DEVELOPMENT OF A HIGH TEMPERATURE SILICON CARBIDE CAPACITIVE PRESSURE SENSOR SYSTEM BASED ON A CLAPP-TYPE OSCILLATOR CIRCUIT

By

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# Table of Contents

List of Figures .......................................................................................................................... 8

List of Tables ............................................................................................................................ 18

Acknowledgements ................................................................................................................... 19

Abstract ................................................................................................................................ 20

Chapter 1: Introduction ............................................................................................................. 22

1.1: Technical Rational .......................................................................................................... 22

1.2: Areas of Potential Impact .............................................................................................. 24

1.2.1: Oil and Gas Extraction Applications ........................................................................... 24

1.2.2: Automotive Applications .......................................................................................... 24

1.2.3: Aerospace Applications ............................................................................................ 25

1.3: Harsh Environment Microsystems ................................................................................... 28

1.3.1 Silicon-based Electronics and Sensors for High Temperature Operations ................. 28

1.3.2: SiC as a Material for High Temperature Microsystems ............................................ 29

1.3.3: MEMS-based Pressure Sensor Technologies ........................................................... 34
1.3.4: High Temperature Pressure Sensor Systems.................................36

1.4: Objectives of this Dissertation..........................................................40

Chapter 2: System Design.........................................................................41

2.1: System Overview.................................................................................41

2.2: Circuit Simulations..............................................................................46

2.3: 6H-SiC MESFET..................................................................................52

2.4: SiCN Capacitive Pressure Sensor.........................................................54

Chapter 3: Substrate and Component Evaluation........................................56

3.1: Benchtop Testing..................................................................................56

3.2: Substrate Evaluation............................................................................57

3.3: High Temperature Inductor Development..........................................62

3.4: Development and Evaluation of High Temperature SiC MIM Capacitors...67

Chapter 4: High Temperature Antenna and Oscillator Circuit Development......84

4.1: Wireless System Characterization......................................................84

4.2: Wireless Oscillator Prototype Design #1: Clapp-type Oscillator Circuits
    with Microfabricated Slot-ring Antennas .............................................86
4.3: Wireless Oscillator Prototype Design #2: Clapp-type Oscillator Circuits with Chip Antennas .................................................................91

4.4: Wireless Oscillator Prototype #3: High Temperature Clapp-type Oscillator Circuit Development ..............................................................101

4.5: Summary.................................................................................................................109

Chapter 5: Development of Wireless Pressure Sensor Systems for High Temperature Operation .................................................................111

5.1: High Temperature Pressure Sensor Testing Apparatus.........................111

5.2: Wireless Pressure Sensor System Prototype Design #1: A Polysilicon-based Capacitive Pressure Sensor System ..............................................112


5.3.1: Temperature and Pressure Testing of a SiCN Capacitive Pressure Sensor.................................................................................................118

5.3.2: High Temperature Testing of Titanate MIM Capacitors ..............120

5.3.3: SiCN-based Pressure Sensor System Prototype Design # 2: Testing........................................................................................................127
5.4: Wireless Pressure Sensor System Prototype Design #3: SiCN-based Capacitive Pressure Sensor System with a Directional Chip Antenna
Antenna…………………………………………………………….130

5.4.1: Pressure Testing of SiCN Capacitive Pressure Sensor……………131

5.4.2: Temperature Testing of Wirewound Inductors…………………..132

5.4.3: Development of a Directional Chip Antenna with Matching Network …………………………………………..137

5.4.4: SiCN-based Capacitive Pressure Sensor System with Directional Chip Antenna Prototype Design # 3: Testing………………..139

5.5: Summary…………………………………………………………………..147

Chapter 6: Development and Qualification Testing of a Packaged Pressure Sensor System…………………………………………………………….149

6.1: Introduction………………………………………………………………………………….149

6.2: High Temperature Testing of Thick Film Resistors…………………150

6.3: Pressure Testing of SiCN Capacitive Pressure Sensor…………………153

6.4: Simulated Response of the Packaged System…………………………155

6.5: Fabrication and Assembly of Packaged Pressure Sensor System………157

6.6: Pressure and Temperature Testing of the Packaged System……………..159
List of Figures

Figure 2.1. Schematic diagram of the wireless sensing circuit with DC bias circuits, antenna and Clapp-type oscillator. The Clapp-type oscillator is denoted by the red box........................................................................................................42

Figure 2.2. Schematic diagram of ADS circuit model used to simulate response of sensing system.................................................................................................................................................47

Figure 2.3. ADS harmonic balance simulation of a 96.40 MHz Group B oscillator design..............................................................................................................................................................49

Figure 2.4. Simulated phase response of a 96.40 MHz Group B oscillator design........50

Figure 2.5. Simulated loop gain of a 96.40 MHz Group B oscillator design.............51

Figure 2.6. Cross-sectional schematic of the Cree MESFET.....................................53

Figure 2.7. Plan-view photograph of the Cree 6H-SiC MESFET.........................54

Figure 2.8. Photograph of the SiCN-based capacitive pressure sensor....................55

Figure 3.1. Photograph of the high temperature probe station (HTPS).....................57

Figure 3.2. Extracted effective permittivity of alumina with a CPW line (s = 40, w = 80 µm) ZO = 50Ω..................................................................................................................................................58
Figure 3.3. Extracted effective permittivity of sapphire with a CPW line (s = 40, w = 80 µm) ZO = 50Ω ................................................................. 59

Figure 3.4. Microphotograph of Ti/Au metallization with wire bonds on an alumina substrate after heating to 400°C .................................................. 60

Figure 3.5. Microphotograph of Ti/Au metallization on a sapphire substrate with wire bonds after heating to 400°C .................................................. 60

Figure 3.6. Photograph of a collection of CoorsTek 996 Alumina Superstrate substrates with different thicknesses .................................................. 61

Figure 3.7. CAD drawing of a 2.5-turn inductor .................................................. 63

Figure 3.8. Equivalent circuit model of the spiral thin film inductors ......................... 65

Figure 3.9. Inductance as a function of frequency and temperature for a 1.5 turn inductor at temperatures between 75 and 475°C ......................... 65

Figure 3.10. Resistance as a function of frequency and temperature for a 1.5 turn inductor ................................................................. 66

Figure 3.11. Extracted inductance, L, as a function of temperature for the 1.5 turn inductor determined by ADS, the Y-parameters, and Sonnet ......................... 66

Figure 3.12. X-Ray photoelectron spectrum of an amorphous SiC film deposited by PECVD at 300°C ................................................................. 68
Figure 3.13. AFM micrograph from a PECVD SiC film deposited at 300°C on a polished Si substrate…………………………………………………………………………………………..69

Figure 3.14. AFM micrograph of a PECVD SiC film deposited at 450°C on a polished Si substrate…………………………………………………………………………………………..70

Figure 3.15. AFM micrograph from a PECVD SiC film deposited at 300°C on an alumina substrate…………………………………………………………………………………………..71

Figure 3.16. AFM micrograph of an as-deposited Ti/Au multilayer deposited on an alumina substrate…………………………………………………………………………………………..72

Figure 3.17. AFM micrograph of Ti/Au multilayer deposited on an alumina substrate and annealed at 650°C for one hour…………………………………………………………………………………………..73

Figure 3.18. AFM micrograph of an as-deposited Ti/Pt/Au/Pt multilayer deposited on an alumina substrate…………………………………………………………………………………………..73

Figure 3.19. AFM micrograph of a Ti/Pt/Au/Pt multilayer deposited on an alumina substrate and annealed at 650°C for one hour…………………………………………………………………………………………..74

Figure 3.20. Micro-photograph of a SiC thin film MIM capacitor……………………………………………………………………………………………………………………………………..75

Figure 3.21. Measured capacitance as a function of temperature for a capacitor consisting of Ti/Pt and Ti/Pt/Au/Pt metallizations and a SiC dielectric film deposited at 300°C…………………………………………………………………………………………..77

Figure 3.22. Measured capacitance as a function of temperature for a capacitor with Ti/Pt/Au/Pt first level metal and a SiC dielectric deposited at 450°C……..77
Figure 23. Capacitance versus voltage from 25 to 350°C for a 144.4 nm² capacitor with a Ti/Au first level metal and a SiC dielectric deposited at 300°C .......... 78

Figure 3.24. Photograph of a capacitor fabricated with a SiC film deposited at 300°C after a measurement performed at 400°C .................................................. 79

Figure 3.25. Measured S-parameters for a 144.4 nm² capacitor with a Ti/Au first level metal and a SiC film deposited at 300°C ........................................ 81

Figure 3.26. Equivalent circuit model of a MIM capacitor .................................. 82

Figure 3.27. Capacitance values from the ADS equivalent circuit model as a function of temperature ................................................................. 82

Figure 4.1. Photograph of the anechoic chamber with absorber lined walls .......... 85

Figure 4.2. Close-up photograph of the high temperature pressure chamber, gain horn and receive antenna in the anechoic room ................................. 86

Figure 4.3. Photograph of a slot-ring antenna integrated with a Clapp-type oscillator ................................. 87

Figure 4.4. Close-up photograph of the Clapp oscillator in Fig. 4.3 ....................... 88

Figure 4.5. Photograph of the off-chip DC bias circuits .................................... 89

Figure 4.6. Measured power versus frequency at 270°C for the Clapp oscillator circuit with slot antenna ................................................................. 90

Figure 4.7. Measured oscillation frequency and received power vs temperature for the Clapp oscillator with slot antenna ........................................ 91
Figure 4.8. Photograph of the 720 MHz oscillator circuit with chip antenna.............93

Figure 4.9. Photograph of the 940 MHz oscillator circuit with chip antenna.............93

Figure 4.10. Measured output power versus frequency from the 720 MHz circuit at 25 and 200°C.................................................................95

Figure 4.11. Measured received power and frequency versus temperature for the 720 MHz circuit.................................................................96

Figure 4.12. Measured output power versus frequency for the 940 MHz circuit at 25 and 250°C.................................................................97

Figure 4.13. Measured received power and frequency versus temperature for the 940 MHz circuit.................................................................97

Figure 4.14. Measured phase noise at 25 and 250°C for the 940 MHz circuit.........98

Figure 4.15. Measured phase noise versus temperature for the 940 MHz circuit at 100 kHz offset frequency..........................................................99

Figure 4.16. Radiation pattern versus temperature for a 720 MHz oscillator with chip antenna.................................................................100

Figure 4.17. Radiation pattern versus temperature for a 940 MHz oscillator with chip antenna.................................................................101

Figure 4.18. Photograph of a 30 MHz oscillator............................................103

Figure 4.19. Photograph of a 90 MHz oscillator............................................103
Figure 4.20. Output power versus frequency at 25 and 450°C for the 30 MHz oscillator

Figure 4.21. Measured peak power and oscillation frequency versus temperature for the 30 MHz oscillator

Figure 4.22. Output power versus frequency at 25 and 470°C for the 90 MHz oscillator

Figure 4.23. Measured peak power and oscillation frequency versus temperature for the 90 MHz oscillator

Figure 4.24. Measured phase noise at 25 and 470°C for the 90 MHz oscillator

Figure 4.25. Measured phase noise versus temperature for the 90 MHz oscillator at the 100 kHz and 10 MHz offset frequencies

Figure 5.1. Photographs of the high temperature pressure chamber

Figure 5.2. Photograph of the wireless pressure sensor with polysilicon capacitive pressure sensor

Figure 5.3. Micrograph of a polysilicon capacitive pressure sensor

Figure 5.4. Output power versus frequency at 300°C for the polysilicon wireless pressure sensor from 0 to 45 psi
Figure 5.5. Frequency vs. pressure for the polysilicon wireless pressure sensor system at 25°C and 300°C………………………………………………………………………117

Figure 5.6. Capacitance versus pressure for the SiCN capacitive pressure sensor……119

Figure 5.7. Measured S-parameters at 25 and 400°C for the 41 pF titanate MIM capacitor………………………………………………………………………………121

Figure 5.8. Equivalent circuit model for the titanate MIM capacitors………………122

Figure 5.9. Measured and simulated S-parameter magnitudes versus frequency for a 14 pF titanate MIM capacitor at 400°C……………………………………………………122

Figure 5.10. Capacitance versus temperature at 1 MHz for a titanate MIM capacitor...125

Figure 5.11. Capacitance vs. frequency for a titanate MIM capacitor………………126

Figure 5.12. Photograph of a SiCN-based wireless pressure sensor system…………127

Figure 5.13. Frequency versus pressure at 25 and 400°C for the SiCN wireless pressure sensor…………………………………………………………………………129

Figure 5.14. Spectral response at 400°C for the SiCN wireless pressure sensor for pressures of 0 and 100 psi……………………………………………………………130

Figure 5.15. Capacitance versus pressure at 25°C for a SiCN pressure sensor……..132

Figure 5.16. Microphotograph of a wirewound inductor on a GSG test fixture………133

Figure 5.17. Equivalent circuit model for a wirewound inductor…………………..133
Figure 5.18. Inductance versus temperature at 1 MHz for a wirewound inductor……135

Figure 5.19. Series parasitic resistance versus temperature at 1 MHz for the wirewound inductor……………………………………………………………………136

Figure 5.20. Inductors response over the frequency range of 10 to 110 MHz……….137

Figure 5.21. Photograph of a chip antenna, matching network and CPW feed line….138

Figure 5.22. Return loss of the chip antenna with and without matching networks…..138

Figure 5.23. Circuit schematic of a Clapp-type oscillator with DC bias circuits……140

Figure. 5.24. Photograph of a SiCN-based wireless pressure sensor with directional chip antenna……………………………………………………………………141

Figure 5.25. Horizontal and vertical radiation patterns for a SiCN wireless pressure sensor with directional chip antenna………………………………………143

Figure 5.26. Wireless pressure sensor Frequency vs. pressure at 25, 100, 200, and 300°C for the SiCN wireless pressure sensor with directional chip antenna……144

Figure 5.27. Output power vs. frequency at 300°C for the wireless SiCN pressure sensor with directional chip antenna at 0, 50 and 100 psi…………………..145

Figure 5.28. Measured output power versus temperature for the wireless SiCN pressure sensor with directional chip antenna at 0, 50, and 100 psi………………146

Figure 5.29. Phase noise versus offset frequency for the SiCN wireless pressure sensor with directional chip antenna……………………………………….147
Figure 6.1. Microphotograph of a thick film resistor on a metal test fixture.............151

Figure 6.2. Resistance versus temperature for a 10 kΩ thick film resistor.............153

Figure 6.3. Pressure vs. capacitance for the SiCN pressure transducer used in the wired packaged system...............................................................154

Figure 6.4. Capacitance versus frequency for the SiCN capacitive pressure sensor from 40 Hz to 110 MHz at 0 psi.........................................................155

Figure 6.5. Circuit schematic of the Clapp-type oscillator used in the packaged SiCN pressure sensor.................................................................156

Figure 6.6. Simulated resonant capacitance versus frequency with respect to changes in pressure.................................................................157

Figure 6.7. Photograph of an assembled SiCN pressure sensor system on alumina substrate.................................................................158

Figure 6.8. Photograph of the packaging for the wired SiCN pressure sensor system...159

Figure 6.9. Photograph of the pressure system characterization fixture....................160

Figure 6.10. Output power versus frequency for the packaged pressure sensor system at 0, 100, 200, 300 and 350 psi.........................................................161

Figure 6.11. Simulated capacitance versus pressure for the packaged SiCN pressure sensor system.................................................................162
Figure 6.12. Simulated frequency response versus pressure using calculated capacitively
pressure sensor values for the packaged SiCN sensor system……………163

Figure 6.13. Photograph of the test fixture and attached packaged sensor positioned
inside tube furnace in preparation for a temperature/pressure test…………164

Figure 6.14. Output power versus frequency at 540°C for a packaged SiCN pressure
sensor system at 0 and 320 psi………………………………………………165

Figure 6.15. Photograph of the vibration test setup……………………………………166

Figure 6.16. Acceleration versus frequency for a 1/4 g sinusoidal sweep profile……167

Figure 6.17. Acceleration versus frequency for a 5.3 Grms random vibration profile...167

Figure 6.18. Output power versus frequency for the packaged SiCN pressure sensor after
vibration testing……………………………………………………………………169

Figure A1. Photographs of the thermoelectric power scavenging measurement
apparatus……………………………………………………………………180
List of Tables

Table 1.1. Relevant properties of common semiconductor materials……………………30

Table 1.2. Mechanical properties of SiC and Si at room temperature…………………33

Table 2.1. Component values summarization for viable sensing systems listed in this
dissertation……………………………………………………………………52

Table 4.1. Component values for the 720 and 940 MHz oscillator designs……………92

Table 4.2. Component values for 30 and 90 MHz oscillator designs………………….104

Table 5.1. Circuit model values for 14 pF MIM capacitor…………………………123

Table 5.2. Circuit model values for 41 pF MIM capacitor…………………………123

Table 5.3. Circuit model values for 390 wirewound inductor………………………134

Table 6.1. Circuit model values for 10 kΩ chip resistor……………………………152

Table 6.2. Tabulated spectrum values of packaged pressure sensor system measurements
taken before and after vibe testing……………………………………….196
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Development of a High Temperature Silicon Carbide Capacitive Pressure Sensor System

Based on a Clapp-Type Oscillator Circuit

Abstract

by

MAXIMILIAN C. SCARDELLETTI

In this dissertation, the development of a packaged silicon carbide (SiC) based MEMS capacitive pressure sensor system that is designed to monitor the pressure of a conventional gas turbofan engine is described. The electronic circuit of the pressure sensor system is based on a Clapp-type oscillator that includes a 6H-SiC MESFET, a SiCN MEMS-based capacitive pressure sensor, titanate MIM capacitors, wirewound inductors, and thick film resistors. The capacitive pressure sensor is incorporated in the LC tank circuit of the oscillator so that a pressure-induced change in capacitance causes a change in the resonant frequency of the oscillator. The MESFET is used to induce oscillation. Individual passive components were evaluated at high temperature to assess their utility in an integrated system. Both wireless and wired variants of the pressure sensor systems for use at high temperature were developed.
In developing the final packaged device, several prototype designs of the Clapp-type oscillator circuit that incorporate wireless capability were explored. Prototype circuits with slot-ring and chip antennas operating between 700 MHz and 1 GHz exhibited a maximum operating temperature of 250°C, limited by the low gain of the MESFET at these frequencies. Prototype circuits operating at 30 and 90 MHz that utilize large spiral inductors in the Clapp oscillator as the radiating element extended stable operation to 470°C.

Several wireless pressure sensor prototypes based on the Clapp oscillator circuit were developed. A prototype incorporating a polysilicon capacitive pressure sensor and a spiral inductor exhibited stable operation up to 300°C. A second prototype that used a SiCN capacitive pressure sensor functioned at temperatures up to 400°C. A third prototype based on the 2nd prototype but incorporating a compact, directional chip antenna had a maximum operating temperature of 300°C, limited by the antenna.

Based on size restrictions, the packaged system utilized a wired configuration. This prototype operates reliably from 0 to 350 psi and from 25 to 540°C, with a sensitivity of 6.8x10^-2 MHz/psi and negligible difference in frequency response. The packaged sensor passed standard benchtop temperature, pressure and vibration acceptance tests required prior to any future test on a flight worthy engine.
Chapter 1

Introduction

1.1. Technical Rationale

The demonstrated utility and economic viability of microsystems technology in applications where silicon-based electronics are well suited to the environmental conditions, such as consumer electronics, healthcare, and telecommunications, has stimulated demand for comparable systems to be used in environmentally demanding applications. Notable harsh environment application areas that would significantly benefit from an infusion of microsystems technologies include: (1) oil and gas exploration/extraction, (2) automotive engine control, and (3) aerospace technologies. Implementation of microsystems in these areas are envisioned to improve efficiency and
extend operational lifetime of key components by enabling closed-loop control through the integration with control electronics. Currently these systems lack the type of on-board control that is possible using microsystems technology due to the extreme operating conditions of system. In situations where sensor-based technologies have been implemented, the sensing part of the system is often physically offset from the position of interest due to the inherent temperature limitations of the electronics, peripheral passive components (capacitors, inductors) and often the sensing elements themselves. Advancements in packaging technologies have not been sufficient to overcome the temperature limitations while maintaining miniaturization, which are ultimately constrained by the temperature stability of the silicon-based electronics. Approaches to locate the temperature-sensitive electronic components in cooler sections of the system have been implemented, but these approaches result in a much larger sensor system, with significantly more wiring and larger packaging. Degradation of the transduced signal due to the displacement of the signal conditioning electronics from the sensor often results from such approaches. However, significant advances in wide-bandgap (WBG) semiconductor materials such as gallium nitride (GaN), silicon carbide (SiC) and diamond (C) have enabled the development of electronic devices with maximum operating temperatures that are compatible with the high temperatures associated with the aforementioned harsh environment application areas. SiC electronic device technologies have advanced to the point that it is now feasible to consider the integration of electronics and microsensors into single, miniature, packaged systems that can achieve accurate real-time data analysis. This dissertation aims to develop such a system for high temperature pressure monitoring suitable for gas turbine engine applications.
1.2. Areas of Potential Impact

1.2.1. Oil and Gas Extraction Applications

One of the oldest and currently the largest users of high-temperature electronics are found in the downhole oil and gas industry. In downhole oil and gas drilling, the operation temperature is a function of underground depth of the well. Worldwide, the typical geothermal gradient is approximately 25°C per km of depth (1°F/70 ft). In the recent past, drilling operations have seen temperatures as high as 150°C to 175°C, but declining reserves of easily accessible oil and natural gas, coupled with the advances in oil extraction technology have motivated the industry to seek deeper oil reserves. These reserves are more difficult to discover and extract, necessitating the utilization of advanced sensor technologies. Drilling deeper results in significantly higher operational temperatures which can reach 300°C, a temperature that exceeds the maximum operating limit for most integrated, Si-based MEMS. To address this issue, an active phase change cooling technique has been developed, but this approach can only be used in situations where the temperatures does not exceed 250°C (E. Pennewitz).

1.2.2. Automotive Applications

The automotive industry is another area that would benefit from microsystems that are suitable for high temperature operation. For much of its history, control systems in automobiles were based on mechanical, hydraulic and vacuum based devices. However, the industry is rapidly migrating towards electromechanical or mechatronic systems. As such, the ability to locate sensors, signal conditioning, and control electronics close to the
heat sources will be an essential requirement. Electronic sensors designed to monitor and maintain on-engine, in-transmission, and on-wheel temperatures must operate at temperatures between 150 and 200°C. At these temperatures Si-based devices can be used. However, in-cylinder pressure sensors must operate reliably at temperatures up to 300°C and exhaust sensing requires sensors that operate as high as 850°C (R. Johnson), temperatures that devices made from materials other than Si.

1.2.3. Aerospace Applications

The aerospace industry is replete with opportunities for harsh environment microsystems. Commercial and military aircraft offer many opportunities for electronic sensing systems that can monitor and access health monitoring conditions to ensure aircraft survivability and function reliably above 300°C. NASA’s VIPR (Vehicle Integrated Propulsion Research) Program has been assigned this task. In this program, electronic systems that can detect variable changes in emissions, temperature, blade tip clearance and pressure are being developed for aircraft engines. These systems must be able to sustain the harsh conditions such as vibration, pressure and temperature (G. Hunter 2013).

Rapid or sudden changes in the emissions produced by combustion can be an indicator that an aircraft engine has sustained significant changes in the propulsion system, the combustion process or the engine health state. These changes are very challenging to monitor on an aircraft engine due to the harsh environment associated with the propulsion system. At NASA Glenn Research Center (GRC), a gas microsensor array was developed for monitoring the emissions produced by aircraft engines (G. Hunter 2013). The sensor
was capable of detecting carbon monoxide (CO), carbon dioxide (CO₂), oxygen (O₂), nitrogen oxide (NOₓ) and other unburned hydrocarbons. Each sensor element in the array was designed to be selective to a targeted chemical species while attempting to minimize cross sensitivity to gases in the environment.

Temperature sensors that can more reliably and that can more accurately determine the temperature in an aircraft engine are also needed. Two temperature sensors, an optical and a thin-film temperature sensor, are being developed at NASA GRC. The optical temperature sensor is based on a Fiber Bragg grating (FBG) and can determine a change in temperature from 25 to 900°C, over a wavelength from 1300 to 1309 nm. The sensor capitalizes on a variation in the refractive index in the fiber core. The fiber optic sensors are immune to electromagnetic interference, are chemically stable and have small weight and size. The sensors use temperature sensitive variations in the wavelength of propagated light as a means to sense temperature and the resultant signals are immune to incidental light variations. This feature makes the temperature measurements acquired by the FBGs much more stable and reliable than those acquired by conventional measurement techniques. A second temperature sensor which consists of a gold-platinum thin-film was demonstrated (Meredith 2014) and was operational up to 960°C. The thin-film sensor is much less invasive to the operational environment, has a minimal impact on the physical characteristics of the supporting components, has much less mass than wires and foils, and will react considerably faster to suitable transient effects.
A microwave blade tip clearance sensor is being developed to actively control and minimize the gap between the rotating blades and the stationary case of gas turbine engines. Maintaining proper blade-case clearance in a gas turbine engine is a critical approach to increasing engine efficiency, reducing fuel consumption, reducing engine emissions and increasing engine service life (M. R. Woike). The microwave blade tip sensor is approximately 14 mm in diameter and 26 mm in length. It contains a transmitting and receiving antenna and can be installed in the case of the engine where it can directly measure the radial clearance between the case and the turbine blade tips. The sensor is functional up to 900°C, operates at 5.8 GHz and can measure a clearance up to 25 mm with an accuracy of 25 µm.

The need for high temperature microsystems is not limited to aeronautic applications. The Venus Exploration Analysis Group (VEXAG) is assessing the feasibility of sending planetary probes, including landers, to Venus to determine the atmospheric formation, evolution, and climate history of the planet (R. Herrick). Such missions will expose the probes to atmospheric conditions with temperatures as high as 480°C. Size and weight restrictions prohibit the use of active cooling, distributed components and conventional packaging approaches to shield Si-based electronics from these prohibitively high temperatures. At this point, the leading approach is to utilize electronics, sensors and peripheral components that are fabricated from temperature tolerant materials, with the leading material being SiC.
1.3. Harsh Environment Microsystems

1.3.1. Silicon-Based Electronics and Sensors for High Temperature Operation

Silicon electronic devices fabricated from bulk silicon generally have a maximum operating temperature of ~200°C, thereby limiting their use as components in harsh environment MEMS. Nevertheless, previous work has been performed to develop Si-based electronics as components in sensors designed for elevated temperature operation. For example, a CMOS analog-to-digital converter (ADC) sensor interface was developed using bulk silicon and can be used for automotive subsystems (T. Watanabe). This ADC system was an all-digital sensor using a time-domain processor TAD (Time A/D Converter). The TAD was developed using a 0.35 µm digital CMOS and operated over a temperature range from -40 to 125°C. The maximum operating temperature of silicon electronics can, however, be extended to ~300°C if the devices are fabricated on silicon-on-insulator (SOI) substrates. A SOI based gate driver has successfully demonstrated operation up to 200°C without any active or passive cooling mechanisms (M. Huque). This prototype has successfully generated 10 to 30V peak-to-peak output voltages with sourcing and sinking currents greater than 4A at 200°C. A SOI gate driver control circuit has been developed to actuate power switches and is operational up to 200°C without a heat sink (R. Greenwell). An ASIC for high temperature signal conditioning for aircraft engine control systems has also been developed (S.V. Solomko). The ACIS was realized on 1 µm SOI CMOS technology and has been shown to function up to 250°C.
Although ultimately limited by temperature, silicon-based sensor systems have also been developed for moderately high temperature applications. In (R. F. Guan), a Si MEMS piezoresistive based pressure sensor based on a Wheatstone bridge configuration, is described. The pressure sensor system operates over a temperature range from -40 to 125°C and from 0 to 0.6 MPa. As the pressure is applied to the diaphragm, the resistive properties of the material change and a resulting change in the output voltage or current can be detected. The sensing system can be used in either constant voltage or constant current mode. Since the pressure system is Si-based and small in size, it can be easily fabricated with typical integrated circuit (IC) fabrication techniques.

A silicon-based temperature sensor for jet engines was developed from commercial-off-the-shelf (COTS) components and is capable of measuring temperatures from -195 to 200°C (R. L. Patterson). The sensing system consists of a temperature-to-frequency relaxation oscillator circuit utilizing a resistance temperature detector (RTD) as the temperature sensing element. The output fluctuates with variation of temperature, with a frequency of about 9 kHz at -195°C and 2.3 kHz at +200°C. The sensor was developed for “hot zones” on the jet engine.

1.3.2. SiC as a Material for High Temperature Microsystems

The key to realizing electronics that are suitable for the high temperatures associated with gas turbine engines is to develop the active components from wide bandgap (WBG) semiconductor materials. The most technically relevant WBG semiconductor materials are gallium nitride (GaN), silicon carbide (SiC) and diamond (C). Some of the
more important properties of these materials for electronic device applications are listed in Table 1.1

<table>
<thead>
<tr>
<th>Material</th>
<th>$E_g$ (eV)</th>
<th>$E_B$ (MV/cm)</th>
<th>$V_{sat}$ ($10^7$ cm/s)</th>
<th>$\sigma_T$ (W/cmK)</th>
<th>$\varepsilon$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Silicon (Si)</td>
<td>1.1</td>
<td>0.6</td>
<td>1</td>
<td>1.5</td>
<td>11.8</td>
</tr>
<tr>
<td>Germanium (Ge)</td>
<td>0.7</td>
<td>0.1</td>
<td></td>
<td>0.58</td>
<td>16.2</td>
</tr>
<tr>
<td>Gallium Arsenide (GaAs)</td>
<td>1.4</td>
<td>0.6</td>
<td>2</td>
<td>0.5</td>
<td>12.8</td>
</tr>
<tr>
<td>Gallium Nitride (GaN)</td>
<td>3.4</td>
<td>3.3</td>
<td>2.5</td>
<td>1.5</td>
<td>8.9</td>
</tr>
<tr>
<td>Silicon Carbide (SiC)</td>
<td>3.3</td>
<td>3.0</td>
<td>2.2</td>
<td>4.9</td>
<td>9.9</td>
</tr>
<tr>
<td>Diamond (C)</td>
<td>5.5</td>
<td>20</td>
<td></td>
<td>22</td>
<td>5.7</td>
</tr>
</tbody>
</table>

SiC has long and notable history in semiconductor technologies going back to 1824 when Jons Jakob Berzelius burnt an unknown compound and observed an equal number of Si and C atoms (J. J. Berzelius). Since then much work has been devoted to the development of SiC to the point where now it is considered the leading WBG semiconductor for electronic applications (M. Shur). Its leadership among WBG semiconductors is due mainly to:

1. 4H-SiC and 6H-SiC are commercially available in large area wafers (> 3 in diam.).
2. SiC can be homoepitaxially grown as thin and thick films, thus avoiding lattice mismatch.
3. Both n- and p-type conductivity can be achieved by in-situ doping during crystal growth or by ion implantation.

4. Silicon dioxide (SiO$_2$) films can thermally be grown on the SiC surface, thus providing a convenient means of surface passivation and dielectric isolation.

Due to a reliable commercial supply of large-area, single crystalline wafers, SiC power devices are now commercially available from a variety of manufacturers including Cree, GeneSiC, Infineon, Panasonic, ROHM, STMicroelectronics, Semelab/TT Electronics, and Central Semiconductor.

As stated previously, Si-based devices, such as metal-oxide-semiconductor field-effect-transistors (MOSFETs) and metal semiconductor field-effect-transistors (MESFETs), have maximum operating junction temperatures of 200°C, which is mainly due to the sizable increase in leakage current at high temperatures, which renders junction-based Si devices unstable. For harsh environment sensor applications, this restriction means that for reliable operation, Si-based devices cannot be placed near a heat source, such as an aircraft engine, that is that is greater than 200°C without a cooling method, which greatly increases cost and complexity. In contrast, SiC devices operate at temperatures up to 500°C (P. Neudeck) due in large part to much reduced leakage currents as a result of the considerably higher band gap (3.2 eV for 6H-SiC versus 1.12 eV for Si).
Junction temperature limitations are not the only reason for using SiC for high temperature applications over Si. SiC junction-based diode have a higher breakdown voltage (i.e., 600V) because of the high electric breakdown field strength of SiC; while similar Si diodes are typically rated at voltages lower than 300V (T. Funaki). SiC has a thermal conductivity of 4.9 W/cm-K as opposed to Si, which has a thermal conductivity of 1.5 W/cm-K. Also, the forward and reverse characteristics of WBG semiconductor-based devices vary only slightly with temperature and time; therefore, they are more reliable at high temperatures.

Although SiC is predominantly used for high voltage, high temperature and high frequency electronics applications, SiC has also been successfully used in a variety of solid state sensors, most notably temperature sensors and gas sensors. Riza, et al., describe a SiC-based temperature sensor designed to monitor the temperature of aircraft engines (N. A. Riza). The temperature sensor is a SiC based blackbody radiator in a two color pyrometry configuration with an operating temperature up to 750°C. Loloee, et al., describe a solid-state MOS hydrogen gas sensor that consists of platinum, SiO₂ and doped SiC (R. Loloee). When in detection mode, the capacitance is held constant using a feedback circuit and the gate bias is monitored as the sensor output signal. The time response of this device to hydrogen containing species at 600°C is in the millisecond range, well within standard operating times for conventional hydrogen sensors.

The mechanical properties of SiC are also highly favorable for harsh environment MEMS applications. Table 1.2 compares the key mechanical properties for MEMS
applications of SiC and Si (T. Kimoto). The hardness and Young’s modulus of SiC is much higher than that of Si while the Poisson’s ratio is very close to other semiconductors. SiC retains its hardness and elastically, even at high temperatures. The fracture strength of the SiC is as high as 21 GPa at room temperature and is estimated to be 0.3 GPa at 1000°C, while the fracture strength of Si falls to 0.05 GPa at 500°C (T. Suzuki).

Table 1.2. Mechanical properties of SiC and Si at room temperature.

<table>
<thead>
<tr>
<th>Properties</th>
<th>SiC</th>
<th>Si</th>
</tr>
</thead>
<tbody>
<tr>
<td>Density (g cm⁻³)</td>
<td>3.21</td>
<td>2.33</td>
</tr>
<tr>
<td>Young’s modulus (GPa)</td>
<td>380-700</td>
<td>160</td>
</tr>
<tr>
<td>Fracture strength (GPa)</td>
<td>21</td>
<td>7</td>
</tr>
<tr>
<td>Poisson’s ratio</td>
<td>0.21</td>
<td>0.22</td>
</tr>
</tbody>
</table>

As a result of its unique and highly favorable ensemble of material properties, SiC is being explored as a microsensor material for an increasing number of harsh environment applications. SiC is widely used as a structural material in MEMS devices due to its robust mechanical properties (M. C. Scardelletti, July 2007) (M. C. Scardelletti, December 2008). SiC thin films almost always exhibit an inherent residual tensile stress, which makes them well suited as structural elements in micromachined beams, cantilevers and diaphragms (M. A. LaBarbera), (R. L. Johnson), (Zhuangde Jaing). Furthermore, its chemical inertness enables it to be resistant to oxidation and other environmental gases (M. C. Scardelletti, December 2006).
1.3.3. MEMS-based Pressure Sensor Technologies

Pressure sensing is one of the most common uses of microsystems technology, in part because pressure sensing can be implemented in MEMS using a wide range of transduction methods. Some sensing modalities, such as surface acoustic wave (SAW) or surface transverse wave (STW), utilize the phase velocity variation of surface waves on piezoelectric substrates when pressure is applied (A-long Kang) (Zhao Yi-yu) (S. C. Moulzolf) (E. Hallynck). Optical pressure sensors capitalize on the pressure modulation of an optical signal, such as intensity, polarization, phase, or spectrum (D. Hazarika) (G. Keulemans) (Yuerui Lu).

The most common type of MEMS-based pressure sensors are diaphragm based devices, in large measure because pressure-sensitive diaphragms are relatively easy to fabricate and both piezoresistive and capacitive transduction mechanisms are straightforward to implement in MEMS structures. Piezoresistive transduction utilizes resistors that exhibit a detectable change in resistance upon application of strain. Si and SiC are piezoresistive materials that, when fabricated into microresistors on a flexible diaphragm, manifest detectable changes in voltage when the diaphragm flexes under an applied pressure. Typically, the piezoresistors are configured in Wheatstone bridges or other resistive networks that operate at DC, which greatly simplifies the system. Zhuangde, et al., describe a silicon-based MEMS piezoresistive pressure sensor was able to detect changes in pressure from 0 to 25 MPa for temperatures up to 200°C (Zhuangde Jiang). Johnson, et al., describe a piezoresistive pressure sensor fabricated on an SOI substrate that was able to operate up to 250°C at pressures from 3,000 to 22,000 psi (R. L. Johnson). The
advantage of the SOI architecture is that it allows the device to operate at temperatures higher than similar devices made in bulk Si due to significantly lower leakage between the piezoresistors and the supporting diaphragm. In (Chien-Hung Wu), a piezoresistive pressure with a diaphragm made of 3C-SiC was constructed. The 3C-SiC crystalline, piezoresistive pressure sensor was configured in a Wheatstone bridge topology and operated from 0 to 400°C over a pressure range of 0 to 80 psi. The pressure sensor resistance increased as a function of temperature causing the voltage output to decrease.

Capacitive pressure sensors are essentially parallel plate capacitors with one of the plates fabricated from a mechanically-flexible diaphragm that deflects upon application of pressure towards the second, fixed and inflexible electrode. The resulting change in separation between the two electrodes results in a measurable change in capacitance. Numerous examples of MEMS-based capacitive pressure sensors can be found in the literature (Shih-Shian Ho), (Kin Fong Lei), (Y. Zhang), including sensors fabricated from silicon. The capacitive pressure sensor design has also been implemented in SiC. A capacitive pressure that employs a single-crystal 3C-SiC diaphragm was operated up to 400°C and was tested over a pressure range from 250 to 2500 Torr (J. Du). The pressure sensor demonstrates a linear response from roughly 1100 to 1760 Torr. The sensor can be used for a wide range of applications including the automobile and the aerospace industries. In (Li Chen), a SiC capacitive pressure sensor was demonstrated up to 574°C and pressures up to 700 psi. Sporian Microsystems has developed a SiC capacitive pressure sensor that is operational up 1000°C and pressures as high as 500 psi. The sensor is fabricated from SiCN which is a high temperature material synthesized by thermal decomposition of
polymeric precursors that possesses excellent mechanical properties, tunable electric properties and superior oxidation/corrosion resistance at temperatures up to 1600°C (L. An, Y. Wang, X. Ren). The sensor systems developed in this dissertation are based on this SiCN capacitive pressure sensor, therefore details regarding this sensor will be presented later.

1.3.4. High Temperature Pressure Sensing Systems

The majority of the work in developing pressure sensors for use in harsh environment applications has focused on the design, fabrication and initial testing of discrete components, such as the pressure transducer and associated electronics. However, there have been a few efforts at component-level integration, both for Si and SiC based sensors. Integration of sensors with Si-based electronics offers the distinct advantage of leveraging a mature and sophisticated integrated circuit technology, however the electronics ultimately limit the maximum operating temperature. Kasten, et al., report on the successful development of a Si capacitive pressure sensor integrated with a monolithic CMOS readout circuit that is capable of measuring changes in pressure at temperatures up to 250°C (K. Kasten). The capacitive pressure sensor has a top circular membrane constructed of polysilicon where the diameter of the membrane defines the range of capacitance the sensor is capable of detecting. The bottom electrode consists of a thin silicon layer that is electrically isolated from the substrate by an oxide layer. Likewise, de Jong et al., describe a Si piezoresistive pressure sensing system with CMOS ASICs was demonstrated up to 275°C (P. C. de Jong). The sensing system has a data accuracy of 15 to 16 bits.
Suster, et al., have successfully implemented wireless data transmission to a Si-based capacitive pressure sensing system (M. Suster). The MEMS pressure sensor had a top Si membrane that is isolated from the Si substrate by a layer of oxide. This wireless pressure sensing system consisted of a low-power silicon-tunnel-based-diode LC-tuned oscillator. The LC tank circuit was comprised of a MEMS capacitive pressure sensor and a spiral inductor, which also acted like a transmitting antenna. A change in pressure induced a change in resonant frequency of the LC-based circuit. The maximum operating temperature of this system was 290°C. However, due to the Si based sensor and diode, the system is limited to operation at temperatures below 300°C.

Development of integrated pressure sensing systems is not limited to silicon-based devices. Okojie, et al., developed a MEMS based 4H-SiC piezoresistive pressure sensor that was operational from room temperature to 800°C (R. Okojie). The piezoresistive sensor was implemented in a Wheatstone bridge configuration. Five packaged sensors were tested in this study. The sensing system demonstrated nearly the same sensitivity at 800°C as was measured at room temperature, indicating that the sensors can be inserted in the hotter sections of a test article without the need for cooling.

Wang, et al., developed a SiC capacitive pressure sensor system that utilized a SiC MESFET-based oscillator and was operational up to 400°C (R. Wang). Like (Suster), the Si-based capacitive pressure sensor was located in an LC tank circuit along with a spiral inductor, which acted as a transmitting antenna. The pressure sensing system operated at 31.5 MHz. The output of the pressure sensor system lacks linearity and will require either
a calibration or look-up table to ensure correct readings for the entire temperature range. This is most likely due to the temperature dependence of the components such as the chip capacitors used which exhibit over a 30% loss in capacitance from 25 to 400°C. The sensing system utilizes a Colpitts oscillator design which, while adequate for initial development, is not the most efficient for high temperature applications that require long-term reliable operation due to the number of temperature-sensitive components in the circuit. Yang, et al., developed a SiC piezoresistive pressure sensor system that operated up to 450°C for a pressure range of 0 to 1000 psi (J. Yang). The SiC piezoresistive sensor was integrated with a custom-designed n-channel SiC JFET. The system included signaling condition circuitry for sensor excitation, signal amplification, DC-AC conversion, and RF transmission. Unfortunately, piezoresistive pressure sensors characteristically exhibit a strong temperature dependence and suffer from contact resistance variations at elevated temperatures over extended periods. This temperature dependence ultimately degrades sensor performance because the contact resistance variation is indistinguishable from the piezoresistance change caused by the change in pressure being sensed.

Lastly, a high temperature, active capacitive sensor was developed at NASA GRC (R. D. Meredith). The capacitive pressure sensor was integrated with SiC JFETs in a ring oscillator configuration. The capacitive pressure sensor was not custom built, but rather was taken from a disassembled MKS Instruments Baratron vacuum gauge and connected to the JFET circuitry. The sensor itself was based on a 4 cm Inconel diaphragm that comes into contact to with a thin ceramic disc with a metal electrode on the bottom. While not technically a MEMS-based sensor, this system demonstrated the efficacy of the SiC JFET
approach for high temperature sensing by operating at 500°C. Unfortunately, the operating range of the pressure sensor was 0 to 3 psi, which is not relevant for the gas turbine engine applications being pursued in this dissertation. At 500°C and 0 psi, the resonant frequency of the active pressure sensor was 21.5 kHz and at 3 psi the resonant frequency shifted to 20 kHz, resulting in a change in frequency of 1.5 kHz. The change in frequency from 0 to 3 psi was not substantial, but the active sensor did operate at 500°C, which was a major breakthrough.

Although not directly translatable to the gas turbine engine applications of interest in this dissertation, the aforementioned examples serve as proof on concept that an integrated SiC pressure sensor system for operation at high temperatures and pressures can, in principal, be developed. However, transition of this technology from the benchtop to the field requires substantial development of key components that were not previously investigated. These include the design of a robust circuit architecture suitable for SiC-based electronics and sensors, development of passive components such as capacitors, inductors and antennas that can be microfabricated on a common substrate and/or integrated onto said substrate with minimal form factor and usable under high temperature conditions, and a package that is compatible with high temperature operation and wireless transmission. Achievements in these areas will enable the fabrication of an integrated pressure sensor system that is suitable for high temperature operation and enclosed in a single package. To the best of our knowledge no such system has currently been demonstrated.
1.4. Objectives of this Dissertation

The ultimate goal of this dissertation is to design, fabricate and characterize a SiC-based MEMs capacitive pressure sensor system that is designed to monitor the pressure of a conventional gas turbine engine and can pass qualification tests required for such sensors. The pressure sensor system will consist of the following components that will be designed specifically for high temperature, high pressure operation:

1. A novel SiCN MEMS-based SiC capacitive pressure sensor
2. A 6H-SiC MESFET
3. High temperature capacitors
4. Spiral inductors and loop, slot and chip antennas
5. Thick film resistors

The electronic circuitry of the sensor system will be based on a Clapp-type oscillator where the capacitive pressure sensor is located in the LC tank circuit that is driven into oscillation by the MESFET. Transduction will be achieved by a pressure-induced change in resonant frequency resulting from a change in capacitance from the sensor. The sensor system will be encased in a custom-built package to enable evaluation of sensor performance using a standard benchtop acceptance test required of such sensors prior to actual engine testing. Conditions for this acceptance test include a maximum system operating temperature of 550°C at tip of borescope plug adaptor, a pressure range of 0 to 300 PSIG, and vibrations of 5.3 Grms along the X-, Y- and Z-axes. The unpackaged sensor will be no larger than 8 x 40 x 4 mm³. To the best of our knowledge, this dissertation is the first attempt to create a pressure sensing system that meets these requirements.
Chapter 2

System Design

2.1. System Overview

The wireless sensor system developed in this dissertation consists of a Clapp-type oscillator that incorporates a MEMS capacitive pressure sensor located in the LC tank circuit of the device. A detailed schematic of the oscillator and the corresponding circuit are shown in Figure 2.1. This particular Clapp-type oscillator design requires one inductor (\(L_T\)), three capacitors (\(C_{\text{SENSE}}\), \(C_1\) and \(C_2\)) and one metal semiconductor field-effect transistor (MESFET). The Clapp-type oscillator design was selected because it requires significantly fewer passive components than the Colpitts oscillator design used previously by others (G. Gonzalez). Under typical operating conditions characteristic of a gas turbine engine (i.e., high temperature and high vibration), fewer environmentally-sensitive components within any particular system reduces the probability of failure. Moreover,
because it requires fewer components, the Clapp-type design can be incorporated into systems with smaller form factors, enabling deployment in confined locations. Another advantage of the Clapp oscillator design over a Colpitts architecture is that the in the Clapp oscillator, the operational frequency is determined only a series LC circuit consisting of the inductor \((L_T)\) and capacitive pressure sensor \((C_{SENSE})\). In contrast, the operational design frequency of a Colpitts oscillator is determined by a parallel LC circuit which functions as a bandpass filter and is dependent on the junction capacitance and the resistive loading of the transistor.

Figure 2.1. Schematic diagram of the wireless sensing circuit with DC bias circuits, antenna and Clapp-type oscillator. The Clapp-type oscillator is denoted by the red box.
The operational frequency (frequency of oscillation) for the Clapp-type oscillator can be found using Eq. 5.1:

\[ \omega_O = \frac{1}{\sqrt{L_i C_{EQ}}} \]  

(5.1)

where \( C_{EQ} \) can be found using:

\[ \frac{1}{C_{EQ}} = \frac{1}{C_{SENSE}} + \frac{1}{C_1 + C_{DS}} + \frac{1}{C_2 + C_{GS}} \]  

(5.2)

where \( C_{DS} \) and \( C_{GS} \) are the transistor capacitances. However \( C_{DS} \) and \( C_{GS} \) are usually negligible so Eq. 5.2 reduces to Eq. 5.3:

\[ \frac{1}{C_{EQ}} = \frac{1}{C_{SENSE}} + \frac{1}{C_1} + \frac{1}{C_2} \]  

(5.3)

Since \( C_{SENSE} \) is usually much smaller than \( C_1 \) and \( C_2 \), Eq. 5.3 reduces to:

\[ \frac{1}{C_{EQ}} = \frac{1}{C_{SENSE}} \]  

(5.4)

and Eq. 5.1 becomes:
\[ \omega_O = \frac{1}{\sqrt{L_T C_{\text{SENSE}}}} \]  

(5.5)

Therefore \( C_{\text{SENSE}} \) and \( L_T \) are used to set the operational frequency. For a tunable inductor, or as in this case a capacitive pressure sensor that varies as a function of pressure, the impedance must remain inductive over the entire \( \Delta C_{\text{SENSE}} \) range and can be verified using:

\[ Z_T = j \left( \omega L_T - \frac{1}{\omega C_{\text{SENSE}}} \right) \]  

(5.6)

The capacitors, \( C_1 \) and \( C_2 \), can be used to control the transconductance (\( g_m \)) condition and can be found using:

\[ \frac{g_m}{\omega^2 \cdot R_S \cdot C_1 \cdot C_2} > 1 \]  

(5.7)

where \( R_S \) is the series resistance within the inductor, \( L_T \).

In this design, the gain condition and operational frequency can be set independently which significantly improves the frequency stability of the circuit. For the Colpitts oscillator, however, the gain and operational frequency are interdependent. Independent control of gain and operational frequency becomes significant at high temperatures, where temperature related instabilities in individual devices of the circuit generally increase. The frequency stability of the Clapp oscillator due to the change in
capacitance is found using:

\[ \Delta \omega_O = -\frac{1}{2} \omega_O \frac{\Delta C_{EQ}}{C_{EQ}} \]  

(5.8)

where \( \Delta C_{EQ} \) is the change in equivalent capacitance. Changes in the equivalent capacitance usually are associated with temperature-induced changes within the active device. However, if \( C_1 \) and \( C_2 \) are relatively large capacitances, \( C_{EQ} \) is approximately \( C_{SENSE} \), making \( \Delta \omega \) independent of the active device. Thus when operating at high temperatures (i.e., 400°C), the pressure sensing system is virtually independent of temperature.

In addition to the Clapp-type oscillator, the oscillator circuit included additional components, most notably the DC bias circuits added to input of the drain and gate of the MESFET as shown in Figure 2.1. The drain bias tee consists of and inductor which acts as an RF choke and prevents RF energy from leaking into the drain source and a capacitor to ground and acts as a DC block.
2.2. Circuit Simulations

The sensing circuit, which consists of the Clapp-type oscillator, DC bias circuits and an antenna, was designed using Keysight™ ADS circuit simulator. A schematic of the circuit model is shown in Figure 2.2. A separate model of the Cree™ SiC MESFET was used in all simulations. This model was developed by Cree and Agilent (now Keysight) and represents the transistor characteristics at room temperature (25°C). The MESFET model can only be used in ADS Version 2009, therefore ADS 2009 was used for all simulations in this dissertation even though more a recent version of the ADS simulator is readily available.

The Clapp-based oscillator circuits explored in this dissertation can be categorized into two distinct groups based on design differences. The first group, hereafter known simply as Group A, utilize designs consisting of the Clapp-type oscillator, DC bias circuits and antenna as shown in Fig. 2.1. The radiating element in the Group A designs consist of slot antennas and chip antennas. Circuit simulations for the Model A designs were performed to determine component values that result in viable oscillator circuits with operating frequencies of 131, 720, 940 and 1000 MHz. In contrast, the second group, known as Group B, utilize designs that are the same as for Group A except that the circuit does not contain an antenna and instead uses a spiral inductor $L_T$ as the radiating element. A schematic of Group B circuit is shown in Fig. 2.2. Circuit simulations for the Group B designs were performed to determine component values that result in viable oscillator circuits with operating frequencies of 30 and 100 MHz.
Figure 2.2. Schematic diagram of the ADS circuit model used to simulate response of sensing system.

In order to evaluate the behavior of a particular circuit design for a given set of device parameters, the aforementioned simulation tools were used to evaluate three key characteristics of the circuit, namely, (1) the oscillatory behavior using a harmonic balance simulation, (2) a simulated phase response near the desired design frequency, and (3) a simulated loop gain analysis at the desired operating frequency. The harmonic balance
simulation is a frequency-domain analysis technique for simulating non-linear circuits and systems. The harmonic balance simulation determines the fundamental frequency of oscillation and the corresponding higher order harmonics which are simply integer multiples of the first harmonic. The absence of peaks would indicate that the circuit fails to operate as an oscillator. The phase response and loop gain provide information about the frequency stability of an oscillator. An oscillator will sustain steady-state oscillation if: (1) the phase of the device at the resonant frequency is a multiple of 360°, and (2) the loop gain at the resonant frequency is greater than or equal to unity (N. Kozlovski). For a particular circuit design, component values were varied until a simulated circuit that met the aforementioned three requirements was obtained. This analysis was performed for all simulations performed in this dissertation, but for brevity, only those performed for the 100 MHz circuit from Group B are presented here.

Figures 2.3 to 2.5 show the results of these simulations for a circuit from Group B designed to operate at 96.4 MHz. For this circuit, the values for $C_T$, $L_T$, $C_1$, $C_2$, $R_G$, $L_D$ and $C_D$ were 3.84 pF, 780 nH, 14 pF, 41 pF, 10 kΩ, 390 nH and 188 pF, respectively. Figure 2.3 presents the harmonic balance simulation for this circuit. This figure plots the output voltage as a function of frequency. Five simulated peaks (represented as vectors in the plot) are observed, with the lowest frequency peak at 96.40 MHz followed by peaks at 192.8, 289.2, 385.6 and 482 MHz. The frequencies of peaks 2 through 5 are integer multiples of the first peak, indicating that the circuit indeed behaves as an oscillator with an operating frequency of 96.40 MHz.
Figure 2.3. ADS harmonic balance simulation of a 96.40 MHz Group B oscillator design.

Figure 2.4 is the simulated frequency response for this circuit, as represented by plotting the phase angle versus frequency for the $S_{11}$ coefficient of the circuit over a frequency range of 80 to 140 MHz. S-parameters, also known as scattering parameters, are simple ratios of input and output powers. In a two-port network $S_{11}$ and $S_{22}$ are known as the reflection coefficients while $S_{21}$ and $S_{12}$ are the transmission coefficients. In Figure 2.4, the phase angle of the $S_{11}$ coefficient is $180^\circ$ at 96.40 MHz, which indicates that the circuit satisfies one of the two requirements for steady-state operation.
Figure 2.4. Simulated phase response for a 96.40 MHz Group B oscillator design.

Figure 2.5 is the simulated loop gain for the 100 MHz, Group B circuit. Figure 2.5 presents a polar plot of $S_{11}$ for the circuit over a frequency range of 80 to 140 MHz. The graph shows that at 96.40 MHz, the $S_{11}$ coefficient has a value of 1.086 at $-180^\circ$. Consequently, the circuit satisfies the second requirement for stable, steady state oscillation. The combined results from Figs. 2.3, 2.4, and 2.5 indicate that the 96.40 MHz, Group B oscillator circuit that incorporates the device parameters described previously is a viable oscillator and worthy of further investigation as a fabricated circuit.
As stated previously, the analysis described above was applied to all simulated circuits for all designs investigated in this dissertation. Table 2.1 summarizes the results of this simulation-based investigation by summarizing the component values that yielded viable Clapp-type oscillator circuits.
Table 2.1. Component values summarization for viable sensing systems listed in this dissertation

<table>
<thead>
<tr>
<th>Oscillator Design Frequency (MHZ)</th>
<th>LT (nH)</th>
<th>Cr (pF)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
<th>Design Group</th>
</tr>
</thead>
<tbody>
<tr>
<td>1000</td>
<td>8.5</td>
<td>10</td>
<td>4</td>
<td>4</td>
<td>A</td>
</tr>
<tr>
<td>940</td>
<td>10.5</td>
<td>50</td>
<td>10</td>
<td>10</td>
<td>A</td>
</tr>
<tr>
<td>720</td>
<td>8.5</td>
<td>10</td>
<td>4</td>
<td>4</td>
<td>A</td>
</tr>
<tr>
<td>30</td>
<td>3000</td>
<td>14</td>
<td>50</td>
<td>50</td>
<td>B</td>
</tr>
<tr>
<td>90</td>
<td>400</td>
<td>14</td>
<td>20</td>
<td>20</td>
<td>B</td>
</tr>
<tr>
<td>103.9 to 101.3</td>
<td>400</td>
<td>8.5 to 9.3</td>
<td>20</td>
<td>20</td>
<td>B</td>
</tr>
<tr>
<td>132.0 to 132.6</td>
<td>400</td>
<td>4.1 to 3.2</td>
<td>14</td>
<td>28</td>
<td>B</td>
</tr>
<tr>
<td>124.7 to 127.5</td>
<td>390</td>
<td>5.15 to 4.41</td>
<td>13</td>
<td>40</td>
<td>A</td>
</tr>
</tbody>
</table>

### 2.3. 6H-SiC MESFET

The active device used in the pressure sensor system is an unpackaged 10W n-type SiC power MESFET (Cree Model #: CRF24010D). The MESFET has features such as a 15 dB small signal gain, a drain to source breakdown voltage of over 100 V and operating frequencies up to 5 GHz. The Cree SiC MESFET was chosen as the active in all the designs because of its previously demonstrated performance at high temperatures [J. B. Casady] coupled with the fact that an accurate SiC Cree MESFET model was previously developed specifically for ADS, thus significantly aiding in system design. A cross sectional schematic of the SiC MESFET is shown in Fig. 2.6, highlighting the doped SiC regions, metal contacts (drain, gate and source regions), the p-type buffer layer, the N-channel and the SiC insulator. Like any conventional MESFET, the gate contact and the n-type channel
layer form a Schottky barrier, and when a negative voltage is applied to the gate, a depletion region is created under the gate that extends into the channel, reducing the current. Because of the Schottky barrier, there is negligible gate current. Increasing the negative gate voltage will eventually cause the depletion region to extend completely through the channel, pinching off the drain current. Once pinch off is reached, a positive voltage can be applied to the drain and then the negative voltage on the gate can be decreased, allowing current to flow through the channel, until the desired operational current, $I_{DS}$, is reached. The p-type buffer layer prevents channel electrons from entering the SiC substrate. A microphotograph of a 10W n-type SiC power MESFET die used in this dissertation is shown in Fig. 2.7. Figure 2.7 shows the drain, gate and source pads. The wire bounds are used to enable an electrical connection between the MESFET and the rest of the circuit.

![Figure 2.6. Cross-sectional schematic of the Cree MESFET.](image)
2.4. SiCN Capacitive Pressure Sensor

The micro-electromechanical systems (MEMS) capacitive pressure sensor used in this dissertation is shown in Fig. 2.8. The sensor, acquired from Sporian Microsystems, is fabricated from SiCN which is a derivative of silicon carbide synthesized by thermal
decomposition of polymeric precursors that possesses excellent mechanical properties, tunable electric properties and superior oxidation/corrosion resistance at temperatures up to 1600°C (L. An, Y. Wang, X. Ren). The sensor and gold metal feed lines are fabricated on a high purity alumina substrate. One electrode of the capacitive sensor is fabricated on the deflecting chamber membrane that forms a sealed cavity, and the second electrode is on the fixed alumina substrate. The sealed cavity is flip-chip bonded onto the alumina substrate with gold contacts such that the two electrodes form a parallel plate capacitor. Thus, as pressure increases the membrane is flexed up into the sealed cavity causing the capacitance of the pressure sensor to decrease.

Figure 2.8. Photograph of the SiCN-based capacitive pressure sensor.
Chapter 3

Substrate and Component Evaluation

3.1. Benchtop Testing

The pressure sensor systems developed in this dissertation consist of transistors, capacitors, inductors, and resistors that must be electrically stable at high temperatures. In order to verify the individual performance of these components, as well as the fully assembled systems, a high temperature probe station (HTPS) was developed. A set of photographs of the probe station is shown in Figure 3.1. The high temperature probe station consisted of a Cascade probe station, custom-made high temperature ground-signal-ground (GSG) probes from GGB Industries, high temperature tungsten/gold tipped DC needle probes, a thermocouple, a high temperature chuck made of shuttle tile and a ceramic heater from Momentive Performance Materials. The temperature of the ceramic heater, which has
a maximum operating temperature of 1000°C, was controlled by a LabVIEW program with a proportional-integral-derivative (PID) controller and 3kW power source using a feedback loop system. S-parameters were measured using the GSG probes in combination with a network analyzer, but the testing capabilities were limited to 450°C due to the maximum operating temperature of the GSG probes, which is approximately 450°C. However, DC or low frequency (less than 100 MHz) measurements can be obtained up to 600°C with the tungsten/gold tip DC needle probes.

Figure 3.1. Photographs of the high temperature probe station (HTPS).

3.2. Substrate Evaluation

The thin film capacitors, inductors and antennas require a substrate that is electrically and chemically stable, electrically insulating, and RF transparent at high
temperatures. Alumina and sapphire were selected for evaluation and characterized by measuring effective permittivity over a frequency range of 45 MHz to 50 GHz at temperatures up to 400°C (M. C. Scardelletti, June 2007) (G. E. Ponchak, February 2009). Coplanar waveguide (CPW) thru-reflect-line (TRL) calibration standards were fabricated on the two substrates and the effective permittivity was extracted. Testing results are shown in Figure 3.2 and Figures 3.3. The CPW TRL standards consisted of a titanium/gold (Ti/Au (0.025/1.23 µm)) metallization and a characteristic impedance (Z₀) of 50Ω. The results indicate that both the sapphire and alumina substrates performance differs by less than 2% between 25 and 400°C.

![Effective Permittivity vs Frequency](image)

Figure 3.2. Extracted effective permittivity of alumina with a CPW line (s = 40, w = 80 µm) Z₀ = 50Ω.
Figure 3.3. Extracted effective permittivity of sapphire with a CPW line \((s = 40, w = 80 \mu m) Z_0 = 50\Omega\).

Figures 3.4 and 3.5 are optical micrographs used to evaluate the effects of exposure to elevated temperature on the Ti/Au metallization. These Ti/Au structures included wire bonds so as to better evaluate their performance as interconnect layers. The Ti/Au metallization on the alumina substrate did not exhibit any degradation after heating to 400°C. In contrast, the Ti/Au metal lines on sapphire showed evidence cracking after heating to 200°C, which became more severe after heating to 400°C.
Figure 3.4. Microphotograph of Ti/Au metallization with wire bonds on an alumina substrate after heating to 400°C.

Figure 3.5. Microphotograph of Ti/Au metallization with wire bonds on a sapphire substrate after heating to 400°C.
Both alumina and sapphire substrates exhibited stable electrical and mechanical performance up to 400°C; however, the excessive cracking in the Ti/Au metallization layers on sapphire at temperatures approaching 400°C indicate that it was not a suitable substrate material for the high temperature sensors to be developed in this dissertation. Moreover, sapphire substrates are about 5 times more expensive than alumina of similar dimensions, consequently alumina substrates were selected for this project. To this end, the CoorsTek 996 Alumina Superstrate (CoorsTek) was chosen as the substrate to be used in all fabricated sensor systems. Substrates of different thicknesses are shown in Fig. 3.6. These substrates come in a variety of convenient thicknesses and can be readily machined into desired shapes and areas. The substrates have a grain size of less than 1 µm and can be polished to a surface roughness of ~25 nm, making them particularly well suited for metallization layers created by thin film deposition and photolithographic patterning.

Figure 3.6. Photograph of a collection of CoorsTek 996 Alumina Superstrate substrates with different thicknesses.
3.3. High Temperature Inductor Development

Thin film spiral inductors were fabricated and characterized over a temperature range of 25 to 475°C. The inductors were fabricated on the 500 µm thick, 99.6 %, double side polished alumina substrate described previously. The metal structures were defined through a conventional lift-off process and consisted of a 0.025 µm-thick Ti adhesion layer, a 0.51 µm-thick first level Au layer, and a 1.0 µm-thick second level Au layer. The entire coil was therefore 1.51 µm thick except under the crossover, which was 0.51 µm in thickness. Because the inductors were to be characterized at temperatures up to 475°C, an air bridge was not used for the crossover. Instead, the crossover was supported by a 0.85 µm-thick SiO₂/Si₃N₄ insulating stack. Five sets of inductors, with each set comprised of a 1.5, 2.5, 3.5, and 4.5 turn inductor with different line widths and spacings, were fabricated for testing. A CAD drawing of a 2.5-turn inductor is shown in Fig. 3.7.
The inductors were characterized from 10 MHz to 30 GHz using an Agilent E8364B Network Analyzer in order to measure the inductance and parasites as a function of temperature. The inductors were measured on the high temperature probe station discussed in Section 3.1 and ground-signal-ground (GSG) high temperature probes (Model P-12-9403 from GGB Industries) were used to facilitate contact. The ground-signal-ground probes had a 150 µm pitch. A short-open-load-thru (SOLT) calibration was performed at the probes tips to ensure accuracy. After the SOLT calibration, which establishes the reference plane at the probe tips, open and a short circuit structures were measured for pad parasitic removal using the Cascade WinCal software. The pad removal step corrected for the capacitance and inductance caused by the probe pads, and is therefore expected to
correct for small variations in the probe parasitic reactance due to a change in temperature. This step is particularly important for accurate testing at temperatures above 300°C, because wafer calibration could not be performed due to the fact that such temperatures exceed the maximum temperature of the calibration substrate.

The corrected S-parameters were optimized against the equivalent circuit model shown in Figure 3.8 using Agilent’s Advanced Design System (ADS) circuit simulator and gradient optimization tool in order to extract the appropriate equivalent circuit parameters. Figure 3.9 is a plot of the extracted inductance versus frequency at temperatures from 75 to 475°C. It is seen that the inductance exhibits the expected relationship with frequency; it is nearly constant at low frequency and increases dramatically as the self-resonant frequency is approached. Therefore, the frequency of high temperature oscillators or other resonant circuits is not expected to vary by more than a few percent due to variations in the inductance. The series resistance, $R$, as a function of temperature is shown in Figure 3.10. The series resistance increases with temperature as expected due to the increase in resistivity with temperature. Furthermore, the extracted inductance, $L$, as a function of temperature for the 1.5 turn inductor as determined using ADS, the Y-parameters, and Sonnet is shown in Figure 3.11.
Figure 3.8. Equivalent circuit model of the spiral thin film inductors.

Figure 3.9. Inductance as a function of frequency for a 1.5 turn inductor at temperatures between 75 and 475°C.
Figure 3.10. Resistance as a function of frequency and temperature for a 1.5 turn inductor.

Figure 3.11. Extracted inductance, \( L \), as a function of temperature for the 1.5 turn inductor as determined by ADS, the Y-parameters, and Sonnet.
3.4 Development and Evaluation of High Temperature SiC MIM Capacitors

Thin film metal-insulator-metal (MIM) capacitors are attractive for microsystems applications because integration is, in principle, straightforward and they exhibit extremely low form factor in the vertical direction. SiC is an appealing dielectric material for high temperature MIM capacitors because the same material can be used as a surface passivation layer in other areas of the circuit, primarily to protect metal interconnects from oxidation during high temperature exposure. Thin film SiC MIM capacitors were fabricated and characterized at temperatures from 25 to 500°C in order to evaluate their utility in the proposed sensor system (M.C. Scardelletti, May, 2011). Five capacitor designs with areas of 0.029, 0.144, 0.292, 0.436 and 0.593 mm$^2$ (170 x 170, 380 x 380, 540 x 540, 660 x 660, and 770 x 770 µm) were fabricated using standard, thin film photolithographic processes. The capacitors were fabricated on four, 2 x 2 inch CoorsTek’s Superstrate TPSTM alumina substrates. The dielectric constant, substrate thickness and loss tangent for the substrate were 9.9, 500 µm and 0.0001, respectively.

Electrically insulating, amorphous SiC was envisioned to be used as the dielectric layer for the MIM capacitors. The amorphous SiC insulator was deposited by plasma enhanced chemical vapor deposition (PECVD) in a PETS-PECVD-SS12-4-LFTM system. The system has a powered upper electrode and water cooled chamber top plate to reduce the formation of particulates during deposition. The system is equipped with a 645 W power supply and a 1500 W resistive heater, enabling a deposition process temperature of up to 500°C. Silane and methane were used as Si- and C- containing precursors. To ensure that insulating films were deposited, no doping gas was used. For the capacitors, the
substrate temperature was held constant at either 300 or 450°C during deposition. The RF power and pressure were held constant at 35 mW and 900 mTorr, respectively. The precursor gas flow rate ratio was fixed at 1:4, using 7.5 sccm of silane and 30 sccm of methane. Figure 3.12 is an X-ray photoelectron spectrum from a SiC film deposited at 300°C. The ratio of Si-to-C is nearly unity, as expected for SiC, with a trace amount of oxygen observed as the main impurity. The Ar in the measurement is residual from the XPS ion milling cleaning process of the sample. Given the growth conditions, it is likely that the films contain hydrogen, which cannot be detected by XPS. XPS spectra from a film deposited at 450°C also showed that the Si-to-C ratio was nearly 1:1.

![X-ray photoelectron spectrum of an amorphous SiC film deposited by PECVD at 300°C.](image)

Figure 3.12. X-Ray photoelectron spectrum of an amorphous SiC film deposited by PECVD at 300°C.
Since interface integrity is a very important characteristic of thin film capacitors, an atomic force microscope (AFM) analysis was performed on the SiC films prior to capacitor fabrication to characterize surface roughness. Initially, SiC was grown on atomically flat Si wafers. Figs. 3.13 and 3.14 show AFM micrographs of the PECVD SiC grown on Si at substrate temperatures of 300 and 450°C, respectively. For depositions performed at 300°C, the amorphous SiC has an average surface roughness of ±2.5 nm, but when deposited at 450°C, the average surface roughness increased to ±10 nm.

Figure 3.13. AFM micrograph from a PECVD SiC film deposited at 300°C on a polished Si substrate.
SiC films were also deposited directly onto alumina substrates using the deposition recipe described above and analyzed by AFM. Figure 3.15 is an AFM micrograph of a SiC film deposited at 300°C on an alumina substrate. It can be seen in this figure that the SiC is much rougher than an equivalent film on Si; however, this roughness is not inherent to the film, but rather is associated with the much rougher alumina substrate. The PECVD SiC film deposited on alumina at 450°C exhibited similar characteristics. Considering surface roughness alone, the better substrate for MIM capacitors would be Si. However, the dielectric properties of alumina are essential for the proposed device, therefore it was decided to fabricate SiC-based MIM capacitors for evaluation purposes in spite of the higher roughness.
For the thin film MIM capacitors, two bottom-level metal electrode configurations were investigated to determine their durability during the high temperature SiC fabrication processing temperatures and on the capacitor characteristics during high temperature measurements. The first configuration consisted of a titanium/gold (Ti/Au: 25/350 nm) multilayer while the second configuration consisted of a titanium/platinum/gold/platinum (Ti/Pt/Au/Pt: 25/10/350/10 nm) multilayer. The metal layers were deposited onto alumina substrates in a conventional electron beam evaporator and defined by a standard lift-off lithography procedure. Half of the samples were subjected to an anneal at 650°C for one hour in nitrogen to evaluate the use of a post deposition high temperature anneal as a means to stabilize the metals for later processing and measurements at high temperature.
As with the SiC dielectric, AFM was used to investigate the surface roughness of the as-deposited and annealed samples. Figs. 3.16, 3.17, 3.18, and 3.19 show AFM micrographs of an as-deposited Ti/Au multilayer, an annealed Ti/Au multilayer, an as-deposited Ti/Pt/Au/Pt multilayer, and an annealed Ti/Pt/Au/Pt multilayer, respectively. For both electrode configurations, it is seen that the metallization layers are relatively smooth prior to annealing, but the surface roughness increases significantly after annealing. The AFM micrographs suggest that the Ti/Pt/Au/Pt multilayer has a higher surface roughness after annealing than its Ti/Au counterpart. This is perhaps a result of the additional Pt films in the 4-layer structure, but more study is required to determine the cause of the roughening.

![Figure 3.16. AFM micrograph of an as-deposited Ti/Au multilayer deposited on an alumina substrate.](image)
Figure 3.17. AFM micrograph of Ti/Au multilayer deposited on an alumina substrate and annealed at 650°C for one hour.

Figure 3.18. AFM micrograph of an as-deposited Ti/Pt/Au/Pt multilayer deposited on an alumina substrate.
Figure 3.19. AFM micrograph of a Ti/Pt/Au/Pt multilayer deposited on an alumina substrate and annealed at 650ºC for one hour.

For the thin film MIM capacitors, the top-level metal layer was a titanium/platinum/gold/platinum multilayer ( Ti/Pt/Au/Pt: 25/10/1300/10 nm). A microphotograph of a thin film SiC MIM capacitor, presented in Fig. 3.20, shows the spatial relationships between the top and bottom layers, the SiC dielectric, and GSG probe feed lines that connect to the top and bottom electrodes. The color variations are indicative of the stacking sequence of the various thin films with red indicative of SiC on the bottom electrode, light brown associated with SiC on alumina and gold uncoated metal.
SiC MIM capacitors were fabricated using both bottom electrode configurations. The capacitors were characterized on a modified version of the high temperature RF probe station detailed in Section 3.1 of this dissertation. The capacitance of the SiC MIM capacitors was measured with a Keithley 590 C-V analyzer at 10 MHz. A LabVIEW program was used to control the C-V meter and the temperature of the ceramic heater. The capacitors were measured from 25 to 500°C in increments of 50°C, over a voltage range from 0 to 20 V in increments of 0.25 V. An open pad calibration standard was used to calibrate the C-V analyzer at each temperature.
The measured capacitance as a function of temperature for capacitors grown at 300ºC and 450ºC is shown in Figs. 3.21 and 3.22, respectively. The capacitance for each growth temperature is nominally independent of the first level metal configuration used, which is expected. It is also seen that the capacitance is essentially independent of temperature up through 350ºC. Using the area of the largest capacitor, which has the smallest fringing field capacitance, the extracted relative room temperature dielectric constant of the SiC grown at 300ºC and 450ºC is 7.7 and 9.4, respectively. Such a difference in dielectric constant is not entirely surprising since the microstructure, mass density and hydrogen concentration of amorphous SiC films deposited at different temperatures by PECVD can vary significantly, especially if the difference in deposition temperature is large, as in this dissertation.
Figure 3.21. Measured capacitance as a function of temperature for a capacitor consisting of Ti/Pt and Ti/Pt/Au/Pt metallizations and a SiC dielectric film deposited at 300ºC.

Figure 3.22. Measured capacitance as a function of temperature for a capacitor with Ti/Pt/Au/Pt first level metal and a SiC dielectric deposited at 450ºC.
Figure 3.23 shows a representative plot of measured capacitance as a function of voltage. It is seen that the capacitance is constant with voltage through 300°C; however at 350°C the variation with voltage increases and the capacitor breaks down at higher temperatures. Furthermore, the SiC was observed to crack as the temperature was increased, with failures observed at 400°C and 450°C for the SiC grown at 300 and 450°C, respectively. Such cracking is visible in the optical micrograph of Fig. 3.24, which is from a SiC MIM capacitor after testing at 400°C. Crack formation could be due to an increase in residual stress and/or densification due to hydrogen effusion from the films. Understanding the mechanisms governing crack formation and methods to mitigate such issues is the subject of current investigation.

Figure 3.23. Capacitance versus voltage from 25 to 350°C for a 144.4 nm² capacitor with a Ti/Au first level metal and a SiC dielectric deposited at 300°C.
During the C-V characterization as a function of temperature, the capacitors fabricated with SiC films grown at 300ºC and a Ti/Pt/Au/Pt first level metal that was annealed at 650ºC failed even at room temperature due to short circuits. The failure locations were visible on the top level metal as small defects that formed as the voltage was increased. The exact cause of this behavior is not well understood, but because the capacitors failed to operate at room temperature, this combination of materials was not considered for further capacitor development.

Figure 3.24. Photograph of a capacitor fabricated with a SiC film deposited at 300ºC after a measurement performed at 400ºC.
S-parameter measurements were made on the capacitors using an Agilent E8364B PNA Network Analyzer. The S-parameters were measured from 10 MHz to 10 GHz over a temperature range 25 to 450°C, at increments of 100°C (except at 450°C). The high temperature probe station described in Section 3.1 was used for the measurements. As with previous high temperature measurements, the 150 µm pitch high temperature probes (Model P-12-9403 from GGB Industries) were used. An on-wafer, short-open-load-thru (SOLT) calibration was performed at room temperature (25°C). A short and open circuit pad removal process was performed using the Probe Pad Removal (PPR) Wizard in Cascade’s WinCal Calibration software at each temperature, which corrects for the capacitances and inductances caused by the pads, and it is therefore expected to correct for the small variations in the probe parasitic reactance due to a change in temperature for the higher temperatures where the wafer calibration was not performed. Figure 3.25 shows a representative plot of the measured S-parameters as a function of frequency and temperature for a capacitor made from a Ti/Au bottom metal electrode and a SiC film deposited at 300°C. It can clearly be seen that the behavior of the $S_{11}$ and $S_{12}$ parameters is essentially independent of temperature between DC and 10 GHz.
Figure 3.25. Measured S-parameters for a 144.4 nm² capacitor with a Ti/Au first level metal and a SiC film deposited at 300ºC.

The lumped element circuit model shown in Fig. 3.26 was optimized against the measured S-parameters using the Agilent ADS software suite. The gradient optimizer was used in all the simulations. The lumped element model consisted of a series capacitance, Cs, parasitic capacitors, C₁ and C₂, a series resistance, Rs, and a series inductance, Ls. Rs and Ls model the feedlines to the capacitor as shown in Fig. 3.26. Note that the model contains no resistance in parallel with the series capacitor to represent the conductance through the dielectric because the modeled values were very large and not accurate. The series resistance was on the order of a few ohms and increased with temperature. The extracted capacitance is shown in Fig. 3.27. It is seen that the capacitance increases slightly with temperature, and that the extracted values are lower than that in Figs. 3.21 and 3.22. This is because the ADS model includes a series inductor, which has the effect of
decreasing the equivalent capacitance in a simple, one element model that consists only of the series capacitor.

Figure 3.26. Equivalent circuit model of a MIM capacitor.

Figure 3.27. Capacitance values from the ADS equivalent circuit model as a function of temperature.
The experiments performed in this dissertation indicate that the capacitance of MIM structures made from SiC films deposited at 300°C were essentially insensitive to temperature up to ~350°C, whereas the capacitance of MIM structures using SiC films deposited at 450°C remained stable up to ~400°C. Measurements as a function of temperature show that the SiC breakdown voltage decreased with temperature, and the SiC developed cracks that induced short circuits at 400°C. These finding indicate that the SiC MIM capacitors could be used in systems that have a maximum operating temperature of nominally 300°C. However, to achieve stable operation at 500°C, further development of SiC as a dielectric for capacitors is required before it can be reliably used at such temperatures.
Chapter 4

High Temperature Antenna and Oscillator Circuit Development

4.1 Wireless System Characterization

All the wireless oscillators developed in this dissertation were characterized in a custom-built anechoic chamber. The anechoic chamber is essentially a room that is lined with absorber structures designed to suppress reflection of RF energy that could potentially cause a disturbance in the emitted radiation pattern of a device under test (DUT). Photographs of the anechoic chamber are shown in Figs. 4.1 and 4.2. The chamber is equipped with an adjustable rotational pedestal, which enables the DUT (which serves as the transmit antenna) to be rotated by 360° in the horizontal plane. The pedestal can also be adjusted in the vertical direction. The receive antenna can be adjusted in the vertical or horizontal plane but does not have rotational capabilities. The chamber is large enough to accommodate a wide variety of test devices as well as equipment to evaluate their
performance under a range of environmental conditions, including a custom-built chamber for high temperature and high pressure testing. Components that comprise portions of this test chamber are also shown in Figs. 4.1 and 4.2.

Figure 4.1. Photograph of the anechoic chamber with absorber lined walls.
4.2. Wireless Oscillator Prototype Design #1: Clapp-type Oscillator Circuits with Microfabricated Slot-ring Antennas

The first prototype wireless design consisted of a Clapp-type oscillator integrated with a slot-ring antenna (G. E. Ponchak, April 2009). A photograph of this prototype is
shown in Fig. 4.3 and a close-up photograph of the oscillator circuit is shown in Fig. 4.4. This prototype was designed to generate and transmit a signal at 1 GHz. The square slot-ring antenna measured 36.72 mm on a side and its geometry was optimized to be impedance matched with the Clapp-type oscillator. This prototype was fabricated on an alumina substrate that was 50.8 x 50.8 mm² using chrome/gold metallization. Lift-off based fabrication techniques were used to define the metallization.

Figure 4.3. Photograph of a slot-ring antenna integrated with a Clapp-type oscillator.
Figure 4.4. Close-up photograph of the Clapp oscillator in Fig. 4.3.

The 1 GHz oscillator used a Cree SiC MESFET as the active device. The capacitors were commercial ceramic capacitors acquired from American Technical Ceramics (ATC) and the inductor was a spiral, thin-film inductor fabricated during the metallization step. Capacitor $C_T$ and inductor $L_T$ comprised the resonant tank circuit of the device and determined the oscillation frequency. The DC bias circuits for the gate and the drain were located off chip and were held at room temperature during testing. The DC bias circuits were bias tees consisting of a capacitor tied to ground which served as a DC block and inductor in series that served a RF choke and are shown in Fig. 4.5. The values of the capacitor and inductor in the DC bias tees were 2.5 $\mu$F and 2.68 mH, respectively. The 1 GHz oscillator was tested in the aforementioned anechoic chamber on a ceramic heater with a PID controller feedback loop to control the temperature (M. C. Scardelletti, June 2007). A wideband horn antenna was placed 1 m above the active antenna, and a spectrum analyzer measured the radiated signal at the horn. It should be noted that the antenna
radiates with a near omnidirectional pattern with slightly more energy radiated into the ceramic heater owing to its higher relative permittivity. The drain voltage was maintained at 10 V and the gate bias was varied to maintain a drain current of 130 mA. The measurements were initiated with the ceramic heater temperature at 25°C and subsequent measurements were made in 10 degree steps. The temperature was held constant for 10 minutes before the radiated signal was measured to ensure that the circuit was at a uniform temperature.

Figure 4.5. Photograph of the off-chip DC bias circuits.

Figure 4.6 shows the measured signal (power versus frequency) at 270°C, and Fig. 4.7 shows the measured oscillation frequency as a function of received power for
temperatures up to 275°C. It is seen that the frequency decreases by less than 4% up to a temperature of 245°C. However, the power dropped by 18 dB over the same range. The maximum temperature that the circuit functioned properly was 270°C, above which the transconductance ($g_m$) of the transistor was no longer sufficient to offset the losses in the oscillator and the antenna.

![Graph showing measured power versus frequency at 270°C for the Clapp oscillator circuit with slot antenna.](image)

Figure 4.6. Measured power versus frequency at 270°C for the Clapp oscillator circuit with slot antenna.
Figure 4.7. Measured oscillation frequency and received power versus temperature for the Clapp oscillator with slot antenna.

4.3. *Wireless Oscillator Prototype Design #2: Clapp-type Oscillator Circuits with Chip Antennas*

In addition to oscillator circuits with slot ring antennas, oscillator circuits with chip antennas were investigated (M. C. Scardelletti, January 2011). Although they are not intended for operation at 400°C, chip antennas are attractive options for wireless applications at moderate temperatures (< 300°C) due to their small form factor and ease of integration. In order to explore routes to reduce the overall size of Prototype Design #1, two oscillator circuits with chip antennas were designed and evaluated. Shown in Figs. 4.8 and 4.9, these circuits were designed to operate at resonant frequencies of 720 and 940
MHz, respectively. The Clapp-type oscillators were fabricated on 99.9 % alumina substrates that were 500 µm thick and a dielectric constant of 9.9. By replacing the slot-ring antenna used the first prototype with a chip antenna, the overall size of the oscillator was reduced from 50.8 x 50.8 mm² to 25.4 x 25.4 mm². The Cree SiC MESFET described previously was used in both oscillator circuits. Because the maximum operating temperature of this design was not expected to exceed 300°C, the previously described thin film SiC MIM capacitors and spiral inductors were used. The thin film spiral inductor for the 720 MHz design incorporated 1.5 turns while that for the 940 MHz design incorporated 1.9 turns, which resulted in inductances of 8.5 and 10.5 nH, respectively according to Sonnet simulations. The values of $C_T$, $C_1$, $C_2$ and $L_T$ are given in Table 4.1 for both the 720 and 940 MHz designs.

The miniature chip antennas were acquired from Johanson Technology (Model # 0920AT50A080). The antennas resonate at approximately 920 MHz, therefore no matching circuit was required for the 940 MHz oscillator circuit design. To provide sufficient matching to the 720 MHz circuit, a 2 pF chip capacitor was placed in series with the input of the antenna to resonate with the series inductance of the antenna. Wire bonds were used to connect the drain, gate and source of the transistor to DC bias pads.

Table 4.1. Component values for the 720 and 940 MHz oscillator designs.

<table>
<thead>
<tr>
<th>Oscillator</th>
<th>$C_T$ (pF)</th>
<th>$C_1$ (pF)</th>
<th>$C_2$ (pF)</th>
<th>$L_T$ (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>720 MHz</td>
<td>10</td>
<td>4</td>
<td>4</td>
<td>8.5</td>
</tr>
<tr>
<td>940 MHz</td>
<td>5</td>
<td>10</td>
<td>10</td>
<td>10.5</td>
</tr>
</tbody>
</table>
Figure 4.8. Photograph of the 720 MHz oscillator circuit with chip antenna.

Figure 4.9. Photograph of the 940 MHz oscillator circuit with chip antenna.
The oscillators were characterized in the previously described anechoic chamber. All of the test equipment was positioned outside the chamber to eliminate unwanted interference. The substrates were clamped to a ceramic heater that was placed on a rotational stage so that a 180° radiation pattern could be measured. The receive antenna consisted of wideband horn with a gain of 2.3 and 4.0 dBi at 720 and 940 MHz, respectively. The distance between the device under test and the receive antenna was 1.7 m. Between the oscillator and ceramic heater were positioned four, 500 µm-thick alumina wafers to minimize electromagnetic effects between the heating element and the circuits. A thermocouple was placed on the substrate near the oscillator to accurately record the temperature of the circuit. The temperature of the ceramic heater was varied with a Labview controlled power source. An Agilent E4440A spectrum analyzer was used to record the frequency spectrum and phase noise at temperature increments of 25°C starting at room temperature and ending when the oscillator failed to function. For both designs, the gate voltage was varied as the temperature was changed to keep the drain current constant at 100 mA. For the 720 MHz circuit, the drain voltage was 6 V, and for the 940 MHz design, the drain voltage was 8.8 V.

The measured spectra for the 720 MHz design at 25 and 200 °C are shown in Figure 4.10. It is seen that the frequency of oscillation and transmitted power decrease with temperature. The oscillation frequency decreases from 726.2 MHz at 25°C to 720 MHz at 200°C, which is less than 1%. The transmitted power decreases from -22.1 dBm at 25°C to -27.9 dBm at 175°C, which is a decrease of only 3 dBm. However, between 175°C and 200°C the transmitted power decreases to -34.4 dBm. At a temperature just above 200°C,
the 720 MHz oscillator ceases to operate. Fig. 4.11plots the frequency and transmitted power for each temperature setting.

Figure 4.10. Measured output power versus frequency from the 720 MHz circuit at 25 and 200°C.
Figure 4.11. Measured received power and frequency versus temperature for the 720 MHz circuit.

Figure 4.12 shows the measured spectra at 25°C and 250°C for the 940 MHz oscillator circuit design. Like the 720 MHz design, the oscillation frequency and transmitted power decrease with increasing temperature. The frequency decreases from 940.2 MHz at 25°C to 932.2 MHz at 250°C, which is less than a 1% change. The transmitted power decreases from -27.6 dBm at 25°C to -32.9 dBm at 175°C. For temperatures above 175°C, the power decreases more rapidly, reaching a maximum of -41.7 dB at 250°C. Just above 250°C the 940 MHz oscillator ceases operation. Figure 4.13 shows the frequency and transmitted power for the 940 MHz oscillator circuit for each temperature setting.
Figure 4.12. Measured output power versus frequency for the 940 MHz circuit at 25 and 250°C.

Figure 4.13. Measured received power and frequency versus temperature for the 940 MHz circuit.
Plots of measured phase noise versus offset frequency for the 940 MHz oscillator circuit at 25 and 250°C are shown in Fig. 4.14. For offset frequencies far from the carrier frequency, the phase noise at 250°C was higher than that at 25°C by 20 dBc/Hz. Figure 4.15 shows the phase noise at an offset frequency of 100 kHz for temperatures between 25 and 250°C. The phase noise increases with increasing temperature by 20 dB/Hz, which is expected. The 720 MHz circuit exhibited similar phase noise characteristics.

Figure 4.14. Measured phase noise at 25 and 250°C for the 940 MHz circuit.
Figure 4.15. Measured phase noise versus temperature for the 940 MHz circuit at 100 kHz offset frequency.

The radiation patterns for the 720 and 940 MHz oscillator circuits are shown in Figs. 4.16 and 4.17, respectively. These radiation patterns represent the transmitted power as a function of the angle between the device under test and the receive antenna. Both graphs clearly demonstrate that the transmitted power decreases as the temperature increases. The transmitted power for the 720 MHz oscillator circuit decreased by 12.3 dB from 25 to 200°C for all measurement angles while that for the 940 MHz design decreased by 14.1 dB from 25 to 250°C. More importantly, however, the radiation patterns for both circuit designs change very little with an increase in temperature.
Figure 4.16. Radiation pattern versus temperature for a 720 MHz oscillator with chip antenna.
4.4. Wireless Oscillator Prototype Design #3: High Temperature Clapp-type Oscillator Circuit Development

The aforementioned oscillator circuits were required to operate at 720 and 940 MHz due to the 920 MHz chip antennas used in the structures (M. C. Scardelletti, January 2011). Unfortunately the Cree SiC MESFET is not well suited for these particular chip antennas, primarily because its gain is substantially lower at these high operational frequencies. This issue is compounded by the fact that as the temperature of the device increases, the gain of
the MESFET decreases substantially due to the increasing channel temperature of the MESFET, which increases the losses of the FET, causing the oscillator circuit to cease operation at some maximum temperature. For the 720 and 940 MHz circuits of Prototype #2, the oscillator failed to resonate at approximately 200 and 250°C, respectively. To address this limitation while still using the Cree SiC MESFET, the design frequency of the oscillator circuit was intentionally decreased. To best assess the trade-offs in lowering the operational frequency, oscillator circuits with resonant frequencies at 30 and 90 MHz were designed and fabricated. Photographs of these circuits are shown in Figs. 4.18 and 4.19 (G. E. Ponchak, December 2010). The oscillators were fabricated on double side polished, 99.6% alumina substrates with a thickness of 508 µm. The values of the various capacitors and inductors are shown in Table 4.2. The inductors were designed based on simulations using Sonnet. The 3000 nH inductor, which was used in the 30 MHz circuit, was a 10 turn, square, planar spiral with a strip width and spacing each of 1 mm and a inner turn side length of 9 mm. The 400 nH inductor, which was used in the 90 MHz circuit, was a 2 turn, circular, planar spiral with 1 mm a line width and spacing of 1 mm and an inner turn diameter of 35 mm. The inductors provided magnetic coupling for wireless signal transmission and therefore were used in place of chip antennas and slot-ring antennas as the radiating element. $C_T$, $C_1$ and $C_2$ are SiO$_2$ MIM capacitors developed at NASA Glenn Research Center as part of a project not related to this dissertation research.
Figure 4.18. Photograph of a 30 MHz oscillator.

Figure 4.19. Photograph of a 90 MHz oscillator.
Table 4.2. Component values for 30 and 90 MHz oscillator designs.

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>C₁ (pF)</th>
<th>C₂ (pF)</th>
<th>Cₜ (pF)</th>
<th>Lₜ (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>30</td>
<td>50</td>
<td>50</td>
<td>14</td>
<td>3000</td>
</tr>
<tr>
<td>90</td>
<td>20</td>
<td>20</td>
<td>14</td>
<td>400</td>
</tr>
</tbody>
</table>

The measured spectra of the 30 MHz wireless circuit at temperatures of 25 and 450°C are shown in Fig. 4.20. To collect these spectra, the gate voltage was varied to maintain a drain current of 50 mA using a 10 V supply. Figure 4.20 shows that lowering the operational frequency of the SiC MESFET based oscillator circuit results in a substantial increase in operational temperature, by at least 200°C. Moreover, the frequency shift over this temperature range was relatively small at 0.5% or 0.14 MHz. Figure 4.21 shows the peak received power and oscillation frequency as a function of temperature between 25°C and 450°C. Figure 4.21 shows that the received power remains relatively constant, decreasing by only 5 dBm between 25 and 400°C. However, the power decreases rapidly with temperature above 400°C. The oscillation frequency varies by less than 0.5% over the temperature range.
Figure 4.20. Output power versus frequency at 25 and 450°C for the 30 MHz oscillator.

Figure 4.21. Measured peak power and oscillation frequency versus temperature for the 30 MHz oscillator.
The measured spectra of the 90 MHz oscillator circuit for temperatures of 25 and 470°C are shown in Fig. 4.22. The 90 MHz design exhibited a maximum stable operating temperature of 470°C. Much like the 30 MHz circuit, the frequency shift over this temperature range was relatively small at 1.7% or 1.53 MHz. Fig. 4.23 shows the peak power and oscillation frequency as a function of temperature. As with the 30 MHz design, the received power is relatively stable and decreases by less than 5 dBm between 25 and 375°C. For temperatures greater than 375°C, the power decreases rapidly. The oscillation frequency varies by less than 2% over the temperature range.

Figure 4.22. Output power versus frequency at 25 and 470°C for the 90 MHz oscillator.
Figure 4.23. Measured peak power and oscillation frequency versus temperature for the 90 MHz oscillator.

The measured phase noise of the 90 MHz oscillator design at 25 and 470°C is shown in Fig. 4.24, and the measured phase noise at an offset frequency of 100 kHz versus temperature is shown in Fig. 4.25. As expected, the phase noise at frequencies far from the carrier frequency increases with temperature by 11 dB/Hz, but interestingly, the phase noise at the 100 kHz offset frequency decreases with temperature by 18 dB/Hz. This behavior at present is not understood and thus requires further study.
Figure 4.24. Measured phase noise at 25 and 470°C for the 90 MHz oscillator.

Figure 4.25. Measured phase noise versus temperature for the 90 MHz oscillator at the 100 kHz and 10 MHz offset frequencies.
4.5 Summary

In this chapter a Clapp-type oscillator circuit that was integrated with a slot ring antenna and designed to operate at 1 GHz was successfully demonstrated. The overall area of the circuit was 50.8 by 50.8 mm\(^2\) and it sustained successful operation up to 270°C. This first prototype showed that a SiC MESFET-based Clapp-type oscillator circuit can be incorporated into a wireless design and will operate at elevated temperature, but the area of the slot-ring antenna significantly compromises miniaturization. To reduce the overall size of the wireless circuit, oscillator circuits with chip antennas that operated at 940 and 720 MHz were developed. The implementation of the chip antennas reduced the overall size of the wireless circuits to 25.4 by 25.4 mm\(^2\), however the maximum reliable operational temperature of these designs did not exceed 250°C, which is well below the target temperature of 400°C. It was determined that the gain of the SiC MESFET at these high frequencies was limiting the maximum operating temperature of these designs. To increase the operational temperature of the device, the frequency of the circuit was intentionally lowered to frequencies where the MESFET gain was high enough to compensate for temperature-dependent losses in the MESFET in the 400°C range. Two oscillator circuits with design frequencies of 90 and 30 MHz were developed. In order to reduce the resonant frequency of the Clapp-type oscillator circuits the inductance of inductor, L\(_T\), was increased. In these designs, the inductor also serves as the radiating element, substituting for the previously used slot-ring and chip antennas. By reducing the resonant frequency of the Clapp-type circuit, the operational temperature of the Clapp-type wireless system was raised to at least 450°C, which exceeds the target operating temperature. Unfortunately, the necessary increase in inductance led to an increase in
substrate area to 50.4 by 50.4 mm$^2$, which was comparable to the first prototype. These findings suggest that to achieve the form factors desired of the proposed pressure sensor system, a wired implementation might be required. Nevertheless, the fact that a Clapp-type oscillator circuit could successfully be integrated into a wireless system that functions above 400°C provides the technical justification to proceed with the incorporation of pressure sensing capabilities within the circuit.
Chapter 5

Development of Wireless Pressure Sensor Systems for High Temperature Operation

5.1. High Temperature Pressure Sensor Testing Apparatus

A custom-built, high temperature pressure chamber (HTPC) was developed to evaluate the performance of the pressure sensor systems. A photograph of this system is shown in Fig. 5.1. This high temperature pressure chamber is made of stainless steel and is capable of operating at temperatures up to 500°C and pressures up to 100 psi. Heating was accomplished by a ceramic heater that is controlled by a PID controller feedback loop (M. C. Scardelletti, June 2007). The chamber is equipped with a 125 mm diameter quartz
sight glass for optical inspection and signal transmission and thermocouples are placed throughout the chamber to ensure accurate temperature readings. The HTPC is equipped with four electrical feedthroughs to facilitate electrical power to devices under pressure. The chamber can accommodate circuits on substrates 50.4 by 50.4 mm$^2$ or smaller. The height from the surface of the ceramic heater to the bottom of the site glass is 50.4 mm. The overall height and width of the HTPC is 0.91 and 0.46 m, respectively. Chamber pressure is applied and controlled via pressurized lines, a cylinder (clean air) and a regulator.

Figure 5.1. Photographs of the high temperature pressure chamber.

5.2. Wireless Pressure Sensor System Prototype Design #1: A Polysilicon-based Capacitive Pressure Sensor System

In order to expedite the development of the Clapp-type oscillator circuit for wireless pressure sensing applications while concurrently developing a SiC-based pressure sensor,
a wireless system that utilizes a polysilicon capacitive pressure sensor was developed (G. W. Hunter, 2010). From a design perspective, this wireless pressure sensor system was very similar to the 90 MHz Clapp-type oscillator that was discussed in Section 4.4 except for the capacitor (CT) located in the tank circuit which was replaced with a MicroFAB™ MEMS-based capacitive pressure sensor. A microphotograph of the pressure sensor system is shown in Fig. 5.2. The MicroFAB polysilicon capacitive pressure sensor is shown in Fig. 5.3. The polysilicon sensor consisted of 16 individual pressure-sensitive plates that were surface micromachined on a silicon substrate. Each transducer had a counter electrode which is rigidly fixed on the substrate surface. All individual transducers of the array were connected in parallel in order to increase the overall sensor output. As a result, the capacitance of the array increases as the pressure increases. The sensor was initially characterized in the aforementioned high temperature pressure chamber using an Agilent Semiconductor Device Analyzer (SDA). The sensor was mounted inside the chamber after which the analyzer was calibrated to the leads of the pressure sensor at room temperature to remove the effects of the chamber and cabling. The sensor was characterize from 0 to 45 psi, exhibiting a change in capacitance from 8.5 to 9.3 pF.
Figure 5.2. Photograph of the wireless pressure sensor with polysilicon capacitive pressure sensor.
The polysilicon-based wireless pressure sensing system was evaluated at temperatures from 25 to 300°C over a pressure range from 0 to 45 psi. Thermoelectric generators (TEGs) were used to generate a portion of the power required to operate the sensor system. Details about the TEGs for power scavenging can be found in Appendix A. As shown in Fig. 5.4, at 300°C and 0 psi, the resonant frequency of the sensor system was 103.9 MHz, while at 45 psi the frequency shifted to 101.3 MHz. The change in frequency from 0 to 45 psi was 2.5%, translating to a sensitivity ($\Delta f/\Delta P$ MHz/psi) of 5.2 % at 300°C.
The output frequency as a function of applied pressure for this sensor system is shown in Fig. 5.5. At a 25°C, the resonant frequency of the pressure sensing system drops from 105.8 to 103.9 MHz as the pressure was increased from 0 psi to 45 psi, which corresponds to a 1.8% decrease over that range of pressures. At 300°C, the resonant frequency drops from 103.9 to 101.3 MHz from 0 psi to 45 psi, which is a 2.5% decrease. This relationship between frequency shift and operating temperature can be attributed to many factors but the magnitude of this shift is relatively small. In contrast, the slope of pressure vs. frequency plot (Fig. 5.5) at 25°C is $38 \times 10^{-3}$ MHz/psi and at 300°C is
$60 \times 10^3$ MHz/psi indicating that the increase in capacitance due to an increase in pressure is more pronounced at the higher temperatures.

![Graph showing frequency versus pressure for the polysilicon wireless pressure sensor system at 25°C and 300°C.](image)

Figure 5.5. Frequency versus pressure for the polysilicon wireless pressure sensor system at 25°C and 300°C.

5.3. **Wireless Pressure Sensor System Prototype Design #2: A High Temperature and Pressure SiCN-based Capacitive Pressure Sensor System**

The maximum operational temperature of the polysilicon-based capacitive pressure sensor system was roughly 100°C below the target temperature for the envisioned system.
This was most likely a result of the polysilicon capacitive pressure sensor becoming lossy at temperatures above 275°C, causing the gain of the MESFET to decrease and eventually leading to cessation of oscillation, which occurred at 300°C. To address this issue, a second prototype design was developed in which the polysilicon capacitive pressure sensor was replaced with a SiCN-based sensor. In anticipation of achieving operation at or above 400°C, the thin film MIM capacitors were replaced with titanate MIM capacitors.

5.3.1 Temperature and Pressure Testing of a SiCN Capacitive Pressure Sensor

The SiCN capacitive pressure sensor used in wireless pressure sensor system Prototype #2 was described in detail in Section 2.4. Prior to assembling the wireless pressure sensor system, the SiCN pressure sensor was individually characterized in isolation using an Agilent B1500A Semiconductor Device Analyzer (SDA) and the high temperature/pressure chamber discussed (HTPC) earlier. The SDA was calibrated to the leads of the pressure sensor inside the high temperature pressure chamber at room temperature to remove the effects of the chamber and cabling. The pressure sensor capacitance was measured at 1 MHz from 0 to 100 psi at 25°C and 400°C and the results are shown in Fig. 5.6. At 25°C, the sensor had an initial capacitance of 4.0993 pF at 0 psi which reduced to 3.2250 pF at 100 psi, resulting in a ΔC of 0.8743 pF or 3.5%. At 400°C, the sensor had a capacitance of 4.2403 pF at 0 psi and 3.3459 pF at 100 psi, which translated to a ΔC of 0.8944 pF or 3.6%. Although the actual capacitance values at a given pressure differed according to temperature, the differential capacitance was essentially independent of temperature. Furthermore, the slopes of the capacitance versus pressure data at 25 and
400°C were $8.7 \times 10^{-3}$ pF/psi and $8.9 \times 10^{-3}$ pF/psi, respectively indicating that the rate of change in capacitance as a function of pressure is insensitive to changes in temperature. These capacitance values could be readily accommodated in the Clapp-type oscillator designs developed previously, therefore the SiCN sensor was selected for use in the high temperature pressure sensor system.

Figure 5.6. Capacitance versus pressure for the SiCN capacitive pressure sensor.
5.3.2. High Temperature Testing of Titanate MIM Capacitors

The titanate MIM capacitors were acquired from Compex. These capacitors consisted of a proprietary titanate insulator and a titanium/platinum/gold metallization layer as both electrodes. The chemical composition of the titanate insulator is proprietary and therefore has not been publically revealed; however the material has a dielectric constant of 40 and a thickness of 0.1016 mm. To determine the potential functionality of the titanate MIM capacitors for use in the proposed sensor system, 14 and 41 pF capacitors with square areas of 4 and 12.25 mm$^2$, respectively, were characterized at temperature and operational frequency, using three methods. In the first method, the S-parameters were recorded with a network analyzer at temperatures from 25°C to 400°C in steps of 50°C, over a frequency range of 10 to 200 MHz. To facilitate the measurement, the high temperature probe station, described Section 3.1 was used. The probe station consisted of a ceramic heater on a chuck made of a high temperature insulating tile, a thermocouple and power source. A LabVIEW program was used to control the temperature settings.

Ground-signal-ground (GSG) high temperature probes with a 150 µm pitch were calibrated with a short-open-load-thru (SOLT) calibration substrate to ensure accuracy to the probe tips. The calibration was only performed at room temperature due to the temperature dependence of the calibration substrate. The MIM capacitors were epoxied to test fixtures on an alumina substrate, which can be seen in the inset of Fig. 3.1. The measured S-parameters of the 41 pF MIM capacitor at 25 and 400°C, shown in Fig. 5.7, are independent of temperature.
The lumped element circuit model, shown in Fig. 5.8, was optimized against the measured S-parameters in Keysight’s Advanced Design System (ADS) software suite. The gradient optimizer was used in all the simulations. The lumped element model consisted of a series capacitance, $C_S$, parasitic capacitances, $C_1$ and $C_2$, series resistance, $R_S$, and series inductance, $L_S$. To demonstrate the accuracy of the optimization method, the S-parameters of the 41 pF MIM capacitor at 400°C were optimized against the capacitor equivalent circuit model and the results are shown in Fig. 5.9. The measured and optimized traces for $S_{11}$ and $S_{21}$ are virtually indistinguishable, indicating the validity of the lumped element.
All of the component values acquired from the optimization method for both the 14 and 41 pF MIM capacitors from 25 to 400°C are listed in Tables 5.1 and 5.2, respectively. Note that this method results in frequency independent component values.

Figure 5.8. Equivalent circuit model for the titanate MIM capacitors.

Figure 5.9. Measured and simulated S-parameter magnitudes versus frequency for a 14 pF titanate MIM capacitor at 400°C.
Table 5.1. Circuit model values for 14 pF MIM capacitor.

<table>
<thead>
<tr>
<th>Temp (°C)</th>
<th>Cs (pF)</th>
<th>Rs (Ω)</th>
<th>Ls (nH)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>13.85</td>
<td>1.934</td>
<td>2.114</td>
<td>0.437</td>
<td>1.204</td>
</tr>
<tr>
<td>50</td>
<td>13.87</td>
<td>2.144</td>
<td>2.145</td>
<td>0.436</td>
<td>1.207</td>
</tr>
<tr>
<td>100</td>
<td>13.911</td>
<td>2.487</td>
<td>2.19</td>
<td>0.435</td>
<td>1.219</td>
</tr>
<tr>
<td>150</td>
<td>13.931</td>
<td>2.623</td>
<td>2.23</td>
<td>0.439</td>
<td>1.228</td>
</tr>
<tr>
<td>200</td>
<td>13.972</td>
<td>3.189</td>
<td>2.314</td>
<td>0.449</td>
<td>1.239</td>
</tr>
<tr>
<td>250</td>
<td>14.003</td>
<td>3.371</td>
<td>2.311</td>
<td>0.452</td>
<td>1.244</td>
</tr>
<tr>
<td>300</td>
<td>14.039</td>
<td>3.796</td>
<td>2.391</td>
<td>0.454</td>
<td>1.253</td>
</tr>
<tr>
<td>350</td>
<td>14.084</td>
<td>4.14</td>
<td>2.406</td>
<td>0.456</td>
<td>1.258</td>
</tr>
<tr>
<td>400</td>
<td>14.132</td>
<td>4.631</td>
<td>2.468</td>
<td>0.44</td>
<td>1.261</td>
</tr>
</tbody>
</table>

Table 5.2. Circuit model values for 41 pF MIM capacitor.

<table>
<thead>
<tr>
<th>Temp (°C)</th>
<th>Cs (pF)</th>
<th>Rs (Ω)</th>
<th>Ls (nH)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>42.804</td>
<td>1.778</td>
<td>2.645</td>
<td>0.352</td>
<td>1.863</td>
</tr>
<tr>
<td>50</td>
<td>42.865</td>
<td>1.916</td>
<td>2.669</td>
<td>0.346</td>
<td>1.874</td>
</tr>
<tr>
<td>100</td>
<td>42.993</td>
<td>2.245</td>
<td>2.716</td>
<td>0.344</td>
<td>1.893</td>
</tr>
<tr>
<td>150</td>
<td>43.082</td>
<td>2.482</td>
<td>2.756</td>
<td>0.352</td>
<td>1.909</td>
</tr>
<tr>
<td>200</td>
<td>43.179</td>
<td>2.834</td>
<td>2.807</td>
<td>0.357</td>
<td>1.919</td>
</tr>
<tr>
<td>250</td>
<td>43.28</td>
<td>3.112</td>
<td>2.847</td>
<td>0.364</td>
<td>1.924</td>
</tr>
<tr>
<td>300</td>
<td>43.368</td>
<td>3.325</td>
<td>2.88</td>
<td>0.367</td>
<td>1.932</td>
</tr>
<tr>
<td>350</td>
<td>43.509</td>
<td>3.668</td>
<td>2.928</td>
<td>0.363</td>
<td>1.939</td>
</tr>
<tr>
<td>400</td>
<td>43.641</td>
<td>4.064</td>
<td>2.965</td>
<td>0.357</td>
<td>1.943</td>
</tr>
</tbody>
</table>

The modeled S-parameter data shows that the values of the two MIM capacitors changed by approximately 2% over the temperature range from 25 to 400°C. Rs increases by approximately 2.5 Ω, which may actually be due the inability to calibrate out the additional loss of the probes as they approach 400°C. The shunt parasitic capacitances, C1 and C2, and the parasitic series inductance, Ls, are negligible for both the 14 and the 41 pF MIM capacitors.
The second method used to characterize the MIM capacitors involved a semiconductor device analyzer (SDA) and high temperature probe station. The GSG probes were replaced with DC needle probes, enabling the temperature range to be extended to 500°C. The measurements were recorded from 25 to 500°C in steps of 50°C. A calibration that consisted of phase compensation to account for the port extension and an open and short was performed to set the reference plane at the probe tips. The measurements were taken at 9.950, 9.975, 1, 1.025 and 1.050 MHz and the average value was recorded. The results are shown in Fig. 5.10.

The 14 and 41 pF MIM capacitors change by nominally 2 to 3% over the temperature range from 25 to 400°C and up to 5% from 400°C to 500°C. Furthermore, the conductance was measured and found to be negligible up to 400°C and then rose to no more than 10 μS for both the 14 and 41 pF capacitors at 500°C. It is not evident if the degradation in electrical performance was due to the temperature dependence of the material or probe contacts beginning to degrade due to the extreme environment.
Lastly, the MIM capacitors were measured from 40 Hz to 110 MHz at room temperature using a four-point probing technique with an impedance analyzer to determine if they were able to operate at the desired frequency range. It is vital to ensure that the passive components do not have a self-resonant frequency (SRF) near the operating frequency range of the sensing system. A calibration was performed, consisting of a phase compensation to account for the port extension, and an open, short and load to set the reference plane at the probe tips. The measured data is shown in Fig. 5.11.
The values of the 14 and 44 pF capacitors were constant across the measured frequency range, varying by less than 2%. The fluctuation in the measured data that occurs at roughly 75 to 90 MHz was due to the calibration routine not properly working at that frequency range, and as a result the data obtained in this frequency range was inaccurate. However, the calibration did recover around 90 MHz and is valid up to 110 MHz indicating that there is no SRF near the operational frequency range of the sensing system.
5.3.3. SiCN-based Pressure Sensor System Prototype Design # 2: Testing

The wireless pressure sensor was fabricated on an alumina substrate that has a dielectric constant of 9.9 and a thickness of 500 µm. A photograph of this sensor is shown in Figure 5.12. The values for $L_T$, $C_1$ and $C_2$ were 500 nH, 14 pF and 28 pF, respectively. $L_T$ is a 2-turn thin film spiral inductor that is also used to provide magnetic coupling for wireless signal transmission. $C_1$ and $C_2$ are titanate MIM capacitors. $C_1$ is a single chip capacitor with a value 14 pF and $C_2$ are two MIM capacitors in parallel to achieve the 28 pF value. $C_T$ is the SiCN capacitive pressure sensor.

Figure 5.12. Photograph of a SiCN-based wireless pressure sensor system.
The wireless pressure sensor was characterized with the HTPC and signal was transmitted to a 20 turn, 20 cm-diameter wire pickup coil over a distance of approximately 1m. The coil was connected to an Agilent E4440A Series Spectrum Analyzer where the transmission data was recorded. The wireless pressure sensor was characterized at 25°C and 400°C, over a pressure range from 0 to 100 psi in steps of 10 psi. Figure 5.13 is a plot of resonant frequency versus pressure at 25 and 400°C. At 25°C and 0 psi the resonant frequency of the wireless pressure sensor is 132.0 MHz and at 100 psi it is 132.6 MHz, equating to a resonant frequency shift of 600 kHz. At 400°C and 0 psi the resonant frequency is 131.3 MHz and at 100 psi it is 131.9 MHz, equating to a frequency shift of 600 kHz. Identical frequency shifts at the two temperatures indicate that the output of the wireless pressure sensor is not temperature sensitive. The slope of pressure vs. frequency plot at 25°C is $6 \times 10^{-3}$ MHz/psi and at 400°C it is $6 \times 10^{-3}$ MHz/psi, also indicating that wireless pressure is not temperature sensitive. The spectral response of the wireless pressure sensor at 400°C for 0 and 100 psi is shown in Fig. 5.14. At 0 psi the resonant frequency is 131.3 MHz with a peak received power of -38.5 dBm and at 100 psi the resonant frequency is 131.9 MHz with a peak received power of 38.2 dBm. These data show that the power output is not sensitive to temperature between 0 and 100 psi. The sensor system has a pressure sensitivity of 0.6 ($\Delta f/\Delta C$ MHz/psi).
Figure 5.13. Frequency versus pressure at 25 and 400°C for the SiCN wireless pressure sensor.
Figure 5.14. Spectral response at 400°C for the SiCN wireless pressure sensor for pressures of 0 and 100 psi.

5.4. Wireless Pressure Sensor System Prototype Design #3: SiCN-based Capacitive Pressure Sensor System with a Directional Chip Antenna

The data presented in the previous section clearly show that the wireless pressure sensor system based on the SiCN capacitive pressure sensor meets the target operating temperature of 400°C, however the total area of the system is 8 times larger than the targeted size of the proposed system. To address this issue, a third prototype wireless capacitive MEMS pressure sensor in which wirewound chip inductors are used to replace the large two-turn spiral inductor in Prototype #2 was developed. Eliminating the large,
two-turn inductor required that Prototype #3 be constructed with a directional chip antenna as the radiating element. Like all previous designs, the system is based on the Clapp-type oscillator design discussed previously. The directional antenna allows for maximum power transfer to the receiver along a proscribed direction. A matching network was added to the input of the chip antenna to shift its resonant frequency to that of the oscillator/pressure sensor.

5.4.1. Pressure Testing of the SiCN Capacitive Pressure Sensor

As with the development of Prototype #2, development of Prototype #3 was initiated by characterizing the SiCN pressure sensor as a discrete device prior to system assembly. The capacitive pressure sensor was characterized using the Agilent B1500A Semiconductor Device Analyzer (SDA) and a high temperature/pressure chamber (HTPC). The SDA was calibrated to the leads of the pressure sensor inside the HTPC at room temperature to remove the effects of the chamber and cabling. The pressure sensor capacitance was measured at 1 MHz at a temperature of 25°C and a pressure range of 0 to 100 psi. Figure 5.15 is a plot of capacitance versus pressure from this test. The pressure sensor was measured twice to examine repeatability. The sensor capacitance was 5.15 pF at 0 psi and decreased to 4.41 pF at 100 psi, resulting in a ∆C of 0.74 pF.
5.4.2. Temperature Testing of the Wirewound Inductors

A SiC MESFET is the active device used in this system and the values for $L_T$, $C_1$, and $C_2$ are 390 nH, 13 pF, and 40 pF, respectively. $L_T$ is a Johanson RF wirewound 390 nH inductor. The Johanson 390 nH wirewound inductor has dimensions of 2 x 1.2 x 1.2 mm$^3$ and is shown in Figure 5.16. As with the MIM capacitors, three methods were used to validate the response of the inductor over the desired frequency and temperature ranges.
The S-parameters of the 390 nH inductor were measured on the network analyser in the same manner as the capacitors. The circuit model is shown in Fig. 5.17. The model consists of a series inductor and resistor and two capacitors in shunt.

![Equivalent circuit model for a wirewound inductor.](image)
The S-parameters were optimized against the circuit model in ADS using the gradient optimizer in all the simulations. The optimization method using the wirewound inductors yielded results that were just as accurate in describing circuit behavior as the model developed for the MIM capacitors. As a result there was close agreement in the \( S_{11} \) and \( S_{21} \) parameters for the measured and model-optimized traces. The values for all the circuit model components are shown in Table 5.3. Just as with the capacitor model components, this method results in frequency-independent component values.

Table 5.3. Circuit model values for 390 wirewound inductor.

<table>
<thead>
<tr>
<th>Temp (°C)</th>
<th>Ls (nH)</th>
<th>Rs (Ω)</th>
<th>C1 (pF)</th>
<th>C2 (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>397.887</td>
<td>4.02</td>
<td>0.531</td>
<td>0.593</td>
</tr>
<tr>
<td>50</td>
<td>398.515</td>
<td>4.426</td>
<td>0.534</td>
<td>0.6</td>
</tr>
<tr>
<td>100</td>
<td>399.779</td>
<td>5.015</td>
<td>0.538</td>
<td>0.605</td>
</tr>
<tr>
<td>150</td>
<td>400.932</td>
<td>5.633</td>
<td>0.55</td>
<td>0.616</td>
</tr>
<tr>
<td>200</td>
<td>401.975</td>
<td>6.195</td>
<td>0.553</td>
<td>0.623</td>
</tr>
<tr>
<td>250</td>
<td>401.381</td>
<td>6.9</td>
<td>0.561</td>
<td>0.627</td>
</tr>
<tr>
<td>300</td>
<td>404.637</td>
<td>7.658</td>
<td>0.568</td>
<td>0.625</td>
</tr>
<tr>
<td>350</td>
<td>405.295</td>
<td>9.281</td>
<td>0.562</td>
<td>0.629</td>
</tr>
<tr>
<td>400</td>
<td>408.321</td>
<td>12.289</td>
<td>0.579</td>
<td>0.634</td>
</tr>
</tbody>
</table>

The value of the inductor, \( L_S \), increased by only 3% from 25 to 400°C, demonstrating its viability through this temperature range. However, the series resistance, \( R_S \), increases from 4.02 to 6.195 Ω from 25 to 200°C, which is an increase of 53%. Furthermore, \( R_S \) increased by nearly 100% from 200 to 400°C, thus suggesting that the material composition of the inductor is beginning to deteriorate and degrade its electrical performance. Slightly above 400°C, the inductor fails, and when taken up to 500°C, the physical damage is irreversible.
The inductors were also characterized using the SDA. The inductors were only characterized up to 400°C due to the information from the aforementioned model that the wirewound inductors would begin to fail above this temperature. Figure 5.18 is a plot of measured inductance versus temperature for the 390 nH inductor. The inductance of the wirewound inductor changes from 408 to 417 nH from 25 to 400°C, roughly 2.5%, over this temperature range. The series resistance was also characterized from 25 to 400°C at 1 MHz. Just as with the inductor characterized using the optimization modeling, the measured series parasitic resistance is nominally linear with a slope of 0.01Ω/°C up to 300°C and a slope of 0.02Ω/°C between 300 and 400°C, as shown in Fig. 5.19, indicating that the temperature is most likely approaching the maximum operating temperature of the inductor.

![Figure 5.18. Inductance versus temperature at 1 MHz for a wirewound inductor.](image)

135
Figure 5.19. Series parasitic resistance versus temperature at 1 MHz for the wirewound inductor.

Finally, the inductor was characterized at room temperature from 40 Hz to 110 MHz using an impedance analyzer. Figure 5.20 is a plot of measured inductance versus frequency for the inductor. From approximately 75 to 90 MHz the measurement is inaccurate due to an error in the calibration routine. However, from 90 to 110 MHz, the calibration was accurate and from the data trace it is evident that the inductor is operational over the design frequency range from 124 to 127 MHz.
Figure 5.20. Inductors response over the frequency range of 10 to 110 MHz

5.4.3. Development of a Directional Chip Antenna with Matching Network

A Johanson Technology ISM chip antenna (P/N: 0169AT62A0010) was used to transmit the RF output signal from the oscillator circuit with SiCN pressure sensor to a receive antenna 10 m away. The nominal resonant frequency of the chip antenna was 169 MHz when using the vendor-suggested circuit board and matching network at the input of the antenna. However, the resonant frequency of the antenna is highly dependent on several factors, such as antenna placement with respect to a ground plane (if used), matching network, and substrate/circuit board material. In this case, the antenna was epoxied to an alumina substrate with no ground plane. The chip antenna with a coplanar waveguide (CPW) feed line and matching network is shown in Fig. 5.21. The return loss ($S_{11}$) of the chip antenna with and without a matching network, shown in Fig. 5.22, was characterized
with the Agilent E8361C Precision Network Analyzer (PNA) and ground-signal-ground (GSG) probes to facilitate connection to the CPW feed lines.

Figure 5.21. Photograph of a chip antenna, matching network and CPW feed line.

![Chip Antenna, Matching Network, and CPW Feedline](image)

Figure 5.22. Return loss of the chip antenna with and without matching network.

![Return Loss Graph](image)
The data in Fig. 5.22 show that since the return loss of the chip antenna has a resonant frequency of 160 MHz, a matching network must be added to the circuit. The matching network designed for this application must decrease the resonant frequency of the chip antenna to that of the original design frequency of the circuit as well as increase the operational bandwidth to accommodate the change in oscillator/pressure sensor frequency due to the change in pressure. The matching network consisted of two 390 nH wire wound chip inductors placed in the shunt at the input to the antenna, which resulted in an equivalent inductance of 195 nH. Accordingly, the reflection coefficient of the chip antenna was at least -5 dB over the operational bandwidth of 125 to 145 MHz.

5.4.4: SiCN-based Capacitive Pressure Sensor System with Directional Chip Antenna Prototype Design #3: Testing

A circuit schematic of the wireless pressure sensor is shown in Fig. 5.23. The DC bias circuits are represented by \( L_d, C_d, \) and \( R_g \). These components are off chip and held at room temperature. The values of \( L_d, C_d, \) and \( R_g \) were 2.5 mH, 1.5 \( \mu \)F, and 5 k\( \Omega \), respectively. At room temperature, the active pressure sensor system was designed to operate at 127 MHz at 0 psi and 130 MHz at 100 psi based on the previously-measured change in capacitance of the SiCN pressure sensor over that pressure range. Although the design frequency was chosen based on currently available chip capacitors and inductors, it can easily be modified for other design frequencies. A photograph of the pressure sensor system with directional chip antenna is shown in Fig. 5.24.
Figure 5.23. Circuit schematic of a Clapp-type oscillator with DC bias circuits.
Figure 5.24. Photograph of a SiCN-based wireless pressure sensor with directional chip antenna.

The chip antenna and matching network were added to the output of the oscillator/pressure sensor via gold wire bonds, as seen in Fig. 5.22. All components were mounted onto alumina substrates using high temperature silver epoxy. The areas of the pressure sensor, oscillator and chip antenna were 15x5 mm², 7x7 mm², 25x5 mm², respectively which, if optimized, can be realized in a footprint of approximately 250 mm². When compared to [Scardelletti 2014], with an approximate area of 2,580 mm², the percentage reduction of area is nearly 96%. For a room temperature, unpressurized test of the system, the sensor was mounted on a Styrofoam block and placed on a rotary stage.
The test was performed in a typical laboratory environment with lab benches and cabinets that acted as reflectors. The signal was transmitted to a Pixel broadband loop antenna over a distance of 10 m. A horizontal and vertical cut of the radiation patterns for the wireless pressure sensing system are presented in Fig. 5.25 and show the system to be directional with a maximum power of -23 dBm at 25°C.

The high temperature/pressure chamber shown in Fig. 5.1 was used to heat and pressurize the wireless pressure sensor up to 300°C and 100 psi, respectively. Because the pressure chamber is metal with a small glass window, far field antenna measurements could not be performed. Instead, a dipole antenna was used as the receive antenna and was placed 1 m above the glass window.
The wireless pressure sensing system was characterized from 25 to 300°C, over a pressure range of 0 to 100 psi in steps of 10 psi, with an Agilent E4440A Series Spectrum Analyzer serving as the receiver. Figure 5.26 plots the resonant frequency versus pressure for temperatures of 25, 100, 200 and 300°C. The output frequency with respect to temperature over the entire pressure range is linear. The sensitivity of the sensor (Δf/ΔP MHz/psi) is essentially independent of temperature, with values of 2.5, 2.6, 2.8 and 2.8 % for 25, 100, 200, and 300°C, respectively.
Figure 5.26. Wireless pressure sensor Frequency versus pressure at 25, 100, 200, and 300°C for the SiCN wireless pressure sensor with directional chip antenna.

Output power versus frequency spectra were recorded at 300°C for pressures of 0, 50 and 100 psi and are shown in Fig. 5.27. The peak power received at 0, 50 and 100 psi is -19.85, -17.87 and -15.95 dBm, respectively. This distribution equates to an increase in received power of 3.9 dBm over this pressure range. The increase in power is due to an increase in the receive antenna gain at this frequency range. The frequency at 0, 50 and 100 psi is 124.66, 126.12 and 127.49 MHz, respectively. This equates to a frequency shift of 2.83 MHz over the pressure range at 300°C. The received power at 0, 50, and 100 psi from 25 to 300°C is shown in Fig. 5.28. The slopes of measured power at 0, 50 and 100 psi are $1.9 \times 10^{-3}$, $1.8 \times 10^{-3}$ and $1.8 \times 10^{-3}$ dBm/°C, respectively, quantities that are virtually the
same indicating that a change in temperature does not affect the response of the wireless system. However the decrease in power with increasing temperature does show that this wireless pressure sensor with integrated chip antenna is effected by temperature, due primarily to a decreasing chip antenna gain due to increasing losses as the temperature increases.

The phase noise of the wireless pressure sensor at 0 psi for temperatures of 25, 100, 200, and 300°C is shown in Fig. 5.29. The data show that the 1 kHz offset is below -30 dBc/Hz and the 10 kHz offset is below -80 dBc/Hz. The phase noise slope is -20 dBc/decade above 10 kHz. The two spurs evident at 2 kHz are likely due to the heater power supply fluctuating to maintain temperature.

![Graph showing output power versus frequency at 300°C for the wireless SiCN pressure sensor with directional chip antenna at 0, 50 and 100 psi.](image)

Figure 5.27. Output power versus frequency at 300°C for the wireless SiCN pressure sensor with directional chip antenna at 0, 50 and 100 psi.
Figure 5.28. Measured output power versus temperature for the wireless SiCN pressure sensor with directional chip antenna at 0, 50, and 100 psi.
5.5. Summary

In Chapter 4 a wireless high temperature oscillator circuit that was operational up to 470°C was described. In this chapter, several designs that integrate pressure sensor capabilities with the high temperature wireless circuits was described. The first prototype, a wireless pressure sensing system utilizing a polysilicon capacitive pressure sensor was developed. The pressure sensor system could operate reliably up to 300°C, over a pressure range from 0 to 45 psi. The polysilicon-based system was unable to achieve an operational temperature of 400°C mainly due to polysilicon becoming very lossy at temperatures above 300°C.
275°C causing the gain of the MESFET to decrease until oscillation was no longer achievable.

A second wireless pressure sensor prototype was developed that incorporated a SiCN capacitive pressure sensor and titanate MIM capacitors. This SiCN-based pressure sensor system was characterized up to 400°C from 0 to 100 psi. The pressure sensing system operated at 131.3 MHz at 0 psi and 131.9 MHz at 100 psi and had a sensor system sensitivity of 0.6 MHz/psi. The second prototype reached the operational temperature goal of 400°C, however the overall area of the system was about 8 times too large. Therefore a third wireless prototype was developed. To reduce the overall size of the pressure sensor system wirewound inductors replaced the 2-turn spiral inductor and a directional chip antenna was added to provide wireless capabilities. The wireless sensor system operated successfully at 300 from 0 to 100 psi. If the position of the components on the alumina substrate were arranged in an optimized pattern, the percentage reduction of area would be nearly 90% of that from wireless pressure sensor system Prototype #2.
Chapter 6

Development and Qualification Testing of a Packaged Pressure Sensor System

6.1. Introduction

Based on the knowledge gained in developing the prototypes described previously, a packaged pressure sensor system was designed and fabricated to specifically measure pressure on a turbofan engine. The sensor system is based on the Clapp-type oscillator design that included components that demonstrated reliable function at 400°C. These components include a SiC-based MESFET, titanate MIM capacitors, wirewound inductors, high temperature thick film resistors and a MEMS-based SiCN capacitive pressure sensor.
This system must also incorporate the DC bias circuits that were used in all previous prototypes but intentionally kept off the substrate to facilitate oscillator development. The entire unpackaged sensor system must be no larger than 8 x 40 x 4 mm³. Based on the findings from the previous prototypes, a wireless version of this sensor would not fit within this area, so this prototype must implement wired connections for power and signal transmission. The sensor system is to be packaged in a custom-fabricated metal housing that is able to pass a standard acceptance qualification tests at a temperature of 500°C for 1 hour, pressures of 0 to 300 psi and vibrations of up to 5.3 G rms along the x-, y-, and z-axis for 20 minutes on each axis. The package will be designed for installation on the engine by way of a borescope plug adaptor fitted to the borescope port of the engine so that the sensor element can be exposed directly to the gas path of the engine. Such a design will make the sensor suitable for on-engine testing, which can involve normal engine operation with adjustments to the fan speed to expose the sensor to steady-state and transient conditions. This dissertation aims to create a sensor system that passes the standard benchtop acceptance tests and thus could be installed on a gas turbofan engine but does not include testing on an actual engine.

### 6.2. High Temperature Testing of Thick Film Resistors

All components for the sensor system were previously characterized at temperature except for the thick film resistors. The resistors considered for use in this prototype are 10 kΩ thick film chip resistors from MiniSystems Inc. These ruthenium alloy chip resistors have dimensions of 1.118 x 0.559 x 0.330 mm³. These resistors have voltage and power ratings of 40 V and 0.04 W, respectively. A photograph of a representative resistor is
presented in Fig. 6.1.

A 10 kΩ resistor is used in the DC bias circuit of the gate side of the SiC Cree MESFET to simplify and reduce the overall size, while maintaining the ability to prevent RF from leaking back into the gate power supply. Since the gate of the FET requires no current, only RF blocking is required and the 10 kΩ resistor is sufficiently large.

![GSG feed lines](image)

Figure 6.1. Microphotograph of a thick film resistor on a metal test fixture.

The resistance equivalent circuit model used to optimize against the measured S-parameters of the resistor was the same circuit model used to evaluate the wirewound
inductors. A schematic diagram of this model was shown previously in Fig. 5.17. The S-
parameters were optimized from 10 to 200 MHz, and the gradient optimizer was used for
all simulations. The results of these simulation, which are presented in Table 6.1, indicate
that the change in resistance from 25 to 400°C is negligible. Also, the parasitics associated
with $L_s$, $C_1$, and $C_2$ are negligible as well.

Table 6.1. Circuit model values for 10 kΩ chip resistor.

<table>
<thead>
<tr>
<th>Temp (°C)</th>
<th>Rs (kΩ)</th>
<th>$L_s$ (nH)</th>
<th>$C_1$ (pF)</th>
<th>$C_2$ (pF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>10.13</td>
<td>1.00E-05</td>
<td>0.578</td>
<td>0.556</td>
</tr>
<tr>
<td>50</td>
<td>10.14</td>
<td>1.00E-05</td>
<td>0.578</td>
<td>0.559</td>
</tr>
<tr>
<td>100</td>
<td>10.07</td>
<td>1.00E-05</td>
<td>0.582</td>
<td>0.561</td>
</tr>
<tr>
<td>150</td>
<td>10.06</td>
<td>1.00E-05</td>
<td>0.588</td>
<td>0.569</td>
</tr>
<tr>
<td>200</td>
<td>10.01</td>
<td>1.00E-05</td>
<td>0.597</td>
<td>0.574</td>
</tr>
<tr>
<td>250</td>
<td>10.08</td>
<td>1.00E-05</td>
<td>0.599</td>
<td>0.578</td>
</tr>
<tr>
<td>300</td>
<td>10.08</td>
<td>1.00E-05</td>
<td>0.606</td>
<td>0.579</td>
</tr>
<tr>
<td>350</td>
<td>10.12</td>
<td>1.00E-05</td>
<td>0.612</td>
<td>0.581</td>
</tr>
<tr>
<td>400</td>
<td>10.12</td>
<td>1.00E-05</td>
<td>0.596</td>
<td>0.579</td>
</tr>
</tbody>
</table>

The resistor was characterized by measuring its resistance from 25 to 500°C using
an Agilent 6.5 digit multimeter, DC needle probes, and the high temperature probe station
described previously. The results of this test are shown in Fig. 6.2. At 25°C, the resistance
is 10.1 kΩ and at 500°C the resistance is 10.7 kΩ, which is a 6% increase in resistance over
the temperature range. Furthermore when the resistor is cooled to room temperature the
resistance returns to its original room temperature value of 10.1 kΩ indicating that this
resistor was not damaged when exposed to temperatures as high as 500°C is thus well
suited for the proposed sensor system.
6.3. Pressure Testing of the SiCN Capacitive Pressure Sensor

The SiCN MEMS capacitive pressure sensor detailed previously was used in this sensor system. Since each particular pressure sensor exhibits slightly different performance values and the variation in these values was not available at the time of this dissertation, it was determined that each SiCN sensor should be tested individually prior to system assembly in order to accommodate for any unexpectedly large differences with respect to previously tested sensors. As such, the pressure sensor used in this prototype was characterized with the Agilent SDA and the high temperature/pressure chamber (HTPC) described previously. The sensor was loaded into the HTPC and the SDA was calibrated to the leads of the pressure sensor at room temperature to remove the effects of the chamber and cabling. The pressure sensor capacitance was measured at 1 MHz from 0 to 100 psi at
$25^\circ\text{C}$, and the results are shown in Fig. 6.3. Over this pressure range, the capacitance changes linearly from 3.84 to 3.3 pF, which is a $\Delta C$ of 0.54 pF. The slope of the capacitance versus pressure line was $5.4 \times 10^{-4}$ pF/psi.

![Graph: Pressure versus capacitance](image)

Figure 6.3. Pressure versus capacitance for the SiCN pressure transducer used in the wired packaged system.

The SiCN pressure sensor was also characterized from 40 Hz to 110 MHz at atmospheric pressure and room temperature with an impedance analyzer, and the results are shown in Fig. 6.4. The response indicates that the pressure sensor capacitance at room temperature does not vary significantly with input frequency and there is no indication of the self-resonant frequency over the operational frequency of the pressure sensor system. The roughness in the trace is mainly due to the calibration.
Figure 6.4. Capacitance versus frequency for the SiCN capacitive pressure sensor from 40 Hz to 110 MHz at 0 psi.

6.4. Simulated Response of the Packaged System

Prior to device assembly, the system was simulated using ADS. The model system, shown schematically in Fig. 6.5, consists of a Clapp-type oscillator with a MEMS capacitive pressure sensor located in the LC tank circuit of the device. To simulate the pressure-induced response of the system, the capacitance of the sensor, $C_T$, is varied and the resonant frequency of the system is calculated. The model showed that as pressure was increased from 0 to 100 psi, the frequency of the pressure sensor system increased from 97 to 103 MHz, which is an increase of 6%. To illustrate this response, a harmonic balance
simulation was performed and the results are shown in Figure. 6.6. The results indicate that when the capacitance of the pressure sensor is 3.84 pF (0 psi), 3.6 pF (50 psi) and 3.3 pF (100 psi), the resonant frequency is approximately 96.7 MHz, 99.2 MHz and 102.8 MHz, respectively.

Figure 6.5. Circuit schematic of the Clapp-type oscillator used in the packaged SiCN pressure sensor.
Figure 6.6. Simulated resonant capacitance versus frequency with respect to changes in pressure.

6.5. Fabrication and Assembly of Packaged Pressure Sensor System

The sensor system was fabricated on an alumina substrate and is shown in the photograph of Fig. 6.7. Figure 6.8 presents a photograph of the stainless steel metal packaging fixture with unpackaged sensor. The Clapp-type oscillator (without the pressure sensor) was fabricated on an alumina substrate that is 3.5 mm wide by 18 mm long by 500 μm thick. DC bias circuitry was added at the gate and drain inputs and consists of a series
10 kΩ resistor on the gate, two 90 pF MIM capacitors in the shunt and a 390 nH wirewound inductor on the drain, as seen in Fig. 6.7. The sensor and oscillator are both mounted to another alumina substrate, a 6 x 35 mm² support substrate, using epoxy. Wire bonds were used to electrically connect the sensor to the oscillator board as well as to make electrical connection to the support substrate. Mineral wire was used to make electrical contact to the support substrate from the metal package. The stainless steel package was 150 mm by 30 mm and is shown in the Fig. 6.8. A cylindrical sleeve that fits over the pressure sensor circuit has an inner diameter of 4 mm.

![Photograph of assembled SiCN pressure sensor system](image)

Figure 6.7. Photograph of an assembled SiCN pressure sensor system on alumina substrate.
6.6. Pressure and Temperature Testing of the Packaged System

The pressure sensor system was characterized in a three-inch tube furnace. To enable pressure sensing at temperature, a custom in-house fixture was developed and is shown in Figure 6.9. The packaged sensor was attached to a test fixture via a Swagelok fitting so the system could be pressurized while in the tube furnace. A thermocouple was positioned inside the test fixture to report the temperature at the tip of the pressure sensor system, which, during on-engine testing, would be located on the outer housing of the engine. Another thermocouple was placed on the outside of sleeve to determine the temperature as close to the electronics on the alumina substrate as is feasibly possible. The packaged pressure sensor was characterized with an Agilent E440A spectrum analyzer at room temperature at pressures of 0, 100, 200, 300 and 350 psi. Figure 6.10 presents the output power versus frequency for the pressure sensor at the five test pressures.

Figure 6.8. Photograph of the packaging for the wired SiCN pressure sensor system.
Figure 6.9. Photograph of the pressure system characterization fixture.
Figure 6.10. Output power versus frequency for the packaged pressure sensor system at 0, 100, 200, 300 and 350 psi.

As shown in Fig. 6.10, the packaged pressure sensor resonates at 96.88, 102.79, 109.54, 116.77 and 119.862 MHz for applied pressures of 0, 50, 100, 200, 300 and 350 psi, respectively. Notice that the simulated sensor response presented previously in Fig. 6.6 and the measured response in Fig. 6.10 for pressures of 0 and 100 psi occur at virtually identical frequencies, verifying the accuracy of the simulation model. Based on this finding, ADS was used to determine the capacitance of the pressure sensor at 200, 300 and 350 psi. Capacitance values of 2.83, 2.43 and 2.30 pF were used to represent the pressure sensor capacitance, $C_T$ at these pressures. Using these values, simulations were performed, resulting in simulated pressure-induced frequency responses that closely match the measured resonant responses at 109.54, 116.77 and 119.86 MHz. A plot of the simulated
capacitance as a function of pressure is shown in Figure. 6.11. The slope of the response is $5.2 \times 10^{-3}$ pF/psi for pressures from 0 to 300 psi, but decreases to $2.6 \times 10^{-3}$ pF/psi for pressures between 300 and 350 psi, suggesting that the capacitive pressure may be operating outside of its linear region. Thus, a second order linear regression model is better suited to represent the entire pressure range. The frequency as a function of pressure for the packaged pressure sensor system is shown in Figure. 6.12. The response is linear over the operational pressure and frequency range with a slope of $6.74 \times 10^{-2}$ MHz/psi.

![Graph of simulated capacitance versus pressure](image)

Figure 6.11. Simulated capacitance versus pressure for the packaged SiCN pressure sensor system.
Figure 6.12. Simulated frequency response versus pressure using calculated capacitive pressure sensor values for the packaged SiCN sensor system.

The packaged pressure sensor was mounted to the test fixture and placed in the 3 inch tube furnace as shown in Fig. 6.13. To emulate the test conditions on an actual turbofan engine, the tube furnace was heated to over 500°C and the pressure was increased from 0 to over 300 psi. The frequency response of the sensor under these testing conditions is shown in Fig. 6.14. The temperature recorded by the thermocouple on the sleeve of the metal sensor packaging was approximately 400°C, which is a better indicator of the circuit temperature than the temperature of the furnace tube. Nevertheless, a significant portion of the package was at the temperature of the tube furnace. To achieve oscillation of the pressure sensor system, the gate bias was -9 V, the drain bias was 8 V, and the drain current
was 100 mA. At 540°C, the sensor system exhibited a frequency shift from 96.3 to 117.8 MHz for pressures of 0 and 320 psi, which translates to a 20% change in resonant frequency and a sensor system sensitivity (Δf/ΔP MHz/psi) of 6.8 x 10^{-2}. It is worth mentioning that the resonant frequencies of the sensor system at room temperature and 540°C (~ 400°C at the sleeve) differ by less than 1%, indicating that the sensor system, specifically the electronics and pressure sensor, are insensitive to changes in temperature.

Figure 6.13. Photograph of the test fixture and attached packaged sensor positioned inside tube furnace in preparation for a temperature/pressure test.
Figure 6.14. Output power versus frequency at 540°C for a packaged SiCN pressure sensor system at 0 and 320 psi.

6.7. Structural Dynamic Testing of the Packaged Pressure Sensor System

Structural dynamic tests were performed to evaluate performance when the system is exposed to the expected vibration environment on a turbofan engine. To conduct the vibration tests, the packaged pressure sensor was connected to a borescope sensor interface adaptor (which is used for on-engine testing) which was attached to a 9 inch aluminum plate on a vibration table. Two control accelerometers were mounted on opposite side corners of the 9 square inch fixture plate for x axis and y axis vibration testing and above and below the test article for z axis testing. Three response accelerometers were located on
the sensor shaft top to provide vibration level data of the sensor. The responses were oriented in each direction of motion. In addition, a reference accelerometer in the direction of motion was mounted to the fixture plate to record any fixture resonances. The testing apparatus is shown in Fig. 6.15.

The acceptance test consisted of a 5 minute, ¼ g sinusoidal sweep before and after a 20 minute random vibration exposure that subjected the system to a maximum vibration of 5.3 G_{rms}. The sinusoidal sweeps were conducted before and after each random vibration level run to see if any significant change in resonances were detected. The sinusoidal level utilized a 0.25 g peak at frequencies from 5 to 2000 Hz at a sweep rate of 2 octaves/minute. The random level sweep utilized an amplitude of 5.3 G_{rms}, and frequencies from 15 to 2000 Hz. Both the sinusoidal and random vibration tests were performed along the x-, y- and z-axes. For brevity only one sinusoidal and one random vibration test profile are presented in this dissertation. They are presented in Figures 6.16 and 6.17, respectively.

Figure 6.15. Photograph of the vibration test setup.
Figure 6.16. Acceleration versus frequency for a $1/4$ g sinusoidal sweep profile.

Figure 6.17. Acceleration versus frequency for a $5.3 \text{ G}_{\text{rms}}$ random vibration profile.
Visual inspections before and after each of the vibration exposures in each of the three different axes were performed as well as a continuous monitoring of the sensor output during each test sequence by manually observing the output frequency spectra. No failures or anomalies were observed in the sensor output during the vibration tests and no damage was observed during the post-test visual inspection.

In addition to the aforementioned tests, the resonant frequency of the packaged pressure sensor system was recorded at room temperature and 0 psi at the beginning and end of each axis test. No change in resonant frequency was observed for each of the three axes. The sensor was then evaluated at temperature and pressure following the procedures detailed previously. Representative plots of output power (represented as magnitude) versus frequency for the sensor at 25°C and 0 psi, 540°C and 0 psi and 54°C0C and 300 psi after vibration testing are shown in Figure. 6.18. This behavior is indistinguishable from the same tests performed before the vibration test (see Fig. 6.11) to within the output uncertainty of the sensor system. In specific, the resonant frequency at 25°C and 0 psi before and after a vibration test were 96.89 and 97.35 MHz, respectively. From these data, the percent change in resonant frequency before and after exposure was 0.5%. With respect to temperature, the resonant frequencies at 25 and 520°C are 97.35 and 96.59 MHz which is a difference of 0.8%. Table 6.2 summarizes the measured spectrum frequencies of packaged pressure sensor system before and after the vibration testing.
Figure 6.18. Output power versus frequency for the packaged SiCN pressure sensor after vibration testing.

Table 6.2. Tabulated spectrum values of packaged pressure sensor system measurements taken before and after vibe testing

<table>
<thead>
<tr>
<th>No. #</th>
<th>Figure</th>
<th>Vibration Test</th>
<th>P (psi)</th>
<th>T (°C)</th>
<th>f (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>6.11</td>
<td>Before</td>
<td>0</td>
<td>25</td>
<td>96.89</td>
</tr>
<tr>
<td>2</td>
<td>6.15</td>
<td>Before</td>
<td>0</td>
<td>540</td>
<td>96.24</td>
</tr>
<tr>
<td>3</td>
<td>6.15</td>
<td>Before</td>
<td>320</td>
<td>540</td>
<td>117.8</td>
</tr>
<tr>
<td>4</td>
<td>6.17</td>
<td>After</td>
<td>0</td>
<td>25</td>
<td>97.35</td>
</tr>
<tr>
<td>5</td>
<td>6.17</td>
<td>After</td>
<td>0</td>
<td>520</td>
<td>96.59</td>
</tr>
<tr>
<td>6</td>
<td>6.17</td>
<td>After</td>
<td>342</td>
<td>520</td>
<td>118.13</td>
</tr>
</tbody>
</table>
6.8. Summary of the Results

The wired pressure sensor prototype described in this chapter successfully met the design and performance requirements in several key areas that must be met for eventual evaluation of the sensor in an on-engine test. In terms of form factor, all the required electronic components, including the Clapp-type oscillator circuit, capacitive pressure sensor and DC biasing circuits were able to be incorporated into a high temperature substrate that fits inside a conventional borescope adapter. Transmission of power to and data from the sensor was successfully accomplished through a robust, wired connection. In terms of performance, the packaged sensor successfully passed a standard benchtop test required of packaged prototypes designed for testing on a flight-worthy turbofan engine. These tests included stable pressure transduction at temperatures as high as 500°C at the borescope adaptor tip, pressures up to 300 psi across the entire temperature range, and vibration tests that involve sinusoidal and random signals with amplitudes as high as 5.3 G\text{gms}. The packaged sensor demonstrated acceptable performance throughout this rigorous battery of tests with no degradation of output signal and no signs of mechanical damage. By passing these tests, the packaged sensor is eligible for future evaluation on a flight-worthy turbofan engine. In summary, this packaged prototype satisfies the objectives of this dissertation research.
This dissertation describes the design, modeling, fabrication and testing of MEMS-based pressure sensor systems for harsh environment applications, specifically the high temperature, high pressure and high vibration environments associated with gas turbofan engines. Of particular focus for this dissertation is an operational temperature of 400°C. The electronic subsystems were based on a Clapp-type oscillator circuit with the pressure sensor located in the LC resonate circuit of the oscillator. The Clapp-type oscillator design was selected because it requires significantly fewer passive components than other oscillator designs. Under conditions characteristic of a gas turbine engine (i.e., high temperature and high vibration), fewer environmentally-sensitive components reduces the probability of system failure. The oscillators consisted of a 6H-SiC MESFET, two
capacitors, one inductor and the capacitive pressure sensor. Because it requires fewer components than alternative designs such as the Colpitts oscillator, the Clapp design can be made into systems of smaller form factor, enabling deployment in confined locations. Another advantage of the Clapp oscillator design is that the inductor ($L_T$) and capacitive pressure sensor ($C_T$) are in series, thus enabling $C_T$ to be used to independently set the operational frequency, while $C_1$ and $C_2$ can be used to independently control the gain conditions. This arrangement improves the frequency stability of the circuit, making the Clapp-type oscillator a better option than other oscillator designs for high temperature applications.

Realization of the proposed sensor system required the development of and/or evaluation of substrates, passive components (capacitors and inductors), active components (MESFETs) and pressure sensors for stable operation at 400°C. Alumina was selected as the substrate material owing to its dielectric properties, compatibility with photolithographic patterning, machinability and low cost. For the capacitors, SiC was investigated as a potential high temperature thin film dielectric. The amorphous SiC films, deposited by plasma enhanced chemical vapor deposition, exhibited suitable and stable dielectric properties when used in thin film MIM capacitors for temperatures up to ~300°C; however, at higher temperatures the capacitance varied strongly with temperature and therefore the SiC-based MIMs were only used in the prototypes that were limited to operating temperatures below 300°C. As an alternative, titanate capacitors were investigated. Originating from a commercial source and therefore the chemical composition was not known, the titanate capacitors exhibited stable operation at
temperatures up to 500°C and therefore were used in the high temperature prototypes. For the inductors, thin film spiral and wirewound inductors were both considered. The thin film inductors exhibited less than a 2% change in inductance up to 500°C while the wirewound chip inductors exhibited a 2% change up to 400°C. At 500°C, the wirewound inductors experienced irreversible damage. The thin film spiral inductors are much larger in area than the wirewound devices, therefore in the prototypes where form factor was not a concern, the thin film inductors were used, whereas when form factor was critically important, wirewound inductors were used. For some wireless prototypes, the thin film inductors could also be used as the radiating element. For the resistors, a thick-film chip resistor was investigated. The resistor exhibited a 6% increase in resistance between room temperature and 500°C. The two active components, a 6H-SiC MESFET and a SiCN capacitive pressure sensor were both acquired from commercial sources. These components were selected in large part because of their demonstrated performance at high temperature (and in the case of the pressure sensor, the maximum detectable pressure), size, and compatibility with chip-based assembly techniques.

Several oscillator designs based on the Clapp-type architecture were explored. A SiC-based MESFET was used as the active device in all the designs. The MESFET is used to facilitate oscillation in the feedback loop while $L_T$ and $C_T$ determine the operational frequency. The first prototype oscillator was designed at 1 GHz and was operational from 25 to 270°C. An on-chip slot-ring antenna was used to transmit the oscillator output signal to a receive antenna a distance of 1 m. In this prototype, DC bias circuits for the gate and drain were located off chip. Over a temperature range of 25 to 245°C the power decreased
by 18 dB and at 270°C the transconductance (gm) of the MESFET was no longer sufficient to offset the losses in the circuit and the oscillator was no longer able to achieve oscillation. The slot ring antenna and its inherently large area ultimately limited the ability for this prototype to achieve a small form factor.

The second prototype oscillator design incorporated on chip SiC MIM capacitors and chip antennas which were selected to reduce the overall size of the circuit relative to the slot antenna based prototype. Like the first prototype, the DC bias circuits were located off chip. Two oscillator designs were investigated, differing by their resonant frequency (720 and 940 MHz). High temperature testing revealed that the 720 MHz design was operational from 25 to 200°C, while the 940 MHz design was operational up to 250°C. The wireless capability of these oscillators was characterized using a receive antenna at a distance of 1.7 m. Radiation patterns for the two designs were essentially invariant with respect to temperature in terms of their angular distribution.

The gain of the SiC MESFET is affected by two important factors, the resonant frequency of the oscillator circuit and the operating temperature of the circuit. For the first two prototypes, resonances were at frequencies where the MESFET gain was relatively low. When superposed onto a declining MESFET gain with increasing temperature, the maximum operating temperature of the first two prototypes was ultimately capped by the gain of the MESFET, which below a certain level results in an oscillator that fails to resonate. Therefore to increase operational temperature two oscillators at 30 and 90 MHz were designed. In order to increase the operating temperature of the Clapp-type resonators,
a third prototype circuit that resonates at a substantially lower resonant frequencies (30 MHz and 90 MHz) were developed. This prototype utilized capacitors and inductors with higher values than the previous designs. The inductors were also able to be used as the radiating element due to their significant increase in size. SiC MIM capacitors were used and the DC bias circuits where located off the substrate. The 30 and 90 MHz oscillators functioned at temperatures up to 450°C and 470°C, respectively. The output power of the 30 MHz design decreased by only 5 dBm at 400°C but then decreased significantly until the circuit failed at 450°C. The resonant frequency shifted by less than 0.5% over the entire temperature range. The output power of the 90 MHz design decreased by only 5 dBm to 375°C, but above 375°C the output power decreased rapidly until the oscillator failed at 470°C. The oscillation frequency varied by only 2% over the entire temperature range.

Three wireless pressure sensor prototypes were designed and characterized. The first prototype used a Clapp-type oscillator circuit, SiC MIM capacitors, two-turn spiral inductors and a polysilicon MEMS-based capacitive pressure sensor. To characterize the performance of this design, a high temperature pressure chamber (HTPC) with quartz sight glass was developed for testing sensors at temperatures between 25 and 500°C and pressures between 0 and 100 psi. The HTPC was equipped with four thermocouples to accurately monitor the temperature and four feedthroughs to facilitate electrical power to devices under pressure. At 25 and 300°C, the sensor system exhibited sensitivities of 38 x 10^{-3} MHz/psi and 60 x 10^{-3} MHz/psi, respectively. Between 25 and 300°C at 0 psi the resonant frequency changed by only 2%. Although these are attractive performance parameters, the sensor system had a maximum operating temperature of ~300°C due in
part to losses associated with the polysilicon capacitive pressure sensor.

A second wireless pressure sensor prototype that utilized the SiCN pressure sensor and titanate MIM capacitors was demonstrated. At 25°C, for pressures of 0 psi and 100 psi, the measured shift in resonant frequency was 600 kHz. At 400°C, the resonant frequency shift over the same pressure difference was also 600 kHz, indicating that this prototype is essentially insensitive to temperature. The sensitivity of this prototype at both 25°C and 300°C is 6 x 10⁻³ MHz/psi. Additionally, power spectrum analysis showed that the output power of the system is independent of pressure.

A third wireless pressure sensor prototype based on the 2nd sensor prototype but incorporating a compact, directional chip antenna was investigated. This design was developed to explore the possibility of creating a high temperature wireless sensor with small form factor. This wireless prototype exhibited a linear frequency versus pressure response for pressures between 0 and 100 psi and temperatures between 25 and 300°C. The antenna enables highly directional transmission over a distance of 10 m. The sensitivity of the wireless pressure sensor (Δf/ΔP MHz/psi) was essentially independent of temperature at roughly 2.5 x 10⁻² and 2.8 x 10⁻² MHz/psi at 25 and 300°C, respectively. Furthermore, the phase noise of the wireless system was -30 dBC/Hz at the 1 kHz offset and decreased to -80 dBC/Hz at the 10 kHz offset frequency, indicating the oscillator performs well over the entire temperature range.

A wired, packaged pressure sensor system was developed specifically to measure
pressures on a turbofan engine. The sensor is based on the Clapp-type oscillator design used in the third sensor prototype which includes the SiC-based MESFET, titanate MIM capacitors, wirewound inductors, high temperature thick-film chip resistors, and a SiCN capacitive pressure sensor all integrated on a single substrate but absent of any antenna. The sensor system is packaged in a custom-fabricated metal housing that enables mounting directly to a functional engine by way of a borescope plug adaptor fitted to a borescope port on the engine. The packaged sensor system exhibits negligible change in frequency between 25 and 540°C (~ 400°C at the electronics) indicating that the system is insensitive to temperature. The packaged system has a 20% change in resonant frequency for pressures between 0 and 320 psi and the sensing system has a sensitivity of 6.8% over the operational pressure and frequency range. The packaged prototype was able to pass standard acceptance qualification testing at a maximum temperature, pressure and vibration of 515°C, 295 psi and 5.3 Grms, respectively. By passing these tests, the packaged sensor is eligible for future evaluation on a flight-worthy turbofan engine.

The major achievements of this dissertation in the area of high temperature pressure sensing include:

1. Development of thin film SiC-based MIM capacitors for use at temperatures up to 500°C and stable operation up to 300°C.
2. Demonstration of titanate MIM capacitors for use at temperatures up to 500°C.
3. Development of thin film spiral inductors for use at temperatures up to 475°C.
4. First to implement a SiC capacitive pressure sensor in a Clapp-type oscillator
design for use at any temperature, in particular at high temperature.

5. Development of a wireless, Clapp-type oscillator utilizing MIM capacitors and thin film spiral inductors for high temperature applications up to 470°C.

6. Development of a wireless pressure sensor system utilizing a Clapp-type oscillator with integrated chip antenna that operate up to 300°C.

7. Development of the first packaged, wired SiC-based pressure sensor system that meets the baseline acceptance qualification tests for pressure sensing on a functional aircraft engine.
Appendix A

Power Scavenging Technique Utilizing Thermoelectric Generators (TEGs)

In order to provide partial DC bias to the wireless pressure sensor Prototype #1, a thermoelectric power scavenging technique was developed. A photograph of this system is shown in Figure. A.1. The high temperature thermoelectric generators (TEG) were composed of silicon germanium. TEGs generate electrical power using a temperature gradient; the greater the temperature difference, the greater the power output. The electrodes were composed of copper on the cold end of the device and nickel foil on the hot end. A heat source was used to warm the hot end of one TEG to 300°C and a cold source chilled the cold end to 19°C, for a temperature differential of 281°C. The transistor
drain voltage, VDS, was held constant at 10 V with 7 V supplied by a DC power supply and the 3 V from the TEGs resulting in 30% of the power being scavenged. The drain current, ID, was held constant at 100 mA and the gate voltage was held at approximately 8.3 V. 7 V supplied by a DC power supply and the 3 V from the TEGs.

TEGs were used also used to generate power in wireless pressure sensor Prototype #2 just as in (G. W. Hunter, 2010) to demonstrate power scavenging. A heat source was used to warm the hot end of the six TEGs to 185°C and a cold source chilled the cold end to 19°C, for a temperature differential of 166°C. The transistor drain voltage, VDS, was held constant at 10 V and the drain current, ID, was held constant at 90 mA. Approximately 2.7 V was supplied by a DC power supply to the drain and the rest from the TEGs which is over 70%.

Figure A.1. Photographs of the thermoelectric power scavenging measurement apparatus.
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