A Highly Efficient, Linear and Broadband GaN MMIC Power Amplifier

By

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Abstract

This thesis presents a novel 1W high PAE Linear Broadband monolithic microwave Asymmetric Distributed Power Amplifier (ADPA) design, which could be suitable for a wide range of 0.5 GHz – 6.5 GHz communication systems, such as LTE, CDMA, WIFI, WiMAX and GPS.

This design is based on GaN500 technology, and was fabricated at the Canadian Photonics Fabrication Centre (CPFC). The ADPA achieves a broadband performance by employing a distributed topology employing asymmetric tapered lines. The ADPA layout area is 2 mm × 2 mm, and consumes a total peak current of 369 mA from a 15 V voltage supply. Over the entire bandwidth, the device delivers more than 1W output power and higher than 20 % PAE; a peak saturated output power of 33.45 dBm and 38.1 % peak PAE exists at 0.5 GHz. By sweeping the bias voltages together with the input power, 52.69% PAE is achieved when output power is 31.96 dBm at 0.5 GHz. With the same device, 45.96 % PAE is obtained at 1 GHz.

In addition, a stage-scaled ADPA is studied in simulation. A 0.17 dBm P_{out} variation and 1.90 % PAE variation is achieved over the 1 GHz – 7.5 GHz bandwidth, while 10 dB-gain and 30 dBm P_{out} are obtained over 0.5 GHz – 10 GHz bandwidth.

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List of Abbreviations

ADPA	Asymmetric Distributed Power Amplifier
AM-PM	Amplitude Modulation-Phase Modulation
BW	Bandwidth
CDMA	Code Division Multiple Access
CPFC	Canada Photonics Fabrication Center, belongs to NRC
DA	Distributed Amplifier (same as TWDA)
DPA	Distributed Power Amplifier
DRC	Design Rule Check
EDGE	Enhanced Data Rates for GSM Evolution
EVM	Error Vector Magnitude
GSM	Global System for Mobile Communications
GaN500	Gallium Nitride, Transistor Gate Length is 500nm
HFET	Heterostructure Field-Effect Transistor
IM	Intermodulation
IIP	Input Intercept Point
IC	Integrated Circuit
ISS	Impedance Standard Substrates
LTE	3rd Generation Partnership Project Long Term Evolution
LPF	Low Pass Filter
LVS	Layout Versus Schematic

MIC	Microwave Integrated Circuit
MMIC	Monolithic Microwave Integrated Circuit
MIM	Metal-Insulator-Metal
MOM	Metal-Oxide-Metal
MTEE	Microstrip Transmission Line Tee
NF	Noise Factor
NRC	National Research Council, Canada
PA	Power Amplifier
PAE	Power Added Efficiency
P_{1dB}	Power Level of 1-dB Compression Point
SSADPA	Stage-Scaled asymmetric distributed power amplifier
TWDA	Travelling Wave Distributed Amplifier (same as DA)
TDA	Tapered Distributed Amplifier
WiMAX	Worldwide Interoperability for Microwave Access

Chapter 1: Introduction

This work describes the design and fabrication of a broadband Distributed Power Amplifier MMIC using GaN500 (CPFC) technology.

1.1 Motivation and Overview

As the demand for high-speed wireless data communication continues to grow, different standards are developed, and wider operating bandwidths are often needed. Depending on the application, different frequencies or bands are assigned. For consumer-end products like handhelds, this becomes a burden because the designers have to use different wireless devices to support different standards, such as CDMA, WiFi or Bluetooth. The more devices used in the system, the more power will be needed if several functions are used simultaneously, which could be a problem for the product's battery life and temperature constraints.

IC designers continue to develop consumer broadband devices which can support multiple standards and work over multi-octave bandwidth. Broadband systems have traditionally employed the Travelling Wave Distributed Amplifier (TWDA) architecture, as they have been extensively investigated and firmly established as being reliable and robust devices that can be realized in MIC and MMIC technology [1]. CMOS, the most popular technology, faces many challenges in trying to deliver large output power. The reason for this is that the breakdown voltage of the CMOS transistor is low, which limits the output voltage swing. In addition, supply voltages

continue to decrease in advanced CMOS technologies, making it even more challenging to design a high power amplifier, which is often required in wireless communication applications such as CDMA and LTE base-stations. This weakness is even more obvious after considering the use of stacked devices and transformer topologies.

Further, in a conventional TWDA, the signals on the drain line are added in the forward direction as they arrive at the output. The waves traveling in the reverse direction are not in phase, and any un-canceled signal is absorbed by the drain-line termination [1]. While this eliminates reflections, it inevitably leads to some power loss which could be used at the amplifier's output.

To address these difficulties, an emerging microwave semiconductor technology, 0.5 μ m Gallium Nitride (GaN500), is adopted in this work. Compared to traditional and enhanced CMOS technologies, it has a much higher breakdown voltage; the gate-to-drain breakdown voltage (BV_{gdo}) measured on the wafer having best performance could potentially go up to 100V. Meanwhile, a tapered line asymmetric distributed topology is used to maximize the output power and PAE as much as possible. The key point is removing the drain-line terminal in the reverse direction through drain line tapering, or in other words, adjusting the different sections of drain lines. Doing so eliminates reflected waves and increase the efficiency and output power of the amplifier.

1.2 Thesis Objectives

This thesis focuses on the research of highly efficient, linear and broadband MMIC Power Amplifier design, which uses the GaN500 process developed and fabricated at the Canadian Photonics Fabrication Center (CPFC), which is a division of the National Research Council (NRC). First, this work starts with the study of GaN500 technology and an assessment of its suitability for broadband PAs. Then, the design of an Asymmetric Distributed Power Amplifier (ADPA) which delivers 1W output power based on this technology is demonstrated, in which the RF choke is not included on chip because of the large size of the inductors. Further, this PA design is fully validated through experimental characterization of fabricated prototypes. The bias points of maximum Power Added Efficiency (PAE) are examined as a function of frequency based on measurements, so as to allow a digital control block in future applications. Finally, possible enhancements to the implemented design that could lead to further increases in linearity, output power, PAE and bandwidth (BW) are investigated.

1.3 Contributions

The major research contribution of this thesis is the implementation of a low supply voltage, high output power, high PAE, wideband linear Class-AB Distributed Power Amplifier. This is the first implementation of a Tapered DPA MMIC using the NRC/CPFC GaN500 process. In this work, inductors with various topologies are investigated, and compared so as to meet different applications. The tapered structure,

stage-scale topology, and cascode stage are well analyzed in different designs. Some limitations of this process are reviewed. Finally, an attempt to flatten the output power variation with frequency is made in this work, challenging the boundaries of maximally flat output power.

1.4 Thesis Organization

Chapter 2 gives an overview of Power Amplifier concepts and key parameters and also, introduces the Gallium Nitride (GaN) technology used here.

Chapter 3 provides the operating principles and investigation of the different distributed amplifier structures such as Conventional DA, Tapered DA and Stage-Scale.

Chapter 4 gives a detailed presentation of the ADPA design and practical implementation, starting with design steps of the prototype chip, followed by extensive practical measured results. Differences between the simulations and measurements are explained.

Chapter 5 investigates via simulation several possible enhancements to the original ADPA design. These include, #1 high Q interconnections, #2 re-tapering of drain line sections, #3 output matching circuit for constant output power, and #4 stage-scaled ADPA.

Chapter 6 presents the conclusion and directions for future work.

Chapter 2: PA Concepts and GaN Technology

Since the 1990s, modern cellular networks have expanded rapidly which gave rise to plenty of commercial applications starting with the second generation (2G) cellular systems. The interest in microwave techniques for communication systems has grown immensely over recent years, and the performance of microwave active and passive circuits for wireless systems technology has become well advanced [1]. Among all the blocks employed in wireless transceiver systems, one of the most critical active circuits is the power amplifier, which is typically used in the transmit side of RF or microwave systems to drive antennas. As a result, the PA has always been the first to benefit from developments in device and semiconductor technologies.

2.1 Power Amplifier (PA)

Recently, with the development of wireless communication, more and more emphasis is placed on the implementation of solid state power amplifiers in different bands. For instance, the WiMAX Forum has published three licensed spectrum profiles: 2.3GHz, 2.5GHz and 3.5GHz. Furthermore, 5GHz is on the way. Analog and digital TV bands (700MHz, and perhaps others for HD digital TV signal) may become available for WiMAX usage. EU commissioner Viviane Reding has suggested re-allocation of the 500–800 MHz spectrum for wireless communication, including WiMAX [2].

In one handheld, multiple wireless communication strands might be used. For example, a state-of-the-art smart phone can have NFC, Wifi, and Bluetooth

co-existing GSM, CDMA, EDGE or LTE. The RF portion always consumes the most power and generates most of the heat among all blocks. The demand of sizing the area and saving system power consumption, or in other words longer battery life, boosts the research of broadband power amplifier implementation from the hybrid (i.e. discrete-component) realm to the monolithic circuit arena by developing new semiconductor technologies.

In practical wireless applications, some of them require the need for linear power amplifiers, while the others do not. This requirement depends mainly on the modulation scheme of the communication system employed. For instance, a linear scheme is required for a linear power amplifier, while a nonlinear scheme can make use of a nonlinear power amplifier. There are also some special conditions, i.e. the non-linear power amplifier can be used in the system which uses a linear modulation scheme. For example, MSK, GMSK (also known as GSM, a well-known protocol used in 2G cell phone communication), and Offset-QPSK with pulse shaping, all have constant envelope characteristics because their constellation does not cross the origin, i.e. no zero crossing. And the information is embedded in the differential phase and not the amplitude change. Since there is no or only minimal amplitude change, the non-linear power amplifier does not generate a spectral regrowth phenomena further degrading channel performance. In other words, because the modulation envelope is constant, there is one single optimum point for each power level. Although the power amplifier is non-linear, the shape of the amplified signal never changes. Consequently, the information will not be lost. That is the reason why a non-linear, high efficiency

power amplifier is attractive in wireless communications, especially in cell phone applications.

When moving to the next generation of wireless communication, non-linear modulation scheme such as QPSK and QAM are popularly used in systems such as 3G WCDMA and LTE. Then, a linear power amplifier is required to be used at the output terminal.

PA Operating Categories

Power amplifiers can be categorized into two groups. The first one is the Linear-Class A, Non-linear Class AB, B, C. Then we have switching type amplifiers, Class E, F, etc. For narrow band amplifier design, a different class amplifier has its advantages and disadvantages, and they are normally trade-offs. For example, the Class A amplifier has the highest linearity among all the classes, at the expense of efficiency. And for switching amplifiers, linearity is traded for efficiency, which is very attractive for some applications which do not require high linearity. Based on the linearity and the efficiency requirements of the wireless communication system discussed before, Class AB is the best choice for this work due to its higher efficiency compared to Class A, and good linearity compared to Class B, C.

Quiescent Point

When we start to design a power amplifier, the first step is to choose the DC bias point or quiescent point (Q-point). In deciding the value of bias voltages, V_{DS} and V_{GS} ,

the swing margins of the input and output are determined. Meanwhile, the class of the amplifier is fixed. We can get a general idea by looking into the following graph, Figure 2.1. V_{th} represents the threshold voltage, and I_{DSS} is the maximum DC drain current the transistor can obtain. Normally, Class A quiescent point settles at the middle of the curve, where $I_D = \frac{I_{DSS}}{2}$. Then, because of its middle position, the transistor can swing with full margin, i.e. without reaching the limitation of I_{DSS} and V_{th} . As can be seen from Figure 2.2, except Class A, the other two clip at negative swing due to reaching the V_{th} limitation of the transistors.

Figure 2.1. Relationship of Q-point and swing margin [1].

Figure 2.2. Drain current swing of different Class amplifier [3].

2.2 PAE and Linearity

Among all the parameters of power amplifier characteristics, the three most important are output power (P_{out}), PAE, and Linearity. And for the low noise amplifier (LNA) normally used in the receiver, Noise Figure (NF) is always the first thumb parameter.

We are familiar with drain efficiency, η , which describes how much power can be converted from DC supply to RF output. It is given by the equation of

$$\eta = \frac{P_{out}}{P_{dc}} \quad (2.1)$$

where

$$P_{dc} = \frac{1}{T} \int_0^T V_{cc} i_{dc} dt = V_{cc} I_{dc} \quad (2.2)$$

Here, P_{out} is the RF output power, and P_{dc} corresponds to the DC power coming from the supply. V_{cc} and I_{dc} are the supply DC voltage and current respectively.

Power-added efficiency (PAE) is another presentation of efficiency, which is given by the equation of

$$PAE = \frac{P_{out} - P_{in}}{P_{dc}} \quad (2.3)$$

where P_{out} and P_{in} are the RF output power and input power respectively. If we introduce the parameter of power gain, G , into the equation above, it can be rewritten as

$$PAE = \frac{P_{out} - \frac{P_{out}}{G}}{P_{dc}} = \eta \left(1 - \frac{1}{G}\right) \quad (2.4)$$

PAE is an important parameter which is widely used in RF/MMIC device design, not only because of the low power tendency of custom chip design, but also due to the temperature control. For instance, in most RF/MMIC custom chips used in handheld devices, the un-converted power still contributes to heating. If the PAE is lower, more

power coming from the DC supply is wasted, and the device heats up. Then, because of the thermal energy, the atoms and electrons could move around randomly, which leads to random current components in the circuit. These products are known as thermal noise, which is a white noise.

Ideally, the circuit operates linearly. In other words, the output of the circuit is linearly related to the input. Unfortunately, in practice, the transfer function of the circuit is complicated. A lot of products other than the fundamental are generated because of the nonlinearity of the active or passive devices, or due to clipping by the rail of the source. Further, recently developed wireless communication schemes such as QAM requires for linear operated electronic circuits. These issues are all important results for circuit linearity research; that is, to find out the region where the circuit can operate linearly. The representative parameter for this characteristic is the 1-dB compression point, or P_{1dB} . And for different applications, this specification is replaced by others, such as 2-dB compression point (P_{2dB}), 3-dB compression point (P_{3dB}), and saturation point (P_{sat}). The choice of parameter is determined based on the different linearity requirements of modulation/demodulation schemes. Here, we only introduce the most commonly used concept, 1-dB compression point (P_{1dB}).

1-dB compression point (P_{1dB}) is simply the power level, specified at both input and output, where the output power is 1dB less than it would have been in an ideal linear device [3]. Normally, 1-dB compression point is measured or simulated with 1-tone. It is different with the other important parameter of linearity, IP3, which will be introduced in this chapter later. As shown in Figure 2.3, we can derive the equation of

actual output voltage v_o and the ideal linear output voltage v_{oi} as

$$20\log_{10}\left(\frac{v_o}{v_{oi}}\right) = -1\text{dB}$$

(Equation 2.7)

then we can have

$$\frac{v_o}{v_{oi}} \approx 0.89125 \quad (2.5)$$

As we know, for an input voltage v_i , the actual output of 1-tone because of nonlinearity is

$$v_o = k_1 v_i + \frac{3}{4} k_3 v_i^3 \quad (2.6)$$

Then, the ideal linear output voltage is

$$v_o = k_1 v_i \quad (2.7)$$

So, the input 1-dB compression point can be found by equation

$$\frac{k_1 v_{1\text{dB}} + \frac{3}{4} k_3 v_{1\text{dB}}^3}{k_1 v_{1\text{dB}}} = 0.89125 \quad (2.8)$$

Because nonlinearity causes compression rather than expansion, k_3 must be negative, we have input $v_{1\text{dB}}$ as

$$v_{1\text{dB}} \approx 0.38 \sqrt{\frac{k_1}{-k_3}} \quad (2.9)$$

In addition to measuring the linearity, there is another important parameter, the third-order intercept point (IP3). The reason for research on the improvement of the IP3 is not only for the fundamental product's linearity, but also for the noise because of the nonlinear products such as IP3. In order to analyze it, we first review the Volterra series output voltage v_{out} to input voltage v_{in}

$$v_{\text{out}} = k_o + k_1 v_{\text{in}} + k_2 v_{\text{in}}^2 + k_3 v_{\text{in}}^3 + \dots \quad (2.10)$$

The distortion and intermodulation products are generated because of the nonlinearity

of the electronic system, which can be summarized as Table 2.1. In the table, HD represents the harmonic distortion, and IM represents the intermodulation.

Table 2.1. Summary of distortion products.

Product	Frequency
fundamental	f_1
fundamental	f_2
HD2	$2f_1$
HD2	$2f_2$
IM2	$f_1 + f_2$
IM2	$f_2 - f_1$
HD3	$3f_1$
HD3	$3f_2$
IM3	$2f_1 - f_2$
IM3	$2f_2 - f_1$

It can be seen from Table 2.1 that the frequency of the intermodulation product IM3 is $2f_1 - f_2$ or $2f_2 - f_1$. Normally, multi-carrier or multi-frequency communication such as cell phone requires for a narrow band. In other words, the spacing between each fundamental is very small. For instance, W-CDMA only transmits on a pair of 5 MHz-wide radio channels, while CDMA2000 transmits on one or several pairs of 1.25 MHz radio channels. As a result, it is impossible to filter out the IM3 products from the fundamentals, which could act as noise to adjacent channels.

The third-order intercept point (IP3) is a theoretical point where the amplitudes of third-order intermodulation tones at $2f_2 - f_1$ and $2f_1 - f_2$ are equal to the amplitudes

of the fundamental tones at f_1 and f_2 [3]. As shown in Figure 2.3, the slope of fundamental and IM3 products are 1dB and 3dB respectively. The input power of third-order intercept point (IP3) is called the input third-order intercept point (IIP3), and the output power of this point is known as the output third-order intercept point (OIP3). Those points cannot be practically measured directly, since then the amplifier will be heavily loaded. When performing the simulation and measurement, the space of the two fundamental tones should be taken care of. There are two rules for choosing the spacing. First, according to the customer chip application, it is better to take the channel space as the spacing of two-tone test. On that point, the measurement simulates exactly the real condition in the application. Second, we need to make sure the two fundamental tones and the IM3 tones are in the same gain bandwidth. Then, f_1 and f_2 , $2f_2-f_1$ and $2f_1-f_2$ components will not have any difference because of the different gain among different frequencies. With the setup described above, we estimate the worst condition which we could encounter with later in applications. Through optimization, the device performance could be improved.

Figure 2.3. P_{out} of fundamental (f_1) and IM3 ($2f_2 - f_1$) as a function of P_{in} [3].

2.3 Gallium Nitride Process and Device Characteristic

Wide bandgap semiconductors have been introduced into power amplifiers design, which made the implementation of high output voltage swing amplifiers become possible. The materials' ability to operate at higher temperatures, higher power densities, higher voltages and higher frequencies make them highly interesting to be used in future electronic system applications [4]. Among all these materials, Gallium Nitride (GaN) turns out to be the optimal choice for RF high power and high frequency applications. First, let us review the reasons that limit traditional silicon device power and frequency performance. CMOS devices have very low breakdown voltages, which means large RF (AC) and DC current must be applied to achieve high output power. However, high current operation is inefficient due to series losses coming from the components and interconnects. Also, from Equation 2.14, high capacitance and low impedance limit the maximum frequency which device can operate at.

$$f_T = g_m / 2\pi C_{gs}, \quad f_{\max} = \frac{f_T}{2} \sqrt{R_{ds} / R_g} \quad (2.11) [4]$$

A comparison of different materials available for the implementation of power amplifiers is presented Table 2.2.

Table 2.2. Material properties of CMOS and wide-bandgap semiconductor bulk

@300K (from [4] [12] [13] [14]).

Material	E_g (eV)	ϵ_r	K (W/°K·cm)	E_c (V/m)
Si	1.12	11.9	1.5	3e5
SiC	3.26	10.1	3.7 - 4.5	3e6
GaAs	1.43	12.8	0.54	4e5
InP	1.35	12.5	0.67	4.5e5
GaN	3.44	9.5	1.3 - 2.1	3e6

As seen in Table 2.2, Gallium Nitride has the largest Bandgap Energy (E_g) which results in the highest Electric Breakdown Field (E_c), yielding the highest breakdown voltage, or largest terminal peak voltage, translating into the highest RF output power. Next, the relative permittivity (ϵ_r) is an indication of the capacitance of a capacitor using the semiconductor material as a dielectric, and it also affects the device terminal impedances. Gallium Nitride has the lowest dielectric permittivity, which means it has the lowest capacitive loading since $C \propto \epsilon_r A$ ('A' represents the area). On the other hand, for a given impedance, GaN will require the largest area among the four technologies studied; however, the highest current is obtained when the area is fixed. Finally, the thermal conductivity (K) of a semiconductor material is extremely important since it is a measure of the ease of extracting the dissipated power from the device. Poor thermal conductivity results in degraded device performance at elevated temperatures. As demonstrated in Table 2.2, Gallium Nitride has the same thermal

conductance as conventional Silicon, which is better than other wide-bandgap semiconductors such as GaAs and InP. Meanwhile, the thermal conductivity of Silicon Carbide (SiC) is normally 3.7 – 4.5 W/°K·cm, which is even higher than Silicon. Recent processes now grow GaN on SiC with much improved thermal capabilities. After considering all of the advantages that GaN has to offer compared to other available technologies for PA design, one might ponder the question if after all, there are any disadvantages of using GaN? Surprisingly, the answer is yes. There are two important effects, namely the trapping effect and thermal effect, which can cause current dispersion (collapse and rolling off) as shown in Figure 2.4.

Figure 2.4. Current dispersion due to trapping effect and thermal effect [4].

During ideal operation, as shown in Figure 2.5 (I), the positive interface charge at the heterointerface results in the negative channel charge in an n-type FET, if we assume polarization doping only. Then, a negative surface charge at the top of the barrier is

caused by the interface positive charge. The change of the surface charge during dynamic device operation will then directly interfere with the channel charge. In the second case, as shown in Figure 2.5 (II), the empty charges in the dielectric layer represent the de-trapping of the surface charges. As a result, the channel charge is modified. If the charge at the surface is not fully compensated statically and dynamically, some atoms will be trapped as empty charges in the channel shown in Figure 2.5 (II). So the electron carriers in the channel are trapped. This dynamic trapping and de-trapping is fundamentally connected to the static and dynamic leakage current at the semiconductor barrier/dielectric interface [4]. As a result, the current collapses. One other reason for the current collapse is due to the hot electron injection. When the temperature is high enough, fast random moving electrons can be injected into adjacent layers. Thanks to the research of process engineers, the trapping effect has been minimized. However, buffer traps in a Gallium Nitride (GaN) transistor may exhibit longer time than those encountered in GaAs. This greatly limits the performance of the device in a high power / high frequency system.

Figure 2.5. Trapping effects at the interface dielectric-semiconductor and impact on FET channel [4].

On the other hand, thermal and self-heating effects are presented in every power semiconductor device. Especially when the device is working in the high voltage and high power region, the self-heating causes electron carriers to accelerate in random directions instead of following the drain to source channel. Some of the electrons eject themselves out of the channel, causing significant roll off in the high voltage and high current region, which is observed in DC and RF current variations, as shown in Figure 2.4.

Today, Gallium Nitride (GaN) High-Electron Mobility Transistors (HEMTs) or Heterojunction FETs (HFETs) are the most advanced electronic nitride devices. The heterostructures offers many advantages such as high breakdown and transport properties of undoped Gallium Nitride (GaN).

Gallium Nitride (GaN) Heterojunction FETs (HFETs) designed and modeled by the Canada National Research Center (NRC) were selected in our work. This GaN HFETs are processed on 3-inch diameter epitaxial layers grown on insulating silicon carbide wafers. Semi-insulating silicon carbide is one of the most attractive substrate materials for electronic applications, because of the favorable combination of lattice mismatch, isolation, and thermal conductivity [4]. The cutoff frequency and maximum oscillation frequency of the measured $2 \times 40 \mu\text{m}$ device is 27 GHz and 75 GHz respectively. Also, a 100 V device breakdown voltage is obtained with gate-drain spacing of $2 \mu\text{m}$ or greater. For details of this process, please refer to the appendix.

2.4 Literature Review

Amplifiers with extremely wide bandwidths and good RF performance have been successfully realized over the past 2 decades using both hybrid and monolithic technologies [1]. After the design specifications are fixed (output power > 1 W, PAE > 20%), various technologies and topologies are reviewed. Commonly used CMOS technologies, including SiGe BiCMOS, can normally deliver lower than 1W RF power. As reported in reference [15], TSMC 0.35 μ m SiGe BiCMOS is used in cascode structure; by biasing the 1 mm $\times$ 0.8 mm device with 3.3 V DC voltage, 17 dBm output power is delivered in 0.6 GHz – 2.4 GHz bandwidth with maximum 45% peak PAE. A commercial chip supplied by Hittite using a Dual SiGe HBT working in DC to 5 GHz band has been reported delivering 20 dBm RF power with approximately 15.5% PAE. Usually, the only way for CMOS PAs to obtain more than 1W output power is using a transformer or balun, as reported in reference [16] and [17]. However, including transformer or balun in the circuit would probably increase the device total size and introduce EM noise into other blocks of the circuit. When considering previous work reported in both GaN and GaAs, ultra-broadband PAs, which exhibit multi-octave bandwidth, have been reported by researchers starting in 2000. Output power larger than 1W can be obtained with a simple structure by using such technologies as [18] – [26] which offer advantages when compared to CMOS. These designs are summarized in Table 2.3. In conclusion, CMOS material is limited by its bandgap which leads to low breakdown voltage. Without using a transformer or balun, the delivered output RF power is usually lower than 1W. On the other hand,

III-V materials such as GaN and GaAs can easily obtain a high output voltage since they feature a wider bandgap. However, self-heating normally limits the performance of devices such as GaAs PAs due to GaAs' low thermal conductivity.

Table 2.3 Power amplifier design using different technologies

Technology	Topology	DC voltage (V)	Frequency Bandwidth (GHz)	Peak PAE (%)	Pout (dBm)	OIP3 (dBm)	Size (mm ²)	Ref
0.35um CMOS	Distributed active-transformer	2	2.4 (single frequency)	41	33	-	1.3×2	16
0.35um CMOS	Single stage active-transformer	2.5	2.4 (single frequency)	31	33.4		1.3×2	17
TSMC 0.35um SiGe BiCMOS	Cascode	3.3	0.6 - 2.4	45	17	-	1×0.8	15
Hittite SiGe HBT	-	5	DC - 4	20.2	18	32	-	Commercial Chip
Hittite Dual SiGe HBT	-	8	DC - 5	15.5	20	34	14.9	Commercial Chip
Hittite InGaP HBT	-	8	DC 4.5	17.9	22	35	-	Commercial Chip
GaN HEMT	Cascade	30	3 - 4/ 9.5 - 12	28	37.5	-	4×2.8/ 3.5×2.8	18
0.2um GaN	Distributed	10	2 - 18	9	31	-	4×2	19
0.2um GaN	Darlington	15	0.05 - 12.3	20	30	-	1.7×1.7	20
0.5um GaN	Distributed	15	0.1 - 5	30	30	-	2.2×4.2 5	21
0.7um GaN	Cascode resistive feedback	19	0.2 - 7.5	15	31.5	-	-	22
0.2um GaN	Cascode Darlington	15	1 - 4	20	31	41	1.7×1.7	23
0.3um GaN	Cascode Distributed	25	DC - 8	31	37.8	43	2.5×1.4	24
GaAs	Non-uniform distributed	10	2 - 18	23	30.5	-	3.6×1.7	25
0.25um GaAs PHEMT	Single stage travelling wave amplifier (SSTWA)	-	1 - 8	50	30	-	1.31×2.93	26
Hittite GaAs MESFET	Distributed	12	DC - 6	16	29	41	2.98×2.48	Commercial Chip

2.5 System Concept

As introduced in Chapter 1, this work is built for multi-standard communications. Because of the broad bandwidth of this work, multiple communication standards can be delivered with only one simple device instead of multiple narrowband solution chips; this means a smaller device size and lower total cost. Furthermore, we need a

switchable matching network for different applications, as shown in Figure 2.6. The block of this work is marked with red rectangle.

Figure 2.6 System View

Chapter 3: Principles of Distributed Power Amplifier

In this chapter, we will focus on the investigation of the popular distributed structure used for ultra-broadband or multi-octave power amplifier design. 3 different types of distributed topologies, i.e. conventional Distributed Power Amplifier (DPA), and Asymmetric Distributed Power Amplifier (ADPA), will be demonstrated in their operating principles with design equations. Then, the advantages and disadvantages in the performance of these three will be presented in the following sections. Finally, some guidance will be concluded for our design in Chapter 4.

3.1 Conventional Distributed Power Amplifier (DPA)

Based on the well-researched and established traveling wave concept, distributed (or traveling wave) amplifiers are immensely popular in the design of power amplifiers operating across ultra-broad bandwidths or multi-octave. In a conventional distributed amplifier, the input and output lines are two transmission lines with a number of transistors connected between them, as shown in Figure 3.1.

Figure 3.1. Distributed amplifier schematic [1].

The problem of achieving a broadband match to the transistor input and output impedances is overcome by incorporating the input and output capacitances of a number of transistors into artificial transmission-line structures [1]. The lossy input transmission line on the transistors' gate side absorbs the input capacitance, and simultaneously, the loss output transmission line on the transistors' drain side incorporates the output capacitance. The input and output capacitances are mainly supplied by transistors. As the signal travels down the gate transmission line, each transistor amplifies traveling voltage wave and transfers the signal to the drain transmission line through its transconductance. The output signals of each transistor are added in the forward direction as they arrive at the output node. At the same time, the waves traveling in the reverse direction are not in phase, and any un-canceled signal is absorbed by the drain line termination, or dummy end. In that, the reflection problem on the drain line is solved by the dummy termination. The same theory can be used in building gate-line dummy termination. In an ideal condition, adding more stages can increase gain. Unfortunately, the transmission lines, which we used as drain lines and gate lines, are heavily loaded by the transistor's resistance, and the transmission lines are actually lossy; therefore, the number of active device sections cannot be added indefinitely. This is because the attenuation of the transmission lines will offset the gain obtained by adding an additional transistor device.

In order to deliver maximum output power, it is essential that the drain currents generated by each transistor stage are added in phase as the signal traveling along the drain line toward the output terminal. Based on the above conclusion, in order to use

the amplifier efficiently, the phase shift between transistor stages along the drain line must be exactly the same as the phase shift between transistor stages along the gate line grid. This is achieved by carefully sizing the lengths and widths of the gate and drain transmission lines in order to obtain the correct propagation constants for constructive phasing of the output signals.

Here, the concept is presented in a simple way. The phase shift of each stage is controlled by the artificial line's capacitance and inductance, essentially. The transistor is considered to be free of loss and has a gate capacitance C_{gs} and a drain current generator I_n with associated drain capacitance C_{ds} as shown in Figure 3.2. After investigating the transistor small signal model of Figure 3.2, two capacitances are of interest in the equivalent circuit building. One is the drain-to-source capacitance C_{ds} , which is shunt-connected with the drain line forming a T-section low pass filter on the drain side; the other one is the gate-to-source capacitance C_{gs} , which is shunt-connected with the gate line forming a T-section low pass filter on the gate side. Then, the whole circuit could be represented as a constant-K low-pass filter ladder network as shown in Figure 3.3 based on constant-K model as shown in Figure 3.4. Other parasitic elements in the more general equivalent circuit are neglected to simplify the analysis.

Figure 3.2. Small-signal model of the transistor [1].

Figure 3.3. Equivalent Constant-K low-pass filter ladder network [1].

Figure 3.4. Constant-K section.

Similarly, the transmission lines, which form the input gate line inductance L_g and the output drain line inductance L_d , are considered to be loss free lumped components. Then the gate and drain lines are terminated in their characteristic impedances Z_g and Z_d , respectively. According to low-pass filter theory, for each stage, the phase shift and cut-off frequency can be simply derived as follows [1]:

$$\text{Phase Shift} \quad \varphi = -\arctan(2\pi f Z C) \quad (3.1)$$

$$\text{Cut-off frequency} \quad f_c = 1/2\pi Z C \quad (3.2)$$

where Z represents the transmission line characteristic impedance Z_g and Z_d for different gate and drain section.

Using the above conclusion to obtain maximum output power, we need to match the phase shifts of the gate line and the drain line in order to add the signals at the output

termination, i.e. for each stage $\varphi_g = \varphi_d$. We must also consider keeping the same cut-off frequency in gate and drain lines simultaneously, which can be derived as $f_{c_g} = f_{c_d}$. These equations can be concluded as

$$Z_g \times C_g = Z_d \times C_d \quad (3.3)$$

According to the conclusion explained above, if the phase velocities in the gate line and drain line of each section are roughly equal, which requires $Z_g \times C_g = Z_d \times C_d$, then the travelling signal from each transistor will add in phase at the output terminal. The characteristic impedance of the transmission line on gate or drain side is given by

$$Z = \sqrt{L/C} \quad (3.4)$$

And compared to the parasitic capacitance of transmission line, transistor parasitic capacitance C_{ds} and C_{gs} is dominant. So the cut-off frequency is given by

$$f_c = 1/(\pi \sqrt{LC}) \quad (3.5)$$

Due to the impedance matching better than -10 dB cannot be achieved for frequencies above approximately 86% of f_c [44]. We can calculate the inductance of the gate line and drain line roughly for transmission line design reference.

$$f_H \approx 86\% f_c \quad (3.6)$$

In all the equations we derived above, we assume the transmission lines are ideal, which means they are lossless for a signal propagating along them. Furthermore, we can derive the n-section distributed amplifier power gain based on ideal lossless constant-K line [5].

For an n section distributed amplifier, the total current I_d in the load Z_O is given by

$$I_d = [I_1 e^{-jn\beta d} + I_2 e^{-j(n-1)\beta d} + \dots + I_n e^{-j\beta d}]/2 \quad (3.7)$$

where β_d represents the phase constant of each drain line section in the forward direction.

The voltage wave travelling down the gate line due to input voltage source E_s produces voltages $V_1, V_2, \dots, V_n$ across each input gate capacitor,

$$V_1 = V_{in}e^{-j\beta g}, V_2 = V_{in}e^{-2j\beta g}, \dots, V_n = V_{in}e^{-jn\beta g} \quad (3.8)$$

Where βg is the phase constant of each gate line section in the forward direction.

Then,

$$I_1 = g_m V_1, I_2 = g_m V_2, \dots, I_n = g_m V_n \quad (3.9)$$

In order to simplify the analysis, if we assume the gate line and drain line to be lossless, then the following is achieved,

$$\begin{aligned} |V_1| &= |V_2| = \dots = |V_n| = |V_{in}|, \\ |I_1| &= |I_2| = \dots = |I_n| = |I_{in}| \end{aligned} \quad (3.10)$$

where V_{in} is equal to $E_s/2$ for a matched line.

Re-writing the expression for I_d yields,

$$I_d = V_{in}G_m \{ e^{-j(n\beta d + \beta g)} + I_2 e^{-j[(n-1)\beta d + 2\beta g]} + \dots + I_n e^{-j(\beta d + n\beta g)} \} / 2 \quad (3.11)$$

where current is converted to voltage via drain terminal or load resistance Z_O . The power dissipated in the load Z_O is therefore given by,

$$P_{out} = \frac{1}{16} E_s^2 g_m^2 \left[\sin \frac{n}{2} (\beta d - \beta g) / \sin \frac{1}{2} (\beta d - \beta g) \right]^2 Z_d \quad (3.12)$$

The input power is given by,

$$P_{in} = E_s^2 / (4Z_s) \quad (3.13)$$

The forward available gain is given by

$$G = \frac{g_m^2 Z_d Z_g}{4} \left[\frac{\sin \frac{n}{2} (\beta d - \beta g)}{\sin \frac{1}{2} (\beta d - \beta g)} \right]^2 \quad (3.14)$$

The function can be simplified in limiting case as $\beta_g \rightarrow \beta_d$. Then a sine function can be expressed by the following series,

$$\sin x = x - \frac{x^3}{3!} + \frac{x^5}{5!} - \dots$$

so,

$$\lim(\beta_d - \beta_g) \rightarrow 0 \left\{ \frac{n - \frac{n^3(\beta_d - \beta_g)^2}{4X3!} + \dots}{1 - \frac{(\beta_d - \beta_g)^2}{4X3!} + \dots} \right\}^2 = n^2$$

Finally, we derive the available power gain expression,

$$G = (n^2 g_m^2 Z_d Z_g) / 4 \quad (3.15)$$

In practice, transmission lines are always lossy; that is the reason why there are limitations of the number of sections in a distributed amplifier. The result of the loss in the gate line is that the successive transistors along the gate line receive a decreasing level of input waveform. After a certain number of gate line sections the input signal becomes so weak, that there is no benefit in adding more transistor stages. Meanwhile, referring to the small signal model of the transistor as shown in Figure 3.2, the resistance R_i of each stage absorbs some of the signal travelling along the gate line, and simultaneously, the resistance R_{ds} of each transistor will consume some of the signal travelling along the drain line. The gate line attenuation can be expressed as follows [5].

$$A_g = \frac{\omega_c}{\omega_g} (\omega / \omega_c)^2 / \sqrt{1 - [1 - \left(\frac{\omega_c}{\omega_g}\right)^2] \left(\frac{\omega}{\omega_c}\right)^2} \quad (3.16)$$

Similarly, the drain line loss is given by

$$A_d = \frac{\omega_d}{\omega_c} / \sqrt{1 - \left(\frac{\omega}{\omega_c}\right)^2} \quad (3.17)$$

where ω_c corresponds to the cut-off frequency of gate line and drain line, ω_g and ω_d represent the transistor intrinsic cut-off frequencies respectively, and are given by

$$\omega_g = 1/(R_i C_{gs}) \quad (3.18)$$

$$\omega_d = 1/(R_{ds} C_{ds}) \quad (3.19)$$

Then the gain equation given above, which assumed lossless transmission lines can be modified follows

$$G = \frac{1}{4} g_m^2 Z_d Z_g [(e^{-nA_g} - e^{-nA_d}) / (A_g - A_d)]^2 \quad (3.20)$$

And the optimum number of sections is given by

$$N_{opt} = \ln(A_g/A_d)/(A_g - A_d) \quad (3.21)$$

In the above demonstration, it has been assumed that the gate and drain lines have equal characteristic impedance for matching and equal phase velocities for maximum output power. In the constant-K ladder network, this can only be achieved if C_{gs} equals to C_{ds} . Unfortunately, in practice, the capacitance C_{gs} in the transistor is much bigger than C_{ds} , so more improvement must be taken to equalize the phase shift of gate line and drain line in order to get optimum output. Then, a m-derived section is introduced on the drain line for possible higher performance, by simply adding a series inductance αL_d connected between the drain node of transistor and drain line, as shown in Figure 3.5. In practice, this small inductance can be achieved by implementing a small section of transmission line. And the parameter α is adjusted based on performance.

Figure 3.5. M-derived section.

However, m-derived sections have limited bandwidths. The high capacitance C_{gs} of the large gate width transistors leads to a reduction in the bandwidth. The capacitive-coupled distributed amplifier is introduced [5], as shown in Figure 3.6. This technology allows larger transistors to be employed without reducing the cut-off frequency of the gate line. With the capacitive coupling, the gain is reduced because of the potential divider action across C_{gs} and the coupling capacitor. The shunted high value resistors are needed to supply the DC path for the transistor's gate. However, this imposes the restriction on the power performance, because the low input impedance is needed for maximum saturated power. Also, the shunt structure of the resistor and capacitor is difficult to implement in layout.

Figure 3.6. Capacitive-coupled distributed amplifier [1].

Compared to other topologies used in the implementation of broadband amplifiers such as feedback and Darlington amplifiers, the distributed topology has several advantages. A distributed amplifier provides ultra-wide bandwidth and a good input

match, so gain modules can be easily be cascaded. Its isolation from output to input is good and this coupled with the non-resonant nature of the input and output impedance results in a stable amplifier configuration with no oscillatory tendency on module cascading. It supplies a higher power level than a single stage amplifier, as the current generated by each stage is added together at the output drain line. And because of the distributed topology, it is relatively insensitive to the tolerance of passive and active device characteristics and temperature. But the distributed amplifier is not perfect, it still has disadvantages. Phase velocity equalization is essential because the total output current is dependent on the phase coherence of the individual current generators. This may require extra capacitance to the drain line in parallel with C_{ds} to compensate for the difference between C_{ds} and C_{gs} . This adds complexity and may cause difficulties with realizing the amplifier. On the other hand, the prefect matched phase velocity between gate and drain lines will deteriorate the reverse isolation between input and output transmission lines and cause stability issues due to the feedback caused by the gate to drain parasitic capacitance C_{gd} . Normally, a staggering technique can be used by making the f_c of the drain lines smaller than the counterpart of the gate lines [45] [46]. The amplifier's gain is limited by the gate line attenuation, which will eventually exceed the added gain obtained by increasing the number of transistor stages. Furthermore, because of using several active device stages and since all the transistors are working during amplifier operation, the total DC power consumption is high. In addition, the total layout size is large due to the transmission line section or inductor forming, as well as the large total required gate width.

3.2 Tapered Line Asymmetric Distributed Power Amplifier (ADPA)

In the conventional distributed amplifier, the waveform propagating along the drain line travels in the forward and backward direction simultaneously. In the forward direction, with careful phase matching, the signal is added up at the output terminal. All the signal components travelling reversely are absorbed by the dummy drain terminal. This topology causes power loss and increases the whole chip temperature, because the current travelling reversely to the dummy drain terminal turns into thermal energy. In order to get higher power at the output terminal, the tapered drain line technique is introduced into the distributed amplifier topology, as shown in Figure 3.7. By eliminating the drain dummy termination, this technique attempts to force the total transistor output current travelling in the reverse direction along the drain line to zero, and ensure all the currents fed from the transistors are propagating in the forward direction only. And according to the load pull theory, we need to provide each transistor with its optimal load impedance for maximum output power or power added efficiency (PAE). This is accomplished by tapering the drain line impedance section by section. Without tapering transmission lines on the drain side of transistors, such as conventional distributed amplifier, each active device would see an impedance only half of the terminal impedance, which is 50Ω in most applications. By connecting the first transistor into a section of drain line in which the characteristic impedance is equal to the terminal impedance Z_0 , all of the current will flow into the next stage in the forward direction. Then, if the next section line has characteristic impedance equivalent to half of Z_0 , one third of the current fed from the second

transistor will cancel the reflection coming from the first transistor at the boundary between the first and second drain line. The remaining two thirds of the current developed by the second transistor and four thirds of the current supplied by the first transistor will add and propagate along the drain line forward to the last transistor stage, as shown in Figure 3.7. The transmission line characteristic impedance of the third section should now be equal to one-third of Z_0 . This will continue where each successive line section has characteristic impedance of Z_0/n , where n is the number of the stages implemented.

Figure 3.7. Current distribution in a correctly tapered drain circuit [1].

To illustrate the above explanation with an equivalent circuit, all the transistors can be placed as current sources in the network, and R_L represents the optimum load impedance of a single stage, as shown in Figure 3.8. In the network, R_L is a real value of the load impedance in order to simplify the design, and the termination is neglected. Then, each of the active devices will be operating under similar conditions, including

the optimum load impedance. Hence, it implies that the last section of drain line must be scaled with a characteristic impedance of R_L/n for an n stage distributed amplifier. Similarly, the second last section should have a characteristic impedance of $R_L/(n - 1)$ [1]. By tapering the drain line in this manner, the optimum load required for maximum output power can be synthesized. Therefore, this implies a load impedance of R_L to each stage.

The condition of our design is not the same as introduced here. In our design, we have the last section of drain line which connects the last transistor with the output terminal. According to the theory introduced here, the characteristic impedance of last section drain line should be Z_{opt}/n , and terminal impedance should be Z_{opt} . Then, each transistor is fed its optimal impedance and can obtain maximum output at each section. Further, each drain line needs to phase-shift match with the parallel gate line in the same direction, which guarantees that the output signal coming from each stage is added in phase; and whole PA delivers the maximum output power. This important conclusion will be used in Chapter 4.

Figure 3.8. Current combining for a three-cell distributed amplifier [1].

We have made several assumptions in the above analysis which should be mentioned here.

First, we assumed that each active device or transistor is fed equal current. Unfortunately, due to unequal drive voltages along the lossy gate line, and the process variations, current equalization is difficult to achieve in practice. The gate and drain lines characteristics vary with frequency, which makes the tapering accurate only at a single frequency. There are also other practical considerations, such that the load impedance R_L/n presented to the amplifier might be too low to be synthesized, and then transform to a nominal system terminal impedance of $50\ \Omega$ in most applications. However, for our design in tapered line asymmetric topology, we have used a tapered drain line to make sure each transistor almost sees its optimal impedance. Or by tapering the drain line, each active device amplifies the signal to almost the maximum power it could deliver. But the tapering is not ideal, because the impedance of transmission line shifts along with the frequency. As a result, the output signal is not perfectly flat as a function of frequency.

3.3 Stage-Scaled Distributed Power Amplifier

Although the tapered line asymmetric distributed power amplifier has improved PAE and output power, by looking into the topology, we find that not all the transistors are working efficiently. This is because all active devices have the same size, and are biased at the same voltage in the above topology, which means that some transistors have been fully driven to work successfully, but the others are not. DC power is

wasted in those transistors which are not fully driven; the current draw from the DC power source is not used properly. As a result of this defect, the value of the PAE as well as the efficiency we get from a tapered line asymmetric distributed power amplifier as introduced above is not the maximum value that can be achieved. In order to improve the amplifiers' efficiency as well as the power added efficiency, stage-scale technology is introduced into conventional distributed amplifier design.

Before presenting the stage-scale topology, we first review the conventional distributed amplifier to analyze the reason of total DC power loss. In the former chapter, we concluded that the current led from each active device should be added in phase in order to deliver maximum power to the output load. There are additional notes on this statement. Because of this distributed synthesis, the largest voltage swing only occurs at the last stage, and as a result, the active device in the last section is the only one that experiences maximum allowed voltage swing when output power saturates, while the other stages have sub-optimal voltage swings. As a result, the transistors have not experienced maximum allowed voltage swings and are not driven efficiently, which is the root cause of the DC power wastage in this condition. Therefore, if the voltage swing of the preceding sections can be increased, less current swing is needed to deliver the same power. Of course, DC power and bias current consumption can be done by directly adjusting the DC supply voltages and gate bias voltages of each transistor very carefully, although keeping each stage the same size. Unfortunately, this will complicate the amplifier physical layout and total cost, since each DC supply and bias requires a separate path and pad which will increase layout

area. Each pad also requires an RF choke. These components will increase the total cost of the circuit.

To realize this concept, the stage-scale distributed amplifier topology is introduced. The drain line characteristic impedance and the active device size are scaled simultaneously from section to section to reduce the total DC power consumption without affecting the AC power delivered to the load [6]. In the topology shown as Figure 3.9, the drain line characteristic impedance needs to scale up from the last section to the first section while the transistor size needs to scale down in the same direction to optimize the PAE. In order to reveal the crucial size scaling required, we assume the active device size scales by a factor of K in the reverse direction which traveling from output terminal to input, while the drain line impedances scales by factor of Z , in the same direction. For an n -section stage-scale distributed amplifier, in the scale direction, i_D is scaled K and the impedance Z is scaled for each section, then the peak power generated by the i_{th} section is $(K^2Z)^{n-i}P_{max}$, where P_{max} is the peak power delivered by each gain cell in a distributed amplifier, and n is the total number of the stages. In order to maintain the same maximum power, K^2Z needs to be kept unity. Normally, K is chosen less than unity but close to “1” such as 0.96 in [6]. This gives us a starting point of how to scale the drain line impedance and the transistor size simultaneously without altering the total power delivered. However, the analysis above neglects reflections and transmission line losses when drain line scaling dominants.

Unfortunately, in a practical implementation of the amplifier, the modeling kit

supplied by NRC is limited, which means the transistor size can only be adjusted in redefined steps making it impossible to compensate the impedance shift in drain line perfectly, reducing the ability to achieve higher and constant output power than tapered line technology.

Figure 3.9. Stage-scaled distributed power amplifier [6].

Chapter 4: Design and Experimental Evaluation

Based on the analyses in the preview chapter, we have established so many basic guidelines for the design of the amplifier in this project. Next, design specifications of the amplifier are shown in Table 4.1. In this chapter, we begin with the selection of the transistor's steady-state operating point and the size, followed by the detailed experimental setup along with measurements and evaluation of the results. Finally, the design of the tapered line asymmetric distributed power amplifier (ADPA), improved tapered line ADPA, and also the stage-scale ADPA (SSADPA) will be introduced in this chapter.

Table 4.1. Design Specification.

Technology		0.5 μ m GaN (CPFC)	
Pout	≥ 30 dBm	PAE	$\geq 20\%$
BW	1GHz – 6GHz	Gain	≥ 10 dB
Zo	50	Size	≤ 2 mm $\times$ 2 mm

4.1 Design of Proposed GaN ADPA

In this section, we will first present the single HFET characteristics, including static operating point (or Q point) selection, transistor size decision, and optimal output resistance simulation. Then the design procedure and all the simulation data of the design are reviewed. Finally, some conclusions are given based on the comparison of the designs.

4.1.1 Single Transistor Characteristics Investigation

Before we start the power amplifier design, we need to investigate the characteristics

of the transistor which use CPFC GaN500 technology. Then based on the investigation, we decide the steady-state operating point and the size of the transistor for our power amplifier's application.

For the purpose of applying the final design to common RF transmitter systems, we need to limit the DC supply voltage to 20 V. Currently, the reported narrow-band GaAs/GaN devices are commonly using 15 – 20 V DC supply, so we assume the device will be biased at 15 V DC voltage in this work. Performing a DC simulation using the available GaN500 model in ADS, we obtain the I-V curve as shown in Figure 4.1 of an $8 \times 50 \mu\text{m}$ transistor. The drain current is seen to vary between approximately 100 mA and 400 mA at $V_{ds} = 15 \text{ V}$. Current collapse is observed, which is attributed to hot electron injection and trapping in the buffer layer. Also after observing the I-V curve, it is noticed that the current keeps increasing in the saturation region; this is because the model of the HFET is built with consideration of channel-length modulation, which means the current in the saturation region is not constant as in the ideal model, but with a slope $1/r_o$. But actually, the current will roll off after a particular voltage, which will be reviewed in Chapter 4. Figure 4.2 shows the I_D versus V_{gs} characteristic, indicating a threshold voltage of -4.3 V, which will then be used to determine the steady-state operating point. Then, according to the design datasheet supplied by National Research Council Canada (NRC), together with the simulation, the transistor shut-down voltage is approximately -4.2 V. So, in order to make sure the power amplifier works in deep class-AB type while having high Power Added Efficiency (PAE), we need to bias the transistor close to -4.2 V. On the

other hand, if we bias the transistor closer and closer to its shut-down voltage, the DC current decreases; which means the amplifier's output power cannot meet our targeted specification. Or in other words, low gate-source bias DC voltage could increase the device size.

Figure 4.1. Simulated I_D versus V_{ds} curve of $8 \times 50 \mu\text{m}$ GaN500 HFET.

Figure 4.2. Simulated I_D versus V_{gs} curve of $8 \times 50 \mu\text{m}$ GaN500 HFET biased at 15V.

Operating Point and Transistor Sizing

Based on our assumption of 10 dB power gain, if our amplifier delivers 1W output with 50 Ω input/output terminal impedance, we need have V_{gs_peak} such as 3.16 V given by the following calculation.

$$P_{in} = V_p I_p / 2 = V_p^2 / (2R) \quad (4.1)$$

$$V_{gs_peak} = \sqrt{2PR} = \sqrt{2 \times 0.1 \times 50} \approx 3.16 \text{ V} \quad (4.2)$$

In order to support the full input swing, we need a positive V_{gs} to peak up to approximately 2 V, where the DC I_{d_peak} is about 410 mA. After these conditions are met, we can draw the load line on the I-V curve to verify the Class-AB steady-state operating point (or Q point) chosen for this design. Class-A Q point is biased at $V_{gs} \approx -1.5$ V, as shown in Figure 4.3. We can select the Class-AB operating point for V_{gs} lower than -1.5 V but higher than -4.2 V. Then we sweep the transistor width to find the size which can provide sufficient output power.

Figure 4.3. Load line of $8 \times 50 \mu\text{m}$ GaN500 HFET.

The transistor size is swept in DC simulation as shown in Figure 4.4. The transistor is biased at $V_{ds} = 15 \text{ V}$, $V_{gs} = -2.5 \text{ V}$. According to the target specification (output more than 1 W RF power, achieving at least 20% of PAE and 10 dB power gain), and using a 3-stage ADPA design (stage number will be explained in the following section), we assume P_{out} as 1.2 W for calculation simplification, i.e. each single stage transistor will deliver 0.4W. Then using the following equation,

$$PAE = \eta(1 - 1/G) \quad (4.3)$$

$$\eta = P_{out}/P_{DC} \quad (4.4)$$

Since the gain $G = 10\text{dB}$,

$$P_{DC} = P_{out} / \eta \approx 0.4/0.2 = 2 \text{ W},$$

$$I_{DC} = P_{DC} / V_{DC} = 2/15 \approx 0.13 \text{ A} = 130 \text{ mA}$$

With the calculation shown above, we need slightly more than 130 mA of DC current for each common-source transistor. Considering the smallest available size in this kit, an $8 \times 50 \mu\text{m}$ transistor was selected.

Figure 4.4. Simulated HFET I-V curves for different transistor sizes, biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$.

Next, the cut-off frequency or unity current gain frequency f_T and the maximum oscillation frequency or unity voltage gain frequency f_{max} of the chosen transistor are checked to make sure it can support the application. The simulated f_T/f_{max} as a function of I_D is shown as Figure 4.5, in which the $8 \times 50 \mu\text{m}$ transistor was biased at $V_{ds} = 15 \text{ V}$ while sweeping the V_{gs} from -4V to 0V . At $V_{gs} = -2.5 \text{ V}$, I_D , f_T and f_{max} were observed approximately 144mA , 35GHz and 90GHz respectively. With such an HFET, we can design a power amplifier which can be applied to a highest frequency of 10GHz without major issues. It should be noted that f_T is the parameter corresponding to the current gain into short circuit load, and f_{max} is used to evaluate the small signal power gain while the device is matched to optimum impedance. For these reasons, neither of these two boundaries (f_T , f_{max}) is the true indicator of maximum useful operating frequency of this power device; it will be at some frequency between the two curves.

Figure 4.5. f_T & f_{max} simulation of $8 \times 50 \mu\text{m}$ HFET biased at $V_{DS} = 15\text{V}$.

Transistor Simple Modeling

Further, we modeled the $8 \times 50 \mu\text{m}$ GaN500 HFET biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$ for the theoretical calculation in this section, in order to derive other important parameters such as C_{gs} and C_{ds} . Using the flowing equations [7], we completed s-parameter simulation with fixed bias condition.

$$C_{ds} = \frac{(imag(Y(2,2)) - \omega \times C_{gd})}{\omega} \quad (4.5)$$

$$C_{gs} = \frac{(imag(Y(1,1)) - C_{gd})^2 + (real(Y(1,1)))^2}{(imag(Y(1,1)) - C_{gd}) \times \omega} \quad (4.6)$$

$$R_i = \frac{real(Y(1,1))}{\omega \times C_{gs} \times (imag(Y(1,1)) - C_{gd})} \quad (4.7)$$

$$R_{ds} = \frac{1}{real(Y(2,2))} \quad (4.8)$$

where $C_{gd} = \frac{imag(Y(1,2))}{\omega}$, $\omega = 2 \times \pi \times \text{frequency}$

It can be seen from the Figure 4.6 that for the $8 \times 50 \mu\text{m}$ HFET biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$, C_{ds} is stable along the frequency change up to 10GHz, which is approximately 72 fF; C_{gs} fluctuates very rapidly when frequency changing, compared to C_{ds} , from about 473 fF to 469fF in 1 GHz – 10 GHz bandwidth. We assume the value of C_{gs} is 470 fF for hand calculations, because of its nearly constant behavior. Then from the plot of Figure 4.7, we extract R_i and R_{ds} of 4 Ω and 370 Ω respectively for manual calculation in chapter 4.1.3.

Figure 4.6. C_{ds} and C_{gs} of $8 \times 50 \mu\text{m}$ GaN500 HFET
(biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$).

Figure 4.7. R_i & R_{ds} of $8 \times 50 \mu\text{m}$ GaN500 HFET biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$.

Load Pull to obtain R_{opt}

It is well known that for large signal transistor operation, maximum output power is obtained when optimum impedance Z_{opt} is presented to the output of the device. A unit stage DA is formed by a common-source $8 \times 50 \mu\text{m}$ HFET and investigated here. First, the stability was checked and input is matched with ideal elements at 3.8 GHz within a 200 MHz band as shown in Figure 4.8. Then, we performed a load pull for the Z_{opt} in the unit stage DA, which will be used as reference for drain line tapering in the following section. As expected, the Z_{opt} point should be close to the negative real axis in the smith chart for phase matching between the output voltage and current, in order to obtain maximum power. So, the optimum impedance Z_{opt} can be represented by its real part R_{opt} . It can be seen from the load pull results at different frequencies shown in Figure 4.9 and Figure 4.10, that the optimal output impedance (R_{opt}) of a single HFET is approximately 13Ω if the input is perfectly matched ($S_{11} < -30 \text{ dB}$), and 20Ω for $S_{11} \leq -10\text{dB}$ which is normal in broadband matching.

In our design, a transmission line on the HFETs' drain and gate sides are mainly represented as inductors. Consider the model of an inductor, whose impedance is modeled as series resistance, R_S . $R_S = \omega L / Q$, where $\omega = 2\pi f$, and the Q factors are normally between 4 and 20, depending on the losses present in the processes. After reviewing the possible impedance values of the inductors available in the GaN500 process, we assume $Q = 10$ in our wideband design, and then R_{opt} is approximately 20Ω .

Figure 4.8. Return loss S_{11} of single $8 \times 50 \mu\text{m}$ HFET
(biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$, 3.8 GHz).

Figure 4.9 Load Pull of $8 \times 50 \mu\text{m}$ HFET biased @ $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$, 3.8 GHz with $|S_{11}| < -30 \text{ dB}$.

Peak Pout (dBm) contour @ 4GHz with $|S_{11}| < -10\text{dB}$ Peak PAE (%) contour @ 4GHz with $|S_{11}| < -10\text{dB}$

Figure 4.10. Load Pull of $8 \times 50 \mu\text{m}$ HFET biased @ $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$, 4 GHz with $|S_{11}| \leq -10 \text{ dB}$.

4.1.2 Cascode Gain Stage

In CMOS processes, such as CMOS 130 nm and CMOS 65 nm, a cascode topology is used to obtain higher voltage swing or voltage gain compared to a single common source transistor structure. Then the question comes, can we use the same technology in the GaN500 process? Investigation is introduced here supported by simulation results, and we also did some analysis in theory.

We use 20V DC voltage as supply for the cascode circuit (all transistors have $8 \times 50 \mu\text{m}$ width), and the common-source HFET should be biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$. It can be seen from the simulation result as shown in Figure 4.11, the gate of the cascode HFET should be biased at 12.6 V.

Figure 4.11. DC bias cascode stage.

By comparing the maximum gain and S_{21} between single HFET amplifier and cascode topology as shown in Figure 4.12 and Figure 4.13, it can be seen that cascode stage has much more voltage gain than single common-source structure. In the simulation, ideal RF chocks and DC blocks were used, and a Z_{22} of $77-j384\Omega$ was observed in cascode PA at 2 GHz. This agrees with the theory, because cascode transistor contributes by increasing its load impedance.

All the parameters are evaluated by using Large Signal S-Parameter with 20dBm RF power. Based on the results we have obtained, it seems that using cascode topology in Asymmetric Distributed Power Amplifier is possible and much more powerful than the common-source stage. But we still need evidence that can prove the worth of complexity trading for better gain performance in ADPA.

Figure 4.12. Maximum voltage gain comparison between cascode and single common-source in which common-source HFET is biased at $V_{gs} = -2.5$ V, $V_{ds} = 15$ V.

Figure 4.13. S_{21} comparison between cascode and single common-source in which common-source HFET is biased at $V_{gs} = -2.5$ V, $V_{ds} = 15$ V.

4.1.3 First Tapered Line Asymmetric Distributed Power Amplifier

For the first design, we chose to use a single common-source HFET to form the unit stage of the asymmetric distributed power amplifier (ADPA), in order to focus on the analysis of the tapered line ADPA topology. The basic ADPA circuit schematic is given in Figure 4.14.

In the previous analysis of the single transistor, we have already fixed the HFET width as $8 \times 50 \mu\text{m}$ and biased at $V_{gs} = -2.5 \text{ V}$, $V_{ds} = 15 \text{ V}$, which corresponds to $I_D = 144 \text{ mA}$, according to the design specifications. Now with all the available information, we first need to calculate gate and drain line inductance, and then determine the number of the stages of the ADPA, and finally optimized the tapering of the drain line based on simulation results and optimization. Simultaneously, we must take into account the DC current handling capability of all transistor interconnects especially on the drain side. Referring to Figure 4.14, the DC current through L_{d1} , L_{d2} and L_{d3} is 144 mA, 288 mA and 432 mA respectively.

Calculation

Normally an artificial transmission line can be used up to approximately 86% of its cut-off frequency due to matching issues [44]. According to our application requirement, we assume the highest frequency is $f_H = 8 \text{ GHz}$, well above the 6GHz operational upper frequency. Then $f_c = 9.3 \text{ GHz}$, which means $\omega_c = 2\pi f_c = 58.4 \text{ Grad/s}$. In theory, the characteristic impedance of transmission line is given by:

$$Z_o = \sqrt{L/C} \quad (4.9)$$

and its cut-off frequency is given by:

$$f_c = 1/(\pi\sqrt{LC}) \quad (4.10)$$

Because the parasitic capacitance of HFET is considerably higher than the one of transmission line, and C_{ds} or C_{gs} is parallel with C_{TL} , the cut-off frequency of the transmission line is mainly determined by C_{ds} or C_{gs} and the inductance of the drain/gate line. The following equation can be derived based on conventional distributed amplifier single section and using a constant-K ladder model,

$$\omega_{c_g} = 2/\sqrt{L_g C_{gs}}, \quad \omega_{c_d} = 2/\sqrt{L_d C_{ds}} \quad (4.11)$$

where L_g and L_d represents gate and drain line impedance respectively. With the simulated results of Section 4.1.1, each $8 \times 50 \mu\text{m}$, each single transistor has $C_{gs} \approx 470 \text{ fF}$, $C_{ds} \approx 72 \text{ fF}$ in the 0.5GHz - 10GHz bandwidth, and every drain/gate line section, except those next to the input and output ports, are connected to C_{ds} and C_{gs} on both sides; then we can calculate an $L_g \approx 1.07 \text{ nH}$. We can directly use $L_g = 1.07 \text{ nH}$ as the inductance of the gate line in our asymmetric distributed power amplifier, but the drain line or L_d needs to be tapered for each section; the L_d value will guarantee the cut-off frequency of the drain line will be higher than 10 GHz.

Furthermore, in order to maximize the output power, phase synchronization for each stage has to be ensured as follows,

$$\theta_g = \beta_g l_g = \theta_d = \beta_d l_d \quad (4.12)$$

where $\beta = \omega\sqrt{LC}$, and l_g and l_d are the length of the gate/drain lines.

Then Eq. 4.12 can be re-written as,

$$\omega\sqrt{L_g C_{gs}} \times l_g = \omega\sqrt{L_d C_{ds}} \times l_d \quad (4.13)$$

Figure 4.14. Phase synchronization.

For the first section, we have $L_{d1} = 2.25$ nH with the assumption that $L_{g2} = 1.1$ nH. Next, we will estimate the tapering required for the artificial line in each stage. In Chapter 3, we have already analyzed and introduced the basic idea of tapering. For an n -section line, we need to obtain the drain line's characteristic impedance of Z_O/n . With the assumption that all the transistors have the same size and neglecting process variation, C_{ds} will remain constant for all sections. Assuming the inductance of the first section drain line on the left corresponds to L_{d1} , then L_{d2} , $L_{d3}, \dots, L_{dn}$, we can derive the following relationship:

$$L_{dn} = L_{d1}/n^2 \quad (4.14)$$

So, we can simply calculate that L_{d2} and L_{d3} are approximately 0.6 nH and 0.2 nH respectively for the 3-stage tapered line distributed amplifier. Meanwhile, this relationship can only be taken as a design reference, since the impedance of a spiral inductor is affected by other parameters such as line width.

After calculating all the inductances of the drain and gate lines, we can finally derive the optimal number of stages which is obtained for maximum gain in the asymmetric power amplifier.

From the simulation plotted as Figure 4.7, we know over the bandwidth of interest, the parameters R_i and R_{ds} of HFET, biased at $V_{gs} = -2.5$ V, $V_{ds} = 15$ V, 6 GHz are approximately 4 Ω and 370 Ω , respectively. We can then calculate the drain/gate line attenuation with the equations given as follows,

$$A_g = \left(\frac{\omega_c}{\omega_g}\right) \left(\frac{\omega}{\omega_c}\right)^2 / \sqrt{1 - \left[1 - \left(\frac{\omega_c}{\omega_g}\right)^2\right] \left(\frac{\omega}{\omega_c}\right)^2} \quad (4.15)$$

$$A_d = \frac{\omega_d}{\omega_c} / \sqrt{1 - \left(\frac{\omega}{\omega_c}\right)^2} \quad (4.16)$$

where HFET's intrinsic cut-off frequencies ω_g and ω_d , cut-off frequency of gate/drain line are given by

$$\omega_g = 1/R_i C_{gs} \quad (4.17)$$

$$\omega_d = 1/R_{ds} C_{ds} \quad (4.18)$$

With the attenuation value calculated and assuming a working frequency of 8 GHz, we can calculate that $\omega_g \approx 5.32 \times 10^{11}$, $\omega_d \approx 3.75 \times 10^{10}$, resulting in $A_g \approx 0.06$, and $A_d \approx 0.84$. Finally, the optimum number of sections N_{opt} is given by

$$N_{opt} = \frac{\ln \left(\frac{A_g}{A_d}\right)}{A_g - A_d} \approx 3 \quad (4.19)$$

Over 10 dB of gain is obtained by using Eq. 4. 20 while g_m , Z_d , Z_g is approximately 0.1S, 20 Ω and 50 Ω respectively.

$$G = n^2 g_m^2 Z_d Z_g / 4 \quad (4.20)$$

By sweeping drain line L_{d2} while fixing the other L_{d1} and L_{d3} as 2.25 nH and 0.2 nH respectively (as shown in Figure 4.15), the maximum gain is plotted at different frequencies. Other L_{Dmid} values are desirable, but 0.5 nH is preferred due to its small

size and easier implementation. So, as a starting point, we choose the tapered lines' inductance as 2.25 nH, 0.5 nH, 0.2 nH from the first to the last stage, the Q of which is assumed to be about 10 for broadband design.

Figure 4.15. Maximum Gain (MaxGain) while sweeping middle drain line's inductance.

After an initial study using lumped elements, we begin to migrate to using real kit components, taking into account the current density limitation of the metal layer, which is 6mA/um as stated in design manual.

Recall that we have analyzed the tapered line distributed amplifier in Chapter 3. In order to obtain the maximum output power, we need to feed each transistor its optimal impedance, which means the drain line's impedance represented to each stage should be Z_{opt} , $Z_{opt}/2$ and $Z_{opt}/3$ respectively. After the load-pull of each unit stage was completed in section 4.1.1 at 4 GHz (with return loss at input port $S_{11} \leq 10$ dB), the Z_{opt} is determined to be 20 Ω . Then the first section of drain line is chosen to be a

spiral inductor due to the large required inductance that arises from $Z_{\text{opt}} = \sqrt{L_{\text{d1}}/C_{\text{drain_total}}}$, where the $C_{\text{drain_total}}$ of 5.6pF is the total capacitance presented at the drain side of the first stage. The impedance of the other sections of drain lines are adjusted according to Fig. 4.14 for maximum output power. Furthermore, each drain line section phase shift matches with the parallel gate line on the input side (C_{gs} and C_{ds} are included in the phase shift evaluation). Hence, the impedances for maximum output power presented by the drain lines from the first to last stage are approximately 20 Ω , 10 Ω and 6 Ω , respectively.

Line Tapering

As, we start to size the tapered drain lines, another question comes about. According to the GaN500 design manual, there are 3 different spiral inductor layouts that can be chosen, as shown in Figure 4.16. Type 1 uses 1ME (for the layer table, refer to the Appendix of this thesis) to build all the internal turns, with a bridge connecting out the terminal on one side. Type 2 uses bridges for internal turns when the sections cross the outlet terminal line. Type 3 is different from the above two, which uses landed metal. As shown in Figure 4.17, the Q factor is investigated among these three inductors; type 1 design has the highest Q, while type 2 has the lowest simulated Q. While it would be desirable to employ the highest Q type of inductor to minimize losses, the low Q type 3 is chosen for the application due to its larger current carrying capability. It will be shown in Section 4.4.1 that the difference in achievable PAE and output power between highest Q type and type 3 inductors is minimal.

Figure 4.16. Three types of allowed spiral inductor layouts.

Figure 4.17. Q comparison among three type layouts (1 nH implemented inductor).

After all of the investigation presented above, the coplanar waveguide at drain line sections were sized accordingly, and the final values are as follows: 30 μm , 54 μm , and 78 μm , the length as 2.58 mm, 0.67 mm, and 0.28 mm, from left first section to right last section; as a result, they can sustain 144 mA, 288 mA and 432 mA DC current required, respectively.

Figures 4.18 – 4.22 show the inductance values and Q factors, of the drain and gate lines' implementations. The low Q type spiral inductor design was chosen for broadband matching due to the application of the device. Figure 4.18 confirms that the first inductor implementation has $L_{d1} = 2.5\text{nH}$, as required, and self-resonates well beyond 6 GHz. In addition, its maximum Q is 4.5.

Figure 4.18. Substrate setup and first section drain line characteristics.

Figure 4.19 shows that the implementation of L_{d2} yields an inductance of 0.48nH as required and a maximum Q of 17, with no indication of self-resonate effects up to 10GHz. Figure 4.20 illustrates the third inductor implementation has L_{d3} of 0.18 nH , as required, and self-resonates well beyond 10 GHz. The maximum observed Q is approximately 14.5.

Figure 4.19. 2nd section drain line characteristics.

Figure 4.20. 3rd section drain line characteristics.

Figure 4.21. first and last section (side) gate line characteristics.

Figure 4.22. 2nd and 3rd stage (middle) gate line characteristics.

Figure 4.21 confirms that the implementations of L_{g1} and L_{g4} present inductances of 0.51 nH and 0.66 nH as expected and maximum Quality Factors of approximately 8.1 and 8 respectively, with self-resonances beyond 10 GHz. Figure 4.22 shows the inductance L_{g2} and L_{g3} of the implemented 2nd and 3rd gate lines yield to 1.06 nH as required, and self-resonates well beyond 8 GHz. In addition, its maximum Q is approximately 8.

From the above plots, it can be concluded that, the inductor implementations are providing the desired Z_g , Z_{opt} , $Z_{opt}/2$ and $Z_{opt}/3$ values at the appropriate stage of the ADPA as expected. Detailed phase matching will be presented in Section 4.4. 1.

All of the elements of this first design are indicated in the final circuit schematic of Figure 4.23. The corresponding layout that results from this schematic is shown in Figure 4.24.

Figure 4.23

Schematic of the first ADPA design

Figure 4.24. Final layout of the first design (HFETs marked by the rectangle).

The last drain line section is directly connected to the 50 Ω ground-signal-ground (GSG) pads required for probe measurements. The GSG pads are designed to obtain 50 Ω impedance with 150 μm pitch. The performance (small and large signal S-parameter, stability, power gain, delivered power, PAE, linearity, harmonic distortion) of this first design is now presented, as captured from post-layout simulation. Each device is biased at $V_{gs} = -2.5$ V and, $V_{ds} = 15$ V. The stability of the ADPA is checked in Figure 4.25. K factor is presented using a logarithmic scale; it can be seen that B1 factor is in the range of 0 and 1, and K is higher than 1, which presents the ADPA is stable over 0.1GHz – 10GHz frequency range.

Figure 4.25. Stability factors, K is presented as $\log_{10}(K)$.

The small-signal s-parameter results over the 0.1GHz to 10GHz range are presented in Figure 4.26. As expected, the gain is above 10dB over the operating bandwidth of the artificial transmission line, which is approximately 8GHz. The low frequency gain is 13dB and the upper 3dB corner frequency is approximately 8GHz as expected in

calculation. Looking at S_{11} , the ADPA is seen to be well matched ($S_{11} < -10$ dB) from 0.5 GHz to 6.5 GHz. As seen from the S_{22} curve, the output is not as well matched to 50 Ω but this is not required as the amplifier is designed for maximum power to the load. The low values of S_{12} below approximately -20 dB across the band confirm the good isolation between input and output lines, as assumed in the design methodology. The large-signal s-parameter performance is shown in Figure 4.27 for an input power, $P_{in} = 20$ dBm (the design specification). Very little change is observed compared to the small-signal results of Figure 4.26. S_{11} remains below -10 dB over a broader bandwidth extending from 0.5GHz to 7.5GHz.

Figure 4.26. Small signal s-parameter biased at $V_{gs} = -2.5$ V, $V_{dd} = 15$ V.

Figure 4.27. Large signal s-parameter biased at $V_{gs} = -2.5\text{V}$, $V_{dd} = 15\text{V}$.

The amplifier achieved a gain greater than 10 dB over the 0.1GHz – 8GHz bandwidth, and has obtained a PAE higher than 20 % over the 0.1GHz – 7GHz frequency range as shown in Figure 4.28. The amplifier delivers a peak RF power of approximately 33.2 dBm to the output load, with 33.8 % peak PAE at 0.9 GHz while consuming 6.4 W peak DC power (15 V DC supply, 427mA total simulated DC current).

Figure 4.28. Power gain, P_{out} , and PAE.

Figure 4.29 illustrates the methodology used for the OIP3 and IIP3 calculation. It can be seen from Figure 4.30 that the OIP3 and IIP3 are above 45 dBm and 31 dBm respectively over the entire bandwidth of interest, which means the ADPA is operating linearly while in band. Figure 4.31 presents the spectrum waveform of the harmonic distribution at $f_0 = 6$ GHz. The fundamental has approximately 32 dBm while the 2nd and 3rd harmonic presents 5.9 dBm and 2.7 dBm respectively, which are over 25 dB below the fundamental. Figure 4.32 shows the output signal distortion as a function of frequency. The fundamental is above 30dBm over the entire bandwidth, while the harmonics are almost 20 dB below the fundamental across the band.

Figure 4.29. Fundamental & 3rd Harmonic versus P_{in} at 3 GHz.

Figure 4.30. Linearity over 0.5 GHz – 6.5 GHz.

Figure 4.31. Harmonic distribution at $f_0 = 6$ GHz.

Figure 4.32. Harmonic distortion over 0.5 GHz – 6.5 GHz.

Figure 4.33. Time domain voltage/current input/output waveform at 3 GHz.

From Figure 4.31 and Figure 4.32, it can be seen that, since no harmonic shorting circuitry is included in the APDA, contribution from the harmonics have resulted in the distortion of output waveform in time domain as shown in Figure 4.33. This performance could be improved for a specific narrow-band application, in which a

reconfigurable network could be added to successfully eliminate the harmonics at arbitrary frequencies. Also, it can be seen from Figure 4.34 that all the gain HFETs in the ADPA are operating as Class-AB amplifiers.

Figure 4.34. I_d through the common-source HFET in the ADPA unit gain stage.

It is interesting to now consider how optimal the fixed $50\ \Omega$ load impedance is over the broad operating range of frequencies of this amplifier. We conducted a load pull analysis at the amplifier's output port at various frequencies: 1 GHz, 4 GHz, and 6 GHz, to obtain the R_{opt} required for delivering maximum output power. This is shown in Figures 4.35 – 4.37. It can be seen from the simulation results, that the R_{opt} is not constant as a function of the frequency. If we need to obtain maximum output power at low frequencies, typical for a cell-phone band at 1GHz, we would need to match to the optimum load, which is approximately $55\ \Omega$, close to $50\ \Omega$. Or, we need to use the PA in high frequency applications such as WiFi and WiMAX, where the PA would be

better to match to approximately $30\ \Omega$ optimum load in order to deliver maximum power. It is ideal to build a reconfigurable matching circuit for each application at a specific frequency; however, in this work we focus on the research of asymmetric distributed topology, so a reconfigurable network is not included. However, it will be recommended for future work.

Figure 4.35.

Load pull simulation at 1 GHz.

Figure 4.36.

Load pull simulation at 4 GHz.

Figure 4.37. Load pull simulation at 6 GHz.

There is another phenomenon that attracts our attention in Figure 4.33. The input waveform is also distorted, not an ideal sinusoidal wave which is supplied by an ideal signal source in ADS. The reason for this will be explained in the next section.

4.1.4 Nonlinear Capacitance Effect Research

In this section, we will discuss the nonlinear effect due to the gate-to-source parasitic element inside the HFET, C_{gs} , as this mechanism is responsible for a part of the signal distortion observed in the previous section.

The analysis of nonlinear properties of voltage-dependent capacitors has a long history, reaching as far back as the 1960s when varactor multipliers were the only means of generating significant power levels at GHz frequencies using solid state components [8]. First, we will review the basic equation of current to voltage in a circuit which has nonlinear capacitance at input terminal, as shown in Figure 4.38.

Figure 4.38. Investigate of nonlinear capacitance effect on current.

The linear and nonlinear capacitance is named as C_{lin} and C_{nonlin} respectively. The nonlinear function describing the differential capacitance variation along with the voltage is given as,

$$C_{nonlin} = f(V) \quad (4.21)$$

where the output voltage is equal to the voltage integral of the capacitance,

$$F(V) = \int f(V) dV \quad (4.22)$$

Then, the total current passing through the circuit is proportional to the sum effect coming from both the linear and nonlinear capacitance part, which is given as

$$I = C_{lin} \frac{dV}{dt} = C_{lin} \frac{dF(V)}{dV} \frac{dV}{dt} = C_{lin} f(V) \frac{dV}{dt} \quad (4.23)$$

In the time domain simulation, the device input voltage is distorted from its original sinusoidal form because of the exit of C_{gs} as shown in Figure 4.39. If the bias point is changed, the change of conduction angle of the output is observed. The reason of that is because of the varactor distortion; in other words, it is caused by the voltage distortion of C_{gs} . Further, the voltage distortion appearing across parasitic capacitance C_{gs} will cause significant changes in the output voltage harmonic products.

Single transistor model and characteristics were investigated in Section 4.1, from which it can be concluded that the parasitic capacitance C_{gs} does not change

significantly over frequency. We use the same methodology here to have a look at the varactor effect of C_{gs} versus variations in its gate voltage. A frequency of 3 GHz is chosen for an $8 \times 50 \mu\text{m}$ HFET with gate voltage V_{gs} varying from -3 V to +3 V.

Figure 4.39. Input voltage distortion in time domain and frequency domain.

The gate-to-source capacitance, C_{gs} , is observed to change when V_{gs} is swept, and the margin is approximately 487 fF to 245 fF as shown in Figure 4.40.

Now it becomes clear after this brief inspection that the input varactor capacitance C_{gs} can have a major impact on voltage and current distortion in the ADPA's operation.

As a possible solution to the harmonic distortion caused by the varactor effect of C_{gs} ,

a harmonic short circuit could be used at the input side of the transistor device. This would be similar to harmonic short circuits placed at the output of the device in narrow band PA design. While the 2nd harmonic is normally shorted via $\lambda/4$ short circuited shunt stub (SCSS) at the transistor device's output in MMIC design, it would seem that minimal distortion on the input side is due to both 2nd and 3rd harmonics (see frequency domain plot in Figure 4.39).

Figure 4.40. C_{gs} as a function of V_{gs} at 3 GHz.

4.2 Experimental Setup

The high efficient linear broadband ADPA chips were fabricated at the NRC using the CPFC GaN500 technology. The fabricated chip is shown in Figure 4.40. In this section we describe the experimental set up intended to measure the following PA performance parameters: DC I-V curve; small-signal s-parameters; full power P_{out} (while input 20dBm RF power); 1-dB compression point (P_{1dB}); saturated output power (P_{sat}). From the measured data, PAE and linearity (IIP3/OIP3) will be derived. Because of the limitation of available apparatuses, Bit Error Rate (BER) and Error

Vector Magnitude (EVM) will not be experimentally measured.

Chip Grounding and Heat-sinking

Test samples were mounted on a metal heat-sink as shown in Figure 4.41 in order to, first, spread the thermal dissipation, and second, create a ground plane which exists in the simulation layer set-up but not included in the implemented substrate. Once the chip/heat-sink assembly is placed on the probe station, the ground will exist for both the probes and the PA chip via the test station's metal chuck. Six DC bias pads are connected to GND (three pads on the top and three pads on the bottom as shown in Figure 4.42) using wire bonding to the metal plate. Therefore, the test samples are well grounded not only through the GSG (Ground-Signal-Ground) probe, but also via the ground metal chuck of the probe station.

Figure 4.41.

Chip under test.

Figure 4.42. 2 mm × 2 mm fabricated chip.

DC & RF Measurement Apparatus

The test components used can be categorized as DC and RF in nature; several test benches were set up using different pieces of equipment available in the Department of Electronics at Carleton University. All required instruments were carefully selected based on the ADPA simulated performance. All the test benches' set-up and equipment brief specifications are presented, and the ICs were tested on a Cascade Microtech 9000TM Analytical Probe Station. As an example, a photograph of the DC test and 1-tone RF measurement (1-tone power test and small signal s-parameter measurement) set-up is shown as Figure 4.43.

Figure 4.43. 1-tone power test and small signal s-parameter test set-up.

First, the DC test is presented. Before starting any RF test, an oscillation test was performed to check whether the ADPA would have any unwanted instability and oscillation. This was done by biasing the PA with the DC supply but no RF signal applied to the input, with the output connected to an oscilloscope via the probes. When DC sources were turned on, no oscillation waveform was observed at the output of the power amplifier. This means that the noise injected by the DC supply does not cause the power amplifier to oscillate. After the oscillation check, a DC I-V (current-voltage) curve was measured. By biasing the device with the DC voltage supply, the current levels were recorded.

Then, the RF measurements are introduced. Due to the limitation of the s-parameter Network Analyzer which is opt001, instead of the high power test option such as opt085, the large-signal s-parameter measurement cannot be performed. In our

measurement, only the small-signal s-parameters were obtained. The set-up diagram is shown as Figure 4.44, and the apparatuses are listed in Table 4.2, where ‘BT’ represents the bias tee and ‘DUT’ represents the device under test. In the diagram, probes are included in paths instead of being represented by a separate block. This rule of not illustrating the probes in the diagram will also be applied to all subsequent figures.

Figure 4.44. Small signal s-parameter test set-up diagram.

Table 4.2. Equipment list for small-signal s-parameter test.

Function Name	Equipment Model	Frequency Band	RF Power Rating	DC Rating
V _g Supply	HP E3630A	DC	N/A	0 V – 6 V, 2.5 A /0 V – ±20 V, 1.5 A
V _{dd} Supply	Agilent E3646A	DC	N/A	0 V – 8 V, 3 A /0 V – 20 V, 1.5 A
BT1 Bias Tee	Picosecond 5541A-104	80 KHz – 26 GHz	1 W max.	50 V, 100 mA max.
BT2 Bias Tee	HP 11590A OPT1001	1 GHz – 18 GHz	24 dBm max.	100 V, 500 mA max.
RF Probes	GGB Model 40A-GSG-150P	DC – 40 GHz	N/A	N/A
Network Analyzer	HP 8722ES	50 MHz – 40 GHz	30 dBm max.	40 V max.

Single-tone power measurements such as power gain, output power, and PAE (Power-Added-Efficiency) were performed by using the test diagram and the equipment shown in Figure 4.45 and Table 4.3 respectively. Two different models of bias tee were used for BT2 (Bias Tee 2) depending on the frequency band. For measurements in the 0.5 GHz – 4 GHz range, the Mini-Circuit bias tee was used as the maximum output levels are in excess of 24 dBm, which is the maximum RF power specification of the Agilent bias tee. Due to the limited available RF power of the signal generator and the loss of input cables, a pre-amplifier was used to obtain the required 20 dBm RF input power. This pre-amp was used for measurements at 2 GHz and higher.

Figure 4.45. 1-tone power test set-up diagram

Table 4.3. Equipment list for 1-tone power test.

Function Name	Equipment Model	Frequency Band	RF Power Rating	DC Rating
Vg Supply	HP E3630A	DC	N/A	0 V – 6 V, 2.5 A /0 V - ± 20 V, 1.5 A
Vdd Supply	Agilent E3646A	DC	N/A	0 V – 8 V, 3 A /0 V – 20 V, 1.5 A
BT1 Bias Tee	Picosecond 5541A-104	80 KHz – 26 GHz	1 W max.	50 V, 100 mA max.
BT2 Bias Tee	Mini-Circuit 15542 ZFBT-4R2GW	0.1 MHz – 4.2 GHz	30 dBm max.	30 V, 500 mA max.
	HP 11590A OPT1001	1 GHz – 18 GHz	24 dBm max.	100 V, 500 mA max.
RF Probes	GGB Model 40A-GSG-150P	DC – 40 GHz	N/A	N/A
Attenuator	Mini-Circuits 15542 VAT-6 6dB Attenuator	DC – 6 GHz	1 W Rating	No DC Feeding
Pre-Amp	HP 8349B	2 GHz – 20 GHz	38 dBm max.	N/A
Signal Generator	Agilent E8257C PSG	250 KHz – 40 GHz	23 dBm Rating	N/A
Spectrum Analyzer	HP 8593E	9 KHz – 26.5 GHz	30 dBm max.	0V

Then, 2-tone measurement, which is used for the linearity evaluation of the IIP3 and OIP3, is presented. The set-up diagram and the equipment list are shown in Figure 4.46 and Table 4.4 respectively.

Figure 4.46. 2-tone measurement set-up diagram.

Table 4.4. Equipment list for 2-tone measurement.

Function Name	Equipment Model	Frequency Band	RF Power Rating	DC Rating
Vg Supply	HP E3630A	DC	N/A	0 V – 6 V, 2.5 A /0 V – ±20 V, 1.5 A
Vdd Supply	Agilent E3646A	DC	N/A	0 V – 8 V, 3 A /0 V – 20 V, 1.5 A
BT1 Bias Tee	Picosecond 5541A-104	80 KHz – 26 GHz	1 W max.	50 V, 100 mA max.
BT2 Bias Tee	Mini-Circuit 15542 ZFBT-4R2GW	0.1 MHz – 4.2 GHz	30 dBm max.	30 V, 500 mA max.
	HP 11590A OPT1001	1 GHz – 18 GHz	24 dBm max.	100 V, 500 mA max.
RF Probes	GGB Model 40A-GSG-150P	DC – 40 GHz	N/A	N/A
Attenuator	Mini-Circuits 15542 VAT-6 6dB Attenuator	DC – 6 GHz	1 W Rating	No DC Feeding
Pre-Amp	HP 8349B	2 GHz – 20 GHz	38 dBm max.	N/A
Signal Generator1 Signal Generator2	Agilent E8257C PSG	250 KHz – 40 GHz	23 dBm Rating	N/A
Spectrum Analyzer	HP 8593E	9 KHz – 26.5 GHz	30 dBm max.	0V

4.3 Measurement Results and Analysis of the Fabricated Chip

As introduced in the above section, practical measurements are conducted at DC and at RF separately. The available DC supply does not allow pulsed-mode measurements for the I-V curves. The RF performance is obtained without any signal modulation (i.e. CW mode only).

The measured $I_d - V_{ds}$ curves are shown in Figure 4.47, along with the corresponding simulated curves of the 3 parallel connected transistors. As can be seen, the measured curves show a roll-off in the saturated region which is not found in simulations. This is believed to be due to the trapping effect and mainly the thermal effect as explained in detail in Chapter 2. The thermal effect is present because of the high temperature obtained through DC power dissipation within the device. With reference to Figure 4.47, when the ADPA is biased at $V_{gs} = -2.5$ V, and $V_{ds} = 15$ V (nominal bias conditions), the total drain current is approximately 300 mA, which results in 4.5 W of internal power dissipation; compared to simulation, the measured current is approximately 100 mA down from the simulated curve in the saturated region. It should be noted that, the gate length of the GaN HFET is 500 nm, which is a short gate device. When such a device operates at high temperature, its electron mobility and saturation velocity are reduced from their corresponding low temperature values. While such an effect might be detrimental to RF performance, it is believed that the RF simulation model takes heating into account and the result is satisfactory (as seen in the following RF measurements).

Figure 4.47. Measured and simulated DC I_{ds} - V_{ds} curves
(V_{gs} values are indicated for simulated curves).

Now the measured RF performance is presented. Small-signal s-parameters are measured by using an s-parameter network analyzer, as introduced in the above section. Figure 4.48 presents the measured and simulated small-signal s-parameter results. The measured gain (S_{21}) is approximately 13 dB over the operating band of 0.5 GHz – 6.5 GHz. It is also approximately 2.8 dB lower than the simulation data over the entire 0.1GHz – 10GHz range, which is believed to be due to the self-heating effect that is not taken into account during small-signal simulations. None the less, the measured gain is above the 10 dB specification over the operating frequency range of the ADPA. From the remaining S_{11} and S_{22} curves shown in Figure 4.48, it can be seen that, the measured performance matches well with the simulation data. The ADPA is found to be well matched at the input port ($S_{11} < -10\text{dB}$) from 0.5GHz –

6.5GHz, and the shifts in S_{11} and S_{22} minima are likely due to some parasitic components of the ADPA which are not fully captured during simulation (such as grounding through bond-wires, probe to pad parasitic, etc.). Also, some jitter in the small signal plots exist at the frequencies above the 0.1 GHz – 6.5 GHz bandwidth, which is due to the high frequency limitations of the cables used in the test system. For instance, the available cables employed in the test set up are limited to testing at frequencies below 6 GHz.

To support the investigation of the jittery behavior, the s-parameters of Thru on the Impedance Standard Substrates (ISS) were measured after the network analyzer was calibrated. The results are shown in Figure 4.49. The small jitter shows up, and the return loss S_{11} and S_{22} are above -40 dB from 6 GHz to 10 GHz, which would not be expected with perfect cables. Meanwhile, there exists a difference between S_{11} and S_{22} as a function of the frequency, which is another indication of the imperfection in the cables. All these phenomena related to the performance of the cables used in the test system confirm that the accuracy is limited beyond 6 GHz.

Figure 4.48. Measured and simulated small-signal s-parameters for three samples
(at nominal bias condition of $V_{gs} = -2.5$ V, $V_{dd} = 15$ V).

Figure 4.49. S-parameters of ISS Thru after the network analyzer calibrated.

Further characterization of the ADPA consisted of sweeping V_{gs} while keeping V_{dd} at 15 V to investigate the relationship between the small signal voltage gain (S_{21}) and V_{gs} . The measured results of one sample are shown in Figure 4.50 (the results from two other samples were virtually identical and are not shown). The maximum gain is obtained when the device is biased at $V_{gs} = -3.5$ V, $V_{ds} = 15$ V in simulation, but at $V_{gs} = -3$ V, $V_{ds} = 15$ V in measurement. The input and output return loss S_{11} and S_{22} present no big difference among various bias conditions over the operating frequency range, and the best matching is obtained while the maximum gain is achieved. The difference between optimum simulated and optimum measured V_{gs} values is due to the DC modeling limitation of the HFETs, and is confirmed by the results in Figure 4.47.

Figure 4.50. Small-signal s-parameters of chip sample 1 while sweeping V_{gs} .

According to the comparison of measured and simulated s-parameters as shown above, cable characterization and power compensation should be done in the following power test, where the apparatus has no test system calibration function. The power meter was used in the cable loss data collection. By comparing the power level of the signal generator direct output and at the end of the cables, the power loss of the cables were derived and recorded as the reference of the ADPA output power correction.

Then, the RF power measurements such as P_{out} , PAE, P_{1dB} , and linearity are presented while input power $P_{in} = 20$ dBm at $V_{gs} = -2.5$ V and $V_{ds} = 15$ V.

Figure 4.51 shows the simulated and measured output power as a function of frequency over the 0.5 GHz – 7 GHz frequency range. As expected, the measured P_{out} is above 30 dBm over the entire 0.5 GHz – 6.5 GHz operating bandwidth, and the peak output power of approximately 33 dBm is achieved. Meanwhile, a slight difference (minimum 0.5 dBm) between the measured results and simulated data is observed, and it is changing as a function of frequency. The larger difference exists at frequencies higher than 5 GHz; this could be due to the self-heating effect and unaccounted for measurement error (e.g. random losses of cables).

Figure 4.51. Measured and simulated P_{out} of the ADPA for $P_{in} = 20$ dBm at nominal DC bias.

Figure 4.52. Measured and simulated PAE of the ADPA for $P_{in} = 20$ dBm at nominal DC bias.

Figure 4.52 presents the PAE performance comparison between the simulation and measurement. Due to the fact that the measured current consumption is lower than the value simulated, the measured PAE is above the simulation results at frequencies lower than 1 GHz, but beneath over 1 GHz – 7 GHz. A peak PAE of 38.1% is achieved at 0.5 GHz, while greater than 20 % PAE is obtained over 0.5 GHz – 6.5 GHz operating frequency of the ADPA. As described in Chapter 2, PAE is calculated as $PAE = (P_{out} - P_{in})_{RF}/P_{DC}$. For example at 2 GHz, $P_{out} = 32$ dBm, $V_{ds} = 15$ V, $I_d = 331$ mA, $P_{in} = 20$ dBm to yield $PAE = 30\%$.

In Figure 4.53, measured output power is given as a function of input power at nominal bias, 2GHz. 1-dB compression point P_{1dB} and saturated power P_{sat} of approximately 31.3 dBm and 32.2 dBm are indicated in the plot respectively, according to the concept introduced in Chapter 2. As shown in Figure 4.54, a peak saturated output power of approximately 34 dBm is measured at about 0.7 GHz, while a saturated power P_{sat} above 30 dBm is obtained over the 0.5 GHz – 6.5 GHz frequency range. The measured P_{sat} is approximately 1 dBm down from the simulation, and the difference is observed as a function of frequency, which could be due to the self-heating effect and the error introduced in the measurement.

Figure 4.55 illustrates the comparison between measured 1-dB output power and simulated P_{1dB} as a function of frequency over 0.5 GHz – 7 GHz range. A peak measured P_{1dB} of approximately 32.6 dBm is obtained at 0.5 GHz, while more than 30 dBm is measured over 0.5 GHz – 6.5 GHz bandwidth. The two curves match each other very well, and the simulated curve shows a higher P_{1dB} over the measured one at

most frequencies. A maximum difference of only approximately 1 dB is observed.

The measured OIP3 and IIP3 are given in Figure 4.56 for the two input tones separated by 10 MHz from the design center frequency. The definitions of OIP3 and IIP3 are given in Chapter 2. The measured OIP3 and IIP3 are above 40 dBm and 25 dBm respectively over the entire 2 GHz – 6.5 GHz frequency range, which indicates that the ADPA is operating in the linear region with a nominal input of 20 dBm RF power over the operating frequency bandwidth. The simulation results are in reasonable agreement.

Figure 4.53.

Measured P_{out} at nominal bias, 2GHz

Figure 4.54. Measured and simulated P_{sat} at nominal bias.

Figure 4.55. Measured and simulated P_{1dB} at nominal bias.

Figure 4.56. Measured and simulated linearity at nominal bias.

As can be seen from the above plots, although there are some differences between the measurement results and the simulation data which are caused by self-heating and systematic errors during testing, the measured data successfully matches the simulated results.

To further, characterize the ADPA and determine the highest possible PAE, the bias point is swept together with input power. For instance, power performance is plotted at $V_{gs} = -3.5$ V, $V_{ds} = 15$ V, when sweeping the input power at 3 GHz, as shown in Figure 4.57; the maximum PAE is obtained at approximately 20.5 dBm of input power, just before reaching the peak P_{out} . When biasing the device into deeper saturation, lower PAE would be expected. At the operating condition of the peak PAE, more than 10 dB gain and 30 dBm P_{out} are obtained.

Figure 4.57. Measured power results biased at $V_{gs} = -3.5$ V, $V_{ds} = 15$ V, $I_d = 189$ mA, 3GHz.

Figure 4.58 shows the power and linearity performances while sweeping bias voltages (both V_{gs} and V_{ds}) along with the input power (P_{in}) at 3 GHz. Since multiple combinations of bias and input power are driven by different drain DC currents, the various power levels are plotted versus drain DC current. A maximum PAE of approximately 40 % is achieved when 20 dBm of input power is supplied to the ADPA which draws about 190 mA DC current from the voltage source at $V_{gs} = -3.5$ V, $V_{ds} = 15$ V (point 1). Also, a gain of 10 dB, more than 30 dBm P_{out} and about 38 dBm OIP3 are obtained at this operating condition. A peak gain of 12.9 dB is obtained when the ADPA is biased at $V_{gs} = -2.5$ V, $V_{ds} = 20$ V ($I_d = 270$ mA), with $P_{in} = 18$ dBm (point 2). The maximum output power of 32.5 dBm is achieved at $V_{gs} = -3$ V, $V_{ds} = 20$ V ($I_d = 245$ mA), $P_{in} = 21$ dBm, indicated as point 3 in Figure 4.57.

Figure 4.58. Measured: power and linearity performances versus drain current (sweeping bias voltage and input power at 3 GHz).

Figure 4.59 to Figure 4.61 illustrates the performance comparison between the original operating condition and the maximum PAE operating condition. The ‘operating condition’ mentioned does not only adjust the gate and drain bias voltages, but also the input power level. It can be seen that the output power for the two conditions (nominal and maximum PAE) described is almost the same over 0.5 GHz – 5GHz range, and the P_{out} of the ADPA when working at maximum PAE is above 30 dBm over the entire 0.5 GHz – 6.5 GHz bandwidth. From Figure 4.60, a maximum PAE of approximately 52.7 % is obtained at 0.5 GHz. The PAE is higher than 40 % up to 2 GHz, and larger than 30 % up to 3.5 GHz, while the PAE curve is above 20% over 0.5 GHz – 6.5 GHz bandwidth. The bias voltage comparison between the original and maximum PAE condition is illustrated in Table 4.5.

The gain and linearity of these two operating conditions are checked as shown in

Figure 4.61. A higher gain of approximately 1 dB above the gain of the ADPA working at the original operating condition is obtained, when the ADPA is operating at the maximum PAE condition over the entire 0.5 GHz – 6.5 GHz bandwidth, while linearity is decreased as a trade-off. However, the IIP3 and OIP3 of the ADPA working at the maximum PAE condition is higher than 20 dBm and 30 dBm respectively over 0.5 GHz – 6.5 GHz frequency range, which means the ADPA will operate in linear mode in applications requiring high PAE.

Figure 4.59. Measured P_{out} comparison of the maximum PAE and the original operating condition.

Figure 4.60. Measured PAE comparison of the max. PAE and the original operating condition.

Table 4.5. Measured bias voltage difference between original and maximum PAE operating condition.

Frequency (GHz)	Max. PAE condition		Original condition	
	$V_{gs}(V)$	$V_{ds}(V)$	$V_{gs}(V)$	$V_{ds}(V)$
0.5	-3.5	15	-2.5	15
1	-3.5	15	-2.5	15
1.5	-3.5	15	-2.5	15
2	-3	15	-2.5	15
2.5	-3.5	15	-2.5	15
3	-3.5	15	-2.5	15
3.5	-3.5	15	-2.5	15
4	-3	15	-2.5	15
4.5	-3	15	-2.5	15
5	-3	15	-2.5	15
5.5	-3.5	15	-2.5	15
6	-2.5	15	-2.5	15
6.5	-3	15	-2.5	15
7	-2.5	20	-2.5	15

Figure 4.61. Measured: linearity comparison of the maximum PAE and the original operating condition.

4.4 Conclusion

In this chapter, an asymmetric distributed power amplifier has been successfully designed and measured using the 0.5 μ m GaN process and results presented in detail. A maximum PAE of 52.7 % is obtained at 0.5GHz with an output power of 31.8 dBm in peak PAE bias condition. Meanwhile, more than 10 dB of gain, a P_{out} of 30 dBm and a PAE better than 20 % are achieved over the 0.5 GHz – 6.5 GHz bandwidth. Also the PAE is higher than 40 % for frequencies up to 2 GHz, and higher than 30% for frequencies up to 3.5 GHz. By comparing the measurement and simulated data, the models are found to be reliable. Some of the findings can be followed up in the future, such as reconfigurable matching network for specific multi-band applications.

Chapter 5: Investigation of Possible ADPA Performance Enhancements

The proposed ADPA was fabricated and successfully measured, and then, some prototype of possible ADPA performance enhancements will be investigated in this chapter. Only simulated results will be presented here.

5.1 Improved Design with High Q Type Interconnections

As discussed in Section 4.1.3, type 3 inductor design of interconnects was chosen because of its larger current carrying capacity, although it is not the highest Q option. The limitation of its Q factor would cause power losses when current passes through the lossy interconnections. The effect of using inductors with higher Q is studied here while keeping the circuitry and parameter values the same as the first tapered ASPA design presented above.

As shown from Figure 5.1 to Figure 5.4, type 1 implementation is proven with simulation to obtain the highest Q among the three layout implementations for the chosen inductances of the different gate and drain sections.

Figure 5.1. Simulated Q comparison of three types of L_{d1} .

Figure 5.2. Simulated Q comparison of three types of L_{g1} .

Figure 5.3. Simulated Q comparison of three types of L_{g2} and L_{g3} .

Figure 5.4. Simulated Q comparison of three types of L_{g4} .

Figure 5.5. Simulated performance comparison of the ADPA implemented with the low Q type interconnections and the high Q type interconnections.

Figure 5.5 presents the difference of the power performance between the original ADPA using the low Q but the largest current carrying capacity type interconnections, and the improved one which uses the highest Q type interconnections. The difference in the gain and output power performance is minimal, while PAE is improved largely from 4 GHz to 8 GHz. A maximum improvement in the PAE of 1% is observed at 6.5 GHz, and a maximum PAE of approximately 25% is obtained at 6.5 GHz.

5.2 Improved Tapered Line Asymmetric Distributed Power Amplifier

The initial tapered lines, based on calculation results in the first ADPA design, caused power loss, as the HFET in the 2nd gain stage has not seen its optimum load of the maximum power (see Section 4.1). In this section, we optimize the drain line sections in the 2nd ADPA design for higher output power.

First, we go back to the lumped element circuit simulation. The load pull simulation set-up used (ADS) can only change the load impedance (R_{load}), but has nothing to do with the drain line tapering. Based on the load pull theory, a manual load pull was performed by sweeping the inductances of the three drain line sections together with the R_{load} at three different frequencies, 6 GHz, 4 GHz and 2 GHz, as shown in Figure 5.6 – 5.8. Following the simulation results, the values of the drain line sections and R_{load} should be $L_{d1} = 2.25$ nH, $L_{d2} = 1.5$ nH, $L_{d3} = 0.2$ nH, and $R_{load_opt} = 50 \Omega$ respectively for the maximum output power.

Figure 5.6. Simulated manual load pull at 6 GHz. (a) Peak P_{out} contour.

Figure 5.6.(continue)

Simulated manual load pull at 6 GHz.

(b) peak PAE contour.

Figure 5.7.

Simulated manual load pull at 4 GHz.

(a) Peak P_{out} contour.

Figure 5.7.(continue)

Simulated manual load pull at 4 GHz.

(b) peak PAE contour.

Figure 5.8.

Simulated manual load pull at 2 GHz.

(a) Peak P_{out} contour.

Figure 5.8.(continue)

Simulated manual load pull at 2 GHz.

(b) Peak PAE contour

Then, the L_{d2} drain line's estimated inductance of 1.5 nH is chosen as the start point of re-sizing. As shown in Figure 5.9, this new implemented L_{d2} drain section has an inductance of 1.5nH as expected, and a maximum Q of approximately 13, with no indication of self-resonating effects below 6GHz. The drain line presents approximately 10Ω impedance to the HFET's drain side as expected.

Figure 5.10 shows the phase shift comparison between the implemented L_{g3} and L_{d2} (according to Figure 4.14) in the 2nd ADPA design. A maximum phase shift difference of approximately 14° at 6.5 GHz is observed, and a phase shift difference lower than 10° is obtained over 0.1 GHz – 4 GHz frequency range.

Figure 5.9. The 2nd drain line section characteristics of 2nd design.

Figure 5.10. Phase matching of 2nd stage of 2nd design.

Figure 5.11. Load Pull simulation of the 2nd ADPA Design at 2 GHz, 4 GHz and 6 GHz.

Next, a load pull analysis is performed as shown in Figure 5.11, in order to optimize both the peak P_{out} and the peak PAE. A $50\ \Omega$ load can be used for the maximum output power over the frequency range we want to improve. Figure 5.12 presents a layout of 2nd ADPA design.

Stability is first checked as shown in Figure 5.13, where B1 values are in the range of 0 to 1, and K is higher than 1, to ensure unconditional stability, meaning the 2nd ADPA design will not oscillate.

Figure 5.14 shows the large-signal s-parameters simulation results over 0.1 GHz – 10 GHz range. The voltage gain (S_{21}) is above 10 dB up to 8 GHz, while return loss S_{11} at the input port is beneath -10 dB over 0.5 GHz – 10 GHz. The S_{12} is lower than -20 dB over the entire 0.1 GHz – 10 GHz band which shows great isolation between the input and output of the ADPA. As seen from the S_{22} curve, the output is not well matched to $50\ \Omega$ but this is not required as the PA is designed for maximum output power delivery.

Figure 5.15 illustrates the simulated power performance of the 2nd ADPA design. It can be seen from the curves that the specification of $P_{out} > 30\ \text{dBm}$, $\text{PAE} > 20\ \%$, and $\text{gain} > 10\ \text{dB}$ are achieved over the entire 0.5 GHz – 6.5 GHz operating bandwidth. The PAE at frequencies above 5 GHz is increased compared to the first ADPA design. In order to investigate the details, the first and 2nd ADPA simulated performance is compared in Figure 5.16.

Figure 5.12. Layout of the 2nd ADPA Design
(HFETs are indicated with the rectangle).

Figure 5.13. Stability.

Figure 5.14. Large signal s-parameters.

Figure 5.15. Power gain, delivered power, PAE.

Figure 4.76 presents the comparison of the simulated power performance between the first ADPA design and the 2nd one over 0.5 GHz – 6.5 GHz bandwidth. It can be seen that at 3 GHz, the 2nd ADPA design obtains higher gain and output power, and especially greater PAE than the first design at frequencies above approximately 2.5 GHz. The P_{out} increases approximately by 0.6 dBm, and a maximum PAE improvement of 4 % is obtained in the 2nd ADPA design at frequencies higher than 2.5 GHz, while the performance at lower frequencies has minimal differences. Such a performance improvement acquired at high frequencies is in the trade-off of the linearity.

Figure 5.16. Simulated linearity and power performance comparison between first and 2nd design over 0.5 GHz to 6.5 GHz frequency range.

As shown in Figure 5.17, the fundamental element of the output power is above 30 dBm over the entire 0.5 GHz – 7 GHz frequency range as expected. Meanwhile, the

2nd and 3rd harmonics are approximately 18 dB down from the fundamental. The need for harmonic shorting circuitry would depend on the application.

Figure 5.17. Simulated harmonic distortion over 0.5 GHz – 7 GHz bandwidth.

5.3 Constant Output Broadband ADPA

In the preceding sections, we have reviewed the first ADPA, high Q interconnection effect, and the 2nd (improved) ADPA design. The drain lines are tapered for maximum output power, and all optimizations are based on the Z_{opt} data obtained from simulation of the unit gain stage at 4 GHz, which is the middle of the bandwidth.

In the optimization of the first design section, load pull simulations at 3 different frequencies were carried out (i.e. 2 GHz, 4 GHz and 6 GHz), and R_{opt} was measured as a function of frequency. If we terminate the output of the ADPA to the optimum impedance R_{opt} , which can balance the performances at low and high frequencies in the entire 0.5 GHz – 6.5 GHz bandwidth, a constant output power and PAE could be implemented.

In this section, a design based on constant output and implemented using the first ADPA circuitry is presented. As shown from Figure 5.18 to Figure 5.20, the R_{load} value is swept as a function of the frequency while checking the power amplifier performance in terms of its P_{out} , PAE and power gain. The load value of $20\ \Omega$ is indicated for the constant output power performance, and the output power, PAE and gain are above 32 dBm, 20% and 10 dB respectively over the 0.5 GHz – 7GHz frequency range.

Figure 5.18. P_{out} : sweeping R_{load} as a function of the frequency.

Figure 5.19. PAE: sweeping R_{load} as a function of the frequency.

Figure 5.20. Power Gain: sweeping R_{load} as a function of the frequency.

Then a broadband matching network is implemented to transform $20\ \Omega$ to $50\ \Omega$ load. Again, tapered line technology is chosen. However, the long length of the tapered line broadband matching network should be considered in practice because of substrate cost. In other words, LTCC (Low Temperature Co-fired Ceramic) or other packaging substrate could be a low cost choice for output matching network. An $19\text{ mm} \times 1.15\text{ mm}$ output matching network is implemented in simulation, as shown in Figure 5.21. The small signal s-parameter performance of this matching network is shown in Figure 5.22. This optimized network has a return loss S_{11} and S_{22} lower than -10 dB over the $0.1\text{ GHz} - 11\text{ GHz}$ range and an insertion loss lower than 2 dB .

Figure 5.21. Tapered line broadband matching network.

Figure 5.22. S-parameter performance of tapered matching network.

The simulated performance of the first ADPA design with output broadband matching network is presented from Figure 5.23 to Figure 5.25.

Figure 5.23 shows that the stability factor K and $B1$ ensure unconditional stability. Large signal s-parameters are illustrated in Figure 5.24. Gain (S_{21}) is above 10 dB over 0.1 GHz – 7.8 GHz frequency range, while S_{12} is below -20 dB across the bandwidth confirming the good isolation between input and output ports. The ADPA is seen to be well matched ($S_{11} < -10$ dB) from 0.5 GHz to 8 GHz, when the output is matched to 20 Ω instead of 50 Ω load. It can be seen from Figure 5.25 that the APDA obtains a maximally flat output power and PAE compared to the first ADPA design, in the trade-off of power decreasing at frequencies lower than 2 GHz. A PAE difference of 2.2% is observed while the output power difference is only 0.5 dBm, with 0.3 dB power gain variation over 2 GHz – 6 GHz frequency range. The

performance observed here is better than that of commercial devices normally having 2 dBm P_{out} difference over the operating frequency range, although causing losses in the output power at low frequencies. Based on the investigation, it is believed that the best way to optimize this design is to separate the entire 7 GHz band into several small bands, and present a reconfigurable matching network for individual applications.

Figure 5.23. Stability.

Figure 5.24. Large signal s-parameters.

Figure 5.25. Simulated power performance comparison between first and constant output design (first ADPA + broadband matching network).

5.4 Cascode Tapered Line ADPA

The cascode ADPA is investigated for potential enhancement of the power amplifier. The inductances of the three drain line sections and the load impedance are swept for maximum P_{out} at 4 GHz as shown in Figure 5.26. The maximum simulated PAE is lower than 16 % due to the existence of an upper cascode transistor, although a cascode topology could obtain higher power gain (maximum achievable gain of 15.6 dB) than the common-source amplifier (10 dB – 13 dB). The maximum achievable output power is approximately 31dBm. Because of the low PAE and not so attractive output power performance, a cascode structure is not chosen for this work. However, successful implementations of cascode or dual gate GaN devices are reported by other researchers as shown in reference [9], [10] and [11].

Figure 5.26. Available power performance of cascode stage at 4 GHz.

5.5 Stage-Scaled ADPA

As presented in Chapter 3, a Stage-Scaled uniform distributed power amplifier in 0.13 μm SiGe BiCMOS was reported [6] [43], and the key principle is sizing the transistors together with the drain lines in order to maintain the same output power as the unit gain stage. In this section, we will introduce the implementation of this technology in 0.5 μm GaN process.

First, the drain lines and HFETs are re-tapered. Because of the modeled HFET size limitation in the available kit, the drain lines are tapered using a factor $z = 3/2$, and the HFETs are sized using a factor $k = 3/5$ in the same direction simultaneously to yield a product k^2z of 1.35 which is slightly different from the $k^2z = 1$ requirement in the uniform stage-scale DPAs [6] [43]. Meanwhile, the phase shift of the first and 2nd drain line sections are adjusted to match the gate lines, which is about 18° at 4GHz. The L_{d3} of the first design is kept constant as the last stage in the stage-scaled ADPA design, and the new L_{d1} and L_{d2} are tapered as shown in Figure 5.27. Approximately 1.5 nH and 1 nH inductance are obtained for L_{d1} and L_{d2} respectively, while a maximum Q of 12.5 and 10 are achieved. In that, R_{opt} of the maximum output power is presented at the drain side of each unit gain stage. A maximum of 14° and 2° phase shift gap is observed between L_{d1}/L_{d2} and L_{g2}/L_{g3} respectively over the 0.1 GHz – 6 GHz frequency range as seen in Figure 5.28, which means the phase shift is well matched in the 2nd gain stage. The layout of the first stage-scaled ADPA (SSADPA) is presented in Figure 5.29; the width of the HFETs are 150 μm , 250 μm and 400 μm respectively.

Figure 5.27. Characteristics of L_{d1} and L_{d2} in the stage-scaled ADPA design.

Figure 5.28. Phase synchronization in SSADPA design.

Figure 5.29. 2 mm $\times$ 2 mm Stage-Scaled ADPA layout using original uniform principle which presented in [6] [43].

After that, we re-investigate the design principle of the stage-scaled ADPA. The basic principle here is to size all devices to achieve maximum P_{out} at each gain stage. So, the impedance fed by the drain lines in each stage was readjusted, and the sizes of the HFETs were calculated as 400 μm , 150 μm and 250 μm from the last to the first stage, to ensure the device's R_{opt} produces P_{out} of different size HFETs. The layout of the 2nd SSADPA design is shown in Figure 5.30.

Figure 5.30. 2 mm $\times$ 2 mm Stage-Scaled ADPA layout modified for maximum P_{out} of each stage.

A load pull simulation is then performed for both SSADPA designs; the results are similar. The load pull results of the first SSADPA design using the original principle of uniform stage-scaled DPA is illustrated in Figure 5.31 when the SSADPA is biased at $V_{gs} = -2.5$ V, $V_{ds} = 15$ V. It can be concluded from the plots, if the output of the

SSADPA matches to approximately $30\ \Omega$ over a broad-band from 0.1 GHz to 10 GHz, the P_{out} and other power performances could be maximally flat over the -3 dB gain bandwidth.

Figure 5.31. Load Pull simulation of the stage-scaled ADPA design using the original uniform SSDA principle. (a) at 2 GHz; (b) at 4 GHz.

Figure 5.31. (continued). Load Pull simulation of the stage-scaled ADPA design using the original uniform SSDA principle. (c) at 6 GHz.

Finally, the power performance of the two stage-scaled ADPAs, where the output is matched to $30\ \Omega$ via a $30\ \Omega$ to $50\ \Omega$ broadband matching network, is compared to the first ADPA design and is presented in Figure 5.32 and Figure 5.33. The ADPA is normally biased at $V_{gs} = -2.5\text{ V}$, $V_{ds} = 15\text{ V}$. Two designs are studied and both achieve a 3 dB gain bandwidth of 0.5 GHz – 10 GHz (3.5 GHz broader than the fabricated ADPA), and deliver 30 dBm RF power over the entire bandwidth. A difference of approximately 1.31 dB and 8.27 % variations are observed in the P_{out} and PAE respectively over the 0.5 GHz – 6 GHz bandwidth in the first SSADPA design, and 1.02 dB and 7.29 % differences are obtained in P_{out} and PAE over the 0.5 GHz – 9 GHz range in the 2nd SSADPA design. The smallest P_{out} divergence of 0.17 dBm and 1.90 % PAE variation are achieved over 1 GHz – 7.5 GHz frequency range in the 2nd SSADPA design, while P_{out} , gain and PAE are above 30 dBm, 10 dB and 20% respectively over the entire bandwidth.

Figure 5.32. EM simulated power performances comparison between the first SSADPA design (with output matching network) and the first ADPA design (fabricated).

Figure 5.33. EM simulated power performances comparison between the 2nd SSADPA design (with output matching network) and the first ADPA design (fabricated).

5.6 Conclusion

In this chapter, two SSADPA designs were introduced by optimization of simulated results. A difference of 0.17 dB in the P_{out} and a PAE variation of 1.90 % are achieved over the 1 GHz – 7.5 GHz range. The gain was found to offer trade-offs versus broader bandwidth; 10 dB-gain and 30dBm P_{out} are obtained over the 0.5 GHz – 10 GHz bandwidth. In future designs, there is still much work needed to follow up on some of the findings presented here, such as optimizing the SSADPA for higher output power.

Chapter 6: Conclusion and Recommendations

After presentation of the design work, accomplishments, issues in the design and future work of this work are concluded in this chapter.

6.1 Accomplishments

This work has designed a novel high efficient, linear and ultra-broadband GaN MMIC Power Amplifier. This work is the first time an active MMIC device using CPFC GaN500 process is presented. The power amplifier was successfully measured and the results verify the process is reliable. A Tapered Line Asymmetric Distributed structure was successfully implemented, and the ADPA's measured performance demonstrated superior PAE, Gain, BW product for 1 W output power.

Over the frequency range from 0.5 GHz to 6.5 GHz, more than 10 dB of gain is obtained, and more than 1 W linear RF power is delivered. Among all the performance parameters studied, a high PAE is most attractive. A peak PAE of 38.1 % is achieved at 0.5 GHz in linear mode, and up to 52.69 % PAE can be obtained while 31.79 dBm RF power is delivered. The performance of this ADPA is compared with other designs listed in Table 6.1.

Two IEEE journal papers are ready to be submitted based on the work of this thesis.

Table 6.1. Performance Comparison

Tech.	DC supply (V)	Freq. (GHz)	Peak PAE (%)	Pout (dBm)	Size (mm×mm)	Ref
0.35μm CMOS	2	2.4	41	33	1.3 × 2	16
	2.5	2.4	31	33.4	1.3 × 2	17
0.35μm SiGe	3.3	0.6 – 2.4	45	17	1 × 0.8	15
0.2μm GaN	15	0.05 – 12.3	20	30	1.7 × 1.7	20
GaAs	10	2 - 18	23	30.5	3.6 × 1.7	25
0.2μm GaN	15	2 - 18	15	33	4 × 2	19
GaN HEMT	30	3 – 4/ 9.5 - 12	28	37.8	4 × 2.8/ 3.5 × 2.8	18
0.5μm GaN	15	0.1 - 5	30	30	2.2 × 4.3	21
0.7μm GaN	19	0.2 – 7.5	15	31.5	-	9
0.5μm GaN	15/20	0.5 – 6.5	52.7	> 30	2 × 2	This work

Several ADPA enhancements were investigated via simulation. Among those possible enhancements, a stage-scaled ADPA obtained superior constant output power over a 10 GHz bandwidth. A 0.17 dB P_{out} difference and 1.90 % PAE variation were achieved over 1 GHz – 7.5 GHz frequency range, while 10 dB of gain and 30 dBm P_{out} were obtained over 0.5 GHz – 10 GHz bandwidth.

6.2 Issues in the Design

Although loose dies are mounted on a metal plate while under CW test, the self-heating effect caused by GaN semiconductor characteristics is still a challenge in our design. Such effects are under-estimated in the HFET model which leads to over performance in the DC I-V curve and Harmonic Balance simulation compared to

measurements in CW-mode. The current decreases very rapidly during measurements when increasing the bias voltage, which results in performance differences between the measured data and the simulation. We suggested that there are two solutions. The first is to improve the GaN HFET model. The second is to test the device in Pulse-mode; the idea comes from the truth that the PA is only consuming high current in a very short time period.

Further, the RF cables available in the Department of Electronics of Carleton University are limited in frequency, which results in jittery data curves captured in the measurement. Furthermore, the test equipment is limited. For example, the signal generators have not been calibrated for over two years (they should be calibrated annually by the manufacturers). Also, a modulated signal source is not available.

6.3 Future Work

Higher PAE and output power can be obtained by more carefully sizing tapered lines, and linearity can be improved in maximum PAE mode. A stage-scale structure could be optimized especially when much more different size HFETs are modeled. A voltage controlled switchable matching network could be added to the ADPA's input and output side for better performance, such as harmonic distortion, matching and power gain in specific applications. Also, the other blocks such as digital voltage controller, LNA, Up/Down Converter (mixer) and PLL which were presented in Chapter 2.5 should be implemented. And finally, all of those blocks should be combined as a practical transceiver.

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Appendix

GaN500 (CPFC) Process

A simplified cross-section of the HFET is shown as Figure A.1, with approximate dimensions.

Source contact	Gate	Drain contact
AlGa _N		20nm
AlN		1nm
GaN		200nm
GaN buffer layer		2um
Nucleation layer		20nm
Sic substrate		

Figure A.1 Cross-section of HFET, showing contacts and epitaxial structures

The Gallium Nitride (GaN)/ Aluminium Gallium Nitride (AlGa_N) epitaxial layers are grown on Silicon Carbide (SiC) wafers. The thickness of the SiC substrate is 400um with 12.5 % fabrication tolerance. In simulation, we used the recommended model for the substrate MSUB (Microstrip Substrate) in ADS. The set-up of GaN500 process is listed in Table A.1, and the ADS MSUB substrate model structure is illustrated in Figure A.2.

Table A.1 ADS substrate model MSUB definition of GaN500 process

Name	Description	Value	Units
H	Substrate thickness	400	micron
Er	Relative dielectric constant	9.5	none
Mur	Relative permeability	1	none
Cond	Conductor conductivity	4.10E+07	S/meter
Hu	Cover height	3.90E+34	none
T	Conductor thickness	1	micron
TanD	Dielectric loss tangent	0	none
Rough	Conductor surface roughness	0	Mil
DielectricLossModel	Model for calculating dielectric loss	1	none
FreqForEpsrTanD	Frequency at which ϵ_r and TanD are specified	1.00E+09	Hz
LowFreqForTanD	Low roll-off frequency for TanD	1.00E+03	Hz
HighFreqForTanD	High roll-off frequency for TanD	1.00E+12	Hz

Figure A.2 ADS substrate illustration

Nine layers are defined based on the mask levels used at the various steps of the process as Table A.2.

Table A.2 Process layer

Layer Number	Layer Name
1	MESA
2	OHMIC
3	GATE
4	VIA1
5	NICHROME
6	1ME
7	VIA2
8	BRIDGE
9	2ME

The nine masks listed in Table A.2 are applied to the HFET in the processing sequence. First, the OHMIC mask is used to define the contact area and deposit the contacts. Then contacts are alloyed at high temperature to ensure low resistance contact is made. Second, the MESA mask is used to define the active area of the device, the exposed area will be etched. Then, the GATE mask and VIA1 masks are applied to deposit the gate metal, and a thin layer of dielectric1 together with via1 hole which connect contact to metall is implemented. After that, a thin layer of NICHROME ($50\Omega/\text{sq}$) is formed where resistors are needed. Next, 1ME and VIA2 masks are applied to build metall (1 μm of gold) and a thin layer of dielectric2 together with via2 hole which connect contact to metal2. Finally, the BRIDGE mask and 2ME mask are applied in sequence to define a special resist layer which isolates the 1ME and 2ME layers if there is a requirement, and to deposit the 1 μm thick metal2 gold layer. The layout view of a 2-gate HFET ($2 \times 200\mu\text{m}$) with airbridge is shown as Figure A.3.

Figure A.3 Layout view of 2-gate HFET with airbridge

In ADS, we use 3D EM-Momentum simulator for post-layout simulation. The substrate and layout layers set-up is shown in Table A.3.

Table A.3 ADS 3D EM-Momentum substrate and layout layer set-up of GaN500 process

Substrate Layers				Layout Layers		
Name	Thick [μm]	Permittivity [ϵ_r]	Permeability [μ_r]	Name	Thick [μm]	Conductor [Siemens/m]
air	open	1	1	Strip-2ME	1	$4.1\text{e}7$
air_bridge	1	1	1	VIA2	-	$4.1\text{e}7$
SiN_1	0.1	6.8	1	Strip-1ME	1	$4.1\text{e}7$
SiN_0	0.05	6.8	1	NICHROME	-	$4.1\text{e}7$
GaN	500	9.5	1	VIA1	-	$4.1\text{e}7$
GND	closed	Perfect	-	Strip-MESA	0.1	$3000/\text{sq}$

Because of the properties of the material and process introduced above, high frequency and high voltage performances are observed. According to the NRC measurement data of a $2 \times 40 \mu\text{m}$ HFET device with $0.5 \mu\text{m}$ gate length on PCM standard die, the threshold voltage and the maximum g_m is normally -4.1 V and 0.22

S/mm respectively, and maximum J_{ds} is 1 A/mm, which is suitable for high power design.

In RF performance, f_T and f_{max} values are measured from the extrapolation of H21 (Forward Current Gain) and MAG (Maximum Available Voltage Gain) derived from s-parameter measurements when biasing the HFET at $V_{ds} = 10$ V, $V_{gs} = -2$ V. High frequency applications become possible because of the 75 GHz f_{max} and 27 GHz f_T . In addition, the PCM measurement is performed by selecting nine tiles which form a rectangular array centered on the test wafer. The location of the measured die depends on the dimension of the tile. The PCM measurement is implemented in order to expect fabrication variation in device parameters across the wafer, and from wafer to wafer.

Then, the breakdown voltage of the device is investigated, especially BV_{gdo} . Again, thanks to the measurement data which NRC supplied, on the best wafers BV_{gdo} values in excess of 100 V have been obtained with a gate-drain spacing of 2 μm or greater, which strongly support high voltage applications. Slightly lower breakdown voltages were observed on devices from the same wafer with a gate-drain spacing of 1 μm . However, the results are somewhat variably affected by the process and device leakage current.