DC testing parameters for HEMT
The data gathered from DC measurements have an important role in analyzing HEMTs for modeling and simulation.

### 4.2 RF Testing

For RF characteristics, the HEMTs were tested for s-parameter data. Using the s-parameters data, I could extract the following characteristics: short circuit current gain ($h_{21}$), peak transconductance ($G_{mp}$), cutoff frequency ($f_t$), and maximum frequency of oscillation ($f_{max}$). Figure 4.3.1 and 4.3.2 have diagrams showing the variation of the parameters for the 4x50 p15 and 4x100 p35 HEMTs across the wafer.

![Figure 4.3.1. RF testing parameters for HEMT](image-url)
Figure 4.3.2. RF testing parameters for HEMT

Equipment used to test HEMTs for small signal device characterization consists of HP8510c, HP4142, Cascade prober, and computer. The parameter analyzer is automated and makes acquisition of measured data simple. The RF testing plays a critical role in device fabrication. During device development, RF testing helps analyze performance of devices starting at the completion of gate processing.

The s-parameter data assists in developing data sets for device model simulation. These data sets are broken into cold-FET, hot-FET and linear device operation. The data in combination with extraction algorithms provide supporting detail about the device in order to help model and simulate HEMTs. The modeling and simulation help to understand the performance of the device and provides feedback to make improvements to transistor structure and epitaxial development.

4.3 Pulsed IV Testing

The Dynamic I(V) Analyzer (DiVA) is used to complete pulsed IV characterization testing. Testing consists of pulsing RF signals and biasing the HEMT to look at self-heating dispersion and ways to improve heat-dissipation mechanisms. The role pulsed IV testing plays in device modeling and simulation is to understand the device
behavior near isothermal measurements. Pulsed IV measurements help to characterize trapping mechanisms without the effect of self-heating commonly found in DC measurements. Pulsed IV measurements obtained under appropriate quiescent bias conditions are used for investigating trapping induced dispersion characteristic. The pulsed IV measurements offer ways to investigate the characteristic of the HEMT in ranges where damage or deterioration could occur. Also, pulsed IV measurements help to extend the measurement range without harming the HEMT. Figure 4.4 shows pulsed IV measurement for 4x50 p15 device.

![Pulsed IV measurement for 4x50 p15 HEMT](image)

**Figure 4.2** Pulsed IV measurement for 4x50 p15 HEMT

### 4.4 Load-Pull Testing

Load-pull measurement is a large signal technique designed to characterize device properties. Device characterization involves embedding the device into measurement circuitry that can be impedance tuned. The load-pull system simultaneously monitors the tuned impedance of the characterization circuitry and the performance of the device. [5] Device response is then recorded under the variable load conditions. The resulting loci of
impedances required to obtain a constant performance parameter are displayed on a
Smith Chart in form of closed contours. The major performance parameters are a) output
power, b) power added efficiency, and c) gain. The load-pull contours are determined one
frequency at a time. To have a wideband characterization, load-pull measurements can be
obtained over several discrete frequencies within a frequency band of interest. Care must
be taken to match the impedances presented at each frequency.

Load-pull systems are built in two general categories: traditional load-pull and
active-load. For this research effort, the active-load system is used to conduct load-pull
measurements. The active-load system has several advantages over traditional load-pull
systems. One advantage is that the load reflection coefficient is not limited in magnitude.
Independently adjusting the input attenuators in the system, the user can obtain virtually
any reflection coefficient. Another advantage is automation of the output power and
third-order intercept point contours. And, a third advantage is how the measurement
configuration is completely symmetric and allows the devices under test to see a 50 Ω
matched system at all harmonics of the applied signal.

The Maury Microwave Inc. system is used to conduct load-pull measurements. A
purpose of load-pull characterization is to test power devices in a heat free mode of
operation. [6] The self heating of power devices occurs under large drain bias and at high
drain current in a continuous mode of operation. The pulsed load-pull system allows for
measurement of RF power characteristics under various load and source terminations and
under pulsed RF and bias conditions. It gives researchers ability to observe power characteristics of HEMTs in a heat-free operation mode.

Load-pull measurement results can be used directly in the design of power amplifiers. The data can also be used when choosing the optimal devices for specific power applications or to verify large-signal models. Disadvantages for load-pull are a) no easy method for performing parameter extraction of large-signal model parameters, b) simulating load-pull results using computer aided design tools to perform parameter extraction is also computationally intensive. [5]
5. Small Signal Models

Small signal models assist Radio Frequency (RF) Engineers with examining the performance of AlGaN/GaN High Electron Mobility Transistors (HEMTs). Small-signal models are extremely important for active microwave circuit work. The models provide a vital link between measured s-parameters and the electrical processes occurring within the device. The equivalent circuit elements in the model provide a lumped element approximation to some aspect of the device physics. The equivalent circuit elements values can be scaled with gate width, which enables a designer to predict the s-parameters of different size devices. The ability to include device gate width scaling as part of the circuit design process is important in MMIC design applications.

Different methods developed over the past two decades help examine the performance of MESFETs, HFETs, and HEMTs. The Curtice method, which was originally developed for GaAs Field Effect Transistors (FETs), has proved to be a beneficial method in modeling AlGaN/GaN HEMTs small signal models. The majority of today’s models are derived from the Curtice model. Each model uses scattering parameters (s-parameters) to derive the models. For HEMTs, my research used small signal models to find the optimum transistor for X-Band operation. Three models were researched to determine which model would represent the best equivalent circuit for HEMTs. The three models are 1) Eight element cold FET model (Curtice Model), 2) Eight element hot FET model, 3) Twenty-two element distributed model. The advantages and disadvantage of the each model are explained to give you insight into what model best describes the performance of the HEMTs.
5.1. Eight element models

Two types of equivalent circuit models categorize the eight element models. The models are cold FET model and hot FET model. The eight element model accounts for the positive reverse transfer conductance because of the drain to channel feedback capacitance. The Gate and drain pad capacitances $C_{PG}$ and $C_{PD}$ are assumed to be negligibly small as compared to the input and output admittances of the intrinsic FET, which is often the case with medium-power GaAs FET’s. Other research has shown that only s-parameters measurements are enough to analytically extract extrinsic equivalent circuit parameters (ECP) of the eight-element models. The intrinsic equivalent circuit parameters are analytically extracted for the models. Relative errors help determine the difference between the calculated and measured s-parameters.

5.1.1. Eight element cold FET model

The eight element cold FET model is commonly known as the Curtice model. It was originally developed for the GaAs FET. Figure 5.1 shows the Curtice model. The model uses a cold FET method to extract parasitic capacitances of FETs. The control voltage $V_G$ is defined across $R_G$. The model represents the inductances, pad capacitances and resistances extracted for the equivalent circuit. In this model, $C_{PG}$ and $C_{PD}$ are assumed to be small and can be neglected. The intrinsic FET elements are extracted from y-parameters. The extrinsic ECP’s can be analytically obtained from the s-parameter measurements on a pinched-off cold FET. Measurements for the cold FET are biased with a constant $V_{DS}$ voltage and $V_{GS}$ well below pinch-off. Converting the s-parameters
to z-parameters, the extrinsic ECP for $R_G$, $R_S$, $R_D$, and $L_G$, $L_S$, $L_D$ can be obtained through analytical extraction.

\[
\begin{array}{c}
\text{Gate} \quad L_g \quad R_g \quad a.C_B \quad R_d \quad L_d \quad \text{Drain} \\
\mid \mid \\
C_{PF} \quad b.C_B \\
\mid \\
R_s \\
\mid \\
L_s \\
\mid \\
\text{Source}
\end{array}
\]

**Figure 5.1. Eight Element Cold FET Model including $L_g$, $L_d$**

### 5.1.2. Eight element hot FET model

The eight element hot FET model has the control voltage defined across $C_{GS}$.

Figure 5.2 shows the model elements. The hot FET model elements are similar to the cold FET model. S-parameter for the hot FET is biased at typical operating bias conditions constant $V_{DS}$ and $I_{DS} \approx I_{DSS}/2$. [7]

The values of $C_{GS}$, $C_{DS}$, $C_{GD}$, and $R_I$ are the same for both models. $C_{DS}$ and $g_m$ are nearly independent of the models. $\tau$ and $G_{DS}$ are model dependent. In the Curtice model, $\tau$ accounts for all delay effects under the gate. $G_{DS}$ is an important parameter affecting the output impedance level of power FET’s. $G_{DS}$ is frequency independent for hot FET model. Therefore, the hot FET model should be used to calculate $G_{DS}$.
5.2. Twenty-two element distributed model

The twenty-two element distributed model is useful to identify the device parasitic elements. The model is applicable for large gate periphery devices. The main advantages and modeling approach will be discussed in the section on the model extraction method.

5.3. Physical significance of equivalent circuit element values

The equivalent circuit element and its role need some discussion to help understand the device physics behind the model. The inductance, resistance and capacitances, transconductance, output conductance, transconductance delay, and charging resistance make up the equivalent circuit elements.
5.3.1. Parasitic Inductance

The parasitic inductance $L_s$, $L_d$, and $L_g$ arise mainly from metal contact pads deposited on the device surface. The inductance values are dependent on the surface features of the device. $L_g$ is usually the largest of the three values due to short gate lengths. $L_s$ is often the smallest inductance value, especially for devices with via hole grounds. It is also worth mentioning that these inductances exist in addition to any parasitic bond wire inductance or parasitic package inductances, which must be accounted for in the model.

5.3.2. Parasitic Resistance

The parasitic resistance $R_s$, $R_d$, and $R_g$ are included to account for the contact resistance of the ohmic contacts as well as any bulk resistance leading up to the active channel. $R_g$ results from the metallization resistance of the gate Schottky contact. Although measurements indicate a slight bias dependence in the resistance values, they are held constant in the large-signal models commonly available in circuit simulators. The resistance values can be estimated either from forward conduction measurements or directly from s-parameters using an optimization technique.

5.3.3. Capacitances $C_{gs}$, $C_{gd}$, and $C_{ds}$

The capacitances $C_{gs}$ and $C_{gd}$ model the change in the depletion charge with respect to the gate-source and gate-drain voltages respectively. Under typical amplifier or oscillator bias conditions, the gate-source capacitance is the larger quantity because it models the change in depletion charge resulting from fluctuations in gate-source voltage.
Under normal bias conditions, the gate-drain capacitance $C_{gd}$ is considerably smaller in magnitude than $C_{gs}$ but still critical to obtaining accurate s-parameter predictions. The drain-source capacitance $C_{ds}$ is included in the equivalent circuit to account for geometric capacitance effects between the source and drain electrodes. The values of $C_{gd}$ and $C_{ds}$ are about 1/10 the value of $C_{gs}$. Because of symmetry $C_{gs}$ and $C_{gd}$ are approximately equal for $V_{ds} = 0$ V and reverse roles for inverted drain-source bias conditions ($V_{ds} < 0$ V).

5.3.4. Transconductance ($g_m$)

The intrinsic gain mechanism of the HEMT is provided by the transconductance. The transconductance $g_m$ is a measure of the incremental change in the output current $I_{ds}$ for a given change in input voltage $V_{gs}$. Mathematically, $g_m$ is defined as

$$g_m = \partial I_{ds}/\partial V_{gs} \quad (5.1)$$

The value of transconductance varies with frequency below a frequency of about 1 MHz. Transconductance values vary directly with gate width and inversely with gate length for HEMTs.

5.3.5. Output Conductance ($g_{ds}$)

The output conductance is a measure of the incremental change in output current $I_{ds}$ with the output voltage $V_{ds}$. The output conductance is defined mathematically as

$$g_{ds} = \partial I_{ds}/\partial V_{ds} \quad (5.2)$$

27
Values of $g_{ds}$ are on the order of 1 mS/mm gate width at typical amplifier biases. Also, as gate length is reduced, output conductance tends to increase in HEMTs. Even more significant than the transconductance is the low frequency dispersion in the output conductance. The RF output conductance can be more than 100% higher than DC output conductance. The RF values are of primary concern for small-signal modeling applications.

5.3.6. Transconductance Delay

The transconductance cannot respond instantaneously to changes in gate voltage. The delay inherent to this process is described by the transconductance delay ($\tau$). Physically, the transconductance delay represents the time it takes for the charge to redistribute itself after a fluctuation of gate voltage. Typical values of $\tau$ are on the order of 1 psec for microwave HEMTs. Transconductance delays tend to decrease with decreasing gate length.

5.3.7. Charging Resistance ($R_i$)

The charging resistance is included in the equivalent circuit primarily to improve the match to $S_{11}$. For many devices, however, the presence of $R_g$ is sufficient to match the real part of $S_{11}$. In either case, $R_i$ is difficult to extract and is of questionable physical significance. The inclusion of $R_i$ also complicates the large-signal analysis.
6. Small Signal model extraction methods

The 22 element model has an s-parameter extraction method for device modeling. (1) The extraction method developed by Dr. Günter Kompa uses a multi-plane data-fitting and bidirectional search technique. The extraction method helps to overcome extraction problems found in other extraction techniques such as: 1) optimizer based techniques that have a problem of non-uniqueness making them impractical; 2) taking into account additional measurements for the optimizer technique and using a partitioning approach or automatic decomposition technique even have uncertainties with regard to starting value problems. 3) Fast analytical techniques need special test structures or additional measurements steps such as DC and/or RF characterizations of FET’s under forward bias conditions. Such sequentially derived solutions may yield large errors in FET modeling. 4) Using multi-circuit measurements, the model parameters can uniquely be identified, but the number of optimization variables increases rapidly. 5) A purely analytical reverse solution approach can only be applied under the assumption of almost ideal measured s-parameters. Dr. Kompa’s approach helps to overcome the extraction problems mentioned above. [2]

6.1. The 22-element distributed small signal equivalent circuit model

It is important to identify device parasitic elements for further minimization. The 22-element distributed model helps to identify small signal model elements for GaN HEMTs. The model is general and applicable for large gate periphery devices. The main advantages of the model are as follows: 1) It accounts for all expected parasitic elements of the device; 2) It reflects the physics of the device over a wide bias and frequency
range. The advantages of the model make it suitable for scalable large signal model construction. Figure 6.1 shows the 22 element distributed model for active GaN HEMT.

\[ I_{ds} = V_i G_m e^{-jwt} \]

**Figure 6.1. 22-element distributed model for active GaN HEMT**

In this model, \( C_{pgi}, C_{pdi}, \) and \( C_{gdi} \), account for the interelectrode and crossover capacitances (due from air-bridge source connections) between gate, source, and drain. While \( C_{pga}, C_{pda}, \) and \( C_{gda} \) account for parasitic elements due to the pad connections, measurement equipment, probes, and probe tip-to-device contact transitions. [9]

**6.2. Model Parameter Extraction Technique**

Many of the small signal model parameters in figure 6.1 are difficult if not impossible to determine directly from measurements. The method to determine the model parameters is accomplished through an optimization algorithm. The efficiency of the algorithm depends on the quality of the starting values and the number of optimization variables. Under cold pinchoff condition, the equivalent circuit in figure 6.1 simplifies to
a capacitive network. By reducing some elements, you can reduce the number of
unknowns. A further minimization of the number of optimization variables, and only the
extrinsic elements of the small signal will be optimized. The intrinsic elements are
determined from the deembedded z-parameters. The estimated starting values should be
carried out in a way that accounts for the correlation between the bias condition and the
reactive elements. The s-parameter measurements must cover the frequency range where
the correlation is most obvious. The estimated starting values for the gate-source, gate-
drain capacitances was shown in [9] for a 2 x 50 µm HEMT. Higher frequencies show
meaningful values for $C_{gs}$, and $C_{gd}$ under this technique.

The s-parameter measurement correlation effects the starting values for the small
signal optimization. Capacitances at low frequencies may not be physically justifiable.
This is due to inductive effects that have been found to become influential only above
certain frequency ranges. The inductive influence varies by device size. The required
measurement frequency range for reliable starting values generation reduces significantly
for larger devices, e.g., up to 20 GHz for and 8 x 125 µm device. [9] The model
parameter extraction technique for generation of starting values is based on searching for
the optimal distribution of the total capacitances. Scanning the outer capacitance values
within a specified range assists in the search. For each scanned value, the inter-electrode
capacitances are assigned suitable values and then de-embedded from the measured y-
parameter. The rest of the model parameters are then estimated from the stripped y-
parameters. The whole estimated parameters are used to simulate the device s-
parameters, which are compared with the measured ones. Using this systematic searching
procedure, high-quality measurement-correlated starting values for the small signal model parameters can be found. The closeness of the parameter starting values to the real values simplifies the step for parameter optimization since the risk of a local minimum is reduced.

Dr. Kompa has several research papers on the procedures to extract the model parameters. The first procedure breaks the extraction into several steps. The steps generate the starting extrinsic values of the small signal model parameters. The second procedure works to optimize the starting values by finding a minimum error between the measured and simulated s-parameter values. The third procedure works on the extraction of intrinsic model parameters.

6.3. Extrinsic parameter extraction

The procedure for extrinsic parameter extraction begins with determining the pinchoff voltage for the HEMT. The starting values of the extrinsic capacitances and inductances are generated from pinchoff measurements. The extrinsic resistances are generated from forward measurements. Starting with $V_{GS} < V_P$ and $V_{DS} = 0.0$ V. The equivalent circuit in figure 6.2 can represent the model at this point. For the cold pinchoff device the current source and output channel conductance are excluded. At low frequencies (below 5 GHz), the circuit reduces to a capacitive network similar to the network shown in the Figure 6.2.
Figure 6.2. Cold pinchoff equivalent circuit for the GaN HEMT at low frequency

The y-parameters of this equivalent circuit can be described by the following equations:

\[ y_{11} = j\omega(C_{gxo} + C_{gdo}) \]  \hspace{1cm} (6.1)

\[ y_{22} = j\omega(C_{dso} + C_{gdo}) \]  \hspace{1cm} (6.2)

\[ y_{12} = y_{21} = j\omega(C_{gdo}) \]  \hspace{1cm} (6.3)

where

\[ C_{gdo} = C_{gda} + C_{gdi} + C_{gd} \]  \hspace{1cm} (6.4)

\[ C_{gxo} = C_{pga} + C_{pgi} + C_{gs} \]  \hspace{1cm} (6.5)

\[ C_{dso} = C_{pda} + C_{pdi} + C_{ds} \]  \hspace{1cm} (6.6)
The total capacitances for gate-source, gate-drain, and drain-source branches are determined at low frequency range from cold pinch-off s-parameter measurements, which are converted to Y-parameter.

The next step in the procedure calls for search on the optimal distribution of the total capacitances. Through the search, a minimum error between the measured and simulated s-parameters helps to determine the starting values. Scanning $C_{pga}$, $C_{pda}$, and $C_{gda}$ help to determine the minimum error. Through research, the values of $C_{pga}$, and $C_{pda}$ are scanned from 0 to 0.5$Cdso$, while $C_{gds}$ is scanned from 0 to 0.5$C_{gdo}$. During the process, $C_{pga}$ is assumed to be equal to $C_{pda}$.

$$C_{pga} \equiv C_{pda} \quad (6.7)$$

The gate-drain interelectrode capacitance $C_{gdi}$ is assumed to be twice the pad capacitance $C_{gda}$

$$C_{gdi} \equiv 2C_{gda} \quad (6.8)$$

For symmetrical gate-source and gate-drain spacing, the depletion region will be uniform under pinchoff, making

$$C_{gs} \equiv C_{gd} = C_{gdo} - C_{gdi} - C_{gda} \quad (6.9)$$
The value of $C_{\text{pgi}}$ is calculated using the following equation

$$C_{\text{pgi}} = C_{\text{gsi}} - C_{\text{gd}} - C_{\text{pga}}$$  \hspace{1cm} (6.10)$$

For GaN HEMT devices, $C_{\text{pdi}}$ is a significant part of the total drain-source capacitance. It is found that the assumption

$$C_{\text{pdi}} = 3C_{\text{pda}}$$  \hspace{1cm} (6.11)$$

minimizes the error between the simulated and measured s-parameters. For medium and high frequency ranges, the intrinsic transistor of the pinchoff model is represented by a T-network as shown in the following figure.

![T-network representation of a cold pinchoff FET equivalent circuit](image)

**Figure 6.3.** T-network representation of a cold pinchoff FET equivalent circuit
The interelectrode capacitances (C_{pgi}, C_{pdi}, and C_{gdi}) have been absorbed in the intrinsic capacitances (C_{gs}, C_{ds}, and C_{gd}). The following equations represent the C_d, C_s, and C_g for the T-network.

\[
C_d = C_{ds} + C_{gd} + \frac{C_{ds} C_{gd}}{C_{gs}} \quad (6.12)
\]

\[
C_s = C_{gs} + C_{ds} + \frac{C_{gs} C_{ds}}{C_{gd}} \quad (6.13)
\]

\[
C_g = C_{gs} + C_{gd} + \frac{C_{gs} C_{gd}}{C_{ds}} \quad (6.14)
\]

The values for C_{pga}, C_{pda}, and C_{gda} are deembedded from y-parameters and then converted to z-parameters. The stripped z-parameters equations are written as

\[
z_{11} = R_x + R_s + j\omega(L_x + L_s) + \frac{1}{j\omega} \left( \frac{1}{C_g} + \frac{1}{C_s} \right) + \delta Z_x \quad (6.15)
\]

\[
z_{22} = R_d + R_s + j\omega(L_d + L_s) + \frac{1}{j\omega} \left( \frac{1}{C_d} + \frac{1}{C_s} \right) + \delta Z_d \quad (6.16)
\]

\[
z_{11} = z_{21} = R_s + j\omega L_s + \frac{1}{j\omega C_s} + \delta Z_z \quad (6.17)
\]

where

\[
\delta Z_g = \delta R_g + \delta R_s + j\omega(\delta L_g + \delta L_s) \quad (6.18)
\]

\[
\delta Z_d = \delta R_d + \delta R_s + j\omega(\delta L_d + \delta L_s) \quad (6.19)
\]

\[
\delta Z_s = \delta R_s + j\omega \delta L_s \quad (6.20)
\]
\( \delta Z_g, \delta Z_d, \) and \( \delta Z_s \) represent correction terms related to the intrinsic parameters of the model. Ignoring the correction terms and multiplying the \( z \)-parameters by \( \omega \) and then taking the imaginary part gives the following equations to extract \( L_g, L_d, \) and \( L_s \).

\[
\text{Im}[\alpha Z_{11}] = (L_g + L_s)\omega^2 - \left( \frac{1}{C_g} + \frac{1}{C_s} \right)
\]

(6.21)

\[
\text{Im}[\alpha Z_{22}] = (L_d + L_s)\omega^2 - \left( \frac{1}{C_d} + \frac{1}{C_s} \right)
\]

(6.22)

\[
\text{Im}[\alpha Z_{12}] = (L_s)\omega^2 - \left( \frac{1}{C_s} \right)
\]

(6.23)

The values for \( L_g, L_d, \) and \( L_s \) can be extracted from the slope of \( \text{Im}[Z_{ij}] \) versus \( \omega^2 \) curve. The estimated values of the inductances and interelectrode capacitances (\( C_{pgi}, C_{pdi}, \) and \( C_{gdi} \)) are deembedded. One downfall of the step is that the incomplete deembedding of the outer capacitances and the inductances introduce nonlinear frequency dependence in the real part of deembedded \( z \)-parameters. By multiplying the deembedded \( z \)-parameter by \( \omega^2 \), this effect is reduced. Ignoring the correction terms and multiplying the deembedded \( z \)-parameters by \( \omega^2 \) and then taking the real part of this \( z \)-parameter gives equations to extract resistance values. The values of \( R_g + R_s, R_d + R_s, \) and \( R_s \) are extracted using linear regression.

\[
\omega^2 \text{Re}[Z_{11}] = \omega^2 \left( R_g + R_s \right)
\]

(6.24)

\[
\omega^2 \text{Re}[Z_{22}] = \omega^2 \left( R_d + R_s \right)
\]

(6.25)
The resulting estimated parameters are used to simulate the device s-parameters, which are then compared with the measured ones to calculate the residual fitting error ($\epsilon$). The outer capacitances ($C_{pga}$, $C_{pda}$, and $C_{gda}$) are incremented, and the procedure is repeated until $C_{pga}$, $C_{pda}$ are equal to 0.5$C_{dso}$ and $C_{gda}$ is equal to 0.5$C_{gdo}$. The vector of model parameters $P(\epsilon_{\text{min}})$, corresponding to the lowest error $\epsilon_{\text{min}}$, is taken as the appropriate starting value.

Using voltage pinchoff s-parameters causes an unavoidably high measurement uncertainty for $R_g$, $R_d$, and $R_s$, therefore it is not recommended to use these measurements to obtain values for $R_g$, $R_d$, and $R_s$. Determining the starting values for $R_g$, $R_d$, and $R_s$ is nearly impossible using cold pinchoff s-parameters. For a more reliable starting value, cold gate forward s-parameter measurements at high gate voltage (>2.0 V) provides more meaningful results. This is due to the higher conduction band of GaN based HEMT with respect to corresponding GaAs based HEMT. The higher voltages are applied to reach the condition when the influence of the gate capacitance is negligible. The values of extrinsic capacitances and inductances determined from the deembedded measurements are also deembedded from the gate forward measurements. The starting values of extrinsic resistances are then estimated from the stripped forward z-parameters.

Complete starting values for the pinchoff device model parameters will be discussed in the simulated versus experimental small signal model results section.
6.4. Model Parameter Optimization

The model parameter optimization for distributed small signal modeling requires an approach similar to lumped small signal modeling. Due to the extra parasitic elements of the model, the search space is increased. The algorithm for the lumped small signal model can be modified to apply for the distributed model. The closeness of the starting values to the true value allows the searching space to be reduced and only the optimization of extrinsic parameters is required. Iteration of the parameters through the optimization process assigns suitable extrinsic values for deembedding values from measured data. The intrinsic y-parameters are determined from the deembedded measurement data. The intrinsic model parameters are estimated by means of data fitting from the deembedded measurements. Next, estimated model parameters are used to fit measured s-parameters. The optimization problem is a nonlinear multidimensional one, whose objective function could have multiple local minima. Also, the cold pinchoff device measurements have a high uncertainty. The nonlinear multidimensional problem and high uncertainty from cold pinchoff device measurements increase the probability of trapping into a local minimum. Careful formulation for the objective function reduces the local minima problem. The magnitude of the error between measured s-parameters and its simulated value can be expressed with the following equation. [9]

\[
\varepsilon_{ij} = \frac{\left| \text{Re}(\Delta \tilde{Y}_{ij,n}) \right| + \left| \text{Im}(\Delta \tilde{Y}_{ij,n}) \right|}{W_{ij}} \quad i,j = 1,2 \quad (6.27)
\]

\[n = 1,2,\ldots, N\]

where \( W_{ij} = \max \left| S_{ij} \right| \) \( i,j = 1,2; \ i \neq j \) \quad (6.28)

\[W_{ii} = 1 + \left| S_{ii} \right|, \quad i = 1, 2 \quad (6.29)\]
$N$ represents the total number of data points. $\delta S$ is the difference between the measured s-parameter coefficient and the simulated value. The weighting factor ($W$) deemphasizes data region with higher reflection coefficients due to the involved higher measurements uncertainty.

The scalar error $\varepsilon_s$ is determined using the following equation. The 1-norm is summed across the frequency range for each change in parameters.

$$\varepsilon_s = \frac{1}{N} \sum_{n=1}^{N} \left\| \varepsilon(f_n) \right\|_1$$

where

$$\varepsilon(f_n) = \begin{bmatrix} \varepsilon_{11}(f_n) & \varepsilon_{11}(f_n) \\ \varepsilon_{21}(f_n) & \varepsilon_{22}(f_n) \end{bmatrix}$$

The previous equation is defined at each frequency point $n$. The objective function that is based on s-parameters alone to minimize the fitting error may not necessary lead to physically relevant values of the model parameters. It is suggested to enhance the objective function with other performance quantities based on final application requirements. For GaN based HEMTs the requirements are based on these factors: 1) output and input impedance, 2) device gain 3) stability factor. These factors are important for design of matching networks. The factors can be expressed as a function of s-parameters and fitted during the optimization.

The stability factor defined at the output plane of the device at each frequency is expressed by the following equation.

$$K = \frac{1 - \left| S_{22} \right|^2}{\left| S_{22} - S_{11}^* \Delta \right| + \left| S_{12} S_{21} \right|}$$

40
$S_{11}^{*}$ is the complex conjugate and $\Delta_s$ is the determinant of the s-parameter matrix at each frequency. The fitting error of the stability factor is represented by the following equation.

$$\varepsilon_K = \frac{1}{N} \sum_{k=1}^{N} |K_{\text{meas}} - K_{\text{sim}}|$$

(6.33)

$K_{\text{meas}}$ and $K_{\text{sim}}$ are the stability factors from the measured and simulated s-parameters, respectively.

The device gain is another important factor for GaN HEMT. The maximally efficient gain defined by Kotzebue [10] is recommended to define gain of the model. It remains finite even for an unstable device. The gain is defined at each frequency by the following equation.

$$G = \frac{|S_{21}|^2 - 1}{\ln |S_{21}|^2}$$

The gain error is calculated with the following equation.

$$\varepsilon_G = \frac{1}{N} \sum_{n=1}^{N} |G_{\text{meas}} - G_{\text{sim}}|$$

(6.34)

$G_{\text{meas}}$ and $G_{\text{sim}}$ are the gains computed from the measured and modeled s-parameters.

With errors defined for gain, stability and scalar, the objective function is defined by the following fitting error equation. The fitting error consists of three components. Using a modified Simplex optimization algorithm [9], the objective function is minimized to reveal the optimized device parameters.

$$\varepsilon = \sqrt[3]{\frac{1}{3}(\varepsilon_s^2 + \varepsilon_K^2 + \varepsilon_G^2)}$$

(6.35)
6.5. Intrinsic Parameter Extraction

To extract the intrinsic parameters, the extrinsic parameters have to be dembedded from the s-parameter measurements. The intrinsic parameters extraction process is conducted at bias, $V_{GS} = -1.0 \, \text{V}$ and $V_{DS} = 10 \, \text{V}$, versus frequency. The frequency independence of the intrinsic elements occurs over the entire frequency range. At low and medium frequency range, the independence of the intrinsic elements validates the proposed small signal model topology and the developed extraction method.

A frequency-dependent effect in the intrinsic elements arises due to biasing in the linear region. This effect makes the optimal intrinsic parameters extraction difficult. To account for this effect, an efficient technique is developed for extraction of the optimal value for the intrinsic elements. [9] Through this technique, the intrinsic y-parameters are formulated in a way where the optimal intrinsic element value are extracted using simple linear data fitting. The admittance for the intrinsic gate-source branch, $Y_{gs}$ is calculated with the following equation.

$$ Y_{gs} = Y_{i,11} + Y_{i,12} = \frac{G_{gsf} + j \omega C_{gs}}{1 + R_i G_{gsf} + j \omega R_i C_{gs}} $$  \hspace{1cm} (6.36)

Defining a new variable $D$ with the following equation

$$ D = \frac{\left| Y_{gs} \right|^2}{\text{Im} \left[ Y_{gs} \right]} = \frac{G_{gsf}^2}{\omega C_{gs}} + \omega C_{gs} $$  \hspace{1cm} (6.37)

$C_{gs}$ can be determined from the slope of the curve for $\omega D$ versus $\omega^2$ by linear fitting. $G_{gsf}$ can be determined from the real part of $Y_{gs}$ at low frequencies (in the megahertz range). The admittance for the intrinsic gate-drain branch $Y_{gd}$ is calculated with the following equation.
\[
Y_{gd} = -Y_{i,12} = \frac{G_{gdf} + j\omega C_{gd}}{1 + R_{gd} G_{gdf} + j\omega R_{gd} C_{gd}}
\]  
(6.38)

To extract \( C_{gd}, R_{gd}, \) and \( G_{gdf}, \) the same procedure used to extract \( C_{gs}, R_{i}, \) and \( G_{gsf}. \) By defining a new variable \( D \) to conduct linear fitting, the extraction is straightforward. The following equations are used to extract \( C_{gd}, R_{gd}, \) and \( G_{gdf}. \)

\[
Y_{gm} = Y_{i,21} - Y_{i,12} = \frac{G_m e^{-j\omega \tau}}{1 + R G_{gsf} + j\omega C_{gs}}
\]  
(6.39)

\[
D = \left| \frac{Y_{gs}}{Y_{gm}} \right|^2 = \left( \frac{G_{gsf}}{G_m} \right)^2 + \left( \frac{C_{gs}}{G_m} \right)^2 \omega^2
\]  
(6.40)

\( G_m \) can be determined from the slope of the curve \( D \) versus \( \omega^2 \) by linear fitting. By defining \( D \) with the following equation, \( \tau \) can be determined from the plot of the phase of \( D \) versus \( \omega \) by linear fitting.

\[
D = (G_{gsf} + j\omega C_{gs}) \frac{Y_{gm}}{Y_{gs}} = G_m e^{-j\omega \tau}
\]  
(6.41)

The admittance of the intrinsic drain-source branch \( Y_{ds} \) can be expressed with the following equation. Due to the frequency dependent effect in the output conductance, \( G_{ds} \) is determined from the curve of \( \omega \text{Re}[Y_{ds}] \) versus \( \omega \) by linear fitting.

\[
Y_{ds} = Y_{i,22} + Y_{i,12} = G_{ds} + j\omega C_{ds}
\]  
(6.42)

In [3], Jarndal suggests inspecting the admittance curves using only the portion with the most linearity. It improves the optimization of the intrinsic parameters. Implementing a search procedure to determine the portion with the most linearity requires a correlation coefficient as a measure to define the data range. Intrinsic parameter extraction using the proposed procedure gives an accurate determination for the intrinsic
element values. It plays a significant role under bias conditions where intrinsic measured data is increasingly frequency dependent.

In section 8, the extraction method is used to model 12 HEMTs. The HEMTs are four finger devices with various changes made to the topology. The changes in the devices are source-drain spacing, pitch, gate width. The extraction method will be used to choose the best device to operate at X-Band frequency.
7. Large Signal Models

Research of Large Signal Models for HEMTs has been ongoing for several decades. The three main models considered for my research include the Walter Curtice Cold FET model (WRCC C_FET) [5], Triquint’s On Model 3 (TOM3) [12], and the Non-Quasi-Static FET Model [1]. The WRCC C_FET and the TOM3 were originally developed for MESFET modeling. The following table describes the modeling attributes available with the models. The table indicates the available attributes for each model. “A” indicates an attribute, and “X” indicates the model does not perform the function. All models provide gate-to-source and gate-to-drain capacitance models that are representative of charge conservation.

<table>
<thead>
<tr>
<th>Modeling Attribute</th>
<th>TOM3</th>
<th>WRCC C_FET</th>
<th>Non-Quasi-Static FET</th>
</tr>
</thead>
<tbody>
<tr>
<td>I/V Fit to large $V_{DS}$</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>$g_m(V)$ Function</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>$C_{gs}(V)$ Function</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>$G_{ds}$ Dispersion</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>$g_m$ Dispersion</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>G-D Breakdown</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Self-Heating</td>
<td>X</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Sub-Threshold</td>
<td>X</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Harmonic Balance</td>
<td>?</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Convergence</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Time Delay ($\tau$)</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Ambient Temperature Effects</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
</tbody>
</table>

Table 7.1. Modeling attributes

7.1. WRCC C_FET

The WRCC C_FET model is based on the Curtice 3 model originally published as the Curtice-Ettenberg MESFET model in 1985. [11] It incorporates a charge conservation model and all enhancements since the Curtice 3 model. The modifications include the characterization of self-heating, sub-threshold current, physically correct time delay,
transconductance dispersion, and advanced mathematical convergence properties due to coding that calculates pinch-off voltage for each operating point. The WRCC C_FET uses gate-to-drain breakdown functions that reflect observed device behavior. The WRCC C_FET model provides a gate to source capacitance function that characterizes the “overshoot” capacitance observed in GaN HEMTs. The WRCC C_FET implements self-heating in a physical manner, only the power dissipated in the device will increase its temperature. Figure 7.1 shows the topology of the WRCC C_FET model.

![C_FET large signal model equivalent circuit](image)

**Figure 7.1. C_FET large signal model equivalent circuit**

The model is an empirical large-signal representation of a HEMT and includes added circuitry for thermal heating effects. The model is a four port equivalent circuit, with an extra port that provides a measure of temperature rise. The voltage between ports 2 and 3 is equal to the temperature rise in degrees Celsius. This result of the thermal circuit’s current source being numerically equal to the instantaneous power dissipated in
the HEMT, and the resistance, \(R_{\text{TH}}\) being numerically equal to the thermal resistance. The RC product of the thermal circuit is the thermal time constant, \(\tau\). The thermal circuit allows for compensation in the circuit based on actual device operation. For the C_FET model, the large-signal equivalent circuit is looked at from a small-signal representation. This allows large-signal characterization to be conducted from a series of small-signal and DC measurements.

7.2. TOM3

The TOM3 is used as a modeling tool to assist with linking large and small signal MESFET Models in SPICE. TOM3 is an effort to improve the accuracy of the capacitance equations by using quasi-static charge conservation in the implanted layer of a MESFET. The channel current equation has been refined from previous TriQuint Models to include frequency dispersion and other effects. The TOM3 capacitance model calculated the gate charge from the drain current and the gate capacitances from the drain conductances. The initial investigation of the gate-charge function relied on conductance data extracted from measured s-parameters. Implementing the TOM3 capacitance equations in pSpice verified their utility and showed no convergence problems. Since errors in the drain conductances computed from \(I_{DS}\) directly effect the gate charge and gate capacitances, the margin of accuracy is placed upon the dc current equation.

TOM3 uses a linear slope function to estimate the drain current in the high-power region and is linear in the drain voltage. TOM3 calculates the drain current \(I_{DS}\) from the following set of equations:
\[ I_{DS} = AREA \times I_o(1 + \lambda V_{DS}) \]  
\[ (7.1) \]

\[ I_o = \beta V_G^0 f_K(\alpha V_{DS}) \]  
\[ (7.2) \]

\[ f_K(\alpha V_{DS}) = \frac{\alpha V_{DS}}{[1 + (\alpha V_{DS})^\kappa]^{1/k}} \]  
\[ (7.3) \]

\[ V_G = Q_{v_{ST}} \log[\exp(u) + 1] \]  
\[ (7.4) \]

\[ u = \frac{V_{GS} - V_{TO} + \gamma V_{DS}}{Q_{v_{ST}}} \]  
\[ (7.5) \]

\[ v_{ST} = v_{ST}(1 + m_{STO} V_{DS}) \]  
\[ (7.6) \]

As the \( I_{DS} \) equation and model topology evolve and improve, the changes can be incorporated in the gate-charge and capacitance calculations using the combined equation set of TOM3. [12]

### 7.3. Non-Quasi-Static FET Large Signal Model

In [1], Anwar Hasan Jarndal points out a major draw back for using Quasi-static Large Signal Modeling. Quasi-static assumption is a good first-order approximation in modeling of active device, but does not hold in the whole range of different operations conditions. Due to the strong impact of trapping and self-heating induced current dispersions, RF performance cannot be accurately modeled with quasi-static assumption. The non-quasi-static model implementations for large signal drain current help to predict the self-heating and current dispersions. In addition, at high frequency operation, the device channel charge under the gate does not respond immediately to the stimulation signal. This requires a relaxation time to build up. This effect results in quadratic frequency dependency of measured \( Y_{11} \) at high frequency. The large and small signal
model should be modified to simulate this effect. The device channel transconductance, \( G_m \), cannot respond instantaneously to changes in the gate voltage at high frequency. The time delay inherent to this process should be accounted for in the small-signal model. Figure 7.2 shows the intrinsic non-quasi-static linear device equivalent circuit.

![Intrinsic non-quasi-static linear device equivalent circuit](image)

**Figure 7.2. Intrinsic non-quasi-static linear device equivalent circuit**

In this model, the series resistances \( R_i \) and \( R_{gd} \), account for the quadratic frequency dependency of \( y \)-parameters. The transconductance time delay with respect to the applied gate voltage is described by transit time, \( \tau \). The large signal model topology used for my research is based on the model used by Jarndal. It incorporates the small and large signal topology for GaN HEMTs into one model. Figure 7.3 shows the topology of the Non-Quasi-Static FET Large Signal Model. This topology also reflects the symmetrical structure of the device especially at low drain-source voltages, while \( Q_{gs} \approx Q_{gd} \).
Figure 7.3. Non-quasi-static large signal model

If $G_{gsf}$ is only dependent on the gate-source voltage, the gate current sources, $I_{gs}$ and $I_{gd}$, can be obtained by splitting the gate current. The following equations are used to help calculate $G_{gsf}$ and $G_{gdf}$ for model simulation.

\[
I_{gs}(V_{gs}, V_{ds}) = I_{gs}(V_{gs0}, V_{ds0}) + \int_{V_{gs0}}^{V_{gs}} G_{gsf}(V, V_{ds0})dV
\]

(7.7)

\[
I_{gd}(V_{gs}, V_{ds}) = I_{gd}(V_{gs0}, V_{ds0}) + \int_{V_{gs0}}^{V_{gs}} G_{gdf}(V, V_{ds0})dV - \int_{V_{g0}}^{V_{gs}} G_{gdf}(V_{gs}, V)dV
\]

(7.8)

In order to incorporate the effect of $C_{gs}$, $C_{gd}$, and $C_{ds}$ and maintain the consistency of the large-signal model, the charge sources, $Q_{gs}$ and $Q_{gd}$, can be formulated with the following equations:
7.3.1. Traps Induced Dispersion Modeling

In [1], Anwar Hasan Jarndal discusses the different methods for modeling the non-quasi-static effect in the drain current due to traps induced dispersion. The methods discussed are 1) Based on single-pole function to describe the transition between DC and RF currents. It is limited by how well it efficiently stimulates the low frequency transition range between DC and RF. 2) Method developed by Werthof and Kompa looks at modeling the trap induced dispersion. It uses correlated current accounting for DC drain current, current contributed from RF transconductance and output conductance dispersion, and self-biasing effects at high input power operation. 3) The third method which Anwar Hasan Jarndal used in his research is to model the trapping induced current dispersion based on the assumption that under a operating frequency well above the dispersion cut-off frequency, the trapping mechanism is controlled by the DC components of the applied voltages $V_{gs}(t)$ and $V_{ds}(t)$. [1] The equation used to model the drain current under negligible self-heating effect can be modeled as

$$I_{ds}(V_{gs}, V_{ds}) = I_{DC}^{ds}_{iso}(V_{gs}, V_{ds}) + \alpha_G(V_{gs}, V_{ds})(V_{gs} - V_{gs0}) + \alpha_D(V_{gs}, V_{ds})(V_{ds} - V_{ds0})$$  \hspace{1cm} (7.11)
surface trapping and buffer trapping, respectively. Drain current is considered as a summation of dispersionless DC current and others dispersion contributions due to the RF current components. Model fitting parameters $I_{dc}$, $\alpha_G$ and $\alpha_D$ are extracted from pulsed IV characteristics, which do not heat up the device. As a result, the model can accurately describe the trapping induced dispersion. The following figure shows the model that helps to implement improvements in modeling the transition between DC and RF. The RC high pass configuration, in the gate and drain sides, can simulate a smooth transition between DC and RF by feeding $(V_{gs} - V_{gso})$ and $(V_{ds} - V_{dso})$ RF voltages back to the drain current model.

Figure 7.4. Non-quasi-static large signal model including trapping induced dispersion
7.3.2. Self-Heating Induced Dispersion Modeling

Modeling the drain current to include self-heating is defined with the following equation.

\[ I_{ds}(V_{gs}, V_{ds}) = I_{\text{DC}}^{\text{iso}}(V_{gs}, V_{ds}) + \alpha_G(V_{gs}, V_{ds})(V_{gs} - V_{gs0}) + \alpha_D(V_{gs}, V_{ds})(V_{ds} - V_{ds0}) + \alpha_T(V_{gs}, V_{ds})P_{\text{diss}} \]

(7.12)

where \( I_{\text{DC}}^{\text{iso}} \) is an isothermal DC current and \( \alpha_T \) is a fitting parameter accounting for the device thermal resistance and the nonlinear variation of the drain current with the device self-heating. \( P_{\text{diss}} \) is the average dissipated power. Using this approach, there is no need for technology dependent data or special measurements to determine the value of the thermal resistance. The following figure shows the low pass circuit to add on to the large-signal model equivalent circuit in order to include the self-heating simulation. The model helps to simulate the self-heating due to the average dissipated power and “quasi-static” dissipated power produced by slowly time varying voltage.

![Figure 7.5. Equivalent circuit implementation for device self-heating process](image-url)
7.4. Large Signal Model Extraction Procedures

The large-signal model equivalent circuit for AlGaN/GaN HEMT is shown in Figure 7.6. The circuit has two quasi-static gate current sources, $I_{gs}$ and $I_{gd}$, and two quasi-static gate charge sources, $Q_{gs}$ and $Q_{gd}$. The current sources and charge sources help describe the conduction and displacement currents. The non-quasi-static effect in the channel charge is approximately modeled with two bias-dependent resistances, $R_i$ and $R_{gd}$, in series with $Q_{gs}$ and $Q_{gd}$, respectively.

![Figure 7.6. Large-Signal model for AlGaN/GaN HEMT including self-heating and trapping effects](image)

Figure 7.6. Large-Signal model for AlGaN/GaN HEMT including self-heating and trapping effects
As explained in [1], this implementation improves the model simulation at millimeter wave frequencies by accounting for the required charging times for the depletion region capacitances. The charging times are described implicitly in the model by $R_iC_{gs}$ and $R_{go}C_{gd}$ products. A non-quasi-static drain current model accounts for trapping and self-heating effects is embedded in the proposed large-signal model. [1] The drain current value is determined by the applied intrinsic voltages $V_{gs}$ and $V_{ds}$, while the amount of trapping induced current dispersion is controlled with the RF components of these intrinsic voltages. The components are extracted from the intrinsic voltage using RC high-pass circuits in the gate and drain sides. The capacitors, $C_{GT}$ and $C_{DT}$ values are selected for values that provide a macroscopic modeling of small stored charges in the surface and buffer traps. These charges are almost related to the leakage currents from the gate metal edge to the surface or from the channel into the buffer layer. The small leakage currents in the gate and drain paths are realized with large ($\equiv 1M\Omega$) resistances $R_{GT}$ and $R_{DT}$ in series with $C_{GT}$ and $C_{DT}$, respectively. Implementing the model makes the equivalent circuit more physically meaningful. It improves the model accuracy for describing the low frequency dispersion. The frequency dispersion in the channel transconductance and output conductance is related mainly to the surface and buffer traps. The values of $R_{GT}$, $R_{DT}$, $C_{GT}$ and $C_{DT}$ are selected to define trapping time constants in the order of $10^{-5} – 10^{-4}$ sec. The amount of self-heating induced current dispersion is controlled with normalized channel temperature rise $\Delta T$. The normalized temperature rise is the channel temperature divided by the device thermal resistance $R_{th}$. Low pass circuit is added to determine the value of $\Delta T$ due to the static and quasi-static dissipated powers. The value of the thermal capacitance $C_{th}$ is selected to define the transit time constant in
the order of 1 ms. The thermal resistance $R_{th}$ is normalized to one because its value is incorporated in the thermal fitting parameter in the current model expression.

**7.5. Gate Charge Modeling**

The bias-dependent intrinsic elements are extracted as a function of the extrinsic voltages $V_{GS}$ and $V_{DS}$. A B-spline approximation technique is used to provide uniform data for the intrinsic elements [1]. The technique uses polynomial basis functions of degree $k$ to force fitting curves to pass through the fitted data and its higher derivatives up to ($k$-1) derivative. The intrinsic transconductance and gate-source capacitance are fitted using cubic B-spline approximation. The approximation technique preserves the continuity of the data and its derivatives up to the 2\textsuperscript{nd} derivative. As a result, this will improve the model simulation for the harmonics and the intermodulation distortions. The intrinsic capacitances $C_{gs}$, $C_{gd}$, and $C_{ds}$ are integrated using equation (7.9) and (7.10) to determine the values of $Q_{gs}$ and $Q_{gd}$. The determined values of $Q_{gs}$ and $Q_{gd}$ which have an orthogonal set of $V_{gs}$ and $V_{ds}$, can simply be written in CITI-file format to implement in ADS as a table-based model.

**7.6. Gate Current Modeling**

The gate currents $I_{gs}$ and $I_{gd}$ are determined by integration using intrinsic gate conductances $G_{gsf}$ and $G_{gdf}$ with the equations (7.7) and (7.8), respectfully. This extraction method is simpler than obtaining the gate currents from DC measurements. The calculated values of $I_{gs}$ and $I_{gd}$ as a function of the intrinsic voltages show intrinsic gate threshold voltage around the same value as extrinsic gate threshold voltage. The
values of \( I_{gs} \) and \( I_{gd} \) are then written in a CITI-file format to implement in ADS as a table-based model.

### 7.7. Drain Current Modeling

The intrinsic channel conductances \( G_m \) and \( G_{ds} \) cannot satisfy path-independence integral conditions for large dispersive devices. [1] This makes it difficult to derive an accurate model for the drain current, which accounts for the trapping and self-heating effects based on conventional s-parameter measurements. The optimal method is to derive the current model from pulsed IV measurements under appropriate quiescent bias conditions.

The drain current is modeled using the following equation

\[
I_{ds} (V_{gs}, V_{ds}) = I_{dc,iso}^{\text{DC}} (V_{gs}, V_{ds}) + \alpha_G (V_{gs}, V_{ds})(V_{gs} - V_{gso}) + \alpha_D (V_{gs}, V_{ds})(V_{ds} - V_{dso}) + \alpha_T (V_{gs}, V_{ds}) P_{diss}
\]

(7.13)

\( I_{dc,iso}^{\text{DC}} \) is an isothermal DC current after de-embedding the self-heating effect. \( \alpha_G \) and \( \alpha_D \) model the deviation in the drain current due to the surface trapping and buffer trapping, respectively and \( \alpha_T \) models the deviation in the drain current due to the self-heating effect. The amount of trapping induced current dispersion depends on the rate of dynamic change of the applied intrinsic voltages \( V_{gs} \) and \( V_{ds} \) with respect to those average values \( V_{gso} \) and \( V_{dso} \). This helps to understand how current dispersion is mainly stimulated with the RF or AC components of the gate-source and drain-source voltages, which is
described by \((V_{gs} - V_{gso})\) and \((V_{ds} - V_{dso})\) in equation \(I_{ds}(V_{gs}, V_{ds})\). The self-heating induced dispersion is caused mainly by the low frequency components of the drain signals. By this relation, \(P_{diss}\) in equation \(I_{ds}(V_{gs}, V_{ds})\) accounts for the static and quasi-static intrinsic power dissipation.

### 7.8. Drain Current Modeling for Trapping and Self-Heating Effects

Trapping effects in AlGaN/GaN HEMT are mainly related to the surface and buffer traps. These effects can be characterized by pulsed IV measurements at negligible device self-heating. [1] The surface trapping is characterized using pulsed IV measurements at two extrinsic quiescent biases equivalent to

\[
V_{GSO} < V_{P}, \ V_{DSO} = 0 \ \text{V} \ (P_{diss} \approx 0) \quad (7.14)
\]

\[
V_{GSO} = 0 \ \text{V}, \ V_{DSO} = 0 \ \text{V} \ (P_{diss} \approx 0) \quad (7.15)
\]

Under these two quiescent bias conditions, the drain current variation can be assumed to be strongly related to the surface trapping, because this effect is mainly stimulated by the gate voltage.

The buffer trapping is characterized using pulsed IVs at two quiescent biases equivalent to

\[
V_{GSO} < V_{P}, \ V_{DSO} = 0 \ \text{V} \ (P_{diss} \approx 0) \quad (7.16)
\]

\[
V_{GSO} < V_{P}, \ V_{DSO} >> 0 \ \text{V} \ (P_{diss} \approx 0) \quad (7.17)
\]
Under these two quiescent bias conditions, the drain current variation can be assumed to be strongly related to the buffer trapping, because this effect is mainly stimulated by the gate voltage.

The following figures, 7.6 – 7.7 show the pulse IV measurements for a 4x50 p15 gate width HEMT device plotting the effects of surface and buffer trapping.

Figure 7.7. Pulsed IV measurement showing surface trapping effects
Figure 7.8. Pulsed IV measurement showing buffer trapping effects

The 4x50 p15 is also biased at 20% $I_{\text{dss}}$ to account for the self-heating effect at high quiescent power dissipation. Figure 7.8 shows the pulsed IV characteristics at 20% $I_{\text{dss}}$. Using DC IV characteristics along with the pulsed IV characteristics help improve the self-heating characterization for HEMTs.
7.9. Drain Current Model Fitting Parameter Extraction

The drain current model equation in (7.12) has four unknowns $I_{\text{iso} \text{ DC}}, \alpha_G, \alpha_D,$ and $\alpha_T$. To determine the unknowns, the drain current model equation has to be applied to at least four pulsed IV characteristics at suitable quiescent bias conditions. Using MATLAB, the four independent linear equations are extracted from the pulsed IV data. At each state, the drain current can be assumed to be affected by at most one of the dispersion sources (surface trapping, buffer trapping, or self-heating). [1] Solving the four linear equations, corresponding to the four characteristics, at each bias point, the values of $I_{\text{iso} \text{ DC}}$, $\alpha_G$, $\alpha_D$, and $\alpha_T$ are determined. The following figures, 7.9 – 7.12, show the extracted values of the fitting parameters as a function of intrinsic bias voltages for 4x50 p15 HEMT.

Figure 7.9. Pulsed IV measurement showing self-heating effects
Figure 7.10. $I_{ds}$ for 4x50 p15 HEMT

Figure 7.11. $\alpha_G$ independent variable
7.10. Large-Signal Model Implemented with ADS

The derived nonlinear elements $Q_{gs}$, $I_{gs}$, $Q_{gd}$, and $I_{gd}$ are written in tables versus $V_{gs}$ and $V_{ds}$ in an output file for ADS simulation. The fitting bias-dependent parameters
The intrinsic resistances $R_{gd}$ and $R_i$ and transconductance time delay are defined as bias-dependent table-based elements in one output file. The files are written in a CITI-file format in order to input into ADS.

In [1], Jarndal discusses the model implementation. Figure 7.14 shows the model used to conduct the simulation in ADS. The model is a table-based large signal model.

The extrinsic bias-independent elements $C_{pgas}$, $C_{pda}$, $C_{gda}$, $L_g$, $L_d$, $L_s$, $C_{pgi}$, $C_{pdi}$, $C_{gdi}$, $R_g$, $R_d$, $R_s$ are represented by lumped passive elements. The intrinsic nonlinear part is represented by the ten-port Symbolically Defined Device (SDD) component. The first two ports of the SDD represent the gate current and charge sources. The third and fourth ports represent the intrinsic resistances $R_{gd}$, and $R_i$. Port 5 is used to implement the drain current. Port 6 is used to calculate the instantaneous dissipated power. Ports 7 and 8 represent the AC components of the gate and drain voltages, $(V_{gs} - V_{gso})$ and $(V_{ds} - V_{dso})$. Port 9 is connected to a low pass circuit to derive mean values. The transconductance time delay is modeled using a delayed gate voltage at port 10. The values of the nonlinear elements and the drain current fitting parameters are read from Data Access Components (DACs) as imported data files.
Figure 7.14 Large signal model used for ADS
8. Simulated versus experimental small signal model results

Using MATLAB, the small signal parameters were extracted from the s-parameter data. To extract the parameters, the method described in the small signal model extraction section was used to extract the parameters and compare the fitted data to the measured data. Figure 8.1 shows the fitted s-parameter data with the measured data for the 4x50 p15 HEMT. The error for the measured s-parameter magnitude error between the fitted and measured data is calculated at 1.70E-03 for the 4x50 p15.

![Figure 8.1. Comparison of measured data to fitted data](image)

The following spreadsheets give the extrinsic and intrinsic parameter values for each type of HEMT. From a quick observation, one can see that the values of the parameters scale along with the size of the transistor. The information obtained from the small signal model could be useful to determine the topology of the transistor necessary...
for use in power amplifier applications. The 4x85 p35 has a value for $R_g$ that does not scale accordingly as the HEMTs. The data for the 4x85 p35 shows the device has issues that could lead to problems if the transistor is used in a power amplifier application. To further investigate the transistor, I attempted source and load impedance matching on the device to conduct load pull measurements. The device would not provide a good match therefore screened out when conducted load-pull measurements.

Using the 22-element small signal model, the extrinsic and intrinsic parameters of the HEMTs were defined for each device. The data shows how the device topology of HEMTs could be scaled to meet specific power requirements. To further screen out the devices the pulsed IV measurements and Load-Pull measurements were conducted.
## Extrinsic Parameters

<table>
<thead>
<tr>
<th></th>
<th>4x50x15</th>
<th>4x50x25</th>
<th>4x50x35</th>
<th>4x65x15</th>
<th>4x65x25</th>
<th>4x65x35</th>
<th>4x85x15</th>
<th>4x85x25</th>
<th>4x85x35</th>
<th>4x100x15</th>
<th>4x100x25</th>
<th>4x100x35</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Extrinsic Parameters</strong></td>
<td>Values</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Cpga</td>
<td>0.09688 fF</td>
<td>0.10702 fF</td>
<td>0.11742 fF</td>
<td>0.11542 fF</td>
<td>0.12797 fF</td>
<td>0.14116 fF</td>
<td>0.14217 fF</td>
<td>0.15885 fF</td>
<td>0.17587 fF</td>
<td>0.1616 fF</td>
<td>0.18347 fF</td>
<td>0.20199 fF</td>
</tr>
<tr>
<td>Cpda</td>
<td>0.09688 fF</td>
<td>0.10702 fF</td>
<td>0.11742 fF</td>
<td>0.11542 fF</td>
<td>0.12797 fF</td>
<td>0.14116 fF</td>
<td>0.14217 fF</td>
<td>0.15885 fF</td>
<td>0.17587 fF</td>
<td>0.1616 fF</td>
<td>0.18347 fF</td>
<td>0.20199 fF</td>
</tr>
<tr>
<td>Cgda</td>
<td>0.0412 fF</td>
<td>0.04326 fF</td>
<td>0.045007 fF</td>
<td>0.051887 fF</td>
<td>0.053754 fF</td>
<td>0.05602 fF</td>
<td>0.066546 fF</td>
<td>0.068926 fF</td>
<td>0.072859 fF</td>
<td>0.07485 fF</td>
<td>0.081659 fF</td>
<td>0.083122 fF</td>
</tr>
<tr>
<td>Cpdı</td>
<td>0.291 fF</td>
<td>0.32106 fF</td>
<td>0.35227 fF</td>
<td>0.34626 fF</td>
<td>0.38392 fF</td>
<td>0.42347 fF</td>
<td>0.4265 fF</td>
<td>0.47654 fF</td>
<td>0.52761 fF</td>
<td>0.4848 fF</td>
<td>0.55041 fF</td>
<td>0.60597 fF</td>
</tr>
<tr>
<td>Cpgı</td>
<td>60.4 fF</td>
<td>59.915 fF</td>
<td>60.38 fF</td>
<td>70.115 fF</td>
<td>69.415 fF</td>
<td>69.945 fF</td>
<td>84.082 fF</td>
<td>82.547 fF</td>
<td>76.589 fF</td>
<td>94.084 fF</td>
<td>93.008 fF</td>
<td>92.334 fF</td>
</tr>
<tr>
<td>Cgdi</td>
<td>0.08247 fF</td>
<td>0.086532 fF</td>
<td>0.090014 fF</td>
<td>0.1377 fF</td>
<td>0.10751 fF</td>
<td>0.11204 fF</td>
<td>0.13309 fF</td>
<td>0.13785 fF</td>
<td>0.14572 fF</td>
<td>0.15497 fF</td>
<td>0.16332 fF</td>
<td>0.16624 fF</td>
</tr>
<tr>
<td>Lg</td>
<td>28.057 pH</td>
<td>27.338 pH</td>
<td>26.602 pH</td>
<td>31.518 pH</td>
<td>31.265 pH</td>
<td>31.62 pH</td>
<td>34.715 pH</td>
<td>34.181 pH</td>
<td>35.666 pH</td>
<td>35.727 pH</td>
<td>34.944 pH</td>
<td>33.925 pH</td>
</tr>
<tr>
<td>Rg</td>
<td>0.6433 Ω</td>
<td>0.6257 Ω</td>
<td>0.6493 Ω</td>
<td>1.1919 Ω</td>
<td>1.1331 Ω</td>
<td>1.1303 Ω</td>
<td>1.7103 Ω</td>
<td>1.7403 Ω</td>
<td>0.7642 Ω</td>
<td>2.1411 Ω</td>
<td>2.1902 Ω</td>
<td>2.1127 Ω</td>
</tr>
<tr>
<td>Rd</td>
<td>4.77Ω</td>
<td>4.5131 Ω</td>
<td>4.473 Ω</td>
<td>3.4973 Ω</td>
<td>3.2994 Ω</td>
<td>3.3733 Ω</td>
<td>2.5945 Ω</td>
<td>2.4715 Ω</td>
<td>1.6830 Ω</td>
<td>2.1808 Ω</td>
<td>1.9317 Ω</td>
<td>1.9183 Ω</td>
</tr>
<tr>
<td>Rs</td>
<td>4.2 Ω</td>
<td>4.2002 Ω</td>
<td>4.1863 Ω</td>
<td>3.4203 Ω</td>
<td>3.4135 Ω</td>
<td>3.3914 Ω</td>
<td>2.7864 Ω</td>
<td>2.7535 Ω</td>
<td>2.7026 Ω</td>
<td>2.455 Ω</td>
<td>2.4369 Ω</td>
<td>2.1127 Ω</td>
</tr>
</tbody>
</table>

Table 8.1. HEMT Extrinsic Parameters
### Intrinsic Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>4x50x15</th>
<th>4x50x25</th>
<th>4x50x35</th>
<th>4x65x15</th>
<th>4x65x25</th>
<th>4x65x35</th>
<th>4x85x15</th>
<th>4x85x25</th>
<th>4x85x35</th>
<th>4x100x15</th>
<th>4x100x25</th>
<th>4x100x35</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cgs</td>
<td>167.27 fF</td>
<td>170.13 fF</td>
<td>173.98 fF</td>
<td>218.5 fF</td>
<td>225.16 fF</td>
<td>230.73 fF</td>
<td>297.83 fF</td>
<td>303.55 fF</td>
<td>308.34 fF</td>
<td>360.72 fF</td>
<td>361.03 fF</td>
<td>369.35 fF</td>
</tr>
<tr>
<td>Cds</td>
<td>59.44 fF</td>
<td>68.58 fF</td>
<td>78.53 fF</td>
<td>72.048 fF</td>
<td>84.687 fF</td>
<td>97.669 fF</td>
<td>96.021 fF</td>
<td>113.68 fF</td>
<td>131.46 fF</td>
<td>119.12 fF</td>
<td>143.08 fF</td>
<td>163.43 fF</td>
</tr>
<tr>
<td>Cgd</td>
<td>29.51 fF</td>
<td>31.936 fF</td>
<td>33.729 fF</td>
<td>38.7881 fF</td>
<td>41.322 fF</td>
<td>43.699 fF</td>
<td>53.596 fF</td>
<td>56.346 fF</td>
<td>59.277 fF</td>
<td>66.255 fF</td>
<td>70.452 fF</td>
<td>72.899 fF</td>
</tr>
<tr>
<td>Rgd</td>
<td>1.2076e-12 Ω</td>
<td>1.171E-12 Ω</td>
<td>1.1614E-12 Ω</td>
<td>1.3394E-12 Ω</td>
<td>1.3237E-12 Ω</td>
<td>1.3273E-12 Ω</td>
<td>1.6743E-12 Ω</td>
<td>1.6527E-12 Ω</td>
<td>1.6463E-12 Ω</td>
<td>2.0153E-12 Ω</td>
<td>1.9777E-12 Ω</td>
<td>1.9593E-12 Ω</td>
</tr>
<tr>
<td>τ</td>
<td>1.708 ps</td>
<td>1.7588 ps</td>
<td>1.8059 ps</td>
<td>1.8481 ps</td>
<td>1.8873 ps</td>
<td>1.9824 ps</td>
<td>2.0329 ps</td>
<td>2.0858 ps</td>
<td>2.177 ps</td>
<td>2.134 ps</td>
<td>2.2433 ps</td>
<td>2.3153 ps</td>
</tr>
<tr>
<td>Gds</td>
<td>2.3 mS</td>
<td>2.4 mS</td>
<td>2.5 mS</td>
<td>2.4 mS</td>
<td>2.6 mS</td>
<td>2.8 mS</td>
<td>2.3 mS</td>
<td>2.5 mS</td>
<td>2.9 mS</td>
<td>2.0 mS</td>
<td>2.7 mS</td>
<td>2.9 mS</td>
</tr>
<tr>
<td>Fitting Error</td>
<td>9.96E-04</td>
<td>1.10E-03</td>
<td>1.10E-03</td>
<td>1.20E-03</td>
<td>1.20E-03</td>
<td>1.30E-03</td>
<td>1.40E-03</td>
<td>1.60E-03</td>
<td>1.60E-03</td>
<td>1.50E-03</td>
<td>1.50E-03</td>
<td>1.70E-03</td>
</tr>
<tr>
<td>Stability Factor Error</td>
<td>2.84E-05</td>
<td>9.76E-03</td>
<td>7.35E-05</td>
<td>8.61E-05</td>
<td>1.80E-04</td>
<td>1.62E-04</td>
<td>2.35E-04</td>
<td>2.88E-04</td>
<td>2.20E-04</td>
<td>3.05E-04</td>
<td>1.36E-04</td>
<td>4.10E-04</td>
</tr>
<tr>
<td>S-parameter Magnitude Error</td>
<td>1.70E-03</td>
<td>1.80E-03</td>
<td>1.90E-03</td>
<td>2.00E-03</td>
<td>2.10E-03</td>
<td>2.20E-03</td>
<td>2.30E-03</td>
<td>2.40E-03</td>
<td>2.60E-03</td>
<td>2.60E-03</td>
<td>2.70E-03</td>
<td>2.90E-03</td>
</tr>
</tbody>
</table>

Table 8.2. HEMT Intrinsic Parameters
9. Measured DC Response data

Each HEMT was testing at DC to screen out the best performing transistor. The follow chart gives the results of the DC characteristics for each device. The data obtained from the DC measurements along with s-parameter and Load-Pull measurements help to select the optimum device for X-band operation.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>4x50 p15</th>
<th>4x50 p25</th>
<th>4x50 p35</th>
<th>4x65 p15</th>
<th>4x65 p25</th>
<th>4x65 p35</th>
<th>4x85 p15</th>
<th>4x85 p25</th>
<th>4x85 p35</th>
<th>4x100 p15</th>
<th>4x100 p25</th>
<th>4x100 p35</th>
</tr>
</thead>
<tbody>
<tr>
<td>Breakdown Voltage (V)</td>
<td>18.51</td>
<td>18.07</td>
<td>18.6</td>
<td>12.1</td>
<td>17.8</td>
<td>18.46</td>
<td>16.07</td>
<td>17.1</td>
<td>17.88</td>
<td>15.8</td>
<td>15.67</td>
<td>17.32</td>
</tr>
<tr>
<td>Idss (mA/mm)</td>
<td>892</td>
<td>931</td>
<td>939</td>
<td>879</td>
<td>915</td>
<td>925</td>
<td>860</td>
<td>888</td>
<td>907</td>
<td>839</td>
<td>881</td>
<td>897</td>
</tr>
<tr>
<td>Imax (mA/mm)</td>
<td>1044</td>
<td>1088</td>
<td>1104</td>
<td>1025</td>
<td>1068</td>
<td>1086</td>
<td>1005</td>
<td>1039</td>
<td>1064</td>
<td>983</td>
<td>1029</td>
<td>1048</td>
</tr>
<tr>
<td>V_knee (V)</td>
<td>2.24</td>
<td>2.20</td>
<td>2.22</td>
<td>2.19</td>
<td>2.17</td>
<td>2.20</td>
<td>2.15</td>
<td>2.14</td>
<td>2.23</td>
<td>2.12</td>
<td>2.24</td>
<td>2.17</td>
</tr>
</tbody>
</table>

Table 9.1. Measured DC response for HEMTs

The following IV plots, Figures 9.1 – 9.2, show how two of the HEMTs react to DC voltage. The 4x50 p15 has the lowest current response and the 4x100 p35 has one of the highest current responses when biased with DC voltage. The IV curves also show the self-heating effect on transistors. The self-heating effect is addressed in the large signal model section of this report.
Figure 9.1. 4x50 p15 IV measurements

Figure 9.2. 4x100 p35 IV measurements
Each HEMT was tested at RF to screen out the best performing transistor. The following chart gives the results of the RF characteristics for each device.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>4x50</th>
<th>4x50</th>
<th>4x65</th>
<th>4x65</th>
<th>4x85</th>
<th>4x85</th>
<th>4x100</th>
<th>4x100</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gmp (mS/mm)</td>
<td>330</td>
<td>339</td>
<td>341</td>
<td>331</td>
<td>339</td>
<td>341</td>
<td>328</td>
<td>337</td>
</tr>
<tr>
<td>Fmax @ Gmp (GHz)</td>
<td>114</td>
<td>111</td>
<td>109</td>
<td>117</td>
<td>115</td>
<td>114</td>
<td>118</td>
<td>117</td>
</tr>
<tr>
<td>Ft @ Gmp (GHz)</td>
<td>65</td>
<td>65</td>
<td>66</td>
<td>69</td>
<td>69</td>
<td>72</td>
<td>73</td>
<td>74</td>
</tr>
<tr>
<td>h21 @ 1 @ Gmp (Ratio)</td>
<td>63</td>
<td>63</td>
<td>63</td>
<td>67</td>
<td>67</td>
<td>70</td>
<td>71</td>
<td>72</td>
</tr>
</tbody>
</table>

Table 10.1. Measured RF Response

The following smith plots, Figures 10.1 – 10.2 show how two of the HEMTs react to RF signal at from 0 – 26 GHz. The 4x50 p15 has the lowest s-parameter response and the 4x100 p35 has a high s-parameter response. Both s-parameter results are scaled in order to compare all s-parameters measurements.
Figure 10.1. Comparison of measured data to fitted data for 4x50 p 15 HEMT

Figure 10.2. Comparison of measured data to fitted data for 4x100 p35 HEMT
11. Measured Pulsed IV Response Data

Each HEMT was testing for pulsed IV response to screen out the best performing transistor. The pulsed IV response helps understand devices behavior. The self-heating current dispersion and trapping induced current dispersion are investigated with pulsed IV data. The follow figures show the results of the pulsed IV testing for the 4x50 p15 device.

Figure 11.1. Pulsed IV results for 4x50 p15 HEMT showing surface traps response
Figure 11.2. Pulsed IV results for 4x50 p15 HEMT showing buffer traps response
12. Measured Load-Pull Response Data

Each HEMT was testing for load-pull response to screen out the best performing transistor. The load-pull response helps understand device behavior at high power and high frequency. Table 12.1 shows how each transistor performed at Class AB operation using the Maury Load-Pull system. The testing was conditioned to have maximum power out for each transistor. The 4x65 p15 was tested for maximum power out at $V_{DS} = 20$ V.

Figures 12.1 and 12.2 show $P_{out}$ and power added efficiency (PAE) for two devices.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>4x50 p15</th>
<th>4x50 p25</th>
<th>4x50 p35</th>
<th>4x65 p15</th>
<th>4x65 p25</th>
<th>4x65 p35</th>
<th>4x85 p15</th>
<th>4x85 p25</th>
<th>4x85 p35</th>
<th>4x100 p15</th>
<th>4x100 p25</th>
<th>4x100 p35</th>
</tr>
</thead>
<tbody>
<tr>
<td>PAE %  (max)</td>
<td>46</td>
<td>46</td>
<td>45</td>
<td>58</td>
<td>46</td>
<td>32</td>
<td>48</td>
<td>48</td>
<td>49.2</td>
<td>48</td>
<td>48</td>
<td>49.2</td>
</tr>
<tr>
<td>$P_{in}$ (max) (dBm)</td>
<td>17</td>
<td>19</td>
<td>18.9</td>
<td>20.2</td>
<td>21.9</td>
<td>22.1</td>
<td>17.4</td>
<td>18</td>
<td>20.1</td>
<td>19.4</td>
<td>20</td>
<td>19.6</td>
</tr>
<tr>
<td>$P_{out}$ (max) (dBm)</td>
<td>27</td>
<td>27</td>
<td>27.3</td>
<td>29.8</td>
<td>27.2</td>
<td>27.3</td>
<td>29.3</td>
<td>27.9</td>
<td>27.6</td>
<td>29</td>
<td>30</td>
<td>29.7</td>
</tr>
</tbody>
</table>

Table 12.1. Large Signal performance parameters

![Figure 12.1. Plot of $P_{out}$ and PAE for 4x50 p15 HEMT](image)
Figure 12.2. Plot of $P_{\text{out}}$ and PAE for 4x100 p25 HEMT
13. Selection of optimum HEMT at X-Band

Taking into account all the data and simulation for the twelve HEMTs. The selection of the optimum HEMT has to be based on application. The data and simulation results show that as the width of the HEMT increases the available power handling capability increases. The DC and pulsed IV response of the HEMTs show the link between device topology and increased power capability.

The pulsed IV measurements reveal the available load line for each transistor. Using figures 11.1. and 11.2., the plots indicate the increase in the saturation current as the topology of the device increases. The increased load line help RF power amplifiers designers achieve higher performance from the AlGaN/GaN HEMTs.

The s-parameter data and small signal modeling assist in choosing the optimum HEMT by giving researcher the ability to see how various topologies affect the frequency response of the transistor. Figures 10.1. and 10.2. show how the topology changes the magnitude and phase of the HEMTs response to input signals.
14. Future Considerations

Commercial and military applications have a high demand for wide-bandgap HEMTs. Improvements in HEMT structures and modeling will help to increase the understanding of HEMTs. Research on the HEMT structures provides an avenue to investigate failure analysis and reliability. Model improvements assist in bottom up research by examining physical failure analysis. Increased simulation capabilities provide avenues to optimize epitaxial structure, ohmic contact, schottky gates, and topology based on power requirements or required circuit response.
15. Conclusion

Modeling of AlGaN/GaN HEMTs is a research area gaining momentum for application in understanding device behavior. Across the world researchers are streamlining methods to extract extrinsic and intrinsic parameters based on topology and material composition. A usable small and large signal model procedure for AlGaN/GaN HEMTs that can be utilized for a range of device types has been produced for the 22-element model with excellent results.

Evaluation of the performance characteristics for HEMTs gives researchers a better understanding of the current performance and reliability limitations for AlGaN/GaN based technology. The 22-element model permits high-power amplifier design success and provides an initial analysis potential for a material, device, and performance optimization process. This enables an engineering environment where a device designer and RF model/circuit designer can work collaboratively to create reliable device models that can be used to optimize devices for power performance, down to the material level. Having the ability to design and optimize HEMTs, provides engineers cost effective and time efficient advanced designs in the R&D workplace.

The 22-element model reproduces DC, small-signal RF, s-parameter, and large-signal RF for AlGaN/GaN device characteristics accurately. MATLAB, CADENCE and ADS were very useful software applications for efficient model parameter extraction and simulations.
List of References


