DC, RF, AND THERMAL CHARACTERIZATION OF HIGH ELECTRIC FIELD
INDUCED DEGRADATION MECHANISMS IN GAN-ON-SI HIGH ELECTRON
MOBILITY TRANSISTORS

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Matthew Anthony Bloom

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COMMITTEE MEMBERSHIP

TITLE: DC, RF, and Thermal Characterization of High
Electric Field Induced Degradation Mechanisms
in GaN-on-Si High Electron Mobility Transistors

AUTHOR: Matthew Anthony Bloom

DATE SUBMITTED: March 2013

COMMITTEE CHAIR: Dr. Dennis Derickson, Department Chair
Electrical Engineering Department
California Polytechnic State University,
San Luis Obispo

COMMITTEE MEMBER: Dr. Vladimir Prodanov, Assistant Professor
Electrical Engineering Department
California Polytechnic State University,
San Luis Obispo

COMMITTEE MEMBER: Dr. Tina Smilkstein, Assistant Professor
Electrical Engineering Department
California Polytechnic State University,
San Luis Obispo

COMMITTEE MEMBER: Dr. Todd Weatherford, Associate Professor
Naval Postgraduate School
Monterey, California

ABSTRACT

DC, RF, and Thermal Characterization of High Electric Field Induced Degradation Mechanisms in GaN-on-Si High Electron Mobility Transistors

Matthew Anthony Bloom

Gallium Nitride (GaN) high electron mobility transistors (HEMTs) are becoming increasingly popular in power amplifier systems as an alternative to bulkier vacuum tube technologies. GaN offers advantages over other III-V semiconductor heterostructures such as a large bandgap energy, a low dielectric constant, and a high critical breakdown field. The aforementioned qualities make GaN a prime candidate for high-power and radiation-hardened applications using a smaller form-factor. Several different types of semiconductor substrates have been considered for their thermal properties and cost-effectiveness, and Silicon (Si) has been of increasing interest due to a balance between both factors.

In this thesis, the DC, RF, and thermal characteristics of GaN HEMTs grown on Si-substrates will be investigated through a series of accelerated lifetime experiments. A figure of merit known as the critical voltage is explored and used as the primary means by which the GaN-on-Si devices are electrically strained. The critical voltage is defined as the specific voltage bias by which a sudden change in device performance is experienced due to a deformation of the target GaN HEMT's epitaxial structure. Literature on the topic details the inevitable formation of pits and cracks localized under the drain-side of the gate contact that promote electrical degradation of the devices via the inverse piezoelectric effect. Characteristic changes in device performance due to high field strain are recorded and physical mechanisms behind observed degraded performance are investigated.

The study assesses the performance of roughly 60 GaN-on-Si HEMTs in four experimental settings. The first experiment investigates the critical voltage of the device in the off-state mode of operation and explores device recovery post-stress. The second experiment analyzes alterations in DC and RF performance under varying thermal loads and tracks the dependence of the critical voltage on temperature. The third experiment examines electron trapping within the HEMTs as well as detrapping methodologies. The final experiment links the changes in RF performance induced by high field strain to the small-signal parameters of the HEMT. Findings from the research conclude the existence of process-dependent defects that originate during the growth process and lead to inherent electron traps in unstressed devices. Electron detrapping due to high electric field stress applied to the HEMTs was observed, potentially localized within the AlGaIn layer or GaN buffer of the HEMT. The electron detrapping in turn contributed to drain current recovery and increased unilateral performance of the transistor in the RF regime. Thermal experiments resulted in a positive shift in critical voltage, which enhanced gate leakage current at lower gate voltage drives.

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TABLE OF CONTENTS

	Page
LIST OF TABLES	ix
LIST OF FIGURES	x
1 Introduction.....	1
1.1 High Electron Mobility Transistors (HEMTs)	1
1.2 Gallium Nitride (GaN) HEMTs	6
1.3 Research Focus.....	9
2 Device Under Test (DUT) Overview and Test Plan.....	11
2.1 Selected DUT for Experiments	11
2.2 Simulation of the HEMT's DC and RF Characteristics	16
2.2.1 I-V Characteristics	16
2.2.2 Transfer Characteristics	18
2.2.3 S-parameters	20
2.3 DC and RF Characterization of the DUT.....	22
2.4 Test Plans - Overview	29
3 Experiment 1: High Electric Field Stress Analyses	32
3.1 Asymmetric Electric Field Stress.....	32
3.1.1 Setup	32
3.1.2 Results and Analyses	33

3.2	Symmetric Electric Field Stress	34
3.2.1	Setup	34
3.2.2	Results and Analyses	35
3.2.3	Device Recovery Analysis after Symmetric Electric Field Stress.....	44
4	Experiment 2: Temperature Characterization.....	49
4.1	Effect of Temperature on General Device Performance.....	49
4.1.1	Setup	49
4.1.2	Results and Analyses	52
4.2	Effect of Temperature on the Critical Voltage.....	59
4.2.1	Setup	59
4.2.2	Results and Analyses	60
5	Experiment 3: Electron Trapping and Detrapping.....	62
5.1	Trapping Experiment A: On-State Operation	62
5.1.1	Setup	62
5.1.2	Results and Analyses	62
5.2	Trapping Experiment B: High Electric Field Stress	65
5.2.1	Setup	65
5.2.2	Results and Analyses	66
6	Experiment 4: Small-Signal Analysis.....	70
6.1	Small-Signal Model Development in Agilent ADS.....	70

6.1.1	Setup	70
6.1.2	Results and Analyses	73
7	Conclusion	79
	Future Project Recommendations	82
	List of References	85
	Appendices.....	90
	Appendix A: 1/f-Noise System Prototype	90
	Bias Filter Stages	91
	Transimpedance Amplifier Stage	99
	Voltage Amplifier Stages.....	101
	DC Blocks	105
	Design Summary.....	106
	Appendix B: Cadsoft Eagle Tutorial.....	108
	PCB Design: Package Creation	108
	PCB Design: Schematic and Layout.....	116
	PCB Design: Gerber File Generation	119

LIST OF TABLES

Table	Page
Table 1.1 □ Semiconductor materials and associated parameters [1], [7], [9].....	6
Table 2.1 ☐ Overview of HEMT experiments.....	30
Table 2.2 ☐ Key instruments and software used in experiments.....	31
Table 4.1 ☐ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 10V$	52
Table 4.2 ☐ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 15V$	52
Table 4.3 ☐ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 20V$	53
Table 6.1 ☐ Small-signal parameter definitions.	71
Table 6.2 ☐ Major small-signal parameter changes for pre-stress, post-stress,..... and rested HEMT test conditions.....	77
Table 7.1 ☐ Summary of experimental results.....	81
Table A.1 ☐ NPTB00004 DC characteristics using Toro’s drain supply filter design (with parallel resistors).....	94
Table A.2 ☐ NPTB00004 DC characteristics using the 10Ω resistor in the drain filter	96
Table A.3 ☐ NPTB00004 DC characteristics using the diode in the drain filter.....	98
Table A.4 ☐ Component values for the $1/f$ -noise measurement system design	107
Table B.1 ☐ Necessary Gerber and drill files for board fabrication	122

LIST OF FIGURES

Figure	Page
Figure 1.1 □ Performance comparison between vacuum electron devices and solid-state devices [1]	2
Figure 1.2 □ General HEMT device structure [1]	2
Figure 1.3 □ HEMT band diagram (zero-bias gate condition) [8].	4
Figure 1.4 □ Direction of the vertical electric field (white arrow) and induced tensile strain (yellow arrows) [10]	8
Figure 1.5 □ Electron trap sites within an AlGaN/GaN HEMT [10]	8
Figure 1.6 □ TEM scans showing (a) pit formation and (b) crack formation at the drain-side of the gate contact [11]	9
Figure 2.1 □ Cross-sectional geometry of the NPTB00004	12
Figure 2.2 □ NPTB00004 transistor die (a) layout and (b) wire-bonded into an SOIC package [14].....	12
Figure 2.3 □ Preliminary HEMT test board (coplanar waveguide indicated in red).....	14
Figure 2.4 □ HEMT test structure (inset: NPTB00004 package [13])	15
Figure 2.5 □ Heat sink attachment for a HEMT test structure	16
Figure 2.6 □ Agilent ADS schematic used for DC simulations	17
Figure 2.7 □ NPTB00004 family of simulated I-V curves.....	17
Figure 2.8 □ NPTB00004 family of simulated gate current curves	18
Figure 2.9 □ NPTB00004 family of simulated transfer curves	19
Figure 2.10 □ Single transfer curve at a V_{ds} of 20V	19
Figure 2.11 □ Agilent ADS schematic used for RF simulations.....	21

Figure 2.12 □ Simulated S-parameters for the NPTB00004	21
Figure 2.13 □ Initial I-V measurements (0 to 20V V_{ds} sweep)	22
Figure 2.14 □ 200 Ω structure used to dampen oscillations in the gate feed path	23
Figure 2.15 □ Schematic representation of the 200 Ω structure.....	23
Figure 2.16 □ Effect of time constant on system response to an excitation source	24
Figure 2.17 □ I-V measurements using the 200 Ω gate resistor and increased integration time (0 to 20V V_{ds} sweep)	25
Figure 2.18 □ I-V measurements recorded by sweeping V_{ds} from 20V to 0V	25
Figure 2.19 □ Gate current measurements	26
Figure 2.20 □ Measured transfer characteristics	27
Figure 2.21 □ Bias tees configuration for S-parameter measurements	27
Figure 2.22 □ S-parameter measurements.....	28
Figure 3.1 □ Circuit configuration for the horizontal E-field stress experiments	32
Figure 3.2 □ Gate leakage current measurement results during high asymmetric electric field stress.....	33
Figure 3.3 □ Gate current measurements before and after high asymmetric electric field stress	34
Figure 3.4 □ Circuit configuration for the high symmetric electric field stress experiments.....	35
Figure 3.5 □ V_{crit} observations using the high symmetric electric field step-stress procedure.....	36
Figure 3.6 □ Typical gate leakage characteristic as the gate-source voltage approaches the established V_{crit}	37

Figure 3.7 □ Gate leakage current characteristic over time as $V_{gs} = V_{crit}$	38
Figure 3.8 □ Recorded I-V characteristics before and after high symmetric electric field step-stress procedure.....	38
Figure 3.9 □ Gate current measurements before and after high symmetric electric field step-stress procedure.....	40
Figure 3.10 □ Energy band diagram of a metal-semiconductor contact in thermal equilibrium [27]	41
Figure 3.11 □ Transfer characteristics before and after high symmetric electric field step-stress procedure.....	42
Figure 3.12 □ Measured S-parameters before and after high symmetric electric field stress	43
Figure 3.13 □ Change in I-V characteristics over a relaxation period of 6 days.....	45
Figure 3.14 □ Observed gate current characteristics over a relaxation period of 6 days	45
Figure 3.15 □ Observed transfer characteristics over a relaxation period of 6 days	47
Figure 3.16 □ Change in S-parameters over a relaxation period of 6-days.....	48
Figure 4.1 □ Top view of the test fixture designed for the thermal experiments.....	49
Figure 4.2 □ Side view concept drawing of the test fixture designed for the thermal experiments.....	50
Figure 4.3 □ I-V characteristics at 27°C, 50°C, and 100°C	54
Figure 4.4 □ I-V characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C.....	54
Figure 4.5 □ Gate current characteristics at 27°C, 50°C, and 100°C.....	55

Figure 4.6 □ Gate current characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C	56
Figure 4.7 □ Transfer characteristics at 27°C, 50°C, and 100°C	57
Figure 4.8 □ Transfer characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C	57
Figure 4.9 □ S-parameter measurements at 27°C, 50°C, and 100°C	58
Figure 4.10 □ S-parameter comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C	59
Figure 4.11 □ Shifts in the critical voltage (V_{crit}) parameter with varying back- plate temperatures	60
Figure 4.12 □ Change in junction temperature as the gate-source voltage approaches V_{crit}	61
Figure 5.1 □ Drain and gate current transients of a DUT biased in the saturation region over a 1,000 second span	63
Figure 5.2 □ Detrapping experiment drain current transient results for (a) Device A and (b) Device B ($V_{gs} = -1.3V$, $V_{ds} = 2V$)	64
Figure 5.3 □ Detrapping experiment gate current transient results for (a) Device A and (b) Device B ($V_{gs} = -1.3V$, $V_{ds} = 2V$)	65
Figure 5.4 □ Results of symmetric high-field stress at V_{crit}	66
Figure 5.5 □ Drain current characteristics before and after high-field stress for (a) Device C and (b) Device D	67
Figure 5.6 □ Gate current characteristics before and after high-field stress for (a) Device C and (b) Device D	67

Figure 5.7 □ Detrapping experiment drain current transient results for (a) Device C and (b) Device D ($V_{gs} = -1.3V$, $V_{ds} = 2V$)	68
Figure 5.8 □ Detrapping experiment gate current transient results for (a) Device C and (b) Device D ($V_{gs} = -1.3V$, $V_{ds} = 2V$)	68
Figure 6.1 □ Small-signal model circuit with both intrinsic and extrinsic parameters	70
Figure 6.2 □ Cross-sectional geometry of the GaN-on-Si DUT with intrinsic parameters overlaid	71
Figure 6.3 □ Constructed small-signal model using Agilent ADS	73
Figure 6.4 □ Simulated S-parameters prior to fitting routine	74
Figure 6.5 □ Agilent ADS “Tune Parameter” interface	74
Figure 6.6 □ Simulated S-parameters after using the parameter tuning interface in ADS	75
Figure 6.7 □ Measured S-parameters used for small-signal modeling	76
Figure 6.8 □ Simulated S-parameters after performing a small-signal parameter fit to measured data	76
Figure A.1 □ 1/f-noise measurements system for microwave transistors proposed by Clemente Toro	90
Figure A.2 □ Gate supply filter for a FET	91
Figure A.3 □ Drain supply filter for a FET	91
Figure A.4 □ Gate supply filter for the NPTB00004 HEMT	92
Figure A.5 □ Gate supply filter frequency response	92
Figure A.6 □ DC circuit simulation setup for the NPTB00004	94

Figure A.7 □ Drain supply filter frequency response.....	95
Figure A.8 □ Q-point alteration due to resistive components in the drain supply filter....	95
Figure A.9 □ Frequency response using the 10Ω resistor in the drain filter	97
Figure A.10 □ Frequency response using the diode in the drain filter	98
Figure A.11 □ Drain supply filter design	99
Figure A.12 □ LT6200-5 transimpedance amplifier with a sensitivity of 100V/A.....	100
Figure A.13 □ LT6200-5 current to voltage conversion (100V/A sensitivity, voltage left-axis, current right-axis).....	100
Figure A.14 □ Voltage amplifier with a gain of 14dB (5V/V).....	101
Figure A.15 □ Observable peaking using the suggested amplifier configuration with the LT6200-5	102
Figure A.16 □ Simulated circuit for the LT6200-5 voltage amplifier with swept feedback capacitor values	102
Figure A.17 □ Simulated responses for the LT6200-5 voltage amplifier with swept feedback capacitor values	103
Figure A.18 □ LT6200-5 inverting amplifier configuration	103
Figure A.19 □ LT6200-5 inverting amplifier response (14dB of gain)	104
Figure A.20 □ LT6200-5 inverting amplifier cascade.....	104
Figure A.21 □ LT6200-5 inverting amplifier cascade response (28dB of gain)	105
Figure A.22 □ Lossless DC block	105
Figure A.23 □ DC block frequency response.....	106
Figure A.24 □ Final 1/f-noise measurement system design	107

Figure B.1 □ Cadsoft Eagle main window - starting a custom library	108
Figure B.2 □ Cadsoft Eagle library window	109
Figure B.3 □ Package creation work space with typical grid setting	110
Figure B.4 □ SMD and through-hole size/dimension options, respectively	110
Figure B.5 □ Package design example - 8-pin DIP socket for the TL072	111
Figure B.6 □ Schematic symbol creation work space	113
Figure B.7 □ Schematic symbol design example - 8-pin DIP socket for the TL072	114
Figure B.8 □ Device creation work space	115
Figure B.9 □ Schematic and package symbol connection window	115
Figure B.10 □ Starting a new project in Eagle	116
Figure B.11 □ Starting a new schematic in the project directory	116
Figure B.12 □ Schematic workspace	117
Figure B.13 □ Layout workspace	118
Figure B.14 □ CAM processor window	120
Figure B.15 □ Gerber file configuration for a 2-layer board job	121

1 Introduction

1.1 High Electron Mobility Transistors (HEMTs)

Vacuum electron devices (VEDs) such as traveling-wave tubes, klystrons, and magnetrons have been leading technologies in rectifying, amplification, and switching applications for more than half a century. VEDs can be found in a range of applications such as RADAR, satellite and ground-based communication systems, and audio amplification devices. The popularity of VEDs stem from their high-speed and high-power characteristics. The trade-off for the aforementioned qualities, however, are bulkier form-factors, increased thermal dissipation, smaller bandwidths, larger power supplies, and non-instantaneous start-up operation [1].

To address the many drawbacks VEDs present, solid-state microwave transistor devices were created and are constantly being improved to match, if not exceed the exceptional capabilities of VEDs while lessening the negative trade-offs they present. Amongst the many solid-state transistors being experimented with, the most attractive candidate is the high electron mobility transistor (HEMT). Figure 1.1 shows a general comparison between a variety of VEDs and some commercially available solid-state devices such as metal semiconductor field-effect transistors (MESFETs) and pseudomorphic high electron mobility transistors (pHEMTs). While current solid-state microwave transistor technologies do not presently match VED performance, methods are being researched to increase the overall characteristics of the transistors, such as gate-finger paralleling and optimal epitaxial configurations [2].

Figure 1.1 □ Performance comparison between vacuum electron devices and solid-state devices [1].

HEMTs are much different from typical metal-oxide semiconductor field-effect transistors (MOSFETs) based upon the following characteristics: high electron mobility, small source resistance, high unity-gain frequency (f_T), high transconductance (g_m), high output resistance, and Schottky barrier integration [3]. The general structure of an HEMT is shown in Figure 1.2.

Figure 1.2 □ General HEMT device structure [1].

The epitaxial stack-up for the devices generally consists of a semi-insulating substrate with high thermal conductivity upon which two, large bandgap energy semiconductor materials are grown. Due to the nature of the epitaxial stack, HEMTs are also classified as heterostructures. For example, in Figure 1.2, an n-type donor-supply layer of Aluminum Gallium Arsenide (AlGaAs) is grown upon an un-doped layer of Gallium Arsenide (GaAs); AlGaAs possesses a wider bandgap than GaAs due to its smaller lattice constant (atoms that form the crystal lattice in AlGaAs are much closer together) and donates free electrons to the GaAs layer due to GaAs' characteristic high electron affinity. As AlGaAs is grown upon GaAs, the lattice of the AlGaAs material becomes strained due to the lattice mismatch between the two semiconductor materials. The joining of the two semiconductor materials forms a heterojunction and an overall net piezoelectric polarization results due to the tensile strain between the two materials [1], [4]. Once the semiconductor layers are grown upon a chosen substrate, the drain, gate, and source contacts are formed, where the gate contact takes the form of a Schottky contact (metal-on-semiconductor). The benefits of using a Schottky contact over traditional gate-metal/dielectric/semiconductor configurations are reduced device capacitances (higher operating frequencies) and higher transconductances (larger power gains) [1].

A unique characteristic to HEMT devices is the development of a 2-dimensional electron gas (2DEG) at the heterointerface between the two wide bandgap semiconductor materials. A typical band diagram for an HEMT structure is shown in Figure 1.3. As seen in the diagram, band-bending takes place at the heterojunction due to the differences in bandgap energies, which in turn forms a quantum well. The quantum well develops what

are known as “sub-bands” for electrons to occupy (electrons tend to occupy lower energy states) [1]. The semiconductor material with a larger bandgap energy (e.g. AlGaAs) is an n-type material, providing donor. As electrons occupy the sub-bands within the quantum well, their mobility is restricted to the x- and y-dimensions allowing for higher carrier density (more current, more gain) within the 2DEG of the HEMT device [5], [6], [7]. Also, because the lower bandgap semiconductor material is undoped, there is less impurity scattering in the 2DEG, leading to an increase in carrier mobility (higher speeds).

Figure 1.3 □ HEMT band diagram (zero-bias gate condition) [8].

Several different semiconductor materials have been investigated for their desirable contributions to HEMT performance. Figures of merit amongst varying HEMT semiconductor materials include a large bandgap energy, a low dielectric constant, a high thermal conductance, and a high critical breakdown field. Table 1.1 lists popular semiconductor materials used in HEMT fabrication. The previously mentioned parameters are desirable for the following reasons:

- Large Bandgap Energy (E_g): Determines the transistors ability to operate under high levels of temperature, radiation, and electric fields [1], [7].
- Low Dielectric Constant (ϵ_r): Reduces the overall capacitive area within the device, which in turn increases the unity-gain frequency (f_T) and maximum frequency of oscillation (f_{max} , where the unilateral power gain equals unity) of the transistor. Referring to Equations (1.1), (1.2), and (1.3), if the dielectric constant increases, both f_T and f_{max} will decrease proportionally [1].

$$C \propto \epsilon_r A \quad (1.1)$$

$$f_T = \frac{g_m}{2\pi C_{gs}} \quad (1.2)$$

$$f_{max} = \frac{f_T}{2} \sqrt{\frac{R_{ds}}{R_g}} \quad (1.3)$$

- High Thermal Conductance (K): Under high-power operation, if a device cannot properly dissipate heat due to self-heating, the device can fail as a result.
- High Critical Breakdown Field (E_C): Leads to higher realizable breakdown voltages that can provide large power densities when coupled with high current densities within the 2DEG of the HEMT [7].

Table 1.1 □ Semiconductor materials and associated parameters [1], [7], [9].

<i>Material</i>	Material Parameters			
	E_g (eV)	ϵ_r	K (W/°K-cm)	E_c (V/m)
Vacuum	-	1	-	-
Si	1.12	11.9	1.5	3×10^5
GaAs	1.43	12.5	0.54	4×10^5
InP	1.34	12.4	0.67	4.5×10^5
3C-SiC	2.3	9.7	4	1.8×10^6
4H-SiC	3.2	10.0	4	3.5×10^6
6H-SiC	2.86	10.0	4	3.8×10^6
GaN	3.4	9.5	1.3	3.3×10^6
Diamond	5.6	5.3	20-30	5×10^6
Sapphire	10	7.5-11.5	35-40	$(15 \text{ to } 50) \times 10^6$

1.2 Gallium Nitride (GaN) HEMTs

Out of the available semiconductor materials, GaN is rapidly becoming a prime candidate in the fabrication of HEMTs. Based upon the parameters from Table 1.1, GaN possesses an exceptional bandgap energy (3.4eV), dielectric constant (9.5), and critical breakdown field (3.3MV/m) in comparison to GaAs and Indium Phosphide (InP) materials (GaAs and InP being competing materials). While the thermal conductivity of GaN is relatively poor in comparison to other semiconductor materials (1.3 W/°K-cm), other materials such as Silicon Carbide (SiC), Silicon (Si), Diamond, and Sapphire can be used as a thermally conductive substrate for the GaN HEMT.

A question then arises as to why GaN-based HEMTs are not as widely accepted in the commercial and military markets. The two primary concerns revolving the technology are reliability and reproducibility. Because GaN technology has not fully matured, many different process variations exist for the device. For example, a variety of epitaxial stack-

ups exist, which in turn affects several of the HEMT's device characteristics such as I-V parameters, electric field distribution, power efficiency, and so forth. GaN HEMT substrates are also inconsistent between manufacturers; some may choose SiC due to its high thermal conductivity and higher lattice matching, while others may choose Si due to its cost-effectiveness. Performance variation amongst a group of HEMTs fabricated with the same epitaxial stack and substrate may also exist.

One of the major reliability concerns in current GaN technologies is the creation of crystallographic defect formations during fabrication and through the inverse piezoelectric effect [10]. The inverse piezoelectric effect occurs when a high vertical electric field perpendicular to the device's substrate is present across both semiconductor materials, which in turn promotes tensile strain (Figure 1.4). From [10], the lattice dislocations and defect formations were determined to have formed under a specific gate voltage bias known as the critical voltage (V_{crit}). The critical voltage is defined as the bias voltage at which a change in device performance is observed, which in turn is attributed to the sudden formation of defects within the device. The problem surrounding defect formations are their ability to facilitate electron trap states, with varying trap densities scattered amongst the sites of dislocation. Figure 1.5 shows an example of existing electron trap sites within the AlGaIn layer of a GaN HEMT. Electron trapping ultimately degrades the DC performance of the device, more specifically by causing current reduction due to depletion of the 2DEG and increased gate leakage through the Schottky barrier [1], [10].

Figure 1.4 □ Direction of the vertical electric field (white arrow) and induced tensile strain (yellow arrows) [10].

Figure 1.5 □ Electron trap sites within an AlGaN/GaN HEMT [10].

Other research has found the existence of pit and crack formations under the drain-side edge of the gate contact after being exposed to high electric field stresses [11]. The pit formations that were observed through electrical degradation caused a worst-case 70% drop in drain current and a 0.6V positive shift in threshold voltage. Crack formations that were observed signified the onset of abrupt device failure. Figure 1.6 shows the resulting pit and crack formations seen in [11] through the use of transmission electron microscope (TEM) scans.

Figure 1.6 □ TEM scans showing (a) pit formation and (b) crack formation at the drain-side of the gate contact [11].

1.3 Research Focus

The work presented in this paper utilizes the high-field experiment and accelerated lifetime experiment methodologies of [10] and [11] to investigate degradation mechanisms within GaN-on-Si HEMTs (GaN grown upon Si-substrates). GaN-on-Si technology has become attractive due to the low-cost of fabrication associated with using a Si-substrate, thus making research efforts on the technology more desirable. Other advantages to Si-substrate usage are ease of wafer thinning, via-hole formation, and AuSi eutectic die attach [12]. Because a 17% lattice mismatch and 56% thermal expansion coefficient mismatch exists between GaN and Si, utilizing GaN-on-Si devices in accelerated lifetime experiments may be more beneficial for observing sudden degradation events.

The main purpose of the experiments are to determine how the DC-, RF-, and thermal-level characteristics of the target HEMTs degrade under high-field stress, and what the underlying mechanisms are leading to the observed modes of degradation. The

research presented in this paper is an on-going collaboration between Cal Poly, San Luis Obispo and the Naval Postgraduate School. The collaborative work between the two schools involved performing experimental analyses, extracting “black box” results, and then identifying the mechanisms associated with the degradation results at the material level.

2 Device Under Test (DUT) Overview and Test Plan

2.1 Selected DUT for Experiments

The NPTB00004 from Nitronex was selected as the representative GaN-on-Si device to be evaluated. A single NPTB00004 SOIC package is capable of producing 5W of continuous wave (CW) power and operates from DC up to 6GHz [13]. Device specifications and the cross-sectional geometry for the GaN-on-Si HEMT are listed below and shown in Figure 2.1, respectively, and are derived from [14], [15], [16], and [17]. A physical picture of the NPTB00004 die is shown in Figure 2.2.

- 0.5 μ m (length) Ni/Au (Schottky) T-gate electrodes
- 2mm gate periphery
- 1 μ m gate-source spacing and 3 μ m gate-drain spacing
- Ti/Al/Ni/Au ohmic metallization
- 490nm SiN_x passivation layer and SiN_x encapsulation layer
- 1.5nm GaN cap layer
- 18nm Al_{0.26}GaN_{0.74} barrier and 800nm GaN barrier
- AlGaN transition layer and AlN nucleation layer
- Electroplated gold air-bridge interconnect level
- Source-connected field plate (SCFP)
- 150 μ m Si-substrate (10,000 Ω -cm)

Figure 2.1 □ Cross-sectional geometry of the NPTB00004.

(a)

(b)

Figure 2.2 □ NPTB00004 transistor die (a) layout and (b) wire-bonded into an SOIC package [14].

As described in [12], [15], [16], and [17] the NPTB00004 was developed using metal-organic chemical vapor deposition (MOCVD). Before the primary semiconductor barriers were grown, an AlN nucleation layer was added to prevent unwanted Ga-Si interactions and an AlGaN transition layer was included to provide an increased thermal and lattice match between the GaN layer and Si substrate. Each GaN-on-Si die can be paralleled with multiple dies using the interconnects to form an HEMT with a higher power density. The NPTB00004, however, only utilizes a single die.

One key feature of the GaN-on-Si HEMT that improves performance is the source-connected field plate (SCFP). Reliability studies have demonstrated how inducing the inverse piezoelectric effect in the AlGaN barrier facilitates crystallographic defect formation under the drain-side of the gate contact [18], [19]. By including an SCFP in the growth process, the peak vertical electric field under the gate is redistributed, allowing HEMTs to achieve higher breakdown voltages, lower gate leakage currents, higher cut-off frequencies, higher maximum available gains, and lower gate-drain capacitances [7], [15].

Another important feature within the epitaxial stack is the inclusion of a GaN cap layer. The GaN cap serves two purposes. The first key feature is how the GaN cap mitigates the high vertical electric field seen at the drain-side edge of the gate contact when biased in the off-state. Reduction of the vertical electric field was observed to have reduced the amount of current collapse and gate leakage current experienced within the GaN HEMTs [20]. The second key feature improves the overall surface morphology of the epitaxial stack. Researchers in [20] observed a large increase in their devices'

breakdown voltage (100V to 350V) due to the ease of forming the gate electrodes on a much smoother surface, as opposed to the cruder surface of an AlGaIn cap layer.

Before experimenting with the NPTB00004, a standard test structure needed to be created for each DUT to ensure relative consistency between measurements. The preliminary board that was considered was a basic PCB fabricated on FR4 and constructed in a 50Ω coplanar waveguide configuration (Figure 2.3). The board was created using Cadsoft Eagle schematic and layout software (an Eagle tutorial can be found in Appendix B: Cadsoft Eagle Tutorial). The PCB design was ultimately discarded due to high lead-times and cost-effectiveness. Because the study involved testing a large quantity of NPTB00004 HEMTs, a different method needed to be used in order to speed up structure creation at a low-cost.

Figure 2.3 □ Preliminary HEMT test board (coplanar waveguide indicated in red).

A simple copper DUT structure, shown in Figure 2.4, was used as the test structure of choice. Each structure utilizes a 40mm x 55mm double-sided copper FR4 board, which serves as a grounding plane for the DUT. The source-pad of the HEMT

(located on the underside of the SOIC package) was then soldered to the copper using solder paste and a heat gun. 50 Ω Semi-rigid coaxial cables with SMA(M) connectors were used to provide electrical signals to the device. The coaxial line lengths varied from device to device, which did not affect device characterization during DC-level experiments. During RF experiments, however, the recorded lengths were used to reference-plane shift S-parameter measurements up to the device's package leads. To prevent parasitic inductances along the transmission lines, the outer conductors of each coaxial line were soldered to the copper ground plane at the edge of the copper board and near the leads of the DUT. Because the NPTB00004 possesses the ability to provide up to 5W of power, a 40mm x 55mm (6.5mm thickness) heat sink was added to each structure to provide proper dissipation of heat. A single heat sink attachment is shown in Figure 2.5.

Figure 2.4 □ HEMT test structure (inset: NPTB00004 package [13]).

Figure 2.5 □ Heat sink attachment for a HEMT test structure.

2.2 Simulation of the HEMT's DC and RF Characteristics

According to [13], the NPTB00004 can achieve a maximum drain-source voltage (V_{ds}) of 28V, possesses a typical threshold voltage (V_t) of -1.5V, and utilizes a quiescent gate voltage (V_{gsq}) of -1.3V. At a V_{gsq} of -1.3V and V_{ds} of 28V, a drain current (I_d) of 50mA is produced, which in turn allows for 1.4W of power dissipation. Due to power limitations for the instruments used in the laboratory, the DUTs are only evaluated for $V_{gs} \leq -1.25V$ and $V_{ds} \leq 20V$. All simulations for the DUT were performed using Agilent Advanced Design System (ADS). All simulations assumed a device temperature of 27°C.

2.2.1 I-V Characteristics

Figure 2.6 shows the simulation schematic used to analyze the NPTB00004's I-V, gate current, and transfer characteristics. The list below Figure 2.6 shows the DC conditions used during each simulation. Figure 2.7 and Figure 2.8 are the simulated I-V and gate current characteristics, respectively. A V_{gsq} of -1.3V and a V_{ds} of 20V were chosen as the standard biases for all experiments performed on the NPTB00004. At the aforementioned Q-point, an I_{dsq} of approximately 60mA was simulated, resulting in 1.2W

of dissipated power. Power at the gate terminal was simulated to be negligible from the corresponding gate current measurement in Figure 2.8.

Figure 2.6 Agilent ADS schematic used for DC simulations.

- I-V and gate current characteristics: bias conditions
 - $V_{ds} = 0$ to 20V (100mV increments)
 - $V_{gs} = -1.7V$ to $-1.25V$ (50mV increments)

Figure 2.7 NPTB00004 family of simulated I-V curves.

Figure 2.8 NPTB00004 family of simulated gate current curves

2.2.2 Transfer Characteristics

The following is a list of DC conditions used to simulate the transfer characteristics for the DUT. Figure 2.9 shows a family of transfer curves generated from the bias conditions in the list. A second simulation was done to evaluate the transfer characteristics at a V_{ds} of 20V, which was used in as a baseline for proceeding experiments. The results are shown in Figure 2.10.

- Transfer characteristics: bias conditions
 - $V_{ds} = 0$ to 20V (100mV increments)
 - $V_{gs} = -2$ V to -1.25 V (50mV increments)

Figure 2.9 □ NPTB00004 family of simulated transfer curves.

Figure 2.10 □ Single transfer curve at a V_{ds} of 20V.

2.2.3 *S-parameters*

The schematic used to simulate the S-parameters of the NPTB00004 is shown in Figure 2.11. The bias conditions for the S-parameter simulations are listed below.

- S-parameters: bias conditions
 - $P_{in} = 0\text{dBm}$
 - Frequency range: 100MHz to 3GHz
 - $V_{gs} = -1.3\text{V}$
 - $V_{ds} = 20\text{V}$

For the actual S-parameter measurements, two bias tees were used to provide the DC bias network for all DUTs. A HP11612A bias network and a HP33150A bias tee were used as the DC bias networks. A lower limit of 100MHz used during S-parameter simulation was utilized due to the 100MHz lower frequency limit of the HP33150A bias tee. Because the schematic for the HP33150A could not be found, the components used in the HP11612A were used to represent both bias networks in the simulation circuit of Figure 2.11. The schematic of Figure 2.11 does not take into account line lengths from coaxial cables or the bias networks when simulating the S-parameters.

The simulated S-parameters are shown in Figure 2.12. At the selected Q-point, the drain current and gate current were simulated to be 61.47mA and 21.89fA, respectively. A maximum obtainable gain of approximately 25.58dB (19V/V) was simulated at 100MHz while a reverse transmission gain of -27.54dB (0.042V/V) was simulated at 1.2GHz. The drain impedance was observed to be capacitive over the simulated frequency range. The gate impedance, however, becomes inductive at approximately 1.88GHz ($Z_{gate} = 4.95 + j0.02\Omega$).

Figure 2.11 Agilent ADS schematic used for RF simulations.

Figure 2.12 Simulated S-parameters for the NPTB00004.

2.3 DC and RF Characterization of the DUT

All DC measurements were performed using a HP4155A semiconductor parameter analyzer. Measurements were made using similar bias conditions listed in subsections 2.2.1 and 2.2.2. The integration time used between measurements was 80 μ s. Voltages were applied using BNC cables with SMA(F) adapters. Figure 2.13 shows the resulting I-V measurement for a single HEMT.

Figure 2.13 □ Initial I-V measurements (0 to 20V V_{ds} sweep).

As seen above, the measurements do not correlate to the simulated I-V measurements of Figure 2.7. One set of issues that may be linked to the observation above is the presence of oscillations. To dampen the circuit and prevent oscillations, a 200 Ω resistor was required in the gate-feed path to suppress oscillations [21]. A 200 Ω structure, shown in Figure 2.14, was created and used in all proceeding DC measurements.

Figure 2.14 □ 200Ω structure used to dampen oscillations in the gate feed path.

Figure 2.15 □ Schematic representation of the 200Ω structure.

Transients can also become a factor when applying an excitation source to the device during measurements. An example scenario would be to evaluate the transient response of an RC circuit with two known time constants (τ) and a unit-step source using Equation (2.1) [22].

$$v(t) = V_S \left(1 - e^{-\frac{t}{\tau}} \right) u(t), \text{ where } \tau = RC \quad (2.1)$$

As τ becomes smaller, the resulting system waveform reaches V_S at a much faster rate, or in other words reaches the steady-state voltage quicker. As τ becomes larger, $v(t)$ takes much longer to reach the steady-state V_S value. While a smaller τ seems desirable, if the system is underdamped (e.g. due to insufficient compensation resistor), the signal

could overshoot V_s and experience a ringing nature as the signal approaches steady-state. An example of the aforementioned signals are shown in Figure 2.16.

Figure 2.16 □ Effect of time constant on system response to an excitation source.

Based upon the previous analyses, the 200Ω component was added in the gate-feed path and the integration time was increased from $80\mu\text{s}$ to 266ms . Due to the increase in measurement time, the increment in V_{ds} was increased from 100mV to 500mV . The resulting I-V measurements are shown in Figure 2.17. An abnormal linear region was observable in the new I-V measurements, where the drain current appears suppressed by some phenomenon. The current reduction in the linear region, or “kink” effect, was determined to be attributed to electron-traps within the HEMT [23], [24]. A second set of I-V curves were generated by sweeping V_{ds} instead from 20V down to 0V (500mV decrements).

Figure 2.17 I-V measurements using the 200Ω gate resistor and increased integration time (0 to 20V V_{ds} sweep).

Figure 2.18 I-V measurements recorded by sweeping V_{ds} from 20V to 0V.

The new I-V measurements are displayed in Figure 2.18. While the linear region experienced an observable increase in drain current, the characteristics still remain relatively ohmic in nature. Another observation was how the saturation region of each curve tended to curl around $V_{ds} = 20V$. The curling characteristic could potentially be inherent to the way the device turns on when instantaneously biased at a rather large drain voltage value using the transistor parameter analyzer.

Figure 2.19 □ Gate current measurements.

A family of gate current measurements is shown in Figure 2.19 above. As seen in the measurements of Figure 2.19, as the device is set further into the on-state, the effective gate current drawn from the device decreases. It should be noted that the gate current measurements in Figure 2.19 possess a negative polarity because current is being sourced from the HEMT (the same will be seen throughout the entirety of the report). Figure 2.20 shows the measured transfer characteristics for the DUT with a constant V_{ds} bias of 20V.

Figure 2.20 □ Measured transfer characteristics.

Figure 2.21 □ Bias tees configuration for S-parameter measurements.

The S-parameter measurements were extracted using an Anritsu MS4622B Vector Network Analyzer (VNA) using the bias conditions and bias tees described in Subsection 2.2.3. RF measurements did not include the use of matching networks since parameters such as maximum attainable gain (MSG) and voltage standing-wave ratio

(VSWR) were negligible figures of merit. A visual demonstrating the connections between the bias tees and the DUT structure is shown in Figure 2.21. Before conducting measurements, reference plane shifts were applied to both the gate and drain transmission lines (Channels 1 and 4, respectively). Reference plane shifts were performed by entering into the VNA the physical length of each line in millimeters and selecting the correct dielectric material used to insulate the outer and inner conductors. An S2P file was extracted from the VNA and imported into Agilent ADS to plot the results (Figure 2.22).

Figure 2.22 □ S-parameter measurements.

From Figure 2.22, it can be seen how the input and output impedance measurements show a type of wrapping nature at high frequencies. The wrapping nature was determined to be an artifact created at high frequencies due to the inductances inherent in the transmission lines used for each DUT's structure.

2.4 Test Plans - Overview

Table 2.1 lists the test plans for four groups of HEMTs (approximately 60 transistors) involved in the GaN-on-Si research and Table 2.2 lists the instruments and software tools used during the experimental process. The first experiment involved assessing the HEMT's critical voltage, or the gate bias voltage condition at which a large amount gate leakage current is observed. Two transistor configurations are evaluated in the step-stress procedures. DC and RF characteristics were analyzed before and after high voltage biases were applied to the DUTs. After performing the step-stress tests, the devices were then monitored to determine if any altered characteristics were recoverable post-stress.

Experiment 2 involved assessing the DC and RF parameters of the DUTs under increasing back-plate temperatures. Recovery analyses were then performed after thermally stressing each HEMT. A second set of experiments were then conducted to assess how the critical voltage determined in Experiment 1 varies over increasing back-plate temperatures. Experiment 3 investigated the nature of electron trapping and detrapping mechanisms within the GaN devices under normal operation and during high field exposure. Finally, the goal of Experiment 4 was to map alterations in the test transistors' RF performances to their equivalent small-signal models after exposure to

high-field stress. Using tuning parameters in Agilent Advanced Design System (ADS), small-signal models were developed for the stressed HEMTs.

Table 2.1 Overview of HEMT experiments.

Exp. #	Chapter # and Page #	Experimental Details		
		Exp. Name	Sub-Exp. Name	Testing Overview
1	3 (pg. 32)	High Electric Field Stress Analyses	Asymmetric Electric Field Stress	Evaluation of each DUTs DC and RF characteristics before and after a high drain voltage bias is applied while the device is in the off-state ($V_{gs} = -5V$). The high drain bias will create a large vertical electric field parallel to the substrate of the devices. Main analytical focus is the critical voltage with respect to the drain bias.
			Symmetric Electric Field Stress	Similar evaluations done in the asymmetric E-field stress test will be performed while the device is in the off-state ($V_d = 0V$). A high negative gate bias will be applied to induce a large E-field perpendicular to the substrate of the devices. Main analytical focus is the critical voltage with respect to the gate bias. Additional analyses are performed to evaluate the stressed devices over relaxation periods (no excitation).
2	4 (pg. 49)	Temperature Characterization	Effect of Temperature on General Device Performance	Evaluation of DC parameters, RF parameters, and junction temperature under various back-plate temperatures.
			Effect of Temperature on the Critical Voltage	The devices' critical voltages will be measured under thermal strain and tracked to observe whether or not a temperature dependence exists.
3	5 (pg. 62)	Electron Trapping and Detrapping	Trapping Experiment A: On-State Operation	The test will analyze trapping mechanisms after operating the device in the on-state. Thermal excitation is used to aid in the detrapping of electrons.
			Trapping Experiment B: High Electric Field Stress	The test will investigate potential trapping mechanisms induced through high-field stress. Thermal excitation is used to aid in the detrapping of electrons.
4	6 (pg. 70)	Small-Signal Analysis	~	Creation of a small-signal model to evaluate how high-field stress alters the small-signal parameters of the DUTs.

Table 2.2 □ Key instruments and software used in experiments.

Instrument and Software Details	
<i>Name</i>	<i>Purpose</i>
Agilent Technologies 6555A 120V, 4A DC Power Supply	Provide high-voltage bias from 0-120V.
Agilent Technologies 34401A Digital Multimeter, 6 1/2 Digit	Current monitoring.
HP4155A Semiconductor Parameter Analyzer	Measurement of I-V and transfer characteristics.
ILX Lightwave LDC-3744C Laser Diode Controller	Thermoelectric cooler controller and thermistor monitor.
Anritsu MS4622B 0.01-3GHz Vector Network Analyzer	S-parameter measurements.
TE HY3003-3 DC Power Supply	Provides drain or gate bias voltage.
HP11612A Bias Network 45MHz-26.5GHz	DC bias network used in S-parameter measurements.
HP33150A Bias Tee 0.1-18GHz	DC bias network used in S-parameter measurements.
Agilent Advanced Design System	S-parameter analysis and small-signal modeling.
National Instruments LabVIEW	Automation of instruments.

3 Experiment 1: High Electric Field Stress Analyses

3.1 Asymmetric Electric Field Stress

3.1.1 Setup

Figure 3.1 shows the circuit configuration used to apply a high asymmetric electric field across the device. The electric field applied to the device is deemed “asymmetric” because stress is only applied to the drain of the transistor, creating a strong electric field gradient (horizontal field) between the gate and drain regions. The asymmetric field stress procedure first involved setting V_{gs} to a constant -5V to set the device in the off-state (no drain current conduction, no power dissipation). V_{ds} was then stepped from 30V to 120V in 1V increments every minute. Gate leakage current and drain current were measured during the step-stressing procedure. A HP6555A DC Power Supply was used to supply the large V_{ds} voltage biases.

Figure 3.1 □ Circuit configuration for the horizontal E-field stress experiments.

3.1.2 Results and Analyses

Figure 3.2 shows the resulting gate leakage current measurement recorded using the step-stress procedure outlined in Subsection 3.1.1. Over the entire V_{ds} incremental span, the gate leakage current of the DUT increased from 1.5 μ A to 2.6 μ A.

Figure 3.2 □ Gate leakage current measurement results during high asymmetric electric field stress.

While there was an observable (yet negligible) change in the gate leakage current measurements, there were very few changes between each DUT's pre-stress and post-stress DC and RF characteristics. The only noticeable change was to each device's gate current. Figure 3.3 shows how the device's overall gate current reduces after being exposed to high drain voltage biases.

Figure 3.3 □ Gate current measurements before and after high asymmetric electric field stress.

3.2 Symmetric Electric Field Stress

3.2.1 Setup

Figure 3.4 shows the circuit configuration used to apply a high symmetric electric field across the device. In this configuration, the applied electric field is evenly distributed throughout the device due to the drain and source being joined together and occurs vertical to the substrate of the device. The symmetric field stress procedure involved tying the drain and source terminals to ground, which set the device in the off-state (no drain current conduction, no power dissipation). The first test, known as a step-stress procedure, involved stepping V_{gs} from -10V to -70V in 1V decrements every minute. Gate leakage current was measured during the step-stressing procedure. After identifying the critical voltage (V_{crit}) of the HEMT, each DUT was then stressed at a

constant $V_{gs} = V_{crit}$ bias while gate leakage current was monitored. The constant V_{gs} test-scenario was performed until gate leakage current ceased to change or device failure was experienced. A HP6555A DC Power Supply was used to supply the large V_{gs} voltage biases.

Figure 3.4 □ Circuit configuration for the high symmetric electric field stress experiments.

3.2.2 Results and Analyses

Two transistors were used during the initial symmetric field stress analyses, and their resulting gate leakage current measurements are shown in Figure 3.5. As V_{gs} was stepped more negative over time, an observed rise in gate leakage current was apparent around $V_{gs} = -50V$. As V_{gs} approached $-70V$ (estimated V_{crit} , indicated by a dashed red line in Figure 3.5), a drastic increase in gate leakage current was experienced. An absolute increase of the gate-source voltage beyond $-70V$ either resulted in immediate device failure, or gradual device breakdown leading to imminent device failure.

Performing the symmetric field step-stress experiments with a larger pool of HEMTs resulted in similar gate leakage current characteristics around $V_{gs} = -70V$.

Figure 3.6 shows a representative gate leakage trend for the DUTs, where a sudden rise in gate leakage current was experienced as V_{gs} approached V_{crit} . An increase in gate leakage current was determined to be attributed to the conduction of electrons between the gate and the channel, which in turn was a result of possible defect formations induced by the inverse piezoelectric effect [25].

Figure 3.5 □ V_{crit} observations using the high symmetric electric field step-stress procedure.

Figure 3.6 □ Typical gate leakage characteristic as the gate-source voltage approaches the established V_{crit} .

An additional experiment was performed where the gate-source voltage was left at a constant -70V while gate leakage current was monitored over time. A representative data set resulting from the aforementioned experiment is shown in Figure 3.7. As seen, a relatively linear trend in increased gate leakage current was seen over time. A common observation from the experiments was that device failure was found to occur when the gate leakage current was either equal to or greater than 10mA at $V_{\text{gs}} = -70\text{V}$, equating to roughly 0.7W. The time it took to reach this limit, however, varied between devices. All proceeding symmetric step-stress experiments utilized the 10mA current-limit to prevent device failure from taking place.

Because of the linear nature of the rising gate leakage current seen in Figure 3.7, a linear regression could have been used to determine the average time until device failure. However, the time-to-failure varied from device to device. While the DUT in Figure 3.7

reached $I_{g, leak}$ in approximately 6.5 minutes, another device could have taken up to an hour to reach the same gate leakage current value.

Figure 3.7 □ Gate leakage current characteristic over time as $V_{gs} = V_{crit}$.

Figure 3.8 □ Recorded I-V characteristics before and after high symmetric electric field step-stress procedure.

Figure 3.8 shows the resulting I-V characteristics of a representative DUT before and after being subjected to high-field stress. Prior to performing the experiment, it was hypothesized that the saturation drain current, $I_{d, \text{sat}}$, would decrease as a result of 2DEG depletion due to trapped electrons in the AlGaN barrier and defect formation through the inverse piezoelectric effect [10]. The results from Figure 3.8 instead show an overall increase in $I_{d, \text{sat}}$. For example, for $V_{\text{gs}} = -1.25\text{V}$, $I_{d, \text{sat}}$ increased by approximately 14% from the unstressed value.

A uniaxial tensile strain experiment, performed in [26], determined that a sapphire-substrate HEMT with two SiN_x passivation layers undergoing a symmetric distribution of mechanical strain experienced an overall increase in $I_{d, \text{sat}}$. The increase in saturation drain current seen in Figure 3.8 could then be attributed to the inverse piezoelectric effect, which provides a symmetric tensile strain to the heterostructure due to the induced symmetric electric field. Research performed in [24] also found that high-field stress promoted drain current recovery, which was determined to be a result of electrons becoming detrapped within the bulk of the device and aided in the recovery of the 2DEG carrier concentration. Further analysis on electron trapping and detrapping will be covered in Chapter 5.

Figure 3.9 □ Gate current measurements before and after high symmetric electric field step-stress procedure.

A significant change in gate current was observed after performing a symmetric high-field step-stress experiment on each GaN-on-Si DUT (Figure 3.9). The increase in gate current seen in Figure 3.9 correlates to the increase in gate leakage current experienced during the high-field stress experiments (Figure 3.6 and Figure 3.7). As mentioned in Section 2.1, the GaN-on-Si HEMT DUTs possess Schottky gates made from Ni/Au. The Schottky gate is essentially a Schottky diode structure, formed by the metal-on-semiconductor junction between Ni/Au and the GaN cap. The joining of the two materials develops what is known as a Schottky barrier height (ϕ_B), which is the potential difference between the Fermi energy of the metal and the band edge where the majority carriers reside [27]. Figure 3.10 shows an example of Schottky diode band diagram at thermal equilibrium (no external voltage applied). As a voltage bias is applied

to the device and an electric field is induced across the junction, ϕ_B will lower respectively. The change in the barrier height can be found through Equation (3.1).

Figure 3.10 □ Energy band diagram of a metal-semiconductor contact in thermal equilibrium [27].

$$\Delta\phi_B = \sqrt{\frac{qE_{max}}{4\pi\epsilon_s}} \quad (3.1)$$

q : Elementary charge (1.602×10^{-19} C)

E_{max} : Electric field at the metal – semiconductor interface

ϵ_s : Dielectric constant of the semiconductor

Because a high vertical electric field is induced at the gate of the HEMT devices during the symmetrical step-stress experiments, the Schottky barrier height of these devices is lowered substantially. As a result, gate leakage is enhanced due to a reduction in barrier potential. Work performed by Lo, et. al. revealed how GaN HEMTs with Ni/Au Schottky gate contacts (which is what the NPTB00004 possesses) experienced a

reduction in ϕ_B from 1.09V to 0.66V, whereas devices with Pt/Ti/Au contacts experienced negligible changes from 1.23V to 1.21V [28].

Figure 3.11 □ Transfer characteristics before and after high symmetric electric field step-stress procedure.

A negative threshold voltage shift was observed within each device after performing the step-stress experiment. The negative shift was determined to be an indication of electron detrapping events occurring within the AlGaN layer of the GaN-on-Si HEMTs, resulting in a recovery of the 2DEG channel [24]. Before exposing each device to high electric fields in [24], the devices developed process-dependent defects during the growth process which in turn led to localized trap sites within the AlGaN barrier. The inherent electron traps within each fresh device then depleted the 2DEG region via electron trapping, resulting in a negative threshold voltage shift within the HEMTs (positive shifts will be covered later). Carbon tetrafluoride (CF_4) treatment experiments performed in [29] found that increased fluoroine ion concentrations in the

AlGaN layer localized near the 2DEG would deplete the channel of electrons and therefore induce positive threshold voltage shifts in their devices (CF_4 possesses a high electronegativity). Therefore, if a negative threshold voltage shift was observed in Figure 3.11, electron traps in the AlGaN barrier localized near the 2DEG may be detrapping and enhancing electron density in the channel.

Figure 3.12 Measured S-parameters before and after high symmetric electric field stress.

The pre-stressed and post-stressed S-parameters are shown in Figure 3.12. An observable change in the small-signal parameters was measured, more specifically with the reverse transmission gain, $|S_{12}|$. For microwave field-effect transistors (FETs), $|S_{12}|$ is primarily controlled through the gate-drain capacitance, C_{gd} , which acts as the effective

feedback in microwave amplifier designs [30]. As $C_{gd} \rightarrow 0$, $|S_{12}| \rightarrow 0$, which in turn means that the device becomes unilateral, or lacks feedback. Based upon the measurements in Figure 3.12, as the GaN-on-Si HEMT is exposed to high symmetric field stresses, the small-signal parameter C_{gd} decreases and causes the device to become unilateral. A more detailed analysis on the DUT's small-signal characteristics can be seen in Chapter 6, where an equivalent small-signal model is developed to approximate the numerical changes in each of the small-signal parameters.

3.2.3 Device Recovery Analysis after Symmetric Electric Field Stress

An additional study was performed to observe whether or not the changes found within each DUT during the symmetrical electric field stress experiment persisted well after being performed. After stressing a pool of devices using the step-stress procedure, analyses similar to those performed in Subsection 3.2.2 were conducted. 24 hours after performing the stress procedures and taking initial measurements for each device, another round of measurements were made to assess the performance of the device after 24 hours of relaxation. The process was repeated every 24 hours until changes within each device ceased.

After relaxing for a period of 6 days, the drain current of each DUT showed no observable collapse or return to their corresponding current characteristics seen during the unstressed state. The increase and sustained drain current characteristics seen in Figure 3.13 further supported claims about electron trapping being prominent in the devices (process-dependent) prior to any exposure to high-field strain.

Figure 3.13 □ Change in I-V characteristics over a relaxation period of 6 days.

Figure 3.14 □ Observed gate current characteristics over a relaxation period of 6 days.

Figure 3.14 shows a similar increase in gate current, as was seen in Figure 3.9. After relaxation over 6 days, however, the gate current continued to increase (green line). From Equation (3.1), it was found that high-electric fields within the HEMT devices reduce their Schottky barrier height, which in turn promoted gate leakage. Over the course of the 6-day relaxation period, none of the devices were exposed to high-field stress. While the phenomenon seen in Figure 3.14 cannot be attributed to high-field stress, the result may be due to a change in capacitance within the GaN cap. The change in capacitance would then affect the dielectric constant of the semiconductor and would alter the Schottky barrier height.

In reference to Equation (3.2) and holding all variables besides Q as constants, as the charge (Q) in a capacitor decreases, the dielectric constant (ϵ_r) will decrease proportionally [31]. Therefore, the Schottky barrier height of the GaN HEMTs may be inherently decreasing during the relaxation period due to possible electron detrapping events within the GaN cap, altering the charge stored within this region.

$$\epsilon_r = \epsilon_s = C \frac{d}{\epsilon_0 A} = \frac{Q}{V} \frac{d}{\epsilon_0 A} \quad (3.2)$$

Figure 3.15 shows a case where a negative threshold voltage shift was not observed directly after performing the symmetrical step-stress procedure. From Figure 3.11, a negative threshold voltage shift was expected for all GaN-on-Si devices after exposure to high symmetric fields. However, after sampling a larger group of devices, some of the stressed DUTs also experienced a positive threshold voltage shift. Work in [24] attributed the positive voltage shifts to lower electric fields resulting from

stress voltages set well below the critical voltage. The lower fields were determined to further create trap states within the AlGaIn barrier or GaN cap by changing their relative positions. Therefore, some of the stressed devices throughout the studies in this paper may not have experienced electron detrapping via the step-stress experiment. While the induced electric fields may not have promoted electron detrapping in some of the DUTs during the step-stress experiments, a negative voltage shift was found in all devices after the relaxation periods.

Figure 3.15 □ Observed transfer characteristics over a relaxation period of 6 days.

Minor alterations to the input impedance, output impedance, and forward transmission gain of each device were observed during the relaxation periods. The corresponding S-parameter relaxation measurements are shown in Figure 3.16. A noticeable recovery in each device's reverse transmission gain was measured. The

recovery in $|S_{12}|$ after 6 days, however, was not complete. Further investigation of the S-parameters is performed in Chapter 6.

Figure 3.16 □ Change in S-parameters over a relaxation period of 6-days.

4 Experiment 2: Temperature Characterization

4.1 Effect of Temperature on General Device Performance

4.1.1 Setup

A test structure was constructed to interface with an ILX Lightwave LDC-3744B, as seen in Figure 4.1 and Figure 4.2. The setup utilizes the same copper test structure as Figure 2.4 (pg. 15), but utilizes two thermoelectric coolers (TECs) to provide a symmetric distribution of heat onto the source-plane of the HEMT.

Figure 4.1 □ Top view of the test fixture designed for the thermal experiments.

Because TECs develop a temperature gradient, it was necessary to use nylon bolts to provide high thermal resistance between the aluminum heat sink and the copper source-plane (back-plate). In many of the other experiments, metallic bolts were used to secure the aluminum heat sinks to the copper source-plane, which also promoted heat transfer between the source of the DUT and the heat sink. Therefore, it was necessary to prevent the transfer of heat between the two planes to better control the back-plate temperature.

Figure 4.2 □ Side view concept drawing of the test fixture designed for the thermal experiments.

The original design did not incorporate the use of the styrofoam insulation between the heat sink and the thermistor. As a result, preliminary temperature readings from the ILX control unit did not provide accurate source-plane temperature measurements and instead would fluctuate between the heat sink temperature and the source-plane temperature (due to the temperature gradient). After providing thermal insulation, the readings increased in accuracy and the set point temperature for the source-plane was allowed to stabilize much faster. Thermal grease was also used to better conduct heat transfer between the back-plate and the thermistor element.

DC and RF tests were performed in this section under three back-plate temperature conditions: 27°C, 50°C, and 100°C. Once the back-plate temperature of the DUT stabilized to the desired temperature, I-V characteristics, transfer characteristics, and S-parameters were measured (refer to Subsections 2.2.1, 2.2.2, and 2.2.3 for excitation settings). After each DUT was stressed to 100°C, the devices were allowed to relax over a 24 hour span. After relaxation, the characteristics were reevaluated at 27°C and compared to the measurement results at 27°C from the previous day to determine if temperature-dependent degradation had occurred.

To determine the junction temperature for each device under a given back-plate temperature and voltage biasing condition, Equation (4.1) was used below [32]. P_{DISS} in Equation (4.1) is the drain quiescent current multiplied by the drain-source voltage. The same equation can be used in the experiments of Section 4.2. For the devices used in this study, θ_{JC} was given in [13] to be 23°C/W.

$$T_J = T_{Baseplate} + \theta_{JC} * P_{DISS} \quad (4.1)$$

T_J : Junction temperature (°C)

$T_{baseplate}$: Base – plate temperature (°C)

θ_{JC} : Thermal resistance between the junction and case (°C/W)

P_{DISS} : Dissipated power (W)

4.1.2 Results and Analyses

Figure 4.3 shows the resulting I-V characteristics at $V_{gs} = -1.25$ after exposure to back-plate temperatures of 27°C, 50°C, and 100°C while Figure 4.4 shows a comparison between I-V curves taken at room temperature before thermal stress and

24 hours after high thermal stress. From Figure 4.3, it was observed that the overall drain current of the device becomes suppressed with increasing back-plate temperatures while Figure 4.4 shows a negligible reduction in overall drain current after exposure to high thermal stress. The junction temperature of the DUT in Figure 4.3 at $V_{gs} = -1.25V$ and $V_{ds} = 10V$ is shown in Table 4.1. The junction temperature of the same DUT at $V_{gs} = -1.25V$ and $V_{ds} = 15V$, and $V_{gs} = -1.25V$ and $V_{ds} = 20V$ is shown in Table 4.2 and Table 4.3, respectively. Under all conditions, it can be seen that the power dissipated in the device decreases with increasing base-plate temperature, and the temperature difference between the base-plate and the junction becomes smaller.

Table 4.1 □ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 10V$.

<i>Base-plate Temperature, $T_{Baseplate}$ (°C)</i>	<i>Drain Current (mA) $V_{gs} = -1.25, V_{ds} = 10V$</i>	<i>Dissipated Power, P_{DISS} (W)</i>	<i>Junction Temperature, T_J (°C)</i>	<i>ΔT (°C) $T_{Baseplate} - T_J$</i>
27	89	0.890	47.47	20.47
50	78	0.780	67.94	19.94
100	52	0.520	111.96	11.96

Table 4.2 □ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 15V$.

<i>Base-plate Temperature, $T_{Baseplate}$ (°C)</i>	<i>Drain Current (mA) $V_{gs} = -1.25, V_{ds} = 15V$</i>	<i>Dissipated Power, P_{DISS} (W)</i>	<i>Junction Temperature, T_J (°C)</i>	<i>ΔT (°C) $T_{Baseplate} - T_J$</i>
27	88	1.32	57.36	30.36
50	79	1.19	77.37	27.37
100	56	0.84	119.32	19.32

Table 4.3 □ DUT junction temperature at $V_{gs} = -1.25V$ and $V_{ds} = 20V$.

<i>Base-plate Temperature, $T_{Baseplate}$ (°C)</i>	<i>Drain Current (mA) $V_{gs} = -1.25, V_{ds} = 20V$</i>	<i>Dissipated Power, P_{DISS} (W)</i>	<i>Junction Temperature, T_J (°C)</i>	<i>ΔT (°C) $T_{Baseplate} - T_J$</i>
27	91	1.82	68.86	41.86
50	81	1.62	87.26	37.26
100	63	1.26	128.98	28.98

Experimental results from [33], [34], and [35] also revealed a similar reduction in drain current with increasing back-plate temperatures. Aside from possessing a 16-17% lattice mismatch between Si and GaN, a 54-56% thermal mismatch also exists, which promotes native traps within the device [17], [25]. The suppression in drain current was determined to be a direct result of reduced electron mobility and density within the 2DEG as a result of increased vibrational energy of the crystal lattice, and also from trap-assisted tunneling (TAT) effects seen in the AlGaIn barrier. TAT effects show exponential temperature dependence and become more relevant at elevated temperature ranges [33].

Figure 4.3 □ I-V characteristics at 27°C, 50°C, and 100°C.

Figure 4.4 ▢ I-V characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C.

The TAT phenomenon also directly affects the gate leakage characteristics of the GaN-on-Si HEMTs. As seen in Figure 4.5, an increase in gate current is observed as the back-plate temperature of the representative DUT is increased. Trapped electrons in the AlGaIn barrier are provided with the necessary energy to escape trap formations and tunnel through the AlGaIn barrier as a direct result of increased temperatures, which in turn depletes the 2DEG and enhances gate leakage current [33]. The results seen in Figure 4.5 and the increased electron trapping behavior in the AlGaIn barrier further support the current collapse observations seen in Figure 4.3.

Figure 4.5 □ Gate current characteristics at 27°C, 50°C, and 100°C.

Gate current measurements after 24 hours of relaxation are shown in Figure 4.6. The measurements show a prominent reduction in gate current as a result of the relaxation at room temperature. A reduction in gate current would seem to indicate an increased recovery of electrons in the 2DEG, but the results of Figure 4.4 may suggest

otherwise. The work done by del Alamo and Joh in [10] shows that when the trap states in the AlGa_N are filled with electrons, the gate leakage path is suppressed and the gate leakage current reduces as a result. The filled electron states also prevent the recovery of the 2DEG, which in turn suppresses the drain current of the GaN-on-Si HEMTs.

Figure 4.6 □ Gate current characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C.

Increased back-plate temperatures also directly affected each DUT's transconductance and threshold voltage. As temperature increased, transconductance decreased and a positive threshold voltage-shift was observed (Figure 4.7), which coincides with the data presented in Table 4.1 through Table 4.3. The reduction in transconductance and positive-shift in threshold voltage was also determined to be a result of decreased electron density in the 2DEG [33]. Allowing the DUTs to relax for 24 hours resulted in a full recovery of the devices' transconductance characteristics and threshold voltage (Figure 4.8).

Figure 4.7 □ Transfer characteristics at 27°C, 50°C, and 100°C.

Figure 4.8 □ Transfer characteristics comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C.

Figure 4.9 S-parameter measurements at 27°C, 50°C, and 100°C.

A reduction in low frequency gain was observed in the forward transmission gain ($|S_{21}|$) measurements and an overall increase in reverse transmission gain ($|S_{12}|$) was observed with increasing temperature (Figure 4.9). No significant changes to the input and output impedances were observed during the thermal stress tests. As was determined in Subsection 3.2.2, the reverse transmission gain is correlated to the gate-drain capacitance, C_{gd} , where an increase in $|S_{12}|$ direct translates to an increase in C_{gd} . Small-signal experiments performed in [36] found that C_{gd} varied between 0.11%/°C to 0.14%/°C and that an increase in feedback capacitance reduced the overall gain and efficiency of their devices (Miller Effect). The reduction in $|S_{21}|$ also correlates to the reduction in transconductance seen in Figure 4.7 since forward transmission gain is

proportional to transconductance. After 24 hours of relaxation, the S-parameters recovered fully to their nominal values, as seen in Figure 4.10.

Figure 4.10 S-parameter comparison between a DUT at 27°C and at 27°C 24 hours after thermal exposure at 100°C.

4.2 Effect of Temperature on the Critical Voltage

4.2.1 Setup

All experiments performed in this section utilized the same test structures shown in Figure 4.1. The experiment was performed by first assigning a constant back-plate temperature to each device, and then proceeding with the step-stress procedure until the device's V_{crit} value was achieved. Devices were assigned the following temperatures and stressed accordingly: 27°C, 40°C, 50°C, 60°C, 70°C, 80°C, 90°C, and 100°C.

4.2.2 Results and Analyses

The temperature dependent V_{crit} results for eight DUTs are displayed in Figure 4.11 while Figure 4.12 shows the change in junction temperature during the step-stress procedure for the eight devices. An observable reduction in V_{crit} was found as the back-plate temperature was increased and a rise in junction temperature was found as each the gate-source voltage for each device approached V_{crit} . Similar trends seen in Figure 4.11 were observed in [37] and [38], where V_{crit} had also become less negative with increasing temperature. An explanation for the negative temperature dependence of V_{crit} may be due to a breakdown in an oxide layer ($\beta\text{-Ga}_2\text{O}_3$) that forms when thermal oxidation of GaN occurs [37].

Figure 4.11 □ Shifts in the critical voltage (V_{crit}) parameter with varying back-plate temperatures.

Figure 4.12 □ Change in junction temperature as the gate-source voltage approaches V_{crit} .

The ionic bond formed between Gallium and Oxygen can be severed under high thermal strain resulting in a breakdown of the oxide, which in turn promotes gate leakage. Work in [37] utilized a GaN cap layer between the Schottky gate contact of the device and the AlGaN layer. The devices used in this study also possess a GaN cap and may therefore be experiencing the same oxide creation and breakdown mechanisms causing the observed negative temperature dependence of V_{crit} .

An Arrhenius plot can be created from data similar to Figure 4.11 in order to determine the activation energies associated with the negative temperature-dependent critical voltage measurements. Unfortunately, due to the non-monotonic nature of the data set of Figure 4.11 and a lack of finer temperature intervals, accurate Arrhenius plots could not be produced.

5 Experiment 3: Electron Trapping and Detrapping

5.1 Trapping Experiment A: On-State Operation

5.1.1 Setup

Each DUT was assembled into a test configuration similar to that of the devices used in the temperature stress experiments (Figure 4.1, pg. 49). All devices in this experiment were shielded from light sources and the procedures were performed in dark rooms (no illumination) to prevent sunlight from acting as an electron detrapping source. The devices were first run under normal operation at $V_{gs} = -1.3V$ and $V_{ds} = 20V$ to induce electron trapping and to assess the overall nature of the gate and drain current transients over a span of 1,000 seconds, a testing interval established in [39].

After monitoring the current transients for each transistor during normal operation, the devices were then biased at $V_{gs} = -1.3V$ and $V_{ds} = -2V$ to allow for conduction within the 2DEG channel to occur and to also restrict current flow to prevent self-heating in each DUT. The back-plate of each DUT was then increased from room temperature to $80^{\circ}C$ to while the gate and drain current transients were monitored for electron detrapping events. The detrapping procedures were performed until the measured drain current transients stabilized.

5.1.2 Results and Analyses

A decreasing trend in both drain and gate current were measured during the on-state current transient experiments (Figure 5.1). For the device shown in Figure 5.1, the drain current reduced by 6.5% while the gate current decreased by 23.5% over a time period of 1,000 seconds. The collapse in drain and gate current seen in Figure 5.1 results

from hot-electron injection into the AlGa_N layer or the existence of surface traps [39]. Also suggested in [39] was the occurrence of deep-level traps within the GaN buffer, which further contributed to drain current collapse via depletion of the 2DEG.

Figure 5.1 □ Drain and gate current transients of a DUT biased in the saturation region over a 1,000 second span.

Drain and gate current transients for two devices (Device A and Device B) exposed to rising back-plate temperatures are shown in Figure 5.2 and Figure 5.3, respectively. In the graphs of Figure 5.2, both devices displayed two observable peaks in drain current. For Device A, the peaks occur at back-plate temperatures of 30°C and 70°C while for Device B the peaks occur at 35°C and 75°C. Another interesting observation was how the transition to the second detrapping peak for both devices occurred at back-plate temperatures of approximately 50°C (indicated by the dashed green circles in Figure 5.2) The sudden recovery in drain current for both devices may be an indication of electron detrapping events that are thermally activated. In [39] and [40], it was found that

the temperature-dependent detrapping events correlate with electrons trapped within the AlGa_N layer of the HEMT devices.

Figure 5.2 □ Detrapping experiment drain current transient results for (a) Device A and (b) Device B ($V_{gs} = -1.3V$, $V_{ds} = 2V$).

In Figure 5.3, both devices experienced a general rise in gate current with respect to increasing back-plate temperatures. The result could be due in part to the nature of the electron trapping and detrapping events localized within the AlGa_N layer and close to the surface of the device. An electron trapping and detrapping analysis performed in [39] found that the activation energies associated with observed electron trapping and detrapping events was approximately equal to the Schottky barrier height of the experimental HEMT devices. Therefore, if electrons tunneling through the AlGa_N buffer are presented with energy equivalent or greater than that of the Schottky barrier height, the barrier can be overcome. By overcoming the barrier height, trap-assisted tunneling (TAT) can take place and gate leakage current is enhanced as a function of increasing temperature [33].

Figure 5.3 □ Detrapping experiment gate current transient results for (a) Device A and (b) Device B ($V_{gs} = -1.3V$, $V_{ds} = 2V$).

5.2 Trapping Experiment B: High Electric Field Stress

5.2.1 Setup

Similar procedures from Section 5.1 were used in this section to perform the electron detrapping analyses. Rather than inducing electron trapping events, however, the experiment was replaced with the symmetrical high-field stress procedure used in Section 3.2. The devices in this experiment were subjected to gate-source voltages of $V_{gs} = V_{crit} = -70V$ until the gate current leakage limit of 10mA was achieved.

After subjecting each DUT to high-field stresses at $V_{gs} = V_{crit}$, the devices were then biased at $V_{gs} = -1.3V$ and $V_{ds} = -2V$ to allow for conduction within the 2DEG channel to occur and to also restrict current flow to prevent self-heating in each DUT. The back-plate of each DUT was then increased from room temperature to 80°C to while the gate and drain current transients were monitored for electron detrapping events. The detrapping procedures were performed until the measured drain current transients stabilized.

5.2.2 Results and Analyses

A representative measurement from the constant V_{crit} bias experiment is shown in Figure 5.4. A linear trend was observed in the measured gate leakage current, similar to what was seen in Figure 3.7 (pg. 38). However, as explained in Subsection 3.2.2, the rate at which the gate leakage current increases varies between devices. The gate leakage current for the device in Figure 3.7 increased at a rate of $25.641\mu\text{A/s}$ whereas the gate leakage current for the device in Figure 5.4 increased at a rate of $2.368\mu\text{A/s}$. As expected from the results of the symmetric high-field stress experiments in Subsection 3.2.2, each DUT experienced an increase in saturation current similar to what was seen in Figure 3.8 (pg. 38).

Figure 5.4 □ Results of symmetric high-field stress at V_{crit} .

Figure 5.5 and Figure 5.6 show results similar to what was found in the symmetric high-field step-stress experiments in Section 3.2. An increase in both drain current and gate current were measured post-stress. To reiterate, the increased drain current was determined to be attributed to increased tensile strain at the heterojunction due to the

application of the high symmetrical electric field distribution. The increase in gate current was said to be due to a lowering of the Schottky barrier height with respect to increasing vertical electric fields present at the metal-on-semiconductor interface.

Figure 5.5 □ Drain current characteristics before and after high-field stress for (a) Device C and (b) Device D.

Figure 5.6 ▢ Gate current characteristics before and after high-field stress for (a) Device C and (b) Device D.

The thermal detrapping results for two devices are shown in Figure 5.7. Both devices showed a rapid decrease in drain current upon biasing each device in the linear region. The rapid reduction in drain current seen in the two transient plots was an indication of electron trapping, similar to the electron trapping results of Figure 5.1. The sharp decrease in current further supports the claim that there are process-dependent defects already existent within these devices prior to applying excitation sources.

Figure 5.7 □ Detrapping experiment drain current transient results for (a) Device C and (b) Device D ($V_{gs} = -1.3V$, $V_{ds} = 2V$).

Figure 5.8 □ Detrapping experiment gate current transient results for (a) Device C and (b) Device D ($V_{gs} = -1.3V$, $V_{ds} = 2V$).

An interesting observation, however, was how both devices began to experience a recovery in drain current when the back-plate temperatures of each DUT had reached 50°C (indicated by the dashed green circles in Figure 5.7). The same effect was seen during the experiments of Subsection 5.1.2, which may indicate that there is a specific activation energy associated with the electron detrapping event that takes place. The drain current characteristics for each HEMT continued to increase until the back-plate temperatures stabilized at 80°C. Gate current transients were also measured under increasing back-plate temperatures and are shown in Figure 5.8. While the electron

detrapping event is not as noticeable in Figure 5.8, the rate of increase in gate current does increase around a back-plate temperature of 50°C.

6 Experiment 4: Small-Signal Analysis

6.1 Small-Signal Model Development in Agilent ADS

6.1.1 Setup

The small-signal model for transistors usually consists of two parts: intrinsic parameters and extrinsic parameters [30]. The intrinsic parameters are the characteristics inherent to the die-form of the transistor, while the extrinsic parameters take into account the parasitics introduced after contact formation and packaging. A small-signal representation of a FET is shown in Figure 6.1 while the associated parameter definitions are shown in Table 6.1. Because a HEMT is classified as a metal semiconductor field-effect transistor (MESFET) due to having metalized contacts and a Schottky barrier gate contact, the device can also utilize the same small-signal representation. A fit of the intrinsic parameters to the cross-sectional geometry of the GaN-on-Si HEMT is shown in Figure 6.2.

Figure 6.1 □ Small-signal model circuit with both intrinsic and extrinsic parameters.

Table 6.1 □ Small-signal parameter definitions.

Small-Signal Parameters	
<i>Extrinsic</i>	<i>Intrinsic</i>
Gate Resistance (R_g)	Gate-Source Resistance (R_{gs})
Source Resistance (R_s)	Drain-Source Resistance (R_{ds})
Drain Resistance (R_d)	Gate-Source Capacitance (C_{gs})
Gate Inductance (L_g)	Gate-Drain Capacitance (C_{gd})
Source Inductance (L_s)	Drain-Source Capacitance (C_{ds})
Drain Inductance (L_d)	Transconductance (g_m)
	Electron Transit Time (τ)

Figure 6.2 □ Cross-sectional geometry of the GaN-on-Si DUT with intrinsic parameters overlaid.

C_{gs} , R_{gs} , C_{ds} and R_{ds} were estimated using the measured S_{11} and S_{22} data for the transistor and the characteristic impedance of the system (50Ω in this case). The parameters were extracted using data at 100MHz to minimize contributions from

parasitic inductances. The S-parameters were measured using an Anritsu MS4622B (using the test conditions in Subsection 2.2.3) and the input and output impedances were extracted in Agilent ADS by importing an S2P file from the VNA into the program.

R_{gs} and C_{gs} were first calculated using Equations (6.1) and (6.2). C_{dso} in Equation (6.3) is the series capacitance associated with the imaginary part of the S_{22} impedance. C_{dso} will be used in a series to parallel transformation process to solve for the drain-source capacitance. R_{ds} and C_{ds} were calculated using the transformations of Equation (6.4) and Equation (6.5). More information on series to parallel and parallel to series transformations can be found in [41].

$$R_{gs} = Z_o \text{Re}(z_{11}) \quad (6.1)$$

$$C_{gs} = \frac{1}{\omega Z_o |\text{Im}(z_{11})|} \quad (6.2)$$

$$C_{dso} = \frac{1}{\omega Z_o |\text{Im}(z_{22})|} \quad (6.3)$$

$$R_{ds} = Z_o \text{Re}(z_{22}) + \frac{1}{Z_o \text{Re}(z_{22}) * (\omega C_{dso})^2} \quad (6.4)$$

$$C_{ds} = \frac{1}{Z_o \text{Re}(z_{22}) * R_{ds} \omega^2 C_{dso}} \quad (6.5)$$

The value of g_m was estimated using a transistor parameter analyzer and the electron transient time associated with the drain current, τ , is a parameter that was tuned

using Agilent ADS. The remaining parasitic components were tuned in Agilent ADS until the small-signal model S-parameter data coincided with the measured data.

6.1.2 Results and Analyses

A generic small-signal model (Figure 6.3) was first created and replicated three times in order to produce pre-stress, post-stress, and recovery conditions (refer to Chapter 3.2.3). Before performing the S-parameter data fits, an S2P file with the data for the previously mentioned stress conditions were imported into Agilent ADS in order to act as a template by which the simulated S-parameters could be fit to.

Figure 6.3 □ Constructed small-signal model using Agilent ADS.

The intrinsic section for each of the three small-signal models used in the fitting process were developed using initial DC and RF measurements in conjunction with Equations (6.1) through (6.5) (refer to the previous subsection). After placing the estimated values for each intrinsic component, the extrinsic values were then set to arbitrary values. Without tuning, S-parameters plots similar to Figure 6.4 would result. Next, the “Tune Parameters” interface was used to manually adjust simulated small-signal parameters until the associated model began to correlate with the measured data.

The tuning interface is shown in Figure 6.5 while an example of an approximated data fit is shown in Figure 6.6.

Figure 6.4 Simulated S-parameters prior to fitting routine.

Figure 6.5 Agilent ADS “Tune Parameter” interface.

Figure 6.6 Simulated S-parameters after using the parameter tuning interface in ADS.

Small-signal model fits for each of the three DUT conditions were developed using the previously mentioned tuning schemes. Imported S-parameter data for the three DUT conditions are shown in Figure 6.7 while the simulated S-parameters resulting from the small-signal parameter fit to the measured data are shown in Figure 6.8. While not entirely accurate, the resulting data approximations of Figure 6.8 allow for estimations as to how the small-signal parameters for a DUT change between pre-stress, post-stress, and relaxed states. Major changes in small-signal parameters between each of the three DUT conditions are displayed in Table 6.2.

Figure 6.7 □ Measured S-parameters used for small-signal modeling.

Figure 6.8 □ Simulated S-parameters after performing a small-signal parameter fit to measured data.

Table 6.2 Major small-signal parameter changes for pre-stress, post-stress, and rested HEMT test conditions.

<i>Stress State</i>	Major Change in Small-Signal Parameters			
	<i>Source Resistance (Ω)</i>	<i>Gate-Drain Capacitance (pF)</i>	<i>Drain-Source Capacitance (pF)</i>	<i>Drain-Source Resistance(Ω)</i>
Unstressed	1.200	0.216	3.546	211
Stressed	0.78 (35.00%)	0.146 (32.41%)	2.546 (28.20%)	253 (19.91%)
Rested (6 days)	0.890 (25.83%)	0.188 (12.96%)	3.546 (0.00%)	253 (19.91%)

As seen above, the source resistance (R_s), the gate-drain capacitance (C_{gd}), the drain-source capacitance (C_{ds}), and the drain-source resistance (R_{ds}) experienced the most substantial changes amongst the simulated small-signal parameters. As was expected from Subsection 3.2.2, C_{gd} reduced by 32.41% after the HEMT had been exposed to high symmetric-field stress. In general, a capacitor can be defined by Equation (6.6) [42]:

$$C = \frac{Q}{V} \quad (6.6)$$

Given that the voltage distribution across the capacitor is constant, as the charge (Q) stored in the capacitor changes, the capacitance changes proportionally. Because C_{gd} is experiencing a reduction in capacitance after being exposed to a fixed high field stress, the only direct cause would be due to a reduction in charge stored in the AlGaIn barrier. Therefore, the results may indicate how inducing the inverse piezoelectric effect in the GaN-on-Si HEMTs promotes the detrapping of electrons inherent from process-dependent defects created during the growth of the epitaxial layers. Furthermore, a 12.96% recovery in C_{gd} not only indicates how native defects in the device promote electron trapping, but how degradation in the AlGaIn barrier induced by high field stress is not entirely permanent.

DC stress experiments performed in [25] on GaN-on-Si HEMTs have demonstrated non-permanent changes in the drain resistance (R_d) and the source

resistance (R_s) when the devices under test were exposed to voltages below V_{crit} . When V_{gs} was set equal to or greater than V_{crit} , the changes to the parameters were irreversible. While R_d did not appear to experience any significant alterations over the test frequency range after high field stress, R_s increased by approximately 35%. While the DUTs were exposed to extensive periods of high voltage stressing at V_{crit} , the parameter R_s was able to recover by 25.83%, which further indicates that the inverse piezoelectric effect may not be the sole driving factor for permanent device degradation and defect formation in GaN-on-Si HEMTs.

Both the drain-source resistance (R_{ds}) and capacitance (C_{ds}) increase post-stress by 19.91% and 28.20%, respectively. After a relaxation period of 6 days, C_{ds} is able to make a full recovery. The value of R_{ds} , however, remained unchanged after the 6-day rest period after the DUT was exposed to high-field stress. From Equation (1.3) (pg. 5), an increase in R_{ds} may have potentially contributed to a proportional increase in the f_{max} of the transistor. The previously mentioned results may be an indication that there are some defect formations in the GaN barrier following high field stress at the device's V_{crit} occurring concurrently with electron trapping and detrapping events.

7 Conclusion

Over the course of the research, several mechanisms leading to either device degradation or recovery have been recorded and linked to high electric field-related conditions in GaN-on-Si HEMTs. The high-field stress experiments have shown that GaN-on-Si devices with epitaxial structures similar to Figure 2.1 should experience critical voltages around -70V while in a $V_{ds} = V_s = \text{ground}$ (symmetric) configuration and undergo rapid increases in gate leakage current at the onset of V_{crit} . Though the electrical characteristics of the HEMTs change with respect to biasing the devices at V_{crit} , the changes are not permanent and device characteristics are slightly recoverable. While observable changes occur in the symmetric off-state condition, no noticeable alterations were noticeable in the $V_{gs} = -5V$ off-state (asymmetric) scenario.

Variations in ambient temperature were found to have a direct influence on the overall performance of the GaN-on-Si HEMTs. Increasing the back-plate temperatures of the devices alters their corresponding junction temperatures, which in turn degrades their DC and RF characteristics under increasing temperature. The observations are undesirable in a practical sense because slight temperature gradients in an ambient setting can shift the operating point of the transistors. For example, if the HEMTs are utilized in a power amplifier (PA) setting and multiple PAs are required for a specific application, the self-heating of the devices can directly influence an adjacent device's performance. Another issue stemming from temperature variation was the direct correlation between increasing base-plate temperatures and a reduction in the absolute value of V_{crit} . The observed positive shift in V_{crit} as a function of increasing base-plate temperature indicated

that less gate drive is required to alter the characteristics of the devices, promote an increase in gate leakage current, and potentially force them into abrupt failure modes.

The analyses performed in Chapter 5 have further provided evidence of hot electron injection into the buffer layers while the devices are operating in the on-state. The detrapping experiments following the on-state analyses show specific back-plate temperatures at which drain current suddenly begins to recover. Similar detrapping observations were found when performing the same experiment on devices previously exposed to high electric fields. The aforementioned phenomena may indicate the existence of a specific activation energy at which electrons are released from buffer traps, which in turn leads to 2DEG recovery. Another important observation drawn from the studies of Chapter 5 (in conjunction with Chapter 3) was how high-field stress promotes (rather than inhibits) drain current recovery through symmetric tensile strain of the semiconductor materials and also assists in electron detrapping.

The small-signal analyses performed on stressed HEMTs revealed major changes to four specific small-signal parameters: source resistance (R_s), gate-drain capacitance (C_{gd}), drain-source capacitance (C_{ds}), and drain-source resistance (R_{ds}). The R_s and C_{gd} parameters of the transistors experienced reductions up to 30% due to stress and then underwent some partial recovery over a 6-day relaxation period. A reduction in C_{gd} further supports the claim that the symmetric high field stresses applied to the devices promotes electron detrapping in the AlGaN through the inverse piezoelectric effect due to a reduction in localized charge. Enhanced unilateral performance also results from the high-field stress due to a reduction in C_{gd} , which in turn relates to the devices' reverse transmission gain. A reduction in C_{ds} was observed post-stress with full recovery after

relaxation. The only parameter to have experienced permanent alteration was R_{ds} , which may suggest some degradation within the GaN barrier of the HEMTs. Though the change to R_{ds} was permanent, the permanent increase to the resistance proportionally increases the maximum frequency of oscillation for the transistor as well.

Table 7.1 Summary of experimental results.

Exp. #	Chapter # and Page #	Experimental Details		
		Exp. Name	Sub-Exp. Name	Results Summary
1	3 (pg. 32)	High Electric Field Stress Analyses	Asymmetric Electric Field Stress	No significant changes to device characteristics were observed after stressing the devices to 120V on the drain in the off-state.
			Symmetric Electric Field Stress	Critical voltage for the DUT determined to be $V_{gs} = -70V$ in the $V_{ds} = 0V$ off-state. Rapid increase in gate current as V_{gs} approaches the critical voltage. Changes to device characteristics post-stress included the following: an increase in saturation drain current, an increase in gate current, negative threshold voltage shift, and a reduction in $ S_{21} $ (correlating to C_{gd}). Noticeable device recovery after an extended period of no electrical excitation; characteristic changes post-stress are not permanent.
2	4 (pg. 49)	Temperature Characterization	Effect of Temperature on General Device Performance	A reduction in drain current, an increase in gate current, a positive shift in threshold voltage, and a rise in gate-drain capacitance observed. Characteristic changes are non-permanent.
			Effect of Temperature on the Critical Voltage	Critical voltage shows a negative temperature dependence; absolute value of the critical voltage decreases with increasing temperature.
3	5 (pg. 62)	Electron Trapping and Detrapping	Trapping Experiment A: On-State Operation	On-state operation promotes electron trapping, which is observable through the reduction in gate and drain current transients. Observable recovery events during the application of thermal excitations points towards activation energies associated with electron detrapping.
			Trapping Experiment B: High Electric Field Stress	The test indicates that stressing the devices at the critical voltage enhances device performance, indicating possible electron detrapping during stress. Further evidence of a specific activation energy associated with electron detrapping.
4	6 (pg. 70)	Small-Signal Analysis	~	A reduction in source resistance (R_s), a reduction in gate-drain capacitance (C_{gd}), a reduction in drain-source capacitance (C_{ds}), and an increase in drain-source resistance were observed post-stress. Only permanent parameter change was observed with drain-source resistance.

Future Project Recommendations

Several improvements can be made to the research performed in this paper. The first would be to improve upon the temperature studies performed in Chapter 4. The structure that was created for the experiments (Figure 4.1, pg. 49) was very rudimentary and could have potentially been the primary reason why the temperature-dependent critical voltage results in Section 4.2 were not monotonic in nature. Also, an increased pool of HEMTs and experimental temperature points would produce more reliable results. For example, instead of making gate leakage current measurements from 27°C, 40°C - 100°C in 10°C steps, it might be more advantageous to make the measurements in temperature steps of 5°C (even finer if the data acquisition becomes more efficient). Increasing the amount of temperature data points taken during the critical voltage experiments would also aid in the construction of Arrhenius plots, which would help identify the activation energies associated with the temperature dependence of V_{crit} .

Better methods and testing strategies need to be developed for the electron trapping and detrapping experiments performed in Chapter 5. While some detrapping events were observed during experimentation, the detrapping methodology needs to be redefined and improved upon. For example, work done in [39] utilizes pulsed biasing during the electron trapping and detrapping procedures, and also analyzes the time constants associated with the trapping and detrapping events. Knowledge about the time constants can explain where within the device the traps are localized (e.g. deep traps in the GaN buffer). C-V measurements may also be beneficial in this study.

While the tuning method used to extract the small-signal parameters in Chapter 6 produced quick and rough estimates as to how the parameters were changing with respect to high-field stress, the method is by no means perfectly accurate. The tuning performed in Agilent ADS was based off of a “best-guess” methodology where-in simulated parameters were tuned until the simulated S-parameters were comparable to what was measured. A better strategy would be to investigate the mathematical models presented in papers such as [43], [44], [45], [46], and [47]. The papers discuss proper biasing schemes to extract various sets of S-parameters, which are in turn involved in Z- or Y-transformations to map the measured parameters to small-signal parameters. The methods were attempted during the RF studies performed in this paper, but the results from the mathematical models could not be made to converge.

One particular concept that would be beneficial in the investigation of GaN-related defect mechanisms would be a thorough $1/f$ -noise analysis. Work performed in [48] found that higher gate leakage currents in AlGaIn/GaN HEMTs contributed to reduced noise performance. For fundamental background on noise sources, refer to [49]. Preliminary research on a $1/f$ -noise measurement system was performed at the early stages of this research, but unfortunately the efforts could not be carried out in full. Preliminary $1/f$ -noise measurement system designs and simulations are presented in Appendix A: $1/f$ -Noise System Prototype. Because HEMTs are very attractive in power amplification applications that may involve mixers in the system design (such as in a phase-locked loop, or PLL), noise can become a major concern, especially if the noise can be upconverted to the signal of interest (leading to degraded signal to noise ratios). An interesting investigation would be to evaluate the noise performance of GaN-on-Si

HEMTs exposed to high electric field stress and identify mechanisms that could contribute to increased noise levels (e.g. electron detrapping contributing to increased burst noise). To allow for robust noise measurements, the system would need to be encased in a PEC casing to prevent external noise sources from influencing measurements.

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Appendices

Appendix A: 1/f-Noise System Prototype

An improved 1/f noise measurement system was proposed by Clemente Toro, a post-graduate student from the University of South Florida [50]. Toro's thesis involved developing a general 1/f noise system that could provide broad noise analyses from 1Hz to 10MHz for a number of microwave transistors (BJTs, FETs, HBTs, HEMTs, etc.).

Figure A.1 shows Toro's arbitrary noise system configuration.

Figure A.1 □ 1/f-noise measurements system for microwave transistors proposed by Clemente Toro.

The device under test used in the design of the system was the Nitronex NPTB00004 GaN-on-Si HEMT. While flicker noise exists at both the drain and the gate of the device, the prototype was only configured to make measurements from the drain terminal for proof-of-concept purposes. Each proceeding section in this paper will outline the general purpose of each stage followed by a design analysis.

Bias Filter Stages

A majority of power supplies produce relatively noisy DC signals, which can directly affect the signal-to-noise ratio during 1/f measurements [50], [51]. Each of the bias filters were designed to provide sufficient suppression of signals greater than or equal to 1Hz. The recommended supply filters by Toro for a FET are shown in Figure A.2 and Figure A.3.

Figure A.2 □ Gate supply filter for a FET.

Figure A.3 □ Drain supply filter for a FET.

While Toro's design suggestions are practical for arbitrary FETs, the supply filter for the NPTB00004 must be designed with some parameters in mind. The AN-009 application note provided by Nitronex discusses the need for a 200Ω resistor along the

gate bias feed path to dampen oscillations caused by the device [21]. If the filter of Figure 2 is used, the DC signal would only see 150Ω in the path (the capacitor to ground is an open-circuit to DC). By replacing the 50Ω resistor with a 100Ω resistor, the 200Ω value can be achieved. The final circuit design is shown in Figure A.4 and the Agilent frequency response simulation is shown in Figure A.5. The simulation was performed assuming a 50Ω system.

Figure A.4 □ Gate supply filter for the NPTB00004 HEMT.

Figure A.5 □ Gate supply filter frequency response.

The drain supply filter of Figure A.3 was designed for arbitrary FETs, where low supply voltage was used. The GaN HEMT, however, is a high-powered device where nominal power dissipation can be anywhere from 2-3W. Therefore, high current draw will flow through the drain feed path depending on the value of V_{gs} and V_{ds} (the NPTB00004 is a depletion-mode device, so V_{gs} controls the depletion region of the device). With high current draw and resistive elements in the feed path, large voltage drops will be present in the circuit of Figure A.3.

To remove the voltage drops, the filter could be redesigned by removing the resistors and placing several capacitors in parallel along the supply line to filter a broad range of AC signals (large capacitors filter low frequencies, while small capacitors filter high frequencies). There are two reasons why the two series resistors are needed, however. The first reason is that the 500Ω resistor provides a high impedance to the $1/f$ -noise at the drain of the device. If the 500Ω resistor is removed, the $1/f$ noise could potentially see an AC short to ground via the feed line capacitor instead of taking the path through the transimpedance amplifier. Toro suggests keeping the 500Ω resistor as the minimum-valued resistor required to prevent the $1/f$ noise from feeding back into the supply filter. The 100Ω resistor is required to aid in dampening oscillatory effects in the event that the $4700\mu\text{F}$ capacitor resonates with the inherent inductance of the supply.

The drain supply was simulated using Agilent ADS. The first simulation utilizes the circuit configuration of Figure A.3. To provide better power distribution along the feed path, two 200Ω resistors were placed in parallel to form the 100Ω resistor, while two $1\text{k}\Omega$ resistors were placed in parallel to achieve the 500Ω resistor. Figure A.6 shows the

NPTB00004 ADS model in a DC configuration. Table A.1 and Figure A.7 show the resulting DC characteristics and frequency response, respectively.

Figure A.6 DC circuit simulation setup for the NPTB00004.

Table A.1 NPTB00004 DC characteristics using Toro's drain supply filter design (with parallel resistors).

Vgs	Id.i	Vdrain	I1k.i	I200.i
	Vds=10.000	Vds=10.000	Vds=10.000	Vds=10.000
-2.000	2.169 mA	8.699 V	1.084 mA	1.084 mA
-1.950	2.882 mA	8.271 V	1.441 mA	1.441 mA
-1.900	3.787 mA	7.728 V	1.894 mA	1.894 mA
-1.850	4.921 mA	7.047 V	2.460 mA	2.460 mA
-1.800	6.324 mA	6.206 V	3.162 mA	3.162 mA
-1.750	8.043 mA	5.174 V	4.022 mA	4.022 mA
-1.700	10.14 mA	3.917 V	5.069 mA	5.069 mA
-1.650	12.69 mA	2.389 V	6.343 mA	6.343 mA
-1.600	15.46 mA	723.3 mV	7.731 mA	7.731 mA
-1.550	16.00 mA	398.3 mV	8.001 mA	8.001 mA
-1.500	16.17 mA	296.8 mV	8.086 mA	8.086 mA
-1.450	16.27 mA	240.3 mV	8.133 mA	8.133 mA
-1.400	16.33 mA	203.2 mV	8.164 mA	8.164 mA
-1.350	16.37 mA	176.9 mV	8.186 mA	8.186 mA
-1.300	16.40 mA	157.4 mV	8.202 mA	8.202 mA
-1.250	16.43 mA	142.4 mV	8.215 mA	8.215 mA

Figure A.7 □ Drain supply filter frequency response.

As seen in Table A.1, after setting V_{ds} to 10V and stepping V_{gs} more positive, the drain current starts to increase and the effective drain voltage at the device starts to diminish. In other words, the resistive elements alter the quiescent point of the HEMT. The aforementioned effect is shown graphically in Figure A.8.

Figure A.8 □ Q-point alteration due to resistive components in the drain supply filter.

Next, the 200 Ω parallel pair was replaced with a 10 Ω resistor. While the 10 Ω resistor provides subtle dampening effects in comparison to the effective 100 Ω resistor, the power dissipation and voltage drop would be smaller. In the event that the 4700 μ F parallel resonates with the inherent inductance of the supply, the 10 Ω would still aid in lowering the Q-factor of the resonant components (as seen from a series to parallel transformation in [41]). Table A.2 and Figure A.9 show the DC characteristics and frequency response using the altered filter configuration, respectively. While the Q-point shift is similar to the effects seen in Figure A.8, the overall drain current to the device is higher and the voltage at the drain of the device is also higher. The frequency response of the redesigned filter is also similar to that of the original design.

Table A.2 NPTB00004 DC characteristics using the 10 Ω resistor in the drain filter.

Vgs	Id,i	Vdrain	Ihk,i
	Vds=10.000	Vds=10.000	Vds=10.000
-2.000	2.173 mA	8.892 V	1.086 mA
-1.950	2.890 mA	8.526 V	1.445 mA
-1.900	3.800 mA	8.062 V	1.900 mA
-1.850	4.940 mA	7.481 V	2.470 mA
-1.800	6.352 mA	6.761 V	3.176 mA
-1.750	8.079 mA	5.879 V	4.040 mA
-1.700	10.17 mA	4.811 V	5.087 mA
-1.650	12.70 mA	3.525 V	6.348 mA
-1.600	15.73 mA	1.980 V	7.863 mA
-1.550	18.40 mA	618.4 mV	9.198 mA
-1.500	18.84 mA	392.2 mV	9.419 mA
-1.450	19.01 mA	302.7 mV	9.507 mA
-1.400	19.12 mA	250.2 mV	9.559 mA
-1.350	19.19 mA	214.9 mV	9.593 mA
-1.300	19.24 mA	189.5 mV	9.618 mA
-1.250	19.27 mA	170.4 mV	9.637 mA

Figure A.9 □ Frequency response using the 10Ω resistor in the drain filter.

Another design strategy involves replacing the 1kΩ parallel pair with a diode. The diode would not only improve upon the unwanted voltage drops and allow for a more accurate biasing of the Q-point, but would also restrict noise current from filtering through the 4700μF capacitor. The main trade-off with using a diode is that the component is a semiconductor material, which will contribute noise c to the overall spectral response. The DC characteristics and frequency response using a diode are seen in Table A.3 and Figure A.10, respectively. The last two columns in Table A.3 were used to determine the voltage drop across the diode during the sweep.

Table A.3 NPTB00004 DC characteristics using the diode in the drain filter.

Vgs	Id.i	Vdrain	var("VD+")	var("VD-")
	Vds=10.000	Vds=10.000	Vds=10.000	Vds=10.000
-2.000	2.178 mA	9.112 V	9.978 V	9.112 V
-1.950	2.907 mA	9.091 V	9.971 V	9.091 V
-1.900	3.840 mA	9.067 V	9.962 V	9.067 V
-1.850	5.020 mA	9.042 V	9.950 V	9.042 V
-1.800	6.499 mA	9.014 V	9.935 V	9.014 V
-1.750	8.333 mA	8.983 V	9.917 V	8.983 V
-1.700	10.58 mA	8.948 V	9.894 V	8.948 V
-1.650	13.32 mA	8.909 V	9.867 V	8.909 V
-1.600	16.62 mA	8.865 V	9.834 V	8.865 V
-1.550	20.55 mA	8.814 V	9.795 V	8.814 V
-1.500	25.19 mA	8.757 V	9.748 V	8.757 V
-1.450	30.61 mA	8.693 V	9.694 V	8.693 V
-1.400	36.89 mA	8.620 V	9.631 V	8.620 V
-1.350	44.08 mA	8.539 V	9.559 V	8.539 V
-1.300	52.22 mA	8.448 V	9.478 V	8.448 V
-1.250	61.36 mA	8.348 V	9.386 V	8.348 V

Figure A.10 Frequency response using the diode in the drain filter.

From Table A.3 and Figure A.10, not only do the DC characteristics improve drastically, but the filtering response also improves at lower frequencies. The final drain supply filter configuration is shown in Figure A.11. Because no more than 100mA of drain current is expected through the drain path, the power rating for the 10Ω resistor is negligible (100mW worst case). A 1N4001 was chosen for its high power rating of 3W.

The voltage rating for the 4700 μ F electrolytic capacitor is also imperative to the design. A voltage rating of 25V was chosen for the electrolytic.

Figure A.11 □ Drain supply filter design.

Transimpedance Amplifier Stage

A transimpedance amplifier (TIA) was required to convert the noise current created by the drain into a noise voltage. Conversion of the drain noise current to a noise voltage allows for the amplification of the noise voltage using cascaded voltage amplifier stages. Design of the TIA requires the use of a low-noise operational amplifier with a large gain-bandwidth product. The low-noise specification is imperative to the design of the TIA (as well as the voltage amplifiers) because the flicker noise could potentially be mixed-in with or dominated by the noise generated by the op-amp. The gain-bandwidth product (GBP) specification must also be considered due to the general rule of thumb that the GBP of an op-amp must be at least an order or magnitude larger than the closed-loop gain multiplied by the maximum signal frequency input into the system.

The LT6200-5 was chosen for its GBP of 800MHz (500MHz worst-case). Knowing that the system needs to work up to 10MHz and the voltage amplifiers provide a gain of 5V/V, the GBP specification for the LT6200-5 was sufficient. The TIA design using the LT6200-5 is shown in Figure A.12, which is Toro's suggested design. Using

100 Ω as feedback provides a TIA sensitivity of 100V/A. Figure 13 shows an example of the TIA operating with the 100V/A configuration.

Figure A.12 LT6200-5 transimpedance amplifier with a sensitivity of 100V/A.

Figure A.13 LT6200-5 current to voltage conversion (100V/A sensitivity, voltage left-axis, current right-axis).

Voltage Amplifier Stages

Toro designs both inverting amplifier stages to provide 14dB (5V/V) of amplification, resulting in 28dB (25.12V/V) of total amplification. Toro's suggested design for a single amplifier stage is shown in Figure A.14.

Figure A.14 □ Voltage amplifier with a gain of 14dB (5V/V).

After simulating the design using an LT6200-5 op-amp, observable peaking was found in the response (Figure A.15). To compensate the amplifier, capacitors ranging from 1pF up to 10pF were iteratively placed in parallel with the 50kΩ to reduce the peaking seen near the cut-off frequency of the amplifier (adds a pole). The capacitors eliminated the peaking effect, but decreased the cut-off frequency of the amplifier. The amplifier was originally designed to provide 14dB of gain from 1Hz up to 10MHz. Toro used a 3-dB frequency of 17MHz (11dB) as a figure of merit in his design.

To provide sufficient gain up to 10MHz, the 10kΩ and 50kΩ resistors were scaled down to 1kΩ and 5kΩ, respectively. The parallel feedback capacitor was also swept between 1pF to 10pF during LTSpice simulations. The LTSpice simulated circuit is shown in Figure A.16 and the resulting responses are shown in Figure A.17.

Figure A.15 ☐ Observable peaking using the suggested amplifier configuration with the LT6200-5.

Figure A.16 ☐ Simulated circuit for the LT6200-5 voltage amplifier with swept feedback capacitor values.

Figure A.17 Simulated responses for the LT6200-5 voltage amplifier with swept feedback capacitor values.

A 1pF feedback capacitor was chosen in the final design of the voltage amplifier.

The circuit for the voltage amplifier stage is shown in Figure A.18 and the resulting response is shown in Figure A.19.

Figure A.18 LT6200-5 inverting amplifier configuration.

Figure A.19 LT6200-5 inverting amplifier response (14dB of gain).

Next, an additional inverting amplifier was added to the previous amplifier. The circuit is shown in Figure A.20 and the overall frequency response is shown in Figure A.21. Note that in the final cascade implementation, a 5.1k Ω resistor was chosen in place of each 5k Ω due to its availability in metal film form (carbon film resistors are known to produce 1/f noise [52]).

Figure A.20 LT6200-5 inverting amplifier cascade.

Figure A.21 □ LT6200-5 inverting amplifier cascade response (28dB of gain).

DC Blocks

Each of the DC blocks in the system consist of a 2200uF electrolytic capacitor in parallel with a 0.15uF capacitor, which creates a relatively lossless block between each stage. The circuit for the DC block is shown in Figure A.22 and the associated frequency response for the can be seen in Figure A.23. Note that the DC block at the drain of the HEMT must have capacitors with voltage ratings of at least 25V. The DC blocks for the remainder of the system do not require such stringent voltage ratings.

Figure A.22 □ Lossless DC block.

Figure A.23 □ DC block frequency response.

Design Summary

The final $1/f$ noise measurement system design is shown in Figure A.24. The associated component values are shown in Table A.4. Note that the system was designed for through-hole parts on a copper-clad perfboard for proto-typing purposes. If the design is able to demonstrate a proof-of-concept, a second iteration will be done on a PCB using higher-quality components. It should be noted that along with certain resistors, specific electrolytic capacitors also contribute to low-frequency noise. Work done in [53] found that the noise contribution from tantalum-based electrolytic capacitors were bias dependent. Therefore, to increase the robustness of the $1/f$ -noise measurement system, special attention should be made when selecting specific resistors and capacitors in the final design.

Figure A.24 Final 1/f-noise measurement system design.

Table A.4 Component values for the 1/f-noise measurement system design.

Reference	Part Description	Package Type
C7, C10	1pF, ceramic, 50V, $\pm 0.25\text{pF}$	Radial
C4, C6, C9, C12	0.15uF, ceramic, 25V, 10%	Radial
C3	2200uF aluminum electrolytic, 25V, 20%	Radial can - through hole
C5, C8, C11	2200uF aluminum electrolytic, 10V, 20%	Radial can - through hole
C1	4700uF aluminum electrolytic, 10V, 20%	Radial can - through hole
C2	4700uF aluminum electrolytic, 25V, 20%	Radial can - through hole
R3	100hm, 1/4W, 1% Metal Film	Axial
R1, R2, R4	1000hm, 1/4W, 1%, metal film	Axial
R5, R7	1kOhm, 1/4W, 1%, metal film	Axial
R6, R8	5.1kOhm, 0.40W, 1%, metal film	Axial
D1	1N4001 general purpose rectifier	Through-hole
U1, U2, U3	LT6200 - 165MHz, Rail-to-Rail Input and Output, 0.95nV/VHz Low Noise, Op Amp Family	SOIC-8
X1	NPTB00004 5W GaN HEMT	SOIC-8 (with source pad)

Appendix B: Cadsoft Eagle Tutorial

PCB Design: Package Creation

Download the free version of Cadsoft Eagle from <http://www.cadsoftusa.com/download-eagle/?language=en>. There are usually three steps involved in the entire PCB process: package creation, schematic entry, and PCB layout. The package creation step, in some cases, can be skipped if the package/footprint for a specific part is already given.

Start by creating a new library in Eagle (refer to Figure B.1). After a new library window appears, name the library file and save it to the “lbr” directory where Eagle was installed to (example: JohnDoeCustom.lib).

Figure B.1 ☐ Cadsoft Eagle main window - starting a custom library.

During the component creation process, there are three steps the user must proceed through to make the final part. The three steps are Package Creation, Symbol Creation, and Device Creation. Package Creation involves creating a footprint for the component, which will consist of laying out land patterns (surface mount, through-hole,

etc.), silk screens, and so forth. Symbol Creation involves creating the schematic symbol for the component. Component Creation links the package to the schematic symbol. Refer to Figure B.2 for the library window.

Figure B.2 ☞ Cadsoft Eagle library window.

Begin component creation by selecting the package button. When the “Edit” window appears, name the package after the component’s package type, and then save the name. A black background with a grid should appear; the window has now changed to the package creation window. Before setting up the package in the workplace, the user should have consulted the component’s datasheet for recommended land patterns. Once the land pattern is found, start the process by setting up the work space’s grid (Figure B.3). After setting a desired grid area, start placing either surface mount pads (SMDs) or through-hole pads onto the workspace using the recommended land patterns from the datasheet. The size and shape of each pad can be altered by going to the top of the work space and changing the dimensions (Figure B.4). To make a circular SMD, set the roundness to 100%.

Figure B.3 □ Package creation work space with typical grid setting.

Figure B.4 □ SMD and through-hole size/dimension options, respectively.

Note that the SMDs should be on the “Top” (copper) layer, and the through-hole pads should be on the “Pads” layer. To check if certain aspects of the design are on the right layers, select the visible layers drop down and select/deselect layers. After placing pads, silkscreens can be placed to outline the component’s package and display the component’s schematic reference number and manufacturer number. Component outline silkscreens can be created using the *Wire* and *Circle* tools on the left toolbar, while the

reference number (Name) and manufacturer's part number (Value) silkscreens can be created using the **Text** tool. Note that during component creation, the reference number and part number should be set as ">Name" and ">Value." The actual designators will be placed onto the layout workspace once the schematic is created. The component outline silkscreens should be created on the "tPlace" layer while the Name and Value silkscreens should be placed on the "tName" and "tPlace" layers, respectively. The "t" in each layer represents the top layer ("b" would then represent the bottom layer). As an example, a finished part is shown in Figure B.5 (8-pin DIP socket for the TL072).

Figure B.5 Package design example - 8-pin DIP socket for the TL072.

A few tips about the package creation process should be mentioned. The first is the shortcut to place pads down through coordinate commands. At the top of the work

space is a white command prompt bar. Instead of clicking and dragging pads to a specific coordinate, each point can be entered in an (x y) format. For example, if a pad is required 60mils x 100mils from the origin (and the units are in inches), the coordinate command would be (0.060 0.100) (type the coordinates and press enter).

The second tip revolves around the use of the crosshair at the center of the workspace. The crosshair is used as an indicator for how a part will be placed onto the user's cursor when being dragged and placed onto the workspace during the layout stage. Therefore it is a matter of preference for the user on how the part will be oriented onto the workspace during the package stage. A general method used is to center the bottom-left-most pad onto the cross-hair and design the package from there. Once the package is completed, select the **Group** tool and drag a selection box over every single package element. Once the package is highlighted, select the **Move** tool and right-click near the package. At the bottom of the tool-tip that shows up, select **Move: Group**. Center the package around the center cross-hair and save the design.

The third tip is in regards to pad naming. To make connections between the schematic symbol and the package simple, each pad can be given a name. Type "Name" into the command prompt and select each pad. A new window will come up where the respective pin name or designator can be edited. The same process should be performed during schematic symbol creation when handling pin names.

Next, create a schematic symbol by selecting the associated icon. An "Edit" prompt will appear much like during the package process. Name the schematic symbol after the manufacturer's part name and proceed. A window similar to the package creation work space will appear, but the background will be white (Figure B.6). The

default grid configuration should be acceptable for this process, but can also be altered to the user's preferences. The entire schematic process is similar to that of the package process, where each pin will receive a name/designator and the symbol will be labeled with the >Name and >Value conventions. An example symbol can be seen in Figure B.7 (8-pin DIP socket for the TL072).

Figure B.6 Schematic symbol creation work space.

Figure B.7 □ Schematic symbol design example - 8-pin DIP socket for the TL072.

Finally, select the **Device** icon to complete the component creation process. A new window should appear as seen in Figure B.8. First, click the **Add** button and select the schematic symbol that was created. Next, click the **New** button and add the created package symbol. Finally, press the **Connect** button to link the schematic and package symbols together. The connection window should look like Figure B.9. If the pads and pins from the previous processes were given their respective names, they should line up directly with one another as seen in Figure B.9. If not, the user will have to search for and select the connections between each associated pad and pin. By the end of the connection process, the “Connection” section in the last-third of the window should be full. Press “OK” to complete the component process.

Figure B.8 Device creation work space.

Figure B.9 Schematic and package symbol connection window.

PCB Design: Schematic and Layout

Start a new project in Eagle by referring to Figure B.10, and then start a new schematic as seen in Figure B.11.

Figure B.10 Starting a new project in Eagle.

Figure B.11 Starting a new schematic in the project directory.

A new schematic work space should appear, as seen in Figure B.12. All parts should be accessible via the **Add** button (including parts from the custom-made library). Place parts onto the schematic work space and wire them together using the **Wire** tool. All components can have their reference names and values changed by selecting the **Name** and **Value** tools. One useful tool to use is the **Group** tool when having to move or copy multiple components at any given time. For example, group all desired components together by creating a selection box with the **Group** tool. Once all desired components are highlighted, select either the move or copy option, and then right-click the work space. Finally, when the tool-tip appears, select “Group: “<Action>,” where <Action> is either move, copy, or another command. After completing the schematic entry, press the **Board** button at the top of the work space to generate a board layout from the schematic.

Figure B.12 Schematic workspace.

When Eagle generates the board layout, the program will create a component pile at the bottom left corner of the layout work space (Figure B.13). Before routing the board, it is usually wise to specify the design rules for the board, which follow the guidelines established by the board fabricator. To access the design rules, go to **Tools > DRC...** and input the necessary fabrication limitations.

Figure B.13 □ Layout workspace.

As seen in Figure B.13, the white rectangular boundaries within the work space are the board dimensions for the PCB. The boundaries can be adjusted by clicking and dragging the polygon. Each of the components on the screen can also be clicked and dragged onto the PCB area. Note how each connected component from the schematic is now connected by a yellow line known as a “netlist.” All components can either be

manually routed by following each netlist, or an auto-route feature can be used by selecting the **Auto** button on the left toolbox. When routing components, a majority of the routing will take place on the “Top” layer (rustic-colored). However, routing can also take place on the “Bottom” layer (blue-colored). For multi-layer designs, such as a 4-layer board, the top and bottom layers are usually the signal layers, whereas the inner layers are usually dedicated to power and ground. The 2-layer board in this tutorial, however, will utilize the top layer as the component layer and the bottom layer as ground.

To set the bottom layer as ground, select the **Polygon** tool in the left toolbox, and then select the layer to be “Bottom” from the layer pull-down menu at the top of the work space. Create a rectangular polygon around the inner sides of the PCB boundary (keep a small amount of buffer room between the ground plane and board edges). After the polygon is created, right-click on the polygon, select “Properties,” and rename the signal name to “GND.” Once connections are made to the GND netlist, the bottom layer should automatically fill up from the blue polygon GND layer. If not, press the “Ratsnest” button. The same methodology can be used to create polygons dedicated to “power” netlists as well.

PCB Design: Gerber File Generation

To generate the Gerber and drill files, select the “CAM” button at the top of the work space (refer to Figure B.13). When the CAM processor window opens, a job must be selected to generate the Gerber files for a 2-layer board (refer to Figure B.14). Next, select the “gerb274x.cam” job file, which is a default job created specifically for 2-layer boards.

Figure B.14 CAM processor window.

Once the job is selected, the Cam processor window should now look like Figure B.15. Take notice of the tabs that appear at the top of the window. Gerber files should be generated for the component layer, ground layer, component silkscreens, component layer solder-mask stop, and ground layer solder-mask stop. For each layer, the affected board attributes are highlighted in blue. Click “Process Job” when all parameters are configured.

Figure B.15 Gerber file configuration for a 2-layer board job.

After the Gerber Files are saved, select File > Open > Job... and select the “excellon.cam” job to create the drill files for the board. A similar screen like Figure B.15 should appear. Again, once all parameters are set, click “Process Job.” When sending the Gerber and drill files to a board fabrication house, certain files are necessary in the board fabrication process. For this project, the company Advanced Circuits (www.4pcb.com) was used as a board fabrication vendor. Table B.1 lists the files necessary to have Advanced Circuits fabricate a board.

Table B.1 □ Necessary Gerber and drill files for board fabrication.

<i>File</i>	<i>Description</i>
.cmp	Component-side (top) data
.sol	Solder-side (Bottom) data
.plc	Component-side silk screen data
.stc	Component-side solder stop-mask data
.sts	Solder-side stop mask data
.drl	Excellon drill description