GaN HEMT and MMIC Design and Evaluation

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Abstract

Gallium Nitride based devices due to their inherent material properties are considered as one of the most promising devices to realize high power, high frequency transistors with lower power consumption in next-generation applications. Although the technology has been studied since early 1970s, there is still a vast room and expectations in its yet unachieved findings. In present work the GaN technology is explored and state-of-the-art studies of GaN based HEMTs and their application in MMICs are presented. Different designs are presented and evaluated and the results are reported. In particular the HEMT performance is studied in terms of DC in addition to large signal conditions, where the device’s performance becomes function of power levels it is driven with. The peculiarities and challenges of building an automated Load-Pull setup are outlined and analysis for further improvements is presented.
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1. Introduction

Gallium-Nitride (GaN) is a very hard and mechanically stable material with large heat capacity [1]. It is a direct wide-bandgap, compound semiconductor and due to its physical properties is considered as one of the most promising candidates for building high-power and high-frequency active devices. Supported with its inherent material characteristics, devices based on this technology can deliver considerably superior performance as compared to broadly employed silicon or III-V\(^1\) solutions. The key advantages of GaN technology can be outlined as follows:

- High Power density capability that is due to higher charges (i.e. than GaAs) which result in higher current (1 to 1.4 A/mm). Accordingly, reported power densities are with an order higher than that of GaAs based devices [2].
- High breakdown voltage, typically >100V, caused by high breakdown field [2]
- Low Noise [3]
- High Temperature, that is due to low minor carrier density as well as high thermal conductivity [4].
- High tolerance to radiation by reason of high carrier generation energy and small lattice constant [4]
- High speed due to high carrier velocity [4]

The advantage of GaN technology over the challengers is depicted in a diagram on Figure 1.1, where the key characteristics, due to which the semiconductors are valued, are compared.

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\(^1\) III-V (or 13-15) semiconductors are composed of elements from group 13 (B, Al, Ga, In) and from group 15 (N, P, As, Sb, Bi) of the periodic table [47]
As the demand for high speed data communication becomes severely higher, it as a result implies increased power consumption in base stations as well as consumer premise equipments. Along with other characteristics, high break down voltage of the compound allows different biases to be used and this on the other hand gives better performance from the power handling capabilities point of view, so crediting to outstanding high microwave-output-power density features, GaN based devices are considered as one of the most promising devices to realize high power amplifiers with lower power consumption in next-generation high-speed wireless communication systems [5]. Figure 1.2 illustrates power versus frequency capabilities of GaN technology along with other broadly employed semiconductor materials.
GaN’s advantage in this field of high frequency operation is also plain. For high frequency operation, saturated drift velocity is most important factor since the velocity of carriers in the high electric field region is governed by $V_{\text{sat}}$. From recently published experimental results, the bulk GaN has saturation drift velocity of $1.4 \times 10^7 \text{ cm/s}$ \cite{6} whereas for Si it is $1 \times 10^7 \text{ cm/s}$ and for GaAs only $0.74 \times 10^7 \text{ cm/s}$ \cite{7}. As can be seen from figure 1.3, AlGaN/GaN heterostructure has even higher saturation velocity (about $3 \times 10^7 \text{ cm/s}$) \cite{8}.

[Figure 1.2 Power and frequency regions for GaN and other materials]

[Figure 1.3 Saturation drift velocity versus electric field]
The GaN technology is principally interesting for applications where the requirement for withstanding of harsh usage conditions like high temperature or high drain voltages are required, but the main expectations come from its operation in high power and frequency applications. As depicted on figure 1.2, GaN seems to be a valuable tradeoff between these two parameters.

Based on outlined features, gallium nitride based devices may not only deservedly substitute incumbent technologies, but enable a vast series of new applications that are simply unachievable with current solutions [4].

Though the research of GaN technology originates in early 1970s, it’s commercial applications have been delayed due to mainly high costs related to the process as well as material itself. The explosive increase in interest in the AlGaN, and in particular AlGaN family of materials in recent years has been fueled by the application of blue/green/UV light-emitting diodes (LEDs) in full-color displays, traffic lights, automotive lighting and general room lighting using so-called white LEDs [9].

Due to [10] major applications that will be addressed in a near future for GaN-based materials and devices are:

- UV optical sources capable of operation down to 280 nm for use in airborne chemical and biological sensing systems, allowing direct multi-wavelength spectroscopic identification and monitoring of UV-induced reactions.
- Power amplifiers and monolithic microwave integrated circuits (MMICs) for use in high performance radar units and wireless broadband communication links and ultra high power (>1 MW) switches for control of distribution on electricity grid networks.
- Room temperature, ferromagnetic semiconductors for use in electrically-controlled magnetic sensors and actuators, high density, ultra-low power memory and logic, spin-polarized light emitters for optical encoding, advanced optical switches and modulators and devices with integrated magnetic, electronic and optical functionality.

As outlined above, due to material properties such as wide energy bandgap, high two dimensional electron gas (2-DEG) density and high mobility, GaN technology is one of the most
promising semiconductor competing with conventional technologies (e.g. Si, GaAs). The applications of this technology show potential in low noise applications due to low noise figure recently reported as small as 0.7 dB at 10 GHz frequency [11] as well as it’s possible feasibility for high frequency DC/DC converters [12]. Recent reports qualified GaN based devices as promising candidates for high power switches [13] demonstrating maximum output power and upconversion gain of 19.6 dBm/11 dB at 2 GHz and 13 dBm/5 dB at 5 GHz frequencies and attenuators [14], demonstrating broadband operation up to 18 GHz with dynamic range more than 30 dB, thus showing the comparable results to InP- and GaAs-based integrated circuits but provided considerably higher power handling capabilities (15 W/mm). Though the possible applications are much broader then high power performance, the major studies in this thesis addressed the power performance of GaN technology and GaN based HEMTs applications. Considerable effort was put on large signal analysis studies and fully automated Load Pull setup was assembled. For this purpose, the theory behind Large Signal Network Analysis was intensely considered and all the components needed for large signal measurements were brought together. The setup was designed for one tone signal at 6 GHz frequency, while the room for its upgrade for intermodulation measurements was considered.

Chapter 2 - Theoretical Background, describes underling theory which in fact is a review of GaN technology in conjunction with AlGaN/GaN based HEMT’s operation and application in microwave monolithic integrated circuits, employing different measurement techniques such as DC characterization, high frequency small signal and large signal analysis. Also the theory following large signal network analysis is given. As a key component in large signal operation, microwave computer controlled tuner performance and operation is described. Several designs and their impact on overall performance have been investigated and the measurement results are presented in terms of DC as well as large signal conditions in chapter 3 – Results. Summary of the work performed, suggestions for further improvement of the load pull setup and future trend survey conclude the work in Chapter 4 - Discussion and Conclusion.
2. Theoretical Background

GaN Based HEMT Description

Most of the reported state of the art results of GaN-based semiconductor devices are from aluminum-gallium-nitride/gallium-nitride (AlGaN/GaN) High Electron Mobility Transistors, HEMTs\(^2\). The first demonstration of AlGaN/GaN HEMT was made in 1993 by Khan [15]. It took over 10 years of intensive research and developments and as a consequence, recent GaN based HEMTs results demonstrated a very high cut off frequency \(f_T = 181\text{GHz}\) [16], as well as very high power density of 30 W/mm at 4 GHz [17] and 10W/mm at 40 GHz [18]. These power results demonstrate more than 3 times higher power density than state of the art aluminum-gallium-arsenide/gallium-arsenide (AlGaAs/GaAs) HEMTs [8]. Figure 2.1 presents a rate of progress in power performance of AlGaN/GaN HEMTs versus time, which is considered as the main reason of increased interest in this technology for military as well as commercial applications in recent years.

\[\text{Figure 2.1 Historical progress in GaN transistor technology [34]}\]

\(^2\) High Electron Mobility Transistors (HEMTs), also often referred to as Heterojunction Field Effect Transistors (HFETs), Modulation Doped Field Effect Transistors (MODFETs), Two-dimensional Electron Gas FET (TEGFET) and Selectively Doped Heterojunction Transistor (SDHT)
The key matter in HEMTs operation is the heterojunction interface of two semiconductor materials with different bandgap. Silicon-doped aluminum gallium nitride (AlGaN) is grown on top of GaN. AlGaN has an even higher energy gap than GaN. The silicon impurities donate electrons to the crystal that then tend to accumulate in the regions of lowest potential – known as a quantum well – just beneath the AlGaN/GaN interface. This forms a sheet of electrons, which constitutes a two-dimensional electron gas (2DEG). Here, the electrons gain higher mobility because they are physically separated from the ionized silicon donor atoms residing in the AlGaN. Ohmic contacts for source, gate and drain are introduced using photolithography. The modulation-doped heterostructure described thus far is fairly standard in other semiconductor systems. The 2DEG can be contacted with source and drain metals and the depletion region modulated with a gate contact to realize transistor action [19]. Figure 2.2 depicts the general structure of AlGaN/GaN HEMT.

**Polarization**

The general structure of AlGaN/GaN HEMT is depicted on figure 2.1. This structure is same as conventional HEMT. The 2DEG is of crucial importance in heterojunction interface. The origin of 2DEG in AlGaN/GaN heterostructure is spontaneous and piezoelectric polarization. Due to these two individual polarization mechanisms, even without intensive doping a sheet carrier
concentration in 2DEG of 10^{13}\text{cm}^{-2} or higher can be achieved [8]. Spontaneous polarization is polarization on heterojunction interface at zero strain\(^3\), which results from net charge of the growth front. Piezoelectric polarization results from the difference between lattice constants at the heterostructure, and thus increases as the strain at the interface increases. These two polarization effects are quite important in nitride-based HEMTs. The tensile strain cause by the growth of Al\(_x\)Ga\(_{1-x}\)N on GaN results in a piezoelectric polarization \(P_{pz}\) that adds to the spontaneous polarization \(P_{sp}\) and results into net polarization \(P(x)\), that is due to [20] given by

\[
P(x) = P_{pz} + P_{sp} = -(3 \cdot 2x - 1 \cdot 9x^2) \times 10^{-6} - 5 \cdot 2 \times 10^{-6}x \text{ Ccm}^{-2}
\]

This results in a net positive charge at that AlGaN/GaN interface, as depicted on figure 2.3

Figure 2.3 Net positive charge at AlGaN/GaN interface caused by the sum of the net spontaneous and piezoelectric polarization between AlGaN and GaN

where \(Q_{\pi}\) includes contribution from spontaneous as well as piezoelectric polarizations.

Device structure of the AlGaN/GaN HEMTs with the locations of 2DEG and polarization charge is depicted on figure 2.2.

In the growth of GaN materials and the design of AlGaN/GaN HEMT devices research was done in the last years at High Frequency Electronics Department of the Technical University of Darmstadt. HEMTs that are addressed in this thesis are fully in-house fabricated.

\(^3\) Strain is phenomena occurring at the interface due to different lattice constants of materials at the interface.
Al$_{0.25}$Ga$_{0.75}$N/GaN HEMT structure is grown on sapphire by low-pressure metal-organic chemical vapor deposition (MOCVD). The HEMT structures characterized throughout this project have 1μm gate length and different gate widths (50, 75, 100, 125, 150 and 200μm).

**DC Characteristics**

DC characterization of HEMT (any FET) is one of the key parameters to check device’s performance. Three main parameters and their interdependence are usually verified while this kind of measurements. As a rule, HEMT performance is checked for two type of DC measurements, that are I-V and transfer characteristics of device. A regular I-V plot represents drain – source ($I_{DS}$) current as a function of applied gate – source ($V_{GS}$) and drain – source voltages ($V_{DS}$). Depicted on figure 2.4 typical I-V characteristics of a FET are given.

![Figure 2.4 Typical I-V characteristics of FET](image)

Illustrated on figure 2.4, IMAX (Imax) indicates the maximum possible achievable drain – source current, beyond which, if current – limiting resistor is not used, device will be destroyed. IDSS (IDSAT) stands for saturated drain – source current when the voltage over gate – source is 0 volts. Pinch-off voltage, VPO (VPO) is a gate – drain voltage, where the terminals of drain –
source act as open circuit and no current flows though there is high drain – source voltage. VBR (VBR) is a gate – drain breakdown voltage which is indirectly measured as the difference between drain – source and gate – source voltages and causes the unavoidable damage of the device [21].

Another interesting characterization of HEMT in terms of DC is its transfer characteristic that is drain – source current as a function of gate – source voltage and is depicted on Figure 2.5.

A further interesting parameter that can be extracted from DC characterization is transconductance (gm) which is defined in fact as derivative of transfer characteristics and is depicted on Figure 2.6.
High Frequency Characteristics

As was illustrated in previous chapter, DC characterization can be used for quite a significant number of useful parameters revealing when it comes to transistor characterization. But the concept of DC measurements makes sense as long as we are dealing with low frequency signals. As soon as the frequency goes higher, and consequently the reduced wavelength of a signal becomes comparable to the device dimensions, other techniques, called Small Signal and Large Signal Network analysis come to face. In this chapter both of this techniques are described, however, the main emphasize is put on Large Signal Network Analysis

Small Signal Network Analysis

As it is pointed out in the technique name, small signal network analysis is device characterization technique where the measurements are done under small signal power conditions. Under small signal are considered signals that are small enough not to drive the
device into saturation and thus have only liner affect on the device [22]. Scattering parameters (S - parameters) are all what small signal network analysis is about. The fundamental goal of these measurements is presenting the effects of RF energy flow through device4. For any two-port device (i.e. GaN HEMT based amplifier evaluated in scope of current thesis), the scattering matrix consists of four S parameters ($S_{11}$, $S_{21}$, $S_{12}$, $S_{22}$), fully describing device performance. Additional details can be found in [23].

*Large Signal network Analysis*

Although the S-parameters seem to be ideal for network analysis, their application is limited to Small Signal network analysis [24]. These kinds of measurements are reliable and give a complete view of the component performance while the superposition theorem holds meaning that device is operating in linear region of operation. Generally small signal analysis is absolutely ideal for passive components characterization, which are on the whole purely linear. But whenever it comes to active components (i.e. amplifiers) the linearity conditions cannot be preserved longer and the output of the device performance becomes a function of power level it is driven with. For this reason, another technique, large signal network analysis and in particular, Load/Source Pull tuning is used for putting device under realistic large-signal operation conditions and its complete and accurate characterization [25].

Under term Load or Source pull, technique that permits to measure the microwave behavior of a test device under variable load or source impedance conditions is meant. Two major applications of this technique are power measurements using Load Pull and noise measurements using Source Pull, though special interest throughout this work is placed on power measurements, consequently the peculiarities of Load Pull measurement technique will be discussed further.

---

4 Device should not necessarily be two-port.
There are two possible techniques of performing load pull measurements, active and passive load pull. Active Load Pull is measurement technique that permits to generate a "virtual load" to the DUT by injecting in the output terminals a signal synchronized with the signal flowing thru the DUT and not by reflecting it on a passive load as is the case with Passive Load Pull. Both methods certainly have their advantages and disadvantages. As main benefit of Active Load Pull system may be outlined ability to generate close to 1 reflection coefficient at the device level. This feature can be compensated with passive Load Pull through introducing impedance transformers. Usage of transformers will cause the limitation of instantaneous bandwidth of the setup but generates very high reflection factors and improves measurement accuracy. This technique however can only be applied while test fixture measurements and not for "on wafer" tests. In the case of "on-wafer" the only alternative is to lower the loss of the probes and the cables to the tuners.

The typical and of most interest parameters measured under large signal conditions are power added efficiency – PAE, frequency response – GAIN and $P_{outvsP_{in}}$ - 1dB compression point. Figure 2.7 depicts all three values.

![Figure 2.7 Large-signal for an AlN/GaN MISFET [19]](image-url)
Illustrated on the figure 2.8 are the simulated power contours for class-A GaN HEMT based C-band power amplifier, developed in scope of [19]. The contours deviate from circles because the simulations are done under large signal (consequently non linear) conditions.

![Power contours for the 6 GHz amplifier design simulated by Source Pull](image)

*Figure 2.8 Power contours for the 6 GHz amplifier design simulated by Source Pull [19]*

**Load Pull Setup Description**

One of the tasks addressed in scope of current thesis project was establishing a large signal measurement setup for testing in-house fabricated HEMTs and MMICs. The setup was the only lacking component in a series of facilities available at the department of High Frequency Electronics for full evaluation of devices (HEMTs, MISFETs) and MMICs starting from material growth, processing and modeling level up to DC and RF characterization. Figure 2.9 depicts the
The measurement system is based on the computer controlled microwave tuners (CCMT) made by FOCUS MICROWAVES Inc and is automatically controlled by the software package WinCCMT specially developed by FOCUS MICROWAVES Inc for load/source-pull measurements. It can operate two tuners simultaneously and uses various software applications to control the tuners allowing them to execute measurements at different reference planes, setup corrections and calculations. The output power and the available input power at the transistor ports are calculated using the reading from power meter and accounting for the available losses of the input network (coupler, isolator, input bias tee), the real losses of the output network (output bias tee in this case) as well as two probing sections at the device input and output. The latter consists of a semirigid coaxial cables and probes. GPIB\(^5\) interface is adopted for hardware control.

\(^5\) Sometimes referred as HPIB in the documentation of Hewlett-Packard
It should be outlined that described setup was designed for one tone signal measurements at 6 GHz frequency. The possibility of further applications of the setup for intermodulation (two tone) measurements was taken into account so that all passive components were chosen with the broad enough dynamic range to cover the frequency range up to 18 GHz. For details on hardware models and specifications please refer to the appendix A. The final view of the setup is illustrated on figure 2.10

![Figure 2.10 The final view of load-pull setup](image)

**Tuner Performance**

As already mentioned, computer controlled microwave tuners from Focus Microwaves, model 1816, are used for load pull setup. This model can operate in the frequency range 1.6 – 18 GHz. Tuners are controlled with sophisticated software by means of which, operator can cover the Smith Chart with 1,800,000 points at 1.6 GHz and 1,048,000 points at 18 GHz frequency. Tuners
“consist of two precise hardened and grounded parallel shafts accurately guiding a carriage controlled by a center lead screw with trapezoidal thread as illustrated on figure 2.11. The carriage slides on Teflon-coated bearings that reduce friction and eliminate noise. An extremely precise stainless steel vertical axis with a 1.5 to 3µm resolution per motor step is located inside each carriage and is used to control the vertical position of each probe” [26].

Tuners are calibrated for particular number of points (per frequency) and equidistantly distributed over the smith chart. While tuning to some particular position on the Smith Chart, a second order Lagrange interpolation algorithm is applied to 9 neighboring calibrated points to estimate the S – parameters of the tuned point [27]. For detailed description of the interpolation algorithm functionality, please refer to appendix B.
Once the components were purchased and connected, the whole setup was calibrated and further verified using recommended techniques from the supplier. The details on calibration procedure can be seen in Appendix C and verification process is outlined in Appendix D.

**Microwave Monolithic Integrated Circuits**

As already outlined, Microwave Monolithic Integrated Circuits (MMICs) and in particular power amplifiers are expected to be the major area of dominance for GaN technology in a near future. AlGaN/GaN HEMTs are promising devices for high power and high frequency applications. This is due to the superior GaN electronic material properties such as large bandgap, high breakdown field, high peak and saturation carrier velocity and good thermal conductivity. Besides their attractive power features, AlGaN/GaN HEMTs are very promising for low noise applications, because they combine low noise figure performance with high breakdown voltage characteristics \[3\]. AlGaN/GaN HEMTs have been recently explored for use in robust hybrid low noise amplifier circuits \[28\], operating in X-band \[29\] and characteristics over broad band of frequency (3-18GHz) have been reported \[30\].

In both, power as well as low noise amplifiers, the device (transistor) is positioned as depicted on figure 2.12

![Figure 2.12 A general transistor amplifier schematic](image)
Depending on the application, the source and load matching networks have to be designed to provide the proper impedance matching for the device. For noise applications the source matching of the device is of key importance. AlGaN/GaN HEMT devices are the preferred choice in LNA front-ends due to the capability of withstanding high power levels without RF limiting circuitry, which often degrades the noise performance of communication systems [31].

Very high power density of 30 W/mm at 4 GHz [17] and 10W/mm at 40 GHz [18] has already been reported. The outstanding power performance of AlGaN/GaN devices is a result of the high electric breakdown field in wide bandgap nitride semiconductors (AlGaN), high electron saturation velocity in GaN, and relatively good thermal conductivity of sapphire, though the thermal conductivity is much more improved in the case of growth on SiC substrate. GaN HFET based MMIC’s are a noticeable competitor for GaAs for compact microwave applications with high power levels. They can compete with GaAs MMIC’s in high-bandwidth, high power applications, because the smaller size of the device makes it not only easier to fabricate, but also offers much higher impedance making it much easier to match them to the system (e.g. matching ratio of 10 times larger is needed for GaAs HEMT) while maintain the bread band operation [32].

Along with study of GaN technology in HEMT, its applications in actual microwave monolithic integrate circuits (MMICs) was addressed. Load pull setup which was assembled in scope of present thesis and has been described above was as well dedicated for measurements of MMICs based on GaN technology. Several different circuits were designed during last year at our department [19], among which are C-band oscillator (4 GHz), voltage-controlled attenuator, frequency doubler from 6 to 12 GHz, class-A power amplifiers for L-and (1 GHz) and C-bands (6 GHz). C-band Class-A power amplifier was of primary concern in scope of current thesis. For this reason the setup described further was designed for 6 GHz fundamental tone and a room for further enhancements of the setup for intermodulation measurements was taken into account.
HEMT based Power Amplifiers

In general, classification of power amplifiers is due to conduction angle\(^6\) and is defined as class A, AB, B and C for analog designs and class D and E for switching designs [33]. Conduction angle defines the portion of the input signal that is used for amplification. Amplifier class of operation is defined by operational point of gate (bias point). Different class of operations and their bias points for a typical power are shown on Figure 2.13

![Figure 2.13 Class of operation for a typical power amplifier](image)

Power added efficiency (PAE) of amplifiers differs correspondingly to their classes. Output signal of class-A amplifier is an exact replica of an input signal without clipping (angle of flow \(\theta=360^\circ\)) but otherwise it’s the most linear one. The conduction angle 360 degrees is achieved through biasing the amplifier with half of its gate and drain voltages, reaching its theoretical maximum PAE of 50% [19].

As already outlined, the design approach of GaN-based power amplifiers significantly differs from conventional GaAs-based circuits in a sense that both the input and output impedance transformation ratios are drastically lower for given output power. E.g. AlGaN/GaN HEMT with 0.6 µm gate length, breakdown voltage of 80 V and maximum drain current of 1 A/mm has 75 Ωmm output load that is two times more than that of GaAs-based HEMT showing the same input capacitance of 2.7 pF/mm [34]. As a consequence, since for the GaN-based HEMT offers

\(^6\) Sometimes also referred as Angle of Flow [33]
ten times the power density for the same output power, the input transformation ratio is ten
times less, while the output is twenty times less than that of GaAs HEMT thus making this
technology preferable for power amplifiers, especially for high frequency operations, where the
simplicity of impedance transformation translates to circuit matching simplicity.

Another important feature of GaN based HEMTs as compared to conventional ones is high
operation voltage of these devices due to wide band gap and high breakdown voltage. In
commercial systems such as wireless base stations operating at 28 Volts [34], use of GaN
HEMTs avoids need for voltage step down converters, since they can easily operate at 28 V and
potentially up to 42 Volts. Figure 2.14 summarizes the advantages of GaN technology’s power
performance.

As previously stated, AlGaN/GaN HEMT based MMICs were designed at the department of HFE
at TU Darmstadt. The process of MMIC fabrication was conducted along with current studies.
The circuits are on their final stage of fabrication (air bridge) and will be characterized using
setup in the outlook. The amplifier to be measured is non-offset 7 GaN HEMT based C-band
amplifier with cutoff frequency of 10 GHz, distributed elements for matching network. The

---

7 The distances between gate–source and gate-drain are equal [8].
amplifier gain reaches 5 dB. The final chip size is 4.08x2.99 mm². The layout of amplifier is illustrated on figure 2.15.

Figure 2.15 Layout of the 6 GHz power amplifier [19]
3. Results

DC Characterization and Different Design Effects

The measurement results of four different designs, such as Gate offset, Passivation effect, Field-plate Gate and different gate width have been studied and their effects are outlined in the following subchapters.

Gate Offset

The impact of offset versus non-offset design on the overall performance was studied. In offset devices, as depicted on figure 3.1, the distance between gate and drain is different (three times in this case) from the distance between source and gate. AlGaN/GaN HEMTs normally operate under high electric field and the electric field has its peak value between the gate and drain ohmic contacts.

Therefore if the distance between gate and drain increases, the on-breakdown voltage is expected to be increased due to the decreased field strength at the same drain bias. However, due to the increase of gate to drain length, the parasitic inductance and resistance increase and this leads in DC and RF performance degradation. As in the case with passivation, the designs
were identical with exception for the offset distance. Their DC characteristics are shown in Fig. 3.2 and 3.3 [8].

Passivation

Another study addressed the passivation effect. Passivation is the process of making a material "passive" in relation to another material prior to using the materials together [35]. Thin layer of Si$_3$N$_4$ was depositing on the top of AlGaN barrier to passivate the transistor during the process prior to the final, aribridge process. The rest of the process steps were performed identically so
that while characterization, the Passivation effect could be studied. Figure 3.4 shows the $I_{DS}$-$V_{GS}$ curve and the transconductance ($g_m$) of the passivated and non-passivated devices. The threshold voltage ($V_{th}$) was not affected and was found to be around -2V.

The maximum saturation drain current $I_{DS}$ at $V_{GS} = 2$ V was much higher for the passivated devices. The maximum $I_{DS}$ for the non-passivated devices was only $\sim 370$ mA/mm but for passivated device was $\sim 430$ mA/mm, thus maximum drain current increase of 16 % by passivation was observed. Several publications, addressing the same (passivation effect) issue were studied and the increase in maximum drain current was estimated to be comparable, as
for example Al$_{0.28}$Ga$_{0.72}$N HEMT, presented in [36], was grown on SiC substrate and 22% of increase in maximum current was observed for undoped system. As expected, due to the studies performed at High Frequency Electronics Department of TU Darmstadt and due to studies conducted and presented in [36] the surface charge reduction by passivation that causes increases the 2DEG density and thus current density in the channel was considered as main reason for increment in maximum drain current.

**Field-plate Gate**

Field-plate gate topology is an efficient technique used to reduce the high electric field in the drain access region (region between gate and drain) and was efficiently employed at recent research carried on at our department. As described in [37], the field-plate topology of gate implies an design, where drain-side edge of the metal gate overlaps on a high breakdown and high dielectric constant dielectric. The overlapping structure reduces the electric field at the drain-side gate edge, thus increasing the breakdown of the device. As can be seen from figure 3.6, the current density increase of more than 100mA/mm was observed.

![Figure 3.6 Maximum output current with and without field-plated gate vs different gate width](image-url)
Gate Width Effect

The effect of a varied gate width was examined using DC characterization technique available at the load-pull setup, assembled as a part of current thesis. HEMTs with 6 different channel widths were measured. The channel width varied from 50µm for the narrowest one, through 75, 100, 125, 150 µm up to 200µm. All of them were characterized and evaluated. The software, used for controlling the whole setup gave an opportunity to measure input as well as output I-V characteristics. Two samples were measured simultaneously. Each of them contained more than 300 GaN based HEMTs. It should be mentioned that the voltage source was limited to +6 volts output, though to modulate gate channel with -25volts was used. Based on this limitation from hardware the voltage over drain-source was limited to 6 volts; however the measurements showed that in absolutely most cases this condition was sufficient for getting a clear picture of device’s performance, though the breakdown voltage has not been investigated. At first, HEMTs with 50 microns channel width were examined. Figures 3.7 and 3.8 depict the I-V characteristics for input as well as for output curves respectively.

Figure 3.7 Output I-V characteristics of GaN with 50µm channel width
In general, gate width is directionally proportional to the output power of the device [8]. The maximum difference in drain to source current achieved was 20mA. Following figure (figure 3.9) depicts the output I-V characteristics for HEMT with longest available channel width, 200µm.
As noticed HEMTs with different channel widths were studied, but most of them showed approximately the same maximum output characteristics. For details about I-V performance of other four transistors (channel width of 75µm, 100 µm, 125 µm and 150 µm) please refer to Appendix E. The small variation in maximum output power of different channel width transistors is in compliance with previous studies performed at our department [8], and as can be noticed on figure 3.6, that expected variation for this widths were in the range of 25 mA.

Based on the measurement data illustrated above, changes were performed in the wafer structure as well as design of the device while processing. The device principle base concept was kept same as described throughout current thesis and reported in [8] [16] and [38], but new methods have been introduced into processing of the device (i.e. in-situ SiN passivation,
different percentage of Al in AlGaN barrier layer). Devices with changed material composition are being processed and will be characterized on their compliance with above described techniques (passivation, gate offset, field-plate gate).

**Large signal characterization**

As DC characterization of transistor was accomplished, large signal network analysis were performed. 75µm channel width HEMT was biased at VGS=-1 and VDS=2.5 volts and the load pull setup was run. As can be seen from figure 3.10, the maximum output power was achieved at reflection coefficient \( \Gamma = 0.56 \angle -163.6 \) (corresponding to impedance \( Z = 14.27 - j6.58 \)). The input power was set up at -11.3 dBm, and maximum output power level was measured to be -26.82 dBm meaning that output was 15.5 dB less than input at DUT level, meaning that the non-reasonable response was caused by the setup calibration files and the cascaded S-parameters files, because the amplification at transistor level was achieved as turning on and off DC power supply manually showed the gain of 14 dB.

![Figure 3.10 Load Pull contours for 75µm HEMT biased at VGS=-1 and VDS=2.5 volts](image)
Several other transistors were characterized with this setup but there was no improvement in overall picture. Amplification was achieved at device level but the main problem identified during this analysis was the lower output power level compared to that of input. As depicted on figure 3.10, the output power changed with tuned impedance, meaning that the problem was originating in calibration files and setup configuration file, rather than in setup itself. Figure 3.11 is a 3-D plot of output power contours that are depicted on figure 4.7 and

3.11 3-D view of output power contours for 50µm channel width HEMT biased at VGS=-0.5 and VDS=3 volts

Figure 3.12 Output power contours for 50µm channel width HEMT biased at VGS=-0.5 and VDS=3 volts
4. Discussions and Conclusions

AlGaN/GaN based HEMTs have been investigated through theoretical studies and evaluated in scope of current thesis. The major semiconductor features were studied in terms of their influence on device performance and due to most material properties and effects achieved at AlGaN/GaN heterojunction, GaN technology is deservedly proposed as one of the most promising compound for high power and frequency applications. Its potential was revealed in MMICs, i.e. power as well as low noise amplifiers. Though comparably high casts delay its commercialization for broad applications, it is largely used in defense systems and intensified research that is carried out within recent years will certainly fully utilize its potential in commercial applications. Based on the results presented, the heterostructures with different material composition (percentage of aluminum in AlGaN) were prepared and methods described in this work will be used to characterize the devices after processing.

Considerable effort was put on large signal analysis studies and fully automated Load Pull setup was set up. Ongoing issues with hardware lack or software upgrade were solved in parallel to theoretical studies and all the components needed for large signal measurements were brought together. The setup was designed for one tone signal 6 GHz frequency, while the room for its upgrade for intermodulation measurements was considered and the components were chosen broad enough to cover frequency range up to 18 GHz.

As was identified through measurements, the main problem designed Load Pull system suffered from was a reduced output power as compared to the input signal level. The reasons were studied and following proposals were made that could improve the performance.

One of the reason contributing to attaining decrease output power level as compared to input of setup was no chance of power source calibration. As can be noticed in appendix B, HP8341 B synthesized sweeper was used as a power source. On the other hand, the same sweeper is a part of HP8510 vector network analyzer system and is used for signal generation for VNA measurements. Due to this reason, synthesizer does not support front panel output and the only output is on rare panel, which is fed through circular waveguide to switch which in turn
directs the signal either to VNA or to output of a test set through 3 dB power divider. Figure 4.1 depicts scheme described.

Subsequently, it is not possible to use the same synthesizer as power source for setup and for VNA simultaneously. Thus calibration file for signal source, as described in section “Setup Calibration” in Appendix C included in to setup was accounted as perfect through line with corresponding S -parameters ($S_{11}=S_{22}=0$, $S_{12}=S_{21}=1$) though direct measurements with power sensor attached to power divider’s output showed 5 dB loss.

An additional important contributor to errors were flexible coaxial cables, used in-between the setup components. Most of them showed loses up to 3 dB and slight change in the angle of bend or position was significantly altering the VNA reading meaning that the calibration files for this part of the setup (due to which the setup accounts for losses) cannot be considered as reliable since there is quite a possibility that the actual performance while measurements differs from the one while calibration.

AlGaN/GaN HEMTs were characterized in terms of DC. The setup was automated and fully controllable through software. The output I-V characteristics of different channel width HEMTs were studied and the results (drain-source current vs channel width) were in agreement with previously performed characterizations in terms of [39].

Another vital factor that will contribute to setup performance improvement is calibration technique used for VNA calibration. As underlined in section “VNA Calibration” (Appendix C), SOLT calibration method was used instead of TRL, strongly recommended by software and
CCMT manufacturer, Focus Microwaves Inc. This issue definitely has an influence on setup’s overall performance, since the software uses interpolation method, which is optimized for TRL calibration data. It was demonstrated in “VNA Verification” section (Appendix D) that proposed verification methods (direct SHORT, offset SHORT and THRU) showed deviations from expected result, although the trends were kept in margin. Also tuner measurements at VNA reference plane showed deviation from expected results and instead of 0 or very low reflections coefficient, $\Gamma = 0.24 \angle -100^\circ$ was observed.
## Appendix A

<table>
<thead>
<tr>
<th>Hardware</th>
<th>Manufacturer</th>
<th>Model</th>
</tr>
</thead>
<tbody>
<tr>
<td>Network analyzer</td>
<td>Hewlett – Packard</td>
<td>HP8510B 5+</td>
</tr>
<tr>
<td>Signal source</td>
<td>Hewlett – Packard</td>
<td>HP8341</td>
</tr>
<tr>
<td>Directional coupler</td>
<td>Krytar</td>
<td>1821</td>
</tr>
<tr>
<td>Isolator</td>
<td>WESTERN MICROWAVE</td>
<td>PMI-4402</td>
</tr>
<tr>
<td>Input bias tee</td>
<td>Hewlett – Packard</td>
<td>HP 11612A/B</td>
</tr>
<tr>
<td>Fundamental source tuner</td>
<td>FOCUS MICROWAWES</td>
<td>1816</td>
</tr>
<tr>
<td>Probe station</td>
<td>The GGB Industries, Inc.</td>
<td>40A</td>
</tr>
<tr>
<td>A dual channel power meter</td>
<td>R&amp;S</td>
<td>NRV</td>
</tr>
<tr>
<td>A programmable power supply</td>
<td>Hewlett – Packard</td>
<td>HP E3631A 0-6V</td>
</tr>
</tbody>
</table>

[Back to Load Pull Setup Description]
CCMT tuners are calibrated for each frequency at a fixed number of impedance points covering as equidistantly as possible the whole Smith Chart. The number of calibration points can be selected between 95, 181 and 361. Optionally CCMT tuners can be calibrated at 385 points [27]. The MTS system software provides 181 calibration points. This does not mean that only these points can be tuned. In reality any point of the Smith Chart can be tuned to, the number of tunable states varying between 500,000 and 11 million (table 1). Obviously all these tunable states are useless if the system software is unable to generate the four S-parameter of the tuners accurately for each of these points. Focus' system software accomplishes this task using nonlinear numerical inter- and extra-polation routines selected to generate maximum accuracy in shortest computing time [40]. These routines have been transformed to work in a coordinate system which best fits the nature of our tuners: Phase and amplitude of the reflection factor are controlled practically independently: Horizontal movement controls the phase whereas vertical movement controls almost exclusively the magnitude of the reflection factor.

The interpolation algorithm works as follows: Each time the tuner is moved to a certain physical position X (horizontal) and Y (vertical) [motor steps] the software first identifies the closest calibrated vertical levels. It takes into account two calibration levels above (Y2 and Y3) and one level (Y1) below the actual Y value. Then it searches the S-parameter of the calibration points of the three closest horizontal levels (X1, X2, X3) to the actual X value. This generates a grid of a total of 9 calibration points around the actual point: [X1,Y1], [X1,Y2], [X1,Y3], [X2,Y1], [X2,Y2], [X2,Y3], [X3,Y1], [X3,Y2] and [X3,Y3].
In a first step virtual calibration points are generated at the vertical level $Y$, but at calibrated horizontal levels $X_1$, $X_2$ and $X_3$: $[X_1,Y]$, $[X_2,Y]$ and $[X_3,Y]$ using 2nd order Lagrange interpolation polynomials in vertical direction:

$$S_{ij}(X_k,Y) = L_0(Y)S_{ij}(X_k,Y_1) + L_1(Y)S_{ij}(X_k,Y_2) + L_2(Y)S_{ij}(X_k,Y_3)$$

where $\{k\} = \{1, 2, 3\}$

$$L_0(Z) = \frac{Z - Z_2}{Z_1 - Z_2} \times \frac{Z - Z_3}{Z_1 - Z_3}$$

$$L_1(Z) = \frac{Z - Z_1}{Z_2 - Z_1} \times \frac{Z - Z_3}{Z_2 - Z_3}$$

$$L_2(Z) = \frac{Z - Z_2}{Z_3 - Z_2} \times \frac{Z - Z_1}{Z_3 - Z_1}$$

$Z = X$ or $Y$ and $S_{ij}(X_k,Y_m)$ are the real and imaginary parts of the four measured S-parameter of the tuner at the calibrated points $[X_k,Y_m]$ for $\{k,m\} = \{1,2,3\}$ and $\{i,j\} = \{1,2\}$.

In a second step the calculated S-parameter at these virtual calibration points are used for a second interpolation in horizontal direction using the same formulas as before.

$$S_{ij}(X,Y) = L_0(X)S_{ij}(X_1,Y) + L_1(X)S_{ij}(X_2,Y) + L_2(X)S_{ij}(X_3,Y)$$

These two interpolation steps permit to calculate real and imaginary part of all four S-parameter of the tuner Twoport at every physical position $X,Y$ (in motor steps). As figure 6.1 shows the dependence of the magnitudes of $S_{11}$ and $S_{22}$ of a tuner on vertical movement of the RF slug are nonlinear but still quite smooth; the phases of $S_{11}$ and $S_{22}$ also change with $Y$ but quite slowly. The magnitudes of $S_{11}$ and $S_{22}$ do not change with horizontal movement. $S_{12}$ and $S_{21}$ change only

---

Figure 6.1. Tuner Calibration points
with vertical movement and remain constant for all horizontal positions. We found that this natural behavior of Focus' tuners can be described very accurately with the formulas given.

All experimental evidence up to now shows that well adjusted and calibrated tuners permit total interpolation and tuning error between 40 and 50 dB up to 18 GHz. At higher frequencies overall accuracy slightly drops to 35-40 dB. This in general includes network analyzer drift between calibration and test of the tuners, mechanical instability and temperature effects on the tuners and the test cables. The pure mechanical resetability of the tuners is easier to test in short time and permits this way to eliminate network analyzer drift and temperature change effects. We found the pure mechanical resetability of S-parameter of the tuners, set to arbitrary impedances at different frequencies, to be between 50 and 60 dB up to 18 GHz dropping slowly at higher frequencies [41]. The difference between total error and pure mechanical resetability error can be attributed to some extend to the interpolation mathematics.

We also found some accuracy improvement when increasing the number of calibrated points from 95 to 181; further increase of calibration points to 361 also enhances accuracy but to a lesser extent.

(Back to Tuner Performance)
7. **Appendix C**

**Calibration**
Calibration is in fact a characterization of a component where the two port S-parameters are measured by means of previously calibrated vector network analyzer (VNA). HP 8510B vector network analyzer with software upgrade 5+ was used for measurements and calibrations of the setup components. From the nature of measurements to be performed, it was plain that inaccurately calibrated vector network analyzer would have a significant impact on the operation of the whole setup and consequent measurements which could lead to erroneous results such as gain of passive components. Figure 7.1 depicts the scheme how controller computer and vector network analyzer were employed to make S-parameter measurements.

Calibration data (S – parameter files in .s2p format) created while calibration except for the tuners was included in the fixed setup calibration file.
VNA calibration

It is evident that accurate calibration is a key issue in creating well functional (or even just functional) load pull setup. The requirements for calibration accuracy and consequent minimization of measurement uncertainties increase with frequency of operation. Our operational frequency, 6 GHz, is a point where the errors start increasing drastically with improper calibration [42].

The whole calibration procedure as well as building setup was due to recommendations available from the CCMT and control software manufacturer, Focus Microwaves. Along with different propositions, it was strongly suggested to make use of TRL (Thru – Reflect - Line) calibration that requires only three standards for full two port calibration of VNA [42]. Due to the fact that TRL calibration standards with needed interface (SMA) were not available at the department, the calibration was performed employing SOLT (Short – Open – Load - Thru) method. Software upgrade 5.0 available on VNA 8510B gave an opportunity to enable Frequency List mode instead of Ramp mode. With this function enabled, during the calibration the list of frequencies was generated. After completing the SOLT calibration, it was saved to particular calset, which was further used while tuner calibration.

Tuner calibration

In general, the calibration of the tuners can be performed in two ways. Tuners can be either pre-calibrated and further the calibration files used during the measurements for discrimination of different positions of the slugs of the tuners with corresponding impedances and losses or the measurements can be performed in so called “in-situ” mode, where the vector network analyzer is a part of measurement system and can be switched in and out [42]. The first mentioned technique, tuner pre-calibration was used for tuner characterization so tuners were attached to VNA reference plane as shown on figure 7.2.
As mentioned in section Tuner Performance, software uses Lagrange interpolation algorithm with respect to distinct states of the tuner (impedance points) regularly scattered over the Smith Chart [43]. Figure 7.3 illustrates distribution of calibration points over the Smith Chart for 6 GHz frequency, resolution of 361 points/frequency, and with gamma maximum of 0.88 (VSWR~10:1). The value of maximum reflection coefficient could of course be taken higher, but since the measurement uncertainties (which in this case were of superior interest) were growing with increasing VSWR and the calibration of the VNA was done with SOLT method instead of proposed TRL, the calibration was carried out with VSWR~10:1 value.
Apart of VNA and CCMT calibrations already performed, the whole setup consisting of cables, adapters, passive components had to be calibrated. As an outcome of this calibrations, a setup configuration file was created, which in fact accounted for all the cascaded S-parameter files. All components except probe station were calibrated at VNA coaxial reference plane. Figure 7.6 represents a dialog window which was used for calibration and loading of the setup file.
The frequency was defined due to loaded tuner parameter files. Each block could contain several cascaded S2P files depending on their actual configuration in setup. Since the setup was designed for one tone measurements, the output coupler (Coupler 2 on the figure 7.4) was left blank. In such a case, when the block is not defined, the software assumes that component is a perfect through line and the S - parameters of the component for any frequency are accounted to be $S_{11}=S_{22}=0$, $S_{12}=S_{21}=1$

**Probe calibration**

For contacts at device level, microwave probes from GGB Industries Inc with tip footprints Ground-Signal-Ground (G.S.G) configuration were used. Calibration data from both halves together with short semirigid coaxial cables was generated and included into setup configuration under Fixture part of the setup, depicted on figure 7.4.
There are two calibration techniques applicable for microwave probe characterization, adaptor removal technique [44] or direct calibration techniques using TRL calibration standards [43]. For all calibration procedures, as stated, Focus Microwaves Inc strongly recommends TRL method, but taking into account the significance of probe calibration and contribution to possible errors, TRL calibration was the only possible technique to be employed. Figure 7.5 depicts the probe station.

Figure 7.5. Picoprobe Probe Station by GGB INDUSTRIES INC
8. Appendix D

VNA Verification

At first performance of VNA was reviewed. The calibration accuracy was checked using recommended measurements [45]. VNA was calibrated for two frequency bands, one for narrower band (3.5 – 8.5 GHz) and one for wider band (3 – 20 GHz) and the results were compared. CalSets were dedicated as follows:

**CalSet 1**  
3.5 – 8.5 GHz (51 points and)

**CalSet 2**  
3 – 20 GHz (101 points)

At first residual reflection (Source Match) was verified using a through line for both calsets. MS Excel was used to generate the plots indicating results and trends. Figures 8.1 and 8.2 indicate results for reverse reflections for both CalSets whereas figure 8.3 depicts the values as recommended in [45]. It has to be mentioned that figures for CalSet 2 depict the values over 3.5 – 8.5 GHz frequency range (Although the calibration was performed for 3 – 20 GHz)

![Figure 8.1 Reverse reflection for CalSet 1](image1)

![Figure 8.2 Reverse reflection for CalSet 2](image2)

It has to be mentioned that figures for CalSet 2 depict the values over 3.5 – 8.5 GHz frequency range (Although the calibration was performed for 3 – 20 GHz)
After, SHORT standard attached to port one was measured with and without offset. Figures 8.4 and 8.5 depict the forward reflection over the band while figure 8.6 illustrates the recommended values [45].
From these results it’s apparent that the reflections were higher comparing to the expected. The test with SHORT standard showed nearly the same results, whereas while measuring the reverse reflection, the difference was in the range of 15 – 20dB for CalSet 1 and 30dB for CalSet 2.
Setup Verification
After creating setup file, before loading it was visually verified. Embedded function of the software gives an opportunity to display all setup parameters in one window and visually review if all of them are reasonable and make sense. Figure 8.7 shows the setup parameters display window where S-parameters and corresponding losses of all building blocks of setup are depicted.

Visual inspection clarified there were several points, where attention had to be concentrated. First of all it was loss of input block, 3.28 dB. This block consisted of three coaxial cables, isolator and input bias tee. The connections were reset several times and premeasured but given loss was the lowest achieved. Reviewing input block’s S-parameters also indicated that especially forward transmission parameter, S21 was not of in satisfactory margin. The components were checked separately (using direct connection to VNA coaxial reference plain).
and along with lousy flexible coaxial cables the problem was identified in isolator performance.

Figure 8.8 depicts the $S_{21}$ parameter of input isolator.

![Graph](image_url)

It can be underlined that the due to isolator specifications, the dynamic range of this component was 6–18 GHz so there was a possibility that since our operational frequency was the lower edge of the dynamic range, with increased frequency performance could improve. The component was analyzed at VNA over all its range but there was no improvement in performance (as can be seen from figure 3.15, there is no trend towards improvement with increase in frequency).

The visual inspection of the parameters showed some other errors. Due to system manufacture (FOCUS MICROWAVES Inc), both reverse and forward transmission parameters of all the blocks except that of isolator have to be equal [43]. The error that is tolerable and may be due to imperfections in VNA calibration is $\pm 0.01 / \pm 1$ degree (magnitude/phase). Input and output fixtures (both halves of probes with attached semirigid coaxial cables) were within acceptable margins. The output block was violating both, magnitude as well as phase shift conditions.
At this instance the decision was made to further proceed with current setup since the connections between different components were established through numerical adapters (not all of them of faithful quality) on the one hand and on the other all this described losses had to be taken into account by the software since all this measurements were dedicated for bringing the reference planes of measurements to the tips of the microwave probes and though to device that had to be measured.

Tuner performance verification

“No question, the tuners are the heart of the system, and the accuracy of the measurement is most affected by the repeatability of the tuners” [46]. After calibration of tuners with highest possible resolution (361 points/frequency) they were attached to VNA reference plane and measured under small signal conditions. Although the calibration errors were evident with just inspecting the calibration pattern⁸ the check was performed for tuners initialization state (both X and Y at 0 position) and a significant mismatch was on detected. Figures 8.9 and 8.10 show the impedance pattern of the tuner while calibration and measured (at VNA) initial position (ideally 50 ohm) respectively. This result was expected since calibration as well as tuner characterization were performed with the same Calibration file of VNA.

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⁸ Coverage of the smith chart with calibration points
Figure 8.9 Source tuner measured at initial position

Figure 8.10 Source tuner measured at initial position
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[41] (1993) Appendix to CCMT 4.0 operation Manual: RF-Resetability Test of CCMT’s".


