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UMI®
ELECTRONIC TEMPERATURE SENSOR ARRAYS FOR GAS TURBINE COMPONENTS

CAPTEURS DE DISTRIBUTIONS DE TEMPÉRATURE ÉLECTRONIQUES POUR COMPOSANTES DE TURBINES À GAZ

Applied Sciences Master's Thesis
Field: Mechanical Engineering

Sherbrooke (Québec), Canada

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September 2004
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Abstract

The current master’s thesis presents the development of a new temperature sensing technology for gas turbine components. The proposed sensor array allows real-time simultaneous measurements of temperature at multiple locations, using only two communication leads. Frequency modulation is used to multiplex the signals of more than ten temperature sensors through common wires. At every point of reading, silicon carbide (SiC) microelectronic oscillators generate the required waveforms, at frequencies that are temperature dependent. Those oscillators are fed with a common DC power source, and add their signals together by current addition into the power supply leads. The multiplexed signal can be recorded using only one acquisition channel, and be analyzed in the frequency domain to deduce temperatures.

The resulting sensor array can be seen as a temperature sensitive wire, including two leads, and multiple integrated microscopic oscillators. It is compact, and alleviates the problem of lead routing, which is especially cumbersome in small business or regional aircraft engines. SiC microelectronics promising to be operable at temperatures above 700°C, the proposed sensor array could be used inside cooled turbine airfoils and shrouds, in moderate testing and flight cruising conditions, or in compressor components. It offers new possibilities in ground testing, vehicle health monitoring, and engine control.

Validation tests were conducted using macroscopic high temperature oscillator prototypes. Two oscillators were built using high temperature discrete components, and were tested in an oven up to 180°C. The results of those tests are presented in this thesis.
Résumé

Le présent mémoire de maîtrise décrit le développement d’une nouvelle technologie de mesure de température pour composantes de turbines à gaz. L’assemblage de capteurs proposé permet de mesurer simultanément la température à plusieurs endroits d’une même pièce, en temps réel, en n’utilisant que deux fils de transmission. Les signaux de plus de dix capteurs sont multiplexés à travers des fils communs, par modulation fréquentielle. À chaque point de mesure, un oscillateur microélectronique à base de carbure de silicium (SiC) génère un signal sinusoïdal, à une fréquence qui est influencée par la température. Tous les oscillateurs sont alimentés par une source commune de tension continue et combinent leurs signaux sinusoïdaux par addition de courant dans les fils d’alimentation. Le signal multiplexé peut être enregistré en n’utilisant qu’un seul canal d’acquisition et ensuite être analysé dans le domaine fréquentiel pour en déduire une distribution de température.

L’assemblage de capteurs ainsi créé peut être visualisé comme un câble sensible à la température, incluant deux fils de transmission et plusieurs oscillateurs microscopiques. Cet assemblage est compact et amoindrit l’encombrement causé par l’acheminement des fils, un problème particulièrement embarrassant dans les petits moteurs d’avions d’affaires ou de transport régional. La microélectronique à base de SiC promettant d’être utilisable au-delà de 700°C, le système proposé pourrait être utilisé à l’intérieur d’ailerons et de segments de carénage de turbines, dans des conditions modérées de test ou de vol, ou dans des pièces de compresseurs. Il offre de nouvelles opportunités en termes d’essais statiques, de suivi de la détérioration en vol et de contrôle actif des moteurs.

Des essais ont été réalisés pour prouver la validité du concept présenté ici, en utilisant des prototypes d’oscillateurs macroscopiques. Deux oscillateurs ont été fabriqués en utilisant des composantes électroniques discrètes résistant à haute température et ont été testés dans un four jusqu’à 180°C. Les résultats de ces essais sont présentés dans ce mémoire.
Chapter 1

Introduction

In the aircraft propulsion industry, optimization is essential to meet both customers' targets, such as thrust, specific fuel consumption, range, and cost, and safety requirements dictated by the certification organizations. Within such a stringent frame, building a successful engine imposes continuous innovation in the fields of materials, aerodynamics, controls, etc. Unfortunately, any improvement in performance is oftentimes counterweighted by a loss in durability, and vice versa.

Microtechnologies represent one promising path for simultaneous improvements in performance and durability, inspiring smarter engines that will adapt to changing circumstances. High temperature smart micro-sensors and actuators will soon become available, using recently developed wide band gap semiconductor materials. In the short-term future, using proper sensing devices, maintenance programs will be adapted to specific engine operation, and performances will be managed with regard to the state of deterioration of specific critical parts, on a flight-to-flight basis. In an ideal future, to obtain the best compromise between performance and durability, every part of an engine will actively adapt itself to its own changing conditions, using both micro-sensors and actuators.

Within the general aim of promoting this concept of smart engine by concrete micro-device development, the current thesis elaborates on the development of a new generation
of “smarter” high temperature sensors. The proposed sensors are intended to measure metal temperature distributions, in transient as in steady state, to enable health monitoring of turbine parts that are exposed to combustion products. The ultimate objective is to determine, at every instant of a ground test or flight mission, the temperature distribution over the high pressure cooled turbine airfoils, and to react accordingly through design modifications or active control.

The concept offers several advantages over the existing temperature measuring techniques. Specifically, it uses active silicon carbide (SiC) devices to multiplex the signals of about ten sensors onto the same leads. This avoids overcrowding of communication lines in low access locations like blade and vane cooling passages, and minimizes the number of required data transfer slip-rings when monitoring on rotating parts. A pressure sensor array could readily be developed along the same principles, which adds fundamental value to the idea.

SiC microelectronics promises to be operable at temperatures above 700° C, which largely exceeds the temperature of the air exiting the last compressor stage of modern aircraft engines, in the worst ground testing conditions. This will allow the use of the proposed sensors inside cooled turbine airfoils and shrouds, in moderate testing conditions and in flight cruising condition.

This thesis will present the new concept of sensors in details, and summarize the validation work that has been conducted so far to prove its feasibility. The research context will be presented first, with an overview of the gas turbine application and the existing temperature measuring techniques. Then, the design methodology will be detailed, before presenting the first concrete results of the research project.
Chapter 2

Research Context

2.1 Gas Turbine Environment

2.1.1 General Overview

Figure 2.1 shows a cross section view of the hot end of a PW100 turboprop engine from Pratt & Whitney Canada. The high pressure turbine of the engine is shown here, constituted of two stages of vane (stator) and blade (rotor) assemblies. Those two stages are together dedicated to power the upstream high pressure compressor, from which only the last (centrifugal) stage is shown. Previous (axial) high pressure compressor stages were cut off from the left of the figure.

The numbers in figure 2.1 indicate the thermodynamic states of the gases flowing through the engine. The compressor first pressurizes air up to state 3. The main portion of this pressurized air transpires through the combustion chamber liner, following the arrows in the figure, and participates to the combustion process to reach state 4. The combustion products, mainly composed of heated air, are thereafter expanded through the two stages of the high pressure turbine, from state 41 to state 45. The remaining portion of state 3 air
Table 2.1: Typical business jet engine running conditions during hot day block test take-off (HDBT-TO)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Typical value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_{03}$</td>
<td>500-600°C</td>
</tr>
<tr>
<td>$p_{03}$</td>
<td>250-300 psi</td>
</tr>
<tr>
<td>$T_{b4}$</td>
<td>1300-1550°C</td>
</tr>
<tr>
<td>$p_{b4}$</td>
<td>0.95-0.98$p_{03}$</td>
</tr>
<tr>
<td>$N_{HPT}$</td>
<td>22000-28000 RPM</td>
</tr>
</tbody>
</table>

bypasses the combustion chamber, and is used to cool turbine components.¹

The core of the engine, including the high pressure compressor, the combustion chamber, and the high pressure turbine, is often referred to as the hot gas generator. It generates state 45 hot gas that is used in further turbine stages, i.e., in the power turbine, to power the aircraft propeller. The power turbine stages were cut off from the right of figure 2.1. The power turbine rotors and the propeller are assembled onto a common shaft that extends through the entire length of the engine, for power transmission. In the same way, the high pressure turbine and compressor are linked together using another shaft, which is shorter. The two shafts are concentric and can be seen at the bottom of the cross section view.

The high pressure turbine blades are fed with cooling air that flows outward (from the engine centerline out), through passages located along the discs, whereas high pressure vanes are generally fed from the exterior of the turbine casing. In the case of the PW100 engine, only the two first turbine airfoils are cooled, with cooling air flowing along the arrows of figure 2.1. The cooling air is at stagnation conditions $T_{03}$ and $p_{03}$, before going through pressure losses and heat pick-up against turbine components.

The hot day block test take-off (HDBT-TO) condition is the most demanding ground testing condition required by the certification authorities, in terms of turbine durability. Various engine parameters are approximated in table 2.1 for a typical modern business jet engine operated at this condition. Station numbering is as shown in figure 2.1. $N_{HPT}$ is the rotation speed of the high pressure turbine spool.

¹ Other uses of this air, e.g., de-icing of the nacelle leading edge, will not be discussed here.
Figure 2.1: Combustion chamber and high pressure turbine of the PW100 turboprop engine (reprinted with permission of Pratt & Whitney Canada).
The airfoil of figure 2.2 belongs to a first stage turbine stator. Its external wall is partly cooled by air jets coming from an inserted perforated tube and impinging on its inside face. As it exits through holes across the external wall, the cooling air provides a protective film which also contributes to lower the metal temperature. The first stage stator of the PW100 engine (figure 2.1) employs this type of cooling scheme. Its vane airfoils are approximately 1.5 in in span by 0.5 in thick. Inside such an airfoil, only one or two thermocouples can be installed to measure metal temperatures, for ground testing only. It would definitely be useful to measure temperature at more than two locations, in the same airfoil, which is currently not practical considering lead routing problems, as will be described in a later section.

Figure 2.3 shows a first stage cooled turbine blade. In this case, a serpentine cooling scheme is preferred. In the PW100, the largest serpentine passages are approximately 0.150 in by 0.150 in, which provides rather limited space for the installation of a sensor. In more recent engines, the second stage vane and blade can also use a serpentine cooling scheme, but their passages are generally wider and less numerous than they are for the first stage blade.
In a modern aircraft engine, 20% to 30% of the air that penetrates into the hot gas generator is used to cool parts that are exposed to combustion products, such as the airfoils of figures 2.2 and 2.3. Since this cooling air does not get heated in the combustion chamber, its work contribution through subsequent turbine stages is outweighed by the energy previously deployed by the compressor to pressurize it. Consequently, the overall engine performance is reduced whenever a turbine cooling engineer decides to increase the cooling airflow through any turbine airfoil, to improve its durability.

In order to minimize the cooling airflows, and hence to improve the engine efficiency, it is of primary importance to optimize the different turbine cooling patterns. This optimization process represents somewhat a challenge, for reasons that are exposed next.

2.1.2 Turbine Temperature Variability

Part-to-Part Variations

Inside the combustion chamber, fuel is injected by discrete nozzles that are equally spaced around the circumference of the engine. To each of those fuel nozzles is associated a hot temperature streak that propagates through the first turbine stages. The static components
of turbines, e.g., the vanes and the shroud segments (see figure 2.1), are exposed to those streaks, and their temperature therefore vary along the circumference of the engine.

At the combustor exit (station 41 in figure 2.1), the temperature distribution of the hot gases is characterized by the Overall Temperature Distribution Factor:

$$OTDF = \frac{\hat{T}_{0(41)p} - T_{0(41)}}{T_{0(41)} - T_{03}}$$  (2.1)

where $\hat{T}_{0(41)p}$ stands for the local maximum gas temperature that the first stage vane can be exposed to. In this notation, the hat sign denotes the circumferential peak and the subscript $p$ denotes the radial peak. The OTDF is indeed the combination of Circumferential and Radial Temperature Distribution Factors given by:

$$CTDF = \frac{\hat{T}_{0(41)} - T_{0(41)}}{T_{0(41)} - T_{03}}$$  (2.2)

$$RTDF = \frac{\hat{T}_{0(41)p} - T_{0(41)p}}{\hat{T}_{0(41)}} \approx \frac{T_{0(41)p}}{T_{0(41)}}$$  (2.3)

For a modern engine using a direct flow combustor\(^2\), the OTDF and RTDF can be as high as 30% and 1.07, respectively. A local cooling effectiveness can relate the peak metal temperature $\hat{T}_{mp}$ to the local gas temperature for the hottest airfoil:

$$\epsilon_p = \frac{\hat{T}_{0(41)} - \hat{T}_{mp}}{\hat{T}_{0(41)} - T_{03}}$$  (2.4)

This effectiveness can be approximated to be constant from one airfoil to another, which leads to

\(^2\) The PW100 engine of figure 2.1 has a reverse flow combustor, which means that the combustion products first flow in the forward direction before turning and entering the turbine. In a direct flow combustor, combustion products do not turn, as they are produced to the left of the turbine and always flow in the direction of the turbine gaspath. Fuel nozzle hot streaks propagates further in the turbine in the case of the direct flow combustor.
\[ \varepsilon_p \approx \frac{T_{0(41)} - T_{mp}}{T_{0(41)} - T_{03}} \] (2.5)

Using typical engine parameters and metal maximum allowable temperature, for the first stage stator, the peak metal temperature of the hottest airfoil (\(T_{mp}\)) could be about 90°C higher than the peak metal temperature of the average airfoil (\(T_{mp}\)), in take-off condition. In terms of oxidation life, this means that the parts that are directly exposed to fuel nozzle streaks could deteriorate more than four times faster than the average (hypothetical) airfoil. On the other hand, airfoils that are perfectly aligned midway between two nozzle streaks will be slower to deteriorate.

The temperature distribution factors are hard to assess in a precise manner, either analytically or experimentally. Mainly because of the tolerances on the fuel nozzle flows, the OTDF can vary by \(\pm 3\%\) (if the predicted value is 30\%, it can vary from 27\% to 33\%). From one engine to another, the life of the turbine parts will vary partly for this reason. Furthermore, within the same engine, if one fuel nozzle flows more than its neighbors, it will cause static turbine airfoils and shrouds located in its streak to deteriorate faster.

Finally, turbine components are affected by their own manufacturing tolerances, e.g., by the tolerances on their cooling airflows. They can also be affected by upstream components and gaspath gaps, which further adds to the unpredictability of their exact metal temperatures and lives.

**Engine-to-Engine Variations**

Some higher level variations related to the general operation of an engine can influence turbine component lives. Overall efficiencies of compressor, combustor, and turbine modules can all affect the engine cycle positively or negatively. With less efficient modules, an engine will need to be operated at higher temperatures to produce the required thrust or power. In a general way, for this reason, older engines provide a harsher environment than new engines for turbine parts. Even at the assembly line exit, two new engines demonstrate
slightly different operating cycles, again due to manufacturing and assembly tolerances.

A good way to capture engine-to-engine variability problems is to monitor average gas temperatures of every engine, in flight. Currently, the engine condition is monitored by measuring the temperature of the combustion products at the exit of the high pressure turbine ($T_{45}$\textsuperscript{3}, in figure 2.1), using thermocouples. For constant thrust or power, $T_{45}$ increases as the engine deteriorates. However, this health monitoring method gives no specific information on the status of any part of the high pressure turbine, which explains why some engines experience durability problems even before the temperature limit is reached at station 45.

**Aircraft-to-Aircraft Variations**

Despite all the efforts of the engine constructor to reduce its engine variability, turbine durability is still affected by mission variations. It is well recognized that different aircraft operators will in average get different lives out of the same engine models. Missions can be more or less demanding depending on the location of use, flight durations, take-off conditions, etc. Motorists can assess component lives for each mission declared by aircraft operators analytically, but unexpected variations are often observed. The currently available mean of capturing those variations, from an overall engine condition point of view, is to monitor $T_{45}$, as described previously.

**2.1.3 Safety Margins**

Even if they are exposed to hotter flows than their neighbors, parts located in fuel nozzle streaks are not designed in a special way. In fact, all static components of the turbine are designed as if they were subjected to the worst circumferential conditions. The worst manufacturing tolerances also need to be accounted for, as they can often combine together to result in overall or local variations of temperature. In the same order of idea, despite the changing flight conditions, turbine parts remain passive and generally pass the same

\textsuperscript{3} $T_{45}$ is the so called Inter-Turbine Temperature, or ITT.
cooling airflow all the time. At the design level, the cooling passages of an airfoil are often optimized to withstand the worst take-off condition, and as a result the durability in cruise condition is often better than required\(^4\).

It is becoming obvious that designing turbine components requires bookkeeping of multiple safety margins, especially in the temperature prediction tools. Those safety margins are added to the uncertainty of the analytical models, to ensure reliability in all conditions. Engine manufacturers are reluctant to use probability calculations: if one engine out of one hundred suffers from durability problems, it is one too many engine. If, in the aerospace industry, structural design is optimized to safety margins down to 10% or 5%, it cannot be the case in turbine cooling, considering all the uncertainty factors described so far.

2.2 Needs of the Aircraft Engine Industry

2.2.1 Benefits of Local Instrumentation

It was explained in section 2.1.1 why the high pressure turbine cooling schemes need to be optimized to meet both performance and durability requirements. On the other hand, it was also outlined, in section 2.1.2, that cooled components are subjected to many variations in operating conditions, some of those variations being random. As a consequence, the turbine optimization process requires considerable ground testing and flight experience, and despite all efforts still carries considerable safety margins, as mentioned in section 2.1.3. In this context, the engine industry would get considerable benefits from better instrumenting high pressure turbines, specially from local real time temperature sensing. At NASA Glenn Research Center, three types of applications are envisioned for high temperature electronics and sensors: ground testing, vehicle health monitoring, and engine control [2].

First of all, local temperature sensing is essential to extract all possible advantages of

\(^4\) This depends on the engine application. In some cases, e.g., when the take-off duration is relatively short, and when the operator requires longer and higher speed flight, the cruise condition can be the limiting condition in terms of durability.
expensive ground testing. Ground tests can cost several hundreds of dollars each, accounting for engine manufacturing, assembly and disassembly costs, and test cell operation costs. Such expenses justify any push toward using as many sensors as possible at one time. In order to validate analytical heat transfer models, turbine cooling engineers need more than post-testing observations of component deterioration: they need metal temperature measurements at every running condition of the tests. This detailed lower level information allows to find the root flaws of models. Better temperature monitoring also offers the possibility of shorter tests, avoiding the necessity of reaching high deterioration levels to make useful observations. Experimental parts can therefore be reused for multiple tests.

Secondly, in-flight monitoring allows adaptive maintenance plans. By periodically communicating engine data to its service center, the aircraft operator benefits from recommendations from maintenance experts. Such practices are already in place at Pratt & Whitney Canada: in addition to $T_{45}$ (see section 2.1.2), the rotation speeds and the number of cycles are monitored. These data are used to quantify creep and fatigue deterioration. With additional local turbine temperature sensors, maintenance tasks could focus on specific vulnerable parts, rather than reacting to the general behavior of the engine. Design engineers could also anticipate specific problems before they happen, and start working on preemptive design modifications. It is well recognized that development work spreads over the useful life of an engine, as some unexpected durability problems, often related to cyclic operation, arise only during aircraft operation.

Thirdly, for the aircraft operator, performance gains are to be expected from better instrumentation associated with feedback control. Indeed, since engines are currently designed with safety margins, to ensure durability in all conditions (see section 2.1.3), some engines are not operated to the full capabilities of their turbine components. Knowing about the severity of the operating conditions of the most vulnerable turbine parts, it becomes possible to assess the actual thrust capability of a particular engine. Throttle pushes, and hence increases in operating temperatures, can then be applied to healthier engines, without sacrificing safety. In addition to pure power gains, reductions in fuel consumption are also possible in the same way, as increasing operating temperatures can improve engine cycle
efficiency.

2.2.2 Temperature Sensing Requirements

Within the general aim of bringing the previous advantages to the aircraft industry, the current thesis focuses on a novel temperature sensing technology. A preliminary Product Definition and Requirements Specification (PDRS) document was elaborated to describe the need for a temperature sensing system intended for local monitoring in high pressure turbines. The complete document is provided in appendix A, and table 2.2 draws a summary list of all specifications. The levels to be targeted for each specification were established based on the experience of the author in the Turbine Cooling & Static Structures department of Pratt & Whitney Canada, from January 2002 to September 2003, and upon discussions with colleagues in this same department.

None of the existing temperature measuring techniques normally used in turbines satisfies the specifications presented here. In section 2.3, the most popular techniques will be surveyed and criticized against those specifications. The new concept presented in this thesis will be criticized in the same way in section 4.2.

2.3 Current Temperature Sensing Techniques

2.3.1 Thermocouples

Thermocouples are the only turbine temperature sensors currently used for real time measurements on aircraft. Since for each measuring location, a communication line must reach a data acquisition module, the number of thermocouples is mainly limited by the size of the leads that must exit the turbine. Only one or two thermocouples can usually be installed in an airfoil such as the one of figure 2.2, to minimize the obstruction of the cooling air inlet by leads. Such thermocouples are only used for ground testing, however. In flight, only
Table 2.2: High pressure turbine temperature sensing system specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Level</th>
<th>Flexibility</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Temperature sensing</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Temperature range(^{a,b})</td>
<td>820-1260°C (1500-2300°F) for parts exposed to combustion products(^a)</td>
<td>N/A</td>
</tr>
<tr>
<td></td>
<td>480-820°C (900-1500°F) for parts not exposed to combustion products</td>
<td></td>
</tr>
<tr>
<td>• Accuracy at 1000°C</td>
<td>±6°C (±10°F)</td>
<td>maximum</td>
</tr>
<tr>
<td>• Spatial resolution</td>
<td>5 mm (0.200 in)</td>
<td>maximum</td>
</tr>
<tr>
<td>• Temperature resolution at 1000°C</td>
<td>6°C (10°F)</td>
<td></td>
</tr>
<tr>
<td>• Response time</td>
<td>1 s</td>
<td></td>
</tr>
<tr>
<td><strong>Integration in engine environment</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Ambient temperature(^{b,c})</td>
<td>480-760°C (900-1400°F)</td>
<td>N/A</td>
</tr>
<tr>
<td>• Ambient pressure(^{b,c})</td>
<td>120-450 psi</td>
<td>N/A</td>
</tr>
<tr>
<td>• Resistance to contaminants</td>
<td>To be determined (humidity, dust)</td>
<td></td>
</tr>
<tr>
<td>• Excitation frequencies(^b)</td>
<td>250-500 Hz (engine spool rotation speeds: 15,000-30,000 RPM)</td>
<td>N/A</td>
</tr>
<tr>
<td></td>
<td>10,000-20,000 Hz (~40 blades per rotor)</td>
<td></td>
</tr>
<tr>
<td>• Fixation surface temperature(^b)</td>
<td>Concept dependent</td>
<td></td>
</tr>
<tr>
<td>• Fixation surface material</td>
<td>nickel-iron alloy (with or without coating)</td>
<td>N/A</td>
</tr>
<tr>
<td>• Volume of one sensor</td>
<td>1 mm x 1 mm x 1 mm, if within airfoil cooling scheme</td>
<td>maximum</td>
</tr>
<tr>
<td>• Weight(^d)</td>
<td>500 g (1 lb) (for the complete system, if installed on aircraft)</td>
<td></td>
</tr>
<tr>
<td><strong>Data acquisition</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Data acquisition frequency</td>
<td>1 Hz</td>
<td>minimum</td>
</tr>
<tr>
<td>• Data storage capacity</td>
<td>4 Mb per flight mission (onboard storage)</td>
<td>nice-to-have</td>
</tr>
<tr>
<td>• Data transmission</td>
<td>One pair of leads per airfoil cooling scheme feed passage</td>
<td>maximum</td>
</tr>
<tr>
<td><strong>Durability</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Oxidation life</td>
<td>10,000-20,000 h in flight conditions</td>
<td>minimum</td>
</tr>
<tr>
<td></td>
<td>50-100 h in ground (block) test conditions</td>
<td></td>
</tr>
<tr>
<td>• Low cycle fatigue (LCF) life</td>
<td>10,000-20,000 cycles in flight conditions</td>
<td>minimum</td>
</tr>
<tr>
<td></td>
<td>50-100 cycles in ground (block) test conditions</td>
<td></td>
</tr>
<tr>
<td>• High cycle fatigue (HCF) life</td>
<td>10,000-20,000 h in flight conditions</td>
<td>minimum</td>
</tr>
<tr>
<td></td>
<td>50-100 h in ground (block) test conditions</td>
<td></td>
</tr>
<tr>
<td>• Creep life</td>
<td>10,000-20,000 h in flight conditions</td>
<td>minimum</td>
</tr>
<tr>
<td></td>
<td>50-100 h in ground (block) test conditions</td>
<td></td>
</tr>
<tr>
<td><strong>Cost</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Acquisition cost</td>
<td>40$ (Can.) per airfoil(^e)</td>
<td>maximum</td>
</tr>
<tr>
<td>• Operation cost</td>
<td>10$ (Can.) per airfoil at every engine rebuilt(^e)</td>
<td>maximum</td>
</tr>
<tr>
<td><strong>Manufacturability</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Time for assembly</td>
<td>2 man-hours for the complete system(^e)</td>
<td>maximum</td>
</tr>
</tbody>
</table>

a. The upper limit is the maximum temperature of surfaces directly exposed to the gaspath, whereas the lower limit is typical of coolant side surfaces.
b. Turbine parameters are specified for a typical hot day block test take-off (HDBT-TO) condition. Environment parameters are specified as a range because different levels can be encountered in different engine models or at different locations within the same engine, at this same HDBT-TO condition.
c. Ambient conditions are defined as the stagnation conditions of the cooling air, in the stream of which the system is more probable to be installed. Combustion product conditions are not mentioned here.
d. A more rigorous weight criterion must be established for rotor sensors, considering centrifugal force contribution to rotor stresses. Also, experimental engines might necessitate heavier instrumentation.
e. Production and overhaul costs and assembly time. Experimental engines might require more resources.
the $T_{45}$ gas temperature is monitored, using less than ten thermocouples that plunge into the gaspath. Those are type K thermocouples (Chromel-Alumel), and are shielded with ceramic, which provides them with an acceptable useful life.

Thermocouples can also be used on rotating parts, using slip-ring electrical contacts. Once again, however, this technique is employed only for ground testing.

At NASA Glenn Research Center, researchers are developing thin film thermocouples that can be directly deposited onto turbine blades [3]. Those thermocouples are platinum-13% rhodium versus platinum (type R), and are only a few micrometers thick. Figure 2.4 shows such thermocouples deposited on a nickel alloy blade of the space shuttle main engine fuel turbopump. In order to isolate the sensor from the conductive blade alloy, a MCrAlY coating is first deposited (where M stands for Ni, Fe, or Co)\textsuperscript{5}, followed by an alumina ($\text{Al}_2\text{O}_3$) layer. The thermocouple films are deposited over the alumina, and protected with an additional alumina layer. Those thin film thermocouples were tested in both laboratory environment and real engine conditions. Problems were experienced with the contacts between leads and metallic thermocouple films, resulting in unstable signal.

Thin film thermocouples do not overcome the problem of signal routing from the blades down to the rotor disc hub, and through slip-rings to a static component, but they allow multiple reading locations on a single airfoil. Used in association with a multiplexer located on the disc, they would show great potential.

2.3.2 Thermal Paint

Thermal paint can give detailed temperature distributions on parts exposed to combustion products, by permanently changing color when heated. However, this technique is expensive and complex.

\textsuperscript{5} MCrAlY is commonly used as an oxidation protective coating on nickel alloy turbine airfoils, and as a base coating for thermal barrier coating (TBC). Hence the process of depositing such coating is already well-known by the gas turbine industry.
Figure 2.4: Thin film thermocouples deposited on a blade of the space shuttle main engine fuel turbopump [3].
The paint must first be applied on parts, and be protected with a transparent coating especially designed for this application. Since multiple paints are available to react with more or less resolution to different temperatures, several paints must be applied on the same rotor or stator, each on a different turbine segment. In order to get valuable results, the engine must be maintained at the test condition for a period of time corresponding to the thermal paint calibration time. This duration can be about two minutes. The transition periods of the engine test must also have controlled durations.

The post-test examination of painted parts requires an expert's thoroughness. Color transitions may be subtle: on a turbine airfoil, the paint will often take one single color, but in different shades, e.g., light blue, dark blue, and navy blue. The test parts are compared to calibration samples like those of figure 2.5.

The calibration process of thermal paints is often suspected to be a considerable source of errors. Calibration samples are brought to high temperature in static laboratory conditions,
whereas real are conducted in an environment where combustion products sweep the thermal paint with a high velocity. Nevertheless, an accuracy of ±30°C is expected from thermal paint reading.

One variant of the thermal paint test is the silver plating test. In this case, a thin silver layer is applied on turbine parts, and vanishes when reaching 960°C, the melting point of silver.

2.3.3 Crystal Temperature Sensors

Crystal temperature sensors [5], alike thermal paint, measure a unique temperature that is maintained over a controlled period of time during a ground test. They allow local temperature measurements with more accuracy than thermal paint, i.e., within the range of ±10°C.

The sensors are pieces (.020 in maximum length) of single crystal cubic silicon carbide (SiC) that were irradiated with neutrons. By the mean of this radiation, defects were introduced in the crystal, resulting in a net lattice volume increase. When annealed under a temperature exceeding the temperature of the irradiation process, the crystal progressively takes back its original lattice configuration. This effect can be used to measure temperature of turbine parts. Knowing the time of annealing, i.e., the time of exposition during an engine test, it is possible to deduce the exposition temperature by observing the crystal lattice by X-ray diffraction.

From the turbine manufacturer point of view, the crystal temperature sensors have to be inserted within turbine component walls, inside holes that are subsequently plugged with a ceramic compound. After engine testing, the sensors are taken off and sent back to the sensor supplier's laboratory, where they are analyzed by X-ray. Due to the complexity and cost of the process, crystal sensors are used only when excellent accuracy is needed.
2.3.4 Optical Pyrometry

Kerr and Ivey provide an extensive description of optical pyrometry as it is applied to measure turbine blade temperatures, along with current limitations of the technique [6]. A pyrometer collects radiations from the blade surface, using a lens, and routes those radiations via fibre optics to a detector that converts optical power to an electrical signal.

A single static probe is used to monitor the temperature of all the blades of a rotor: the collected signal is intermittent, with a frequency corresponding to the frequency of blade occurrence in front of the probe. The probe can be embedded within the upstream vane outer shroud, as shown in figure 2.6, and give radiation measurements at only one blade radius. In some experimental engines, for ground testing, it can be equipped of a lens motion control system to provide measurements from the entire blade span. In this latter case, the probe is cooled and plunges into the hot gas stream, disturbing the turbine flow to some extent. Any of those two assembly options necessitate a fair amount of space, which is available only in large engines, e.g., turbofans producing over 30,000 lb of thrust.

Pyrometry features many major advantages, like fast response and absence of contact, but remains a complex method that is still not widely used in the aircraft engine industry. Many parameters must be carefully accounted for in selecting and locating probe compo-
ments, and in designing signal treatment algorithms:

1. Blade surface emittance, which depends on temperature, wavelength, angle of emis-
sion, materials chemical composition, surface roughness, and state of oxidation;

2. Intensity of radiations originating from other sources and reflected by the blade;

3. Absorption and emission from hot gases located in the probe line of sight;

4. Emission from particles and flames randomly escaping from the combustor and cross-
ing the probe line of sight;

5. Lens contamination.

The problem of blade emittance can be solved with considerable testing, in multiple
conditions, and detectors that react to specific wavelengths. Nevertheless, some questions
remain unanswered about emittance variations due to continuous blade deterioration. It is
known that emittance varies with blade oxide growth to reach an asymptote early in the
gine life [6], which allows for compensation in data treatment. However, thermal barrier
coating (TBC) spallation, accumulation of dust, and changes in blade surface roughness
could represent major problems, as they occur more randomly and unexpectedly.

Problems 2 to 4 can be overcome using different signal treatment techniques described
by Kerr and Ivey [6].

As for lens contamination, it represents one of the major problems associated with the
use of pyrometry on aircraft. The RB199 Rolls-Royce military engine is uses an in-flight
pyrometry system since the early 1990's, but this system has suffered from lens fouling
problems since its installation. Purge air systems have been developed to prevent soot and
dust from accumulating on pyrometer lens, but those systems have led to modest results.
So far, 150 hours of operation have been demonstrated with negligible lens deposits [7]. Al-
though acceptable for a military application, this delay is too short for commercial aircraft,
for which the maintenance schedules are spread over longer periods.
A variant of optical pyrometry has also been developed, which is phosphor thermometry [8]. It consists of applying phosphorus materials onto turbine parts, and using a pulsed laser to produce fluorescence. The emitted light is then collected with an apparatus similar to a pyrometer. A major advantage of this latter technique is to allow measurements through flames, and in presence of high reflected radiations.

2.3.5 Critics of Existing Techniques

Table 2.3 summarily criticizes every existing temperature sensing methods presented previously against the main specifications listed in Table 2.2. Arrows up and down indicate advantages and drawbacks, respectively.
<table>
<thead>
<tr>
<th>Specifications</th>
<th>Thermocouples</th>
<th>Measuring techniques</th>
<th>Crystal temperature sensors</th>
<th>Optical pyrometry</th>
</tr>
</thead>
<tbody>
<tr>
<td>Temperature sensing</td>
<td>† Entire range covered</td>
<td>† Entire range covered</td>
<td>† Entire range covered</td>
<td>† Entire range covered</td>
</tr>
<tr>
<td></td>
<td>† ±2°C</td>
<td>† ±15°C</td>
<td>† ±6°C</td>
<td></td>
</tr>
<tr>
<td></td>
<td>† Local measurements only</td>
<td>† Temperature distribution sensing</td>
<td>† Local measurements only</td>
<td></td>
</tr>
<tr>
<td></td>
<td>† Good repeatability</td>
<td>† Depends on test sequence</td>
<td>† Depends on test sequence</td>
<td></td>
</tr>
<tr>
<td>Integration in engine environment</td>
<td>† Heavy wire and slip-ring arrangements</td>
<td>† Colors influenced by engine contaminants</td>
<td>† Meet all requirements</td>
<td>† Cannot be used in small engines</td>
</tr>
<tr>
<td></td>
<td></td>
<td>† Usable on rotors as easily as on stators</td>
<td>† Usable on rotors as easily as on stators</td>
<td>† Lens fouling with soot and dust</td>
</tr>
<tr>
<td>Data acquisition</td>
<td>† Real time measurements</td>
<td>† No real time acquisition (maximum temperature measurement only)</td>
<td>† No real time acquisition (maximum temperature measurement only)</td>
<td>† Real time measurements</td>
</tr>
<tr>
<td></td>
<td>† One channel per sensor</td>
<td></td>
<td></td>
<td>† One channel only</td>
</tr>
<tr>
<td>Durability</td>
<td>† Meet all requirements</td>
<td>† Ground testing only</td>
<td>† Ground testing only</td>
<td>† Good for ground testing</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>† Lens fouling after ~150 hours in flight</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>† Influenced by blade deterioration</td>
</tr>
<tr>
<td>Cost</td>
<td>† Cheap</td>
<td>† Expensive process</td>
<td>† Expensive sensors and process</td>
<td>† Expensive apparatus</td>
</tr>
<tr>
<td>Manufacturability</td>
<td>† Common methodology</td>
<td>† Fastidious process</td>
<td>† Fastidious process</td>
<td>† Easy to install</td>
</tr>
</tbody>
</table>
Chapter 3

Methodology

3.1 Concept Selection

3.1.1 Frequency Modulation

More than twenty years ago, NASA Glenn Research Center recognized the industry needs exposed in section 2.2.1 [9]. As one possible solution, the authors proposed the development of analog to digital converters, multiplexers, and telemetry systems that could withstand 500 to 600°C. These electronic devices would be installed on turbine discs to allow simultaneous readings from the order of one hundred sensors per rotor. In instrumenting a turbine rotor, lead routing represents a considerable problem, with conventional sensing technology. Each thermocouple, pressure sensor, or strain gauge to be installed on a blade requires a pair of wires going down along the disc, and a dedicated data transfer slip-ring (as previously exposed in section 2.3.1). In small engines like those developed by Pratt & Whitney Canada, even routing leads out from static parts can be a cumbersome problem. Clearly, multiplexing shows to be essential to increase the number of sensors in turbines. NASA researchers are still working to develop high temperature electronic devices that will lead to local signal conditioning, using silicon carbide (SiC) materials [2].
Various multiplexing strategies could be employed, e.g., amplitude modulation (AM), frequency modulation (FM), or digital multiplexing (as suggested in ref. [9]). In the current project, FM emerged as the simplest and most efficient solution, as it is less sensitive to noise than AM, and simpler to implement than digital multiplexing. It can also be easily implemented in a sensor array, by making the sensors to generate waveforms by themselves, each within a dedicated frequency band.

This is precisely the concept that is proposed here for temperature sensing. A common pair of leads is used to communicate the temperature readings of about ten sensors using frequency modulation. Each sensor is by itself an oscillator which generates a quasi-sinusoidal signal maintained within one of ten reserved frequency bands. The frequency of each signal varies with the local temperature, but stays within its reserved band. Only one channel is used to acquire the multiplexed signal, and frequency analysis is then used to isolate the ten readings from one another.

Since the local sensors are intended to be very small, the resulting sensor array can be seen as a sensitive wire, as shown in figure 3.1. Commercially, this wire would become available with various sensor spacings, and various numbers of frequency bands. After transferring it to the frequency domain, the multiplexed signal from such a wire would resemble the one of figure 3.2.
Figure 3.2: Typical multiplexed signal obtained from a temperature sensor array.
Figure 3.3: Phase-shift oscillator used for temperature sensing.

3.1.2 Oscillator

At each measuring point of the sensor array, various means could be employed to generate an oscillating signal. One of them would be to use a microscopic vibrating beam which natural frequency would be influenced in some way by temperature. Another one, the one that is being developed in the current thesis, is to use a microelectronic oscillator, which resonating frequency is temperature dependent. This latter concept is preferred because of its simplicity, and because of its ease of miniaturization and packaging.

The selected microelectronic oscillator scheme is sketched in figure 3.3. It is a phase-shift oscillator, and its oscillating frequency is dictated by the values of the capacitors $C$ and resistors $R$ of its feedback network. Specifically, the resistors $R$ are made to change values under temperature variations, which provides the temperature dependence of the frequency. Constant voltage is provided to the oscillator, and an oscillating current signal is picked up by the power supply leads. The design process used to reach this result will be presented in section 3.2, along with the oscillator operating principles.

Multiple oscillators can be connected in parallel, and consequently add their different current signals together, as will be described in details in section 3.2. Figure 3.4 shows such an arrangement with two phase-shift oscillators.

Building such oscillators for high temperature operation, including active devices such
as transistors and diodes, has become possible lately with the advent of wide-band gap semiconductor materials. Silicon carbide (SiC), in particular, has been recognized to be promising for aircraft engine applications [10]. SiC electronics has become available, and offers opportunities for smart micro-devices operating in harsh environments.

3.1.3 Silicon Carbide Technology

Temperature Limit

Doping a semiconductor material allows to modulate its resistivity and to create p-n junctions, which are the basis of any active electronic device. The doping elements introduce electron energy levels that are close to the limits of the band gap (filled for n-type, empty for p-type), which favors the emergence of current carriers (holes and free electrons), and hence increase conductivity. Controlling the concentration of carriers is essential to control device properties.
Apart from these extrinsic carriers that are promoted by the doping elements, intrinsic carriers are generated as the temperature is increased. In this case, electrons are freed by vibration from the covalent bonds of the lattice and act as free carriers. To maintain device characteristics at high temperature, the intrinsic carrier concentration needs to be negligible compared to the extrinsic carrier concentration, or compared to the doping level.

In any semiconductor material, the intrinsic carrier concentration is given by

\[ n_i = 2 \left( \frac{2\pi kT}{\hbar^2} \right)^{3/2} (m_e m_h)^{3/4} \exp \left( \frac{-E_g}{2kT} \right) \] (3.1)

where \( k \) is the Boltzmann constant, \( T \), the temperature, \( \hbar \) the Planck constant, \( m_e \) and \( m_h \), the electron and hole effective density-of-states masses, and \( E_g \), the band gap. Using parameters from Ruff et al. [11] \( n_i \) is plotted in figure 3.5 for silicon, 6H-SiC, and 4H-SiC.

The temperature dependent band gap is given by

\[ E_g(eV) = \begin{cases} 
1.17 - \frac{3.69 \times 10^{-4} T^2}{T(K) + 1108} & \text{for silicon [11]} \\
3 - 3.3 \times 10^{-4} (T(K) - 300) & \text{for 6H-SiC [11]} \\
3.3 - 3.3 \times 10^{-4} (T(K) - 300) & \text{for 4H-SiC} 
\end{cases} \] (3.2)

For 4H-SiC, the band gap at 300 K is given in ref. [12], and the temperature dependence is approximated by the one of 6H-SiC. 4H-SiC wafers have recently been commercialized, and some devices have been fabricated using them [13]. Nevertheless, only sparse data are available on the characterization of this polytype.

The effective masses are approximated to be \( 1m_0 \), i.e., to be equal to the electron rest mass. From various experiments and theoretical calculations (see ref. [14] and [15]), the Si and 6H-SiC electron and hole effective masses are within the range of 0.2 to 1.2\( m_0 \) at 300 K. Furthermore, using the temperature dependence for Si [15], effective masses would increase by less than 0.3\( m_0 \) from 300 to 1200 K. Consequently, using effective masses of \( 1m_0 \) for both Si and 6H-SiC results in errors within one order of magnitude for \( n_i \). It is assumed that the same masses are appropriate for 4H-SiC.
Figure 3.5: Intrinsic carrier concentration calculated from equation 3.1.
Since typical semiconductor devices can be fabricated using doping levels of $10^{16}$ cm$^{-3}$ and higher, it is common to use $10^{15}$ cm$^{-3}$ as the maximum tolerable intrinsic carrier concentration. Following this rule, silicon devices would be operable up to 255°C, 6H-SiC devices, up to 993°C, and 4H-SiC devices, up to 1103°C.

However, Shenai et al. [16] outlined that the reverse bias current is also an important limiting parameter, in addition to the intrinsic carrier concentration. At high temperatures, reverse currents become unacceptably high, increasing proportionally with $n_i$. Fixing the reverse current density limit to $10^{-2}$ A/cm$^2$, Ruff et al. [11] stated that 1000 K (727°C) is the limit of operability of 6H-SiC. In order to reach the temperature limit suggested by the calculation of $n_i$, active devices will first need to tolerate high reverse currents.

**Active Devices**

The highest temperature ever reached by a transistor is 650°C. Using β-SiC, Palsom et al. [17] fabricated a MOSFET that showed an acceptable behavior at this temperature. An attempt to reach 700°C resulted in gate oxide breakdown. The semiconductor was probably able to withstand a higher temperature, in spite of the high leakage currents due to the high intrinsic carrier concentration. Palsom et al. later tested a 6H-SiC MOSFET up to 650°C, again with satisfying results [18]. To date, the highest temperature reached by a 4H-SiC transistor is 400°C, as reported by Scozzie et al. for a JFET [13].

SiC is beginning to find many other high temperature electronic applications, one of the most interesting being the development of a 6H-SiC EEPROM\(^1\) device by Li et al. [19]. The memory device operates using n-channel MOSFET principles, and to date has been tested at temperatures up to 200°C.

---

\(^1\)Electrically erasable programmable read-only memory
3.2 Electronic Oscillator Design

At each measuring point of the sensor array, various means could be employed to generate an oscillating signal. The basic principle of the oscillator used here is schematized in figure 3.6. The phase-shift oscillator of figure 3.3 operates along this principle. Three fundamental components are necessary: an amplifier, a feedback network, and a limiter. The amplifier and the feedback network are connected together to form a loop. Around this loop, in order to produce oscillation, the total gain must be greater than unity, and the total signal phase shift must correspond to a finite number of cycles. The limiter is used to maintain the oscillator within a predetermined amplitude range, and hence to protect the looped components from overloading. Without this limiter, the signal would be amplified until the amplifier reaches saturation, or until one of the components fails.

The feedback network can be any physical component that react preferentially to different excitation frequencies, e.g., a vibrating beam or a quartz piezoelectric crystal characterized by a natural frequency, or an electronic circuit. The frequency response of the network should vary with temperature. For example, the feedback network can be a structure which natural frequency is influenced by temperature. This can be achieved by using a
mismatch in thermal expansion coefficients to make the structure pre-stress level to vary with temperature. The pre-stress level influences the frequency response of the structure, and, consequently, the frequency response of the structure (the feedback network) depends on temperature. In another concept, which is developed here, an electronic feedback network is used, and the electronic components of this network vary with temperature.

The amplifier is made of active electronic components, and it is the power consumed by this amplifier that is used for multiplexing. In operation, each amplifier of a sensor array consumes a current that alternates at the frequency of the corresponding oscillator, around a certain DC value. The currents necessitated by all the oscillators of the array are all supplied through the same two wires, as shown in figure 3.7, and hence these wires carry the frequency content from all the oscillators. Measuring the current through one of these two leads and performing a frequency analysis allows to recuperate the oscillating frequency of all the oscillators of the array. In figure 3.7, current is read by measuring the voltage drop across resistor $R_M$.

Finally, the limiter principle can either be mechanical, in the case of vibrating devices, or electrical. For example, mechanical stoppers can limit the displacements of vibrating structures, whereas diode bridges or heating elements can limit voltages and currents, respectively. The limiter can operate on any component of the oscillator, or along any of its connection lines. Consequently, its location is not limited to the one shown in figure 3.6, where it is simply wired in series within the feedback line. It could also act directly onto feedback network components, or somewhere else in the oscillator loop, e.g., between the feedback network and the amplifier.

In this thesis, only electronic components are used, for simplicity of integration. Many types of electronic oscillators are available, e.g., RC and LC oscillators, negative resistance oscillators, multivibrators, etc. [20]. The following criteria were considered in the selection of one type of electronic oscillator:

- Inductors have not yet reached an acceptable level of miniaturization in integrated
Figure 3.7: Oscillator multiplexing principle based on addition of amplifier currents.
circuits, and thus are discarded from the oscillator design;

- In order to do efficient frequency modulation (as described in section 3.1.1), the generated signal must be as close as possible to a perfect sinusoidal waveform;

- To control the temperature dependence of the oscillation frequency, it is easier to use the thermal behavior of passive components, such as resistors, rather than that of active components, like transistors or diodes (The sensitivity of a resistor to temperature variations can easily be adjusted, using different materials, which is not possible with semiconductor active devices).

Based on those criteria, RC oscillators were considered as valuable candidates. In particular, phase-shift and Wien bridge oscillators were investigated. They both do not need inductors, and can generate almost sinusoidal waveforms. Their oscillation frequency are independent of transistor or diode behavior, and can be adjusted by changing resistor values only.

3.2.1 Phase-shift Oscillator

The phase-shift oscillator was already introduced in figure 3.3. It is composed of a high-pass feedback network, an amplifier, and a diode limiter. Ideally, the amplifier produces a signal phase-shift of 180° that is independent of frequency. The feedback network in figure 3.3 contains four high-pass filters connected in series that together provide an additional 180° phase-shift at one particular frequency. If the gain of the amplifier is sufficiently high to compensate for the losses through the feedback network, at this particular frequency, oscillation occurs. The oscillation frequency simply corresponds to the frequency for which the feedback network produces a 180° phase-shift, with the result of 360° total phase-shift around the loop.
Feedback Network

Since every elementary high-pass filter can provide a phase-shift between 0 and 90° (see Bode plot of figure 3.8), at least three of those filters are necessary in the feedback network in order to reach the required 180° shift. The phase-shifts are additive, such that three filters will produce a 0 to 270° phase-shift, and four filters, a 0 to 360° phase-shift. The advantage of using four filters instead of three is to obtain lower total losses. Using more filters would lead to even lower losses, but would become more cumbersome in an integrated circuit, for a relatively negligible advantage in signal losses.

Table 3.1 summarizes the RC feedback network theoretical results. The third network to appear in this table is the one that is preferred. The grounded resistors ensure that the exit signal is purely alternative (AC), with no DC offset. The option of a low-pass filter cascade such as the first network of the table was disregarded because, compared to a high-pass network, six times bigger components are necessary, for the same oscillating frequency. It is also less convenient to control the DC offset of the exit signal, with such a low-pass network.

Platinum resistors are used in the network to modulate the oscillating frequency of the oscillator. This metal was chosen because of its linear resistance-temperature behavior over the entire temperature range to be covered here. Other resistance temperature detector (RTD) metals, such as copper and nickel, show discontinuities in their thermal resistivity coefficient at relatively low temperature, as shown in figure 3.9. Those discontinuities are unacceptable in the frequency modulation concept presented here, as it would become more fastidious to manage the limits between the frequency bands of each sensor, especially if transient measurements from ambient to operating temperatures are needed.

Amplifier

The requirements for the amplifier of the phase-shift oscillator of figure 3.3 are to provide, in all operating conditions, a gain of at least 18.4 and a phase-shift of 180° at the oscillating
Figure 3.8: Elementary high-pass filter.

Table 3.1: RC feedback network theoretical results

<table>
<thead>
<tr>
<th>Network</th>
<th>$f \mid \Delta \phi = 180^\circ$</th>
<th>$G \mid \Delta \phi = 180^\circ$</th>
</tr>
</thead>
<tbody>
<tr>
<td><img src="image1" alt="Network" /></td>
<td>$\frac{\sqrt{3}}{2\pi RC}$</td>
<td>$\frac{1}{3}$</td>
</tr>
<tr>
<td><img src="image2" alt="Network" /></td>
<td>$\frac{1}{2\pi \sqrt{6} RC}$</td>
<td>$\frac{1}{29}$</td>
</tr>
<tr>
<td><img src="image3" alt="Network" /></td>
<td>$0.837 \frac{2}{2\pi RC}$</td>
<td>$\frac{1}{18.4}$</td>
</tr>
<tr>
<td><img src="image4" alt="Network" /></td>
<td>$1.353 \frac{2}{2\pi RC}$</td>
<td>$\frac{1}{15.4}$</td>
</tr>
</tbody>
</table>
Figure 3.9: Electrical resistance-temperature curves of resistance temperature detector (RTD) metals [21].

As will be explained in section 3.3.2, amplifier gain gets lower at higher temperatures, because of the reduction of transistor transconductance. Thus excess gain must be provided at room temperature.

Bipolar transistors may be used in high gain amplifiers. However, this thesis focuses on MOSFET amplifiers, for their ease of microfabrication, and because more data is available at this time to understand the behavior of SiC MOSFETs, compared to SiC bipolar transistors. CMOS technology is available with SiC, so that n-channel and p-channel MOSFETs can both be used in the same integrated circuit.

The oscillator of figure 3.3 uses a CMOS inverting high gain amplifier stage, followed by a PMOS non-inverting low gain output stage. This amplifier is isolated and shown again in figure 3.12. Before presenting the details of this amplifier, the basic passive load amplifier will be introduced. This simple amplifier was used in many validation tests throughout the project, as will be described in section 4.1.
Passive Load Amplifier Stage  Figure 3.10 shows two passive load MOSFET amplifier stages. Detailed description of the operation of such amplifier stages is made in basic electronics textbooks, e.g., by Sedra and Smith [22]. The input bias, resistor $R_D$, and voltage $V_{DC}$ are chosen to ensure operation in the saturation regime of the transistor, where maximum gain can be reached. The resulting gain is given by

$$G = \frac{V_{out}}{V_{in}} = \pm g_m \left( R_D \parallel Z_{Load\;in} \parallel r_o \right)$$ (3.3)

where $g_m$ is the transistor transconductance. The symbol $\parallel$ means parallel, the number in parentheses being the equivalent impedance of the parallel network of $R_D$, $Z_{Load\;in}$, and $r_o$. $Z_{Load\;in}$ stands for the input impedance of the load circuit being driven by the amplifier, whereas $r_o$ is a transistor parameter given by $\frac{dV_{DS}}{dI_D}$ in the saturation region. For an ideal transistor, $r_o$ approaches infinity, which means that in the saturation region, I-V characteristics are almost horizontal. To maximize the amplifier gain, $R_D$, $Z_{Load\;in}$, and $r_o$ need all to be maximized.

The gain of the amplifier stage is positive or negative depending on the nature of the transistor channel. The gain is negative in the case of the NMOS amplifier of figure 3.10a, where an n-channel transistor is used. This NMOS amplifier stage is inverting. Increasing
the gate input voltage reinforces the n-channel, causes the transistor current $I_{DS}$ to increase, increases the voltage drop across $R_D$, and hence reduces the output voltage. The AC component of the signal is consequently inverted. The opposite is true for the PMOS amplifier of figure 3.10b, which uses a p-channel transistor to provide a positive gain. The AC signal is not inverted in this latter case.

In order to increase the gain of a passive load amplifier, one might need to increase $R_D$. However, this increase must be accompanied by an increase of $V_{DC}$, to maintain the transistor bias conditions (the DC value of $I_{DS}$ must remain constant, to keep the same transistor operating point). The power dissipated through $R_D$ is consequently increased, which might become unpractical in terms of required power supply resources, or heat generation. Active load amplifiers, like the CMOS amplifier, represent a good solution to this limitation.

CMOS Amplifier Stage  In the CMOS inverting amplifier stage of figure 3.11, n-channel transistor $Q_1$ is still the amplifying device, whereas p-channel transistor $Q_2$ takes place for resistor $R_D$ of the passive load amplifier just described. Transistor $Q_2$ is an active load that provides high equivalent resistance (high $\frac{dV}{dt}$ ratio) without necessitating a high voltage drop across it to maintain proper DC value of $I_{DS}$. Consequently, the gain of this active load amplifier stage is by far higher than that of a passive load amplifier, for the same $V_{DC}$ voltage, or for the same dissipated power. It is given by

$$G = -g_{m1} (r_{o1} \parallel r_{o2} \parallel Z_{Load \ \text{in}})$$  \hspace{1cm} (3.4)

where indices 1 and 2 stand for transistors $Q_1$ and $Q_2$.

Transistor $Q_3$ is used to properly bias the amplifier. Transistors $Q_2$ and $Q_3$ form together a current mirror, so that the current flowing through $Q_2$ (and hence $Q_1$) is equal to the current flowing through $Q_3$.

In the CMOS amplifier, both transistors $Q_1$ and $Q_2$ are operated in their saturation region. Since for most transistors, I-V curves are almost flat in saturation, $r_{o1}$ and $r_{o2}$ are
Figure 3.11: CMOS inverting amplifier with current mirror.

generally very high. However, in order to benefit from the high gain of a CMOS amplifier, as per equation 3.4, $Z_{Load\,in}$ must also be very high, i.e., comparable to $r_{o1}$ and $r_{o2}$.

It is not always practical to increase $Z_{Load\,in}$, especially in the current project. In the oscillator designed here, the feedback network, which is the load driven by the amplifier, needs to be miniaturized as much as possible. Theoretically, when using multilayer capacitors and metallic single-layer resistors, increasing capacitances necessitates less substrate area than increasing resistances. Hence, in order to minimize the size of the feedback network, platinum resistors need to have small values. Capacitors can take larger values, in order to avoid unacceptably high oscillating frequencies. Low resistance and high capacitance values result in low feedback network input impedance, which is not desirable to maximize the amplifier gain. An amplifier output stage is a good solution to this problem, as presented next.

**PMOS Output Stage** In figure 3.12, an output amplifier stage was added to the CMOS amplifier of figure 3.11. The input impedance of this amplifier stage is almost infinite, since
no current is flowing through the gate of transistor $Q_4$. Hence, $Z_{\text{load\ in}}$ is almost infinite in equation 3.4, and the CMOS amplifier provides the highest possible gain.

The output stage is only used to lower the output impedance of the CMOS amplifier. It is not required to provide a high gain. In fact, if the CMOS stage gain is just high enough to ensure oscillations, the output stage must provide a unity gain only, when connected to the driven load.

A PMOS non inverting output stage is preferred in this application, to maintain a total phase-shift of 180° throughout the two consecutive amplifier stages. Two NMOS amplifier stages could have been used in series for a total phase-shift of 540°, which would have theoretically ensured oscillations. However, as will be explained in the following subsection on the ring oscillator, introducing a loop phase-shift of more than 1 cycle in an electronic oscillator can cause undesired oscillation modes to appear. This option was disregarded for that reason.
Operational Amplifiers  Several types of operational amplifiers could be used in the current application, providing the required high gains. However, such amplifiers fall outside the scope of the current master's thesis.

Ring Oscillator

Designing and fabricating an amplifier such as the CMOS amplifier just presented can become rather fastidious. Both NMOS and PMOS transistors are necessary to its construction, which increases the complexity of its design and microfabrication processes, compared to an amplifier that would use only one type of transistors. Moreover, in order to reach the gain of 18.4 that is needed to run the phase-shift oscillator of figure 3.3, transistor $Q_1$ must provide high transconductance, which can be difficult to achieve using today's SiC technology.

The ring oscillator schematized in figure 3.13 would require amplifier stages that would be easier to develop. It uses NMOS transistors only, and relatively low transconductance is required from each of those transistors. In this concept, both the amplifier and the feedback network introduced in figure 3.6 are split in three parts and interleaved. At the theoretical oscillating frequency, which is given by

$$f_o = \frac{1}{2\pi\sqrt{3RC}}$$  \hspace{1cm} (3.5)

each single high-pass filter applies a phase-shift of $60^\circ$. Each amplifier stage adds up a $180^\circ$ phase-shift, for a total loop phase-shift of $720^\circ$, corresponding to two oscillating cycles. A minimum gain of 2 is required from each amplifier stage to ensure oscillation, to compensate for the losses through each filter stage. To provide such a small gain is relatively easy, compared to what is required from the amplifier of a conventional phase-shift oscillator. It can be achieved with a single stage passive load amplifier.

In practice, the ring oscillator is slightly more capricious. Within the current project,
Figure 3.13: Ring type phase-shift oscillator.

Table 3.2: The two oscillating modes of a tested ring oscillator, using Panasonic 2SK1228 transistors

<table>
<thead>
<tr>
<th>Frequency</th>
<th>$\Delta \phi_{\text{amplifier}}$</th>
<th>$\Delta \phi_{\text{filter}}$</th>
<th>$\Delta \phi_{\text{total}}$</th>
<th>$\Delta \phi_{\text{loop}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{e1}$</td>
<td>$\sim 180^\circ$</td>
<td>$\sim 60^\circ$</td>
<td>$240^\circ$</td>
<td>$720^\circ$ (2 cycles)</td>
</tr>
<tr>
<td>$f_{e2}$</td>
<td>$\sim 80^\circ$</td>
<td>$\sim 40^\circ$</td>
<td>$120^\circ$</td>
<td>$360^\circ$ (1 cycle)</td>
</tr>
</tbody>
</table>

it was tested under various conditions, but without success. For example, a ring oscillator was built using three Panasonic 2SK1228 low-signal MOSFET amplifiers like the one of figure 3.14a, and discrete filter components ($C = 1$ nF, $R = 100$ $\Omega$) leading to a theoretical oscillating frequency of 920 kHz as per equation 3.5. Unexpectedly, a signal of almost 4 MHz in frequency was generated by this oscillator, instead of the predicted 920 kHz waveform.

Indeed, a higher frequency mode is obtained with the ring oscillator when built with unideal components. As shown in figure 3.14b, a theoretical ideal amplifier stage would maintain a constant gain and a $180^\circ$ phase-shift throughout the entire frequency domain. In reality, however, the transistor reacts differently at higher frequencies, which results in a reduction of amplifier gain and phase-shift (measured data represented by circles in figure 3.14b). In fact, both filter and amplifier phase-shifts decrease with increasing frequency. At frequency given by equation 3.5, the loop phase-shift totalizes $720^\circ$, as described earlier. On the other hand, as frequency is increased, the total loop phase-shift lowers down and reaches $360^\circ$, which corresponds to another oscillating mode. Table 3.2 describes the two modes of the tested ring oscillator.
Figure 3.14: Real vs theoretical behavior of a single stage passive load amplifier.

Since the high-pass filter produces almost no loss at the frequency of the second oscillating mode, the total loop gain can be higher for the second (undesired) mode than it is for the first (desired) mode. This happened with the tested circuit, explaining why the second oscillating mode was dominant. To ensure the first mode to be the dominant one, high frequency components must be filtered out, or amplifier gain curve must optimized to make the loop gain lower than unity at the frequency of the second mode. The option of filtering was judged to add to many large size components to the circuit. The option of optimizing the amplifier gain was also rejected, considering that in the current project, excess gain is necessary at room temperature for the oscillator to operate at elevated temperatures (the amplifier gain decreases with increasing temperature). This excess gain, in a ring configuration, could cause the oscillator to jump from a second-mode oscillation at low temperature to a first-mode oscillation at high temperature.

Based on those considerations, the ring oscillator concept was discarded. Furthermore, it was decided to avoid all oscillator configurations totalizing more than one cycle phase-shift around their loop. Hence, a maximum of one inverting amplifier stage must be used.
in a phase-shift oscillator, when using high-pass feedback networks.

3.2.2 Wien Bridge Oscillator

A Wien bridge oscillator circuit is schematized in figure 3.15. It is composed of a Wien bridge, and two inverting amplifier stages. At the oscillating frequency \( f_o \), the Wien bridge gives a gain of \( \frac{1}{3} \), and no phase-shift, as shown in the Bode plot of figure 3.16. Each amplifier provides a 180° phase-shift, for a total phase-shift of 360° around the oscillator loop. The oscillating frequency is given by

\[
  f_o = \frac{1}{2\pi RC} \tag{3.6}
\]

As a feedback network, the Wien bridge has the advantage of the ring configuration presented in the last section, but without its drawback. First, the oscillator needs relatively low gain amplifier stages, considering the low losses through the bridge at the oscillating frequency. A total gain of 3 needs to be provided by the two amplifier stages. Secondly, parasitic oscillation modes that may appear due to undesired transistor behavior at high frequency are filtered out by the Wien bridge.
In the event that the CMOS amplifier introduced in section 3.2.1 for the phase-shift oscillator is found to be impossible to realize, the Wien bridge would be the back-up concept. This oscillator would use two low transconductance NMOS transistors only, as sketched in figure 3.15. The phase-shift oscillator remains the prime concept, however, for the reason exposed next. Basically, it can provide a better quality of signal, and consequently ease frequency modulation.

3.2.3 Limiter

Ideally, in order to obtain a perfect sinusoidal signal from any oscillator of the form of figure 3.6, the amplifier gain must be set to reach a loop gain as close as possible to one. In the case of this project, since the amplifier gain is affected by temperature (see section 3.3.2), it is impossible to maintain a loop gain of one in every operating condition, unless some adjustments are continuously made to the amplifier gain. Such adjustments being unpractical, the limiter will be the only controlling feature, at the cost of slightly poorer signal
Figure 3.17: Diode signal clipper.

quality. In most conditions, the amplifier provides higher gain than really necessary, and
the limiter maintains the oscillator within stable operation.

In this project, a diode signal clipper was chosen to limit the oscillating voltage amplitude. This limiter is shown in figure 3.17, along with its effect on signal. Clipping a signal in this way adds up harmonic frequency components above the main signal. For effective frequency modulation, those components must be filtered out as much as possible.

In the case of the phase-shift oscillator described previously (section 3.2.1), the feedback network produces considerable losses, which prevents harmonic components to survive near the main oscillating frequency. In the case of a Wien bridge oscillator (section 3.2.2), however, the feedback network, i.e., the Wien bridge, has only poor filtering close to the oscillating frequency, and the parasitic harmonic components are conserved throughout the oscillator loop. Thus, ideally, another type of limiter must be used with the Wien bridge oscillator, like a non-linear heating element (or lamp), to limit the signal without affecting its shape [20]. Such an element can be connected to limit the current through one of the amplifier lines, as its resistivity increases under heat generated by larger currents.

In order to get clear sinusoidal waveform generation without using a cumbersome non-linear heating element, the phase-shift oscillator was selected to be the prime concept. Nevertheless, the Wien bridge oscillator was used with a diode clipper for macroscopic validation tests described further in section 4.1, and was found to generate relatively clear
sinusoidal waveforms. As mentioned previously, the Wien bridge oscillator is a good back-up concept in case that major problems are faced with the phase-shift oscillator. It is simpler, and could as well be used in a first microcircuit prototype.

3.3 Integrated Circuit Development

3.3.1 Feedback Network

The capacitors and platinum resistors occupy a major portion of the integrated circuit. As the frequency to be generated by the oscillator is reduced, the size of the feedback network components increases. After doing trade-offs to minimize the size of the lowest frequency integrated circuit, and fixing a limit to 1 mm², 60 kHz was found to be the lowest possible oscillating frequency at room temperature, with the phase-shift oscillator concept. Capacitors and resistors of 200 pF and 10 kΩ respectively will be used in the feedback network to produce this frequency.

Figure 3.18 shows one possible integrated circuit layout, using those capacitor and resistor values. Detailed design of this circuit was not completed yet, but the layout shown here uses realistic dimensions. The amplifier and limiter are buried underneath the platinum resistors, to save on substrate space. This can be done readily by depositing and polishing an oxide layer onto the semiconductor components, before depositing the platinum resistors.

Microfabrication limitations of the clean rooms of Université de Sherbrooke were used to determine the smallest possible resistor cross section. Lower oscillating frequencies could be achieved in the future, using finer resolution equipment. Lower frequencies could also be achieved using the Wien bridge oscillator concept, which necessitates less components.
Figure 3.18: Layout of a temperature sensitive integrated circuit phase-shift oscillator generating a 67 kHz signal at room temperature.
Capacitors

Capacitors will be composed of a stack of layers of conducting an dielectric materials. SiO₂ will be used as the dielectric, whereas doped substrate SiC and polysilicon will act as electrodes. Conventional flat plate capacitors do not offer a sufficiently high capacitance to substrate area ratio. In order to reach 200 pF with reasonable size capacitors, multiple layers of lateral flux capacitor geometries will need to be microfabricated. A combination of lateral and transversal electric fields will be used, in a configuration that may resemble the one of figure 3.19. Lee [23] claims that capacitance increases of a factor of 10 can be achieved in some cases, when maximizing perimeter to area ratio of lateral flux capacitors. For the intent of the current project, an increase factor of five is targeted over the flat plate capacitor value. Hence,

\[ C = \frac{\varepsilon_r \varepsilon_0 A}{t_{\text{dielectric}}} \eta \quad (3.7) \]

where \( \eta \) is the multiplying factor taken as equal to five, \( \varepsilon_r \) is the relative dielectric constant of SiO₂ (3.85), \( \varepsilon_0 \), the permittivity of free space, \( A \), the covered substrate area, and \( t_{\text{dielectric}} \), the dielectric thickness between conducting layers. With \( \eta = 5 \) and \( t_{\text{dielectric}} = 50 \) nm, a 200 pF square capacitor will use up 242 \( \times \) 242 \( \mu \)m of substrate area. A slightly different aspect ratio is used in the layout of figure 3.18, with the same area.

Platinum Resistors

Considering the relatively high electrical conductivity of platinum, each resistor of the feedback network will take the shape of a long serpentine wire with a small cross section. The resistance of this serpentine is given by

\[ R = \frac{\rho_{\text{Pt}} L}{A_{\text{CS}}} \quad (3.8) \]
Figure 3.19: Integrated circuit capacitor top view (top layer only) and cross section (two dielectric layers illustrated here).
Figure 3.20: Integrated circuit platinum resistor top view.

where $\rho_{Pt}$ is the resistivity of pure platinum taken as $1.035 \times 10^{-7}$ $\Omega$m at 20°C, $L$, the total length of all serpentine segments, and $A_{CS}$, the cross section area of the wire. The deposition thickness is taken as 100 nm, and the lithography resolution, 2 $\mu$m. Both the width of the resistor segments and the spacing between them are chosen to be 2 $\mu$m, as shown in figure 3.20. For a square resistor of 10 k$\Omega$, which is the maximum value needed here, $278 \times 278 \mu$m of substrate area will be required. To minimize the parasitic inductance of this stack of conducting segment wires, the aspect ratio of the resistor will need to be properly chosen. An aspect ratio of four was arbitrarily used in the layout of figure 3.18, for drawing convenience.

In terms of microfabrication, platinum will be deposited using physical vapor deposition (PVD), over a SiO$_2$ insulating layer. Apart from being insulating, SiO$_2$ represents a chemically stable substrate for platinum. Indeed, direct deposition over SiC could result in film degradation at temperatures above 500°C, due to formation of platinum silicides [24]. Depositing over SiO$_2$ eliminates this risk.

The power dissipated by each platinum resistor causes self-heating. The first resistor of
the feedback network (located to the left in figure 3.3) is the one that dissipates the most power. Based on simulation using PSPICE software, this resistor reaches power peaks of less than 100 $\mu$W. In the event that this resistor suffers from overheating, it may need to be made of wider platinum segments than the other resistors. Any configuration using multiple resistance values for the four resistors of the feedback network may also be investigated in terms of heat transfer.

3.3.2 Amplifier and Limiter

SiC Resistors

The resistors used in the amplifier and limiter (see figures 3.12 and 3.17) will be doped directly into the semiconductor substrate, using nitrogen ion implantation. The resistivity of the resulting n-type SiC is given by

$$\rho = \frac{1}{q\mu_n n}$$  \hspace{1cm} (3.9)

where $q$ is the elementary electronic charge, $\mu_n$ is the electron mobility in SiC, and $n$ is the free (mobile) electron concentration. In this equation, the effect of hole mobility is neglected, since the semiconductor is n-doped. The mobility $\mu$, by definition, relates the carrier velocity to an applied electric field:

$$v = \mu E$$  \hspace{1cm} (3.10)

Roschke and Schwierz [25] summarized the experimental and theoretical work that has been conducted to determine the electron mobility in 3C, 4H, and 6H-SiC at various temperatures. They presented semi-empirical equations for each polytype to describe the temperature dependence of this mobility. The field-dependent mobility is given by
\[ \mu_n = \frac{\mu_{n0}}{\left[1 + \left(\frac{\mu_{n0}E}{v_{sat}}\right)^\beta\right]^\beta} \]  

(3.11)

where the low-field (field-independent) mobility \( \mu_0 \) is given by

\[ \mu_{n0} = \mu_{min} + \frac{\mu_{max} - \mu_{min}}{1 + \left(\frac{N}{N_{ref}}\right)^\alpha} \]  

(3.12)

In the latter equation, each of the parameters follows a temperature dependency. For 4H-SiC,

\[ \mu_{max} = \frac{950 \text{cm}^2}{V_s} \left(\frac{T}{300 \text{K}}\right)^{-2.4} \]

\[ \mu_{min} = \frac{40 \text{cm}^2}{V_s} \left(\frac{T}{300 \text{K}}\right)^{-0.5} \]  

(3.13)

\[ N_{ref} = 2 \times 10^{17} \text{cm}^{-3} \left(\frac{T}{300 \text{K}}\right) \]

\[ \alpha = 0.76 \]

\( \beta \) and \( v_{sat} \) also follow temperature-dependent relations:

\[ \beta = \beta_0 + a \times \exp \left(\frac{T - T_0}{b}\right) \]  

(3.14)

\[ v_{sat} = \frac{v_{max}}{1 + 0.6 \times \exp \left(\frac{T}{600 \text{K}}\right)} \]  

(3.15)

where, again for 4H-SiC, \( \beta_0 = 0.816 \), \( a = 4.27 \times 10^{-2} \), \( b = 98.4 \), \( T_0 = 327 \text{K} \), and \( v_{max} = 4.77 \times 10^7 \text{ cm/s} \).

The last expressions describe the mobility in 4H-SiC in a direction perpendicular to the c-axis of the crystal. The mobility along the c-axis would be about 25% higher, according to Roschke and Schwierz [25].

54
Figure 3.21: Variation of SiC resistor value with temperature.

Mobilities generally decrease as the temperature increases, because of the increased lattice scattering [15]. Collisions between free electrons and vibrating lattice elements are more frequent at higher temperature. This causes semiconductor resistances to get larger at high temperature. In the same order of idea, as the doping concentrations are increased, defects are more numerous in the crystal lattice, and hence mobilities are reduced.

Figure 3.21 shows the variation of resistance of an n-type SiC resistor with temperature, for multiple doping levels. Low electric field is assumed through the resistor: equation 3.12 is used in equation 3.9 to predict this temperature dependence.

According to equation 3.3, increasing the value of resistor $R_D$ in a passive load amplifier (see figure 3.10) increases the gain of this amplifier stage. One might then conclude that the temperature effect on this resistor, making it increase, is beneficial for amplifier gain, and does not represent a problem for the design of an oscillator. This is not exactly true. In fact, as $R_D$ is increased, the current $I_{DS}$ flowing through it an through the transistor
gets lower. In turn, the transistor transconductance \( g_m \) gets affected negatively, since it is proportional to the square root of \( I_{DS} \), as will be seen next in equation 3.16. Furthermore, if the amplifier stage is delivering power to a relatively small load, e.g., if \( R_D \gg Z_{\text{Load in}} \), an increase of \( R_D \) will have low positive impact on \( R_D \ || Z_{\text{Load in}} \ || r_0 \), in equation 3.16, and will mainly affect the amplifier gain negatively through a reduction of \( g_m \).

Considering this, it is essential to design semiconductor resistors of the amplifier stages to reach their design values at their average operating temperature, in order to obtain the desired transistor bias conditions. Moreover, in order to minimize the impact of resistor variations within the operating range, doping levels must be kept high. The drawback of this last suggestion is that higher doping levels reduce resistivity and cause resistors to cover larger substrate area.

SiC MOSFET Technology

Transconductance Carrier mobility does not only modify the properties of semiconductor resistors. It is also a key parameter in calculating the transconductance \( g_m \) of amplifier field effect transistors. \( g_m \) is needed to calculate the gain of amplifier stages, as was seen in equations 3.3 and 3.4. From Sedra and Smith [22], the transconductance of a MOSFET can be approximated by

\[
g_m = \sqrt{2\mu C_{OX} \frac{W}{L} I_{DS}}
\]  

(3.16)

where \( C_{OX} \) is the gate oxide capacitance per unit area, \( W \) and \( L \) are the width and length of the MOSFET channel, respectively, and \( I_{DS} \) is the current flowing from the drain to the source (in the case of an n-channel device).

Here, \( \mu \) stands for the channel carrier mobility (\( \mu_n \) for an n-channel, \( \mu_p \) for a p-channel). This mobility cannot be directly estimated by equation 3.11 for n-channel devices. In reality, equation 3.11 gives values of bulk mobility, which is useful to assess resistivity of a doped
semiconductor with no transverse electric field, far from substrate surfaces. The channel of a MOSFET being created just besides the oxide-semiconductor interface, below the gate contact, interface traps and gate transverse electric field reduce the carrier mobility. Channel mobilities are indeed reported to be much lower than bulk mobilities, in similar conditions (temperature, longitudinal electric field, and doping concentration).

Alike the bulk mobility, the channel mobility is expected to decrease under a temperature increase. Resulting reductions of transconductance have indeed been observed experimentally in many SiC transistors at high temperatures, e.g., by Palmour et al. [18]. However, the same authors reported that for some devices, at low temperatures up to \(200^\circ\text{C}\), \(g_m\) increased with temperature, before reaching a maximum and starting to decrease. This phenomenon, not observed in silicon devices, would be explained by the poor quality of the SiC - SiO\(_2\) interface under the gate, providing higher carrier trap densities. Increasing temperature would tend to reduce the affinity of electrons with those traps, and then increase channel mobility to a certain extent. Up to a certain transition temperature, this freeing of electrons would over-compensate for the reduction in mobility expected from general lattice scattering.

In order to improve the room temperature channel mobility, Lipkin and Palmour [26] suggested to reduce the temperature of the gate oxide growth process and to conduct a re-oxidation anneal. This procedure led to a record channel mobility of 72 \(\text{cm}^2/\text{V}\cdot\text{s}\), and to superior gate oxide breakdown field. In the same conditions, at room temperature (300 K), for a channel length of 14 \(\mu\text{m}\) and \(V_{DS} = 0.1\ \text{V}\), which gives a longitudinal electric field of 71 V/cm, and for a doping level of \(10^{17}\ \text{cm}^{-3}\), equation 3.11 gives a mobility of 600 \(\text{cm}^2/\text{V}\cdot\text{s}\), i.e., about ten times the one measured by Lipkin and Palmour.

n-channel depletion devices benefit from higher channel mobilities than n-channel inversion devices. Indeed, in the former case, the gate transverse electric field tends to deplete an already existing channel from its electrons, hence repulsing electrons out from the interface traps. In the latter case, a positive voltage is applied onto the gate, which causes a channel to form besides the interface. Electrons then get trapped more easily in this case. Conse-
quently, depletion devices (conducting at zero gate voltage) must be preferred to inversion (enhancement) devices (non conducting at zero gate voltage), if high transconductance is targeted. In all cases, channel mobility depends on applied electric field, and hence on applied gate voltage.

**Threshold Voltage** Threshold voltages of transistors can be estimated from theory using various semiconductor, oxide, and gate material parameters. In an amplifier like the one of figure 3.12, threshold voltages need to be carefully set to obtain the desired transistor bias conditions. Gate materials and channel doping levels can be properly chosen to set those threshold voltages.

Adjusting the threshold voltage of SiC devices is quite different from adjusting it for silicon devices. This results from the differences in the energy band distribution between Si and SiC. The electronic affinity is similar for both materials ($\chi_{\text{Si}} = 4.01$ eV [27], and $\chi_{\text{SiC}} = 4.0$ eV [28]), but the band gaps are very different ($E_{\text{gSi}} = 1.125$ eV, and $E_{\text{gH-SiC}} = 3.3$ eV).

Using the energy diagrams, the basic equations are presented here for two types of transistors: the n-channel depletion transistor and the n-channel enhancement transistor. Those two examples illustrate together the particularities of threshold voltage setting for SiC transistors.

**n-channel Depletion Transistor** The metal-oxide-semiconductor (MOS) structure of the considered depletion transistor is schematized in figure 3.22. n-doped SiC substrate composes the channel of the device. A polysilicon gate is deposited onto the gate oxide, and is degenerately doped n. All oxide parasitic charges, including charged interface states, border traps, mobile ions, and fixed charges, are gathered in the net oxide charge $Q_0$ ($\text{C/cm}^2$) located at the SiC-SiO$_2$ interface. Lipkin and Palmour [26] obtained a net charge oxide density $Q'_0$ of $1.0 \times 10^{12}$ cm$^{-2}$ ($Q_0 = qQ'_0$), which number can be used as a good starting point.
Figure 3.22: MOS structure of an n-channel depletion transistor.

Figure 3.23 compares the energy levels of the gate and substrate materials, with respect to the vacuum energy level $E_0$. When not in contact, the two materials have different Fermi levels $E_F$. The polysilicon gate has a higher Fermi level, which means that its electrons occupy higher energy levels. $\phi_f$ characterizes the departure of the semiconductor material from its intrinsic (undoped) state, and depends on the doping level:

$$\phi_f = kT \ln \frac{N_D}{n_i}$$  \hspace{1cm} (3.17)

where $k$ is the Boltzmann constant, $T$ is the temperature, $N_D$ is the donor (or n-dopant) concentration, and $n_i$ is the intrinsic carrier concentration, already calculated in section 3.1.3, using equation 3.1.

The two materials can be brought together to contact a common oxide layer, to form the MOS structure of figure 3.22. If a short circuit is provided between them, by setting $V_{GB} = 0$, electrons naturally migrate from the polysilicon gate to the SiC substrate, using this short circuit, until a constant Fermi level is reached throughout the entire structure.
Figure 3.23: Energy levels of free polysilicon gate and n-SiC materials relative to the vacuum energy level $E_0$.

as illustrated in figure 3.24a. Then an energy equilibrium is reached by the electrons in the structure. However, this electron migration causes a voltage difference $\phi_{MS}$ to arise between the gate and the substrate ($V_{Gate} > V_{SiC}$). This built up potential difference is given by

$$q\phi_{MS} = \Phi_{SiC} - \Phi_{poly-Si}$$

(3.18)

where $\Phi_{SiC}$ and $\Phi_{poly-Si}$ are the work functions of the two materials, as defined in figure 3.23.

Under this voltage difference, electrons accumulate at the oxide-semiconductor interface, in the semiconductor, attracted by the higher voltage of the gate. This produces the semiconductor energy bands to bend as sketched in figure 3.24a. The n-channel is present under $V_{GB} = 0$. The transistor is on, with zero gate voltage (the gate being short-circuited with the substrate).

In order to counteract the voltage difference and band bending naturally built up by the electron migration, a negative $V_{GB}$ voltage needs to be applied (to lower voltage onto the gate, relative to the substrate). The n-channel then starts to deplete, the accumulated electrons being repulsed from the oxide-semiconductor interface. The transistor current is reduced until it reaches the flat band condition, as illustrated in figure 3.24b. At this
point, the transistor is still conducting current, since the channel still contains a majority of electron carriers. Indeed, the semiconductor Fermi level $E_F$ is still higher than the intrinsic Fermi level $E_i$, at the SiC-SiO$_2$ interface, which means that free electrons are still in majority, compared to holes. Those electrons are free to move anywhere in the semiconductor, and are no longer confined close to the interface. No electric field nor voltage variations are present in the semiconductor in this flat band condition.

Throughout the oxide layer, however, an electric field is present, because of the parasitic charges $Q_0$. In order to reach the flat band condition, those positive charges, modelled to be located at the SiC-SiO$_2$ interface, need to be counter-balanced by negative charges at the poly-Si-SiO$_2$ interface, which produces the oxide electric field.

If $V_{GB}$ is further reduced, opposite band bending occurs, as electrons are leaving the channel. Inversion starts to occur, which means that electrons are no longer in majority within the channel: holes are starting to create. Figure 3.24c shows the inversion onset condition, where holes are as numerous as free electrons in the channel. At this point, a voltage drop of $\phi_f$ is present through the semiconductor, from its bulk to the oxide interface. Also, the oxide electric field has increased.

As defined by Pierret [15], the threshold voltage is reached when the semiconductor Fermi level $E_F$ gets $\phi_f$ below the intrinsic level $E_i$ at the oxide-semiconductor interface, as sketched in figure 3.24d. At this stage, the voltage drop from bulk SiC to SiC-SiO$_2$ interface is equal to $2\phi_f$. Holes start to accumulate in the channel, and the transistor is considered to be off. $V_{GB}$ is equal to the threshold voltage $V_T$. $V_T$ is given by

$$V_T = -\phi_{MS} - 2\phi_f - \phi_{ox}$$

(3.19)

In this equation, $\phi_f$ and $\phi_{MS}$ can be obtained using equations 3.17 and 3.18. $\phi_{ox}$ can be obtained as described by Pierret [15]. First, the shape of the semiconductor energy bands in figure 3.24d is approximated by the delta-depletion solution, which leads to the knowledge of the electric field in the semiconductor. Then, a boundary condition is applied
Figure 3.24: Energy band diagrams of an n-channel depletion MOS structure.

to relate the oxide electric field to the semiconductor electric field, at the interface, using the dielectric constants of the two materials ($\varepsilon_r\text{SiC}\varepsilon_0E_{\text{SiC}} - \varepsilon_r\text{SiO}_2\varepsilon_0E_{\text{SiO}_2} = Q_0$). Finally, if the electric field is considered to be constant throughout the oxide layer, as it is the case for an ideal dielectric material, $\phi_{\text{ox}}$ can be calculated as $\phi_{\text{ox}} = -E_{\text{SiO}_2}t_{\text{ox}}$, $t_{\text{ox}}$ being the thickness of the oxide layer. The final result is

$$\phi_{\text{ox}} = \frac{\varepsilon_r\text{SiC}}{\varepsilon_r\text{SiO}_2}t_{\text{ox}}\sqrt{\frac{2qN_D}{\varepsilon_r\text{SiC}\varepsilon_0}(2\phi_f)}$$  \hspace{1cm} (3.20)

where $\varepsilon_r\text{SiC}$ and $\varepsilon_r\text{SiO}_2$ are the relative dielectric constants of SiC and SiO$_2$, and $\varepsilon_0$ is the permittivity of free space. $\varepsilon_r\text{SiC}$ was measured to be 9.66 perpendicular to a 6H-SiC crystal by Patrick and Choyke [29], which may be used as a good first assumption in equation 3.20 for a 4H-SiC device.
Figure 3.25: Inversion layer created in a MOS structure composed of a poly-Si gate and p-SiC semiconductor at $V_{GB} = 0$.

**n-channel Enhancement Transistor** The n-channel enhancement transistor must not conduct current under a zero gate voltage. The channel must be doped p, and accumulate electron carriers only when a positive $V_{GB}$ voltage is applied. As was seen in figure 3.24a, with an n-polysilicon gate and n-SiC, an electron accumulation layer is formed in the channel at $V_{GB} = 0$, because of the $\phi_{MS}$ potential that is naturally built up when electrons are transferred to reach a constant Fermi level. Even if the SiC substrate is doped p, if the $\phi_{MS}$ potential is high enough, it is possible to obtain an electron accumulation layer (which would here be called an inversion layer) in the channel. Figure 3.25 illustrates this situation.

No matter if the polysilicon gate were doped p (higher $\Phi_M$), the $\phi_{ms}$ potential would still be high enough to create an inversion layer. The solution proposed here, to obtain an enhancement transistor that does not conduct at $V_{GB} = 0$, is to use platinum as the gate material. Platinum has the highest work function of all metals ($\Phi_M = 5.65$ eV [30]). Its work function is also similar to the work function of intrinsic SiC, which ensures that no inversion layer will be formed in the proposed MOS structure. The MOS structure is schematized in figure 3.26. Figure 3.27 shows the energy bands of the two materials in free states, and illustrates the comparison between their work functions.

Once again, $\phi_f$ characterizes the doping level of the semiconductor, but is negative in
Figure 3.26: MOS structure of an n-channel enhancement transistor.

this case:

$$\phi_f = -kT \ln \frac{N_A}{n_i}$$  \hspace{1cm} (3.21)$$

where $N_A$ is the acceptor (or p-dopant) concentration.

As was done previously for the depletion device, the two materials can be brought together, with only a gate oxide layer between them. Again, if a short circuit is provided between the gate and the substrate, by setting $V_{GB} = 0$, electrons start to migrate from gate to substrate until a constant Fermi level is reached throughout the structure. The SiC energy bands bend as sketched in figure 3.28a. This bending is not enough to create an inversion layer (the transistor is still off), but it is enough to remove a portion of the holes already present in the channel. The voltage difference $\phi_{MS}$ is still given by equation 3.18, and is still positive.

In order to repulse even more holes from the channel and to build up an electron inversion
layer, a positive $V_{GB}$ voltage needs to be applied. This accentuates the band bending of figure 3.28a until the threshold condition of figure 3.28b is reached. At this threshold condition, the transistor is turned on. Note the similarity between figures 3.25 and 3.28b. The difference is that with the polysilicon gate, inversion is reached without applying any voltage ($V_{GB}$), whereas with the platinum gate, inversion necessitates a positive gate voltage to be applied ($V_{GB} > 0$).

The threshold voltage $V_T$ is once again given by equation 3.19. It is now positive, since $\phi_f$ and $\phi_{ox}$ are negative in this case. $\phi_{ox}$ can still be obtained using the methodology described by Pierret [15], which leads to

$$\phi_{ox} = -\frac{\epsilon_{rSiC}}{\epsilon_{rSiO_2}} \phi_{ox} \sqrt{\frac{2qN_A}{\epsilon_{rSiC}e_0} (-2\phi_f)}$$

(3.22)

In the event that pure platinum does not turn out to be a reasonable gate material choice in terms of thermo-mechanical stress, because of the mismatch between the thermal expansion coefficients of platinum and SiO$_2$ (and SiC), platinum silicides (PtSi, Pt$_2$Si) might represent good alternatives. The work function of such silicides is comparable to the one of pure platinum. Values between 5.15 and 5.75 eV were measured experimentally and are reported by Mohammadi [31].
3.3.3 High Temperature Testing

Palmour et al. [18] have tested their SiC transistors at high temperature using a special apparatus. The wafers were placed on a boron nitride block containing resistance heaters. Temperature was controlled using a feedback controller, with a thermocouple embedded in the heating block near the wafer contact surface. Palladium probe tips were used for measurements. Such an apparatus will be needed here for integrated circuit high temperature characterization.

3.3.4 Integration and Packaging

As presented in section 3.1.1, the resulting product will resemble a sensitive wire, with integrated sensors. Along the same two leads, about ten integrated circuits similar to the one shown in figure 3.18 will be connected in parallel. The leads and the circuits will all be protected using a common flexible ceramic sheathing. The resulting sensor array will be bonded to the metal using common ceramic cement.
3.4 Sensor Array Design

3.4.1 Frequency Bands

In order to ease frequency modulation, the designer of a sensor array may want to make sure that every oscillator keeps its signal within a dedicated frequency band, in all operating conditions, as suggested in section 3.1.1. This imposes a limit to the number of measuring points per sensor array.

Knowing the thermal behavior of platinum allows to organize the multiple frequency bands of a complete sensor array, using this approach of isolated frequency bands. The theoretical calibration curves (frequency vs temperature) of the sensor array are defined at the same time. From experiments, in the range of 0 to 850°C, the resistance of a platinum resistor is given by

\[ R = R_0 \left(1 + AT + BT^2\right) \]  \hspace{1cm} (3.23)

where \( A = 3.9083 \times 10^{-3} \, ^\circ C^{-1} \), and \( B = -5.775 \times 10^{-7} \, ^\circ C^{-2} \) \[32\]. \( R_0 \) is the resistance at a reference temperature \( T_0 \) of 0°C. This relation can be approximated by

\[ R \approx R_0 [1 + \alpha (T - T_0)] \]  \hspace{1cm} (3.24)

where \( \alpha \) is the average temperature coefficient, which depends on reference \( T_0 \) and on the temperature range to be covered.

Each electronic oscillator of a sensing array covers a dedicated frequency range, as shown in figures 3.1 and 3.2. At the middle of this range, each sensor has a design frequency \( f_d \), at a design temperature \( T_d \). In the phase-shift oscillator concept selected here, the oscillating frequency is inversely proportional to \( R \). Hence, for each sensor,
\[
\frac{f}{f_d} = \frac{R_d}{R} \approx \frac{R_0 [1 + \alpha (T_d - T_0)]}{R_0 [1 + \alpha (T - T_0)]}
\]  

(3.25)

leading to

\[
f \approx f_d \cdot \frac{1 + \alpha (T_d - T_0)}{1 + \alpha (T - T_0)}
\]  

(3.26)

One way of spacing the sensors in the frequency domain is to make consecutive sensors to have coincident oscillating frequencies when reaching corresponding opposite temperature limits. Mathematically, if every sensor has the same design temperature \(T_d\), and the same operating range \(\Delta T_{\text{range}}\), this can be expressed by

\[
f_i \left(T_d + \frac{\Delta T_{\text{range}}}{2}\right) = f_{i-1} \left(T_d - \frac{\Delta T_{\text{range}}}{2}\right)
\]  

(3.27)

Using equation 3.26, it follows that the ratio \(r\) of two consecutive design frequencies can be obtained from

\[
r = \frac{f_{di}}{f_{d(i-1)}} = \frac{1 + \alpha \left(T_d + \frac{\Delta T_{\text{range}}}{2} - T_0\right)}{1 + \alpha \left(T_d - \frac{\Delta T_{\text{range}}}{2} - T_0\right)}
\]  

(3.28)

The design frequencies then obey a geometric series of ratio \(r\), such that

\[
f_{di} = f_{d1} \cdot r^{i-1}
\]  

(3.29)

As a consequence, if \(f_{d\text{max}}\) is the maximum possible frequency that can be generated, the number \(n\) of measuring points that can be integrated to a sensor array needs to satisfy

\[
n < \frac{\ln \left(\frac{f_{d\text{max}}}{f_{d1}}\right)}{\ln r} + 1
\]  

(3.30)
Figure 3.29: Calibration curves of the sensitive wire of figure 3.1.

The sensitive wire of figure 3.1 was designed using this methodology, with $T_d = 700^\circ C$, $\Delta T_{range} = 300^\circ C$, $f_{d1} = 30$ kHz, $f_{d_{\max}} = 600$ kHz, and $\alpha = 3.85 \times 10^{-3}$ $^\circ C$. The resulting theoretical response curves are graphed in figure 3.29. Once again, within the range of operation of the sensor array, each oscillator always maintains itself inside its dedicated frequency band. Another valuable approach would be to allow frequency bands to overlap each other, and to use temporal signal evolution to distinguish measurements from one another. More numerous sensors could be integrated in a single array by using this second approach. This approach is not developed further in this thesis, however.

3.4.2 Sensitivity

To appreciate the sensitivity of the sensor, equation 3.26 needs to be differentiated with respect to $T$: 
Table 3.3: Frequency peak width, as obtained by FFT of a perfect sinusoidal signal in Matlab, as a function of signal duration

<table>
<thead>
<tr>
<th>Signal duration (periods)</th>
<th>Peak width $\Delta f_{\text{peak}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>4 %</td>
</tr>
<tr>
<td>100</td>
<td>2 %</td>
</tr>
<tr>
<td>200</td>
<td>1 %</td>
</tr>
<tr>
<td>1000</td>
<td>0.2 %</td>
</tr>
</tbody>
</table>

\[
\frac{df}{dT} = \frac{-\alpha f}{1 + \alpha (T - T_0)}
\]  
(3.31)

The ratio of the relative frequency variation to the temperature variation is then given by

\[
\frac{\Delta f}{f \Delta T} = \frac{-\alpha}{1 + \alpha (T - T_0)}
\]  
(3.32)

This ratio depends only on the measured temperature, and is identical for every sensor of an array. As examples, at 400 and 800°C, it is equal to -0.1 and -0.15 %/°C, respectively.

The resolution of the sensor is in part limited by the resolution of the frequency measuring system. Fundamentally, to improve this resolution, the acquisition time needs to be increased. The widths of the peaks, in the frequency domain, get narrower as the frequency analysis is performed over a longer time interval. Using the FFT command in Matlab, the effect of signal duration over frequency peak width was examined. Table 3.3 shows the results, and suggests that if the FFT is conducted over 200 periods, a frequency resolution better than 1 % can be expected. For a sensor operating at 800°C, this would mean a temperature resolution of less than 6.67°C. This would satisfy the basic turbine requirement of table 2.2.
Chapter 4

Results

4.1 Macroscopic Validation

4.1.1 Circuit Design

In order to validate the concept of temperature sensor array presented here, two high temperature macroscopic oscillators were built and tested. High temperature discrete components were assembled on two separate ceramic substrates, one of the resulting circuits being shown in figure 4.1.

RF high power SiC MESFETs from Cree Inc. [33] were used for amplifying. Those transistors providing relatively low transconductance at low power, the Wien bridge configuration was selected for the oscillator, because it allows for low gain amplifier stages.

The circuit schematics was already presented in figure 3.15. The discrete components were chosen or designed based on their resistance to high temperature, and are listed in table 4.1, using the identification symbols of figure 3.15. The ring configuration of figure 3.13 was also tested using similar high temperature components, but was found to oscillate at an undesired mode of about 18 MHz. This configuration was discarded for the reasons exposed
Figure 4.1: Macroscopic high temperature Wien bridge oscillator prototype.
Table 4.1: Macroscopic high temperature prototype discrete components

<table>
<thead>
<tr>
<th>Component</th>
<th>Symbol</th>
<th>Supplier</th>
<th>Part number or description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wien bridge components</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$R$</td>
<td>Omega</td>
<td>A2132</td>
<td>500 $\Omega$</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>A2142</td>
<td>1000 $\Omega$</td>
</tr>
<tr>
<td></td>
<td>$C$</td>
<td></td>
<td>Handmade, 20mm x 20 mm stacking of 0.004&quot; thick pyrex sheets with 0.062&quot; thick copper sheets</td>
<td>600 pF</td>
</tr>
<tr>
<td>Transistors</td>
<td>$Q_1, Q_2$</td>
<td>Cree</td>
<td>CRF-22010-101 or CRF-24010-101 MESFET</td>
<td>N/A</td>
</tr>
<tr>
<td>Amplifier resistors</td>
<td>$R_D$</td>
<td>Vishay Dale</td>
<td>CL-4125DA</td>
<td>800 $\Omega$</td>
</tr>
<tr>
<td></td>
<td>$R_S$</td>
<td>Vishay Dale</td>
<td>CL-4125DA</td>
<td>150 $\Omega$</td>
</tr>
<tr>
<td>Coupling filter components</td>
<td>$C_P$</td>
<td></td>
<td>Same as Wien bridge capacitor C</td>
<td>600 pF</td>
</tr>
<tr>
<td>Limiter resistors</td>
<td>$R_F$</td>
<td>Huntington Electric</td>
<td>ALSR-5</td>
<td>25 k$\Omega$</td>
</tr>
<tr>
<td></td>
<td>$R_C$</td>
<td>Huntington Electric</td>
<td>ALSR-1</td>
<td>20 $\Omega$</td>
</tr>
<tr>
<td>Diodes</td>
<td></td>
<td>Cree</td>
<td>CSD010060A</td>
<td>N/A</td>
</tr>
<tr>
<td>Wire</td>
<td></td>
<td></td>
<td>Steel AWG 21</td>
<td></td>
</tr>
<tr>
<td>Wire sheathing</td>
<td></td>
<td>Omega</td>
<td>XC4-116 (Nextel 440)</td>
<td></td>
</tr>
<tr>
<td>Connections</td>
<td></td>
<td></td>
<td>Steel fasteners, aluminum terminals</td>
<td></td>
</tr>
<tr>
<td>Substrate</td>
<td></td>
<td>McMaster-Carr</td>
<td>8489K54 (Macor)</td>
<td></td>
</tr>
<tr>
<td>Power supply</td>
<td>$V_{DC}$</td>
<td>-</td>
<td>-</td>
<td>80 V</td>
</tr>
</tbody>
</table>

in section 3.2.1.

The basic amplifier stage used in both circuits was numerically designed using the I-V characteristics of the SiC transistor. In order to minimize the operating voltages, the dissipated power, and hence heating of the components, the Cree transistor was used in the triode region, very close to the threshold voltage. Figure 4.2 shows the operating point of transistor $Q_1$ over the corresponding I-V curves. A standard curve tracer was used to measure the transistor characteristics at room temperature. Parametric analysis was performed by varying the values of amplifier resistors $R_S$ and $R_D$, in order to find an acceptable amplifier design.

The Wien bridges of the two tested oscillators are described in table 4.2. The handmade 600 pF pyrex-copper capacitors demonstrated sufficiently good behavior to keep bridge losses to an acceptable level at the operating frequencies. Figure 4.3 shows the Bode plot of
Figure 4.2: Operation of transistor $Q_1$ in the macroscopic high temperature Wien bridge oscillator, at room temperature.
Table 4.2: Wien bridge characteristics of the two macroscopic high temperature prototypes

<table>
<thead>
<tr>
<th>Oscillator</th>
<th>$R$ (Ω)</th>
<th>$C$ (pF)</th>
<th>Oscillating frequency (kHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1500</td>
<td>600</td>
<td>215</td>
</tr>
<tr>
<td>2</td>
<td>1000</td>
<td>600</td>
<td>258</td>
</tr>
</tbody>
</table>

a. Room temperature frequency.
b. Made from two platinum resistors (1000 Ω and 500 Ω) connected in series.

In order to provide a multiplexed signal, the two circuits were connected as shown in figure 4.4. A resistor $R_M$ of 20 Ω was used to measure the total current going through all the amplifier stages. It was connected in the 80 V power supply line, which was found to be the best way to limit its influence on the behavior of the oscillators. If it were placed between the oscillator lowest voltage line and ground, a very small value resistor $R_M$ would be required in order to keep the disturbances low enough, and hence very low power signal would be collected as the output.

The room temperature signals of the two oscillators are plotted in figure 4.5. All signals were measured using the current measuring method of figure 4.4, except that signals in (a) and (b) were obtained by feeding one oscillator at a time.

The shapes and amplitudes of the two individual signals can be understood by looking closely at the operation of the amplifiers. For each oscillator, two inverting amplifier stages are fed with the same current carrying line. The behavior of those two amplifier stages are quite different, in terms of alternative signal. The alternative currents withdrawn by each of them are different in magnitude and phase, and add up together to compose the measured signal. The fact that the signal shows unsymmetrical waves (about the y-axis) suggests that the two currents do not have a phase difference of exactly 180°, as expected from theory. Furthermore, the current going through the second amplifier stage strongly depends on the Wien bridge input impedance, which explains why the two oscillators show different output waveforms. Oscillator 1 uses a higher impedance bridge, which causes the gain of the second
Figure 4.3: Bode plot of a macroscopic high temperature Wien bridge prototype, at room temperature, using handmade 600 pF capacitors, and 500 Ω resistors.
Figure 4.4: Two multiplexed Wien bridge oscillators.
Figure 4.5: Waveforms of the two tested macroscopic high temperature prototypes, measured at room temperature.
stage to be higher. Hence, the alternative current withdrawn by the second stage is more prominent in the case of oscillator 1. This also explains why oscillator 1 produces a higher amplitude signal than oscillator 2, even if the two oscillators use identical diode limiters.

4.1.2 Data Acquisition and Analysis

In the concept presented here, the sensor array is fed with DC voltage only intermittently, depending on the desired temperature measurement frequency. This minimizes power consumption and overheating due to electric power dissipation. At every reading, real time frequency analysis is performed, using a fast fourier transform (FFT) algorithm, and frequency peaks are detected and recorded. Data acquisition is done over 200 periods of the lowest frequency signal of the sensor array, to ensure acceptable sensor sensitivity (see section 3.4.2). The acquisition board sampling frequency corresponds to twice the frequency of the highest frequency signal, to satisfy the Nyquist theorem. Using a dedicated FFT and peak detection processor, a 1 Hz temperature measurement frequency is expected to be achievable.

This strategy of intermittent oscillator operation was used in the macroscopic validation tests. A National Instrument acquisition board was used in association with Labview to actuate the power supply and to acquire data. Matlab scripts were incorporated to the Labview program to conduct the FFT and peak detection calculations. The measuring sequence is described in table 4.3, for a measurement frequency of one measurement per 15 s.

4.1.3 Results

The following results were obtained using the measuring sequence just presented. Figure 4.6 shows how the frequencies of the two multiplexed oscillators drift in time before stabilizing after about ten minutes. On the one hand, the drift of oscillator 1 can readily be explained by the thermal inertia of the platinum resistors. Those resistors are self-heating before
Table 4.3: Measuring sequence used for macroscopic validation

<table>
<thead>
<tr>
<th>Time (s)</th>
<th>Operation</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Power on</td>
</tr>
<tr>
<td>0.1</td>
<td>Data acquisition over $\Delta t_{da}$</td>
</tr>
<tr>
<td>$0.1 + \Delta t_{da}$</td>
<td>Power off</td>
</tr>
<tr>
<td></td>
<td>Frequency analysis</td>
</tr>
<tr>
<td></td>
<td>Peak detection</td>
</tr>
<tr>
<td></td>
<td>Data display and storage</td>
</tr>
<tr>
<td>15</td>
<td>Power on</td>
</tr>
<tr>
<td></td>
<td>Beginning of another sequence...</td>
</tr>
</tbody>
</table>

reaching an equilibrium average temperature and hence an average equilibrium resistance. This causes the frequency to decrease asymptotically. On the other hand, the drift of oscillator 2 is difficult to explain; e.g., it could be due to progressive accumulation of charges in parasitic capacitances. In both cases, this initial frequency drift stays within 0.8% of the stabilized frequency. This is lower than the 1% frequency measurement accuracy that is needed to meet the required temperature measurement accuracy, as specified in section 3.4.2, and hence does not represent a considerable problem.

The two oscillators were tested in an oven up to a temperature of 180°C. Temperature was monitored inside the oven using a standard thermocouple located under the circuits, as sketched in figure 4.7. In order to avoid direct radiation from the resistive heating elements onto the circuit components, a steel case was used as a shield. The thermocouple and the circuits were all placed inside the steel shield, and are assumed to have been kept within the same ambient temperature. Fiberglass was used for electrical insulation only. Temperature was increased and decreased at a maximum rate of 2°C/min, which is believed to be slow enough to ensure a good synchronization between oscillator temperature evolution and thermocouple reading. Quasi-static conditions were maintained throughout all experiments.

Figure 4.8 shows the temperature response of oscillator 2 when it was tested for the first time. During this first test, the signals of the two oscillators were multiplexed, but only oscillator 2 was placed in the oven and heated up. The results demonstrate that oscillator 1 was not influenced by the frequency variations of oscillator 2. The two frequency signals
Figure 4.6: Frequency drift of two multiplexed macroscopic high temperature prototypes at room temperature.
Figure 4.7: Oven setup for high temperature macroscopic validation (front view).

were independent, as expected.

The theoretical curve was built using the standard relation for the resistivity of platinum as a function of temperature (see section 3.4.1). The experimental points depart from theory, and show the sensor to be less sensitive to temperature than predicted.

Hysteresis was observed when the oscillator was cooled down back to room temperature. This is not believed to be due to thermal inertia of the circuit, considering the quasi-static character of the experiment, but rather to some permanent modifications of electrical properties. Indeed, the oscillator did not recover its exact original oscillating frequency, even after 12 hours of rest at room temperature. It drifted from 256.1 kHz at the beginning of the test to 250.5 kHz at the end of the test, for a 2% reduction in frequency. Annealing of the metallic contacts may be responsible for this drift.

Figure 4.9 shows the results of another test where the two oscillators were both placed in the oven and heated up. Oscillator 1 demonstrated a behavior that is closer to the theo-
Figure 4.8: Behavior of a macroscopic high temperature prototype when heated up and cooled down in an oven. The two sensors (oscillators) were multiplexed by current addition, but only sensor 2 was heated up in the oven.
Figure 4.9: Behavior of the two multiplexed macroscopic high temperature prototypes when heated up in an oven.

A theoretical curve than oscillator 2. During this second test, oscillator 2 experienced a negligible permanent change in room temperature frequency, from 250.5 kHz at the beginning of the test to 250.4 kHz at the end of the test.

During this second test, above 170°C, sensor 2 stopped oscillating, whereas sensor 1 oscillated up to about 184°C. Above those limits, transistor transconductance is too low, and the total loop gain is smaller than unity, preventing oscillation. When the two oscillators were cooled down, oscillating signals came back one after the other.

Since the room temperature frequency of oscillator 2 was not permanently altered during the second test, the results of this test were used as a calibration curve for one subsequent test. A third set of data was collected from oscillator 2, and was used to measure oven temperature, based on this calibration. Figure 4.10 plots this temperature measurement against the thermocouple measurement. Again during this third test, oscillator 2 almost
Figure 4.10: Temperature measured with oscillator 2 plotted against thermocouple reading. Calibration of oscillator 2 uses data of the second test, plotted in figure 4.9.

came back to its original frequency (250.4 kHz at the beginning of the test to 249.8 kHz at the end of the test, for a 0.2% reduction).

A temperature departure of about 20°C can be observed between the two measurements (oscillator and thermocouple), in the worst case. This departure is believed to be caused by several parameters. First, the exact position of the thermocouple in the oven may have varied from the second test to the third test. Important temperature gradients might be present in the oven and play a determinant role on the repeatability of the experiment. Secondly, temperature was not increased at the exact same rate in the second and third tests, which suggest that thermal inertia of thermocouple and oscillator might be important parameters, despite the previous assumption of quasi-static experiments.

With a microfabricated integrated oscillator prototype, accurate calibration will be more readily obtained by sticking a thin thermocouple directly onto the circuit substrate, avoiding
any temperature mismatch between the two sensors. Furthermore, thermal inertia will
play a lesser role, as surface heat transfer processes become dominant over volumetric
accumulation of heat, as dimensions are reduced. Response times will be reduced, and long
stabilization periods like the one of figure 4.6 are expected to disappear.

4.2 Predicted Sensor Performance

Table 4.4 returns to the specifications of table 2.2 to assess the performance of the sensor
array developed in this thesis.

In terms of temperature sensing, it was first highlighted in section 3.1.3 that 4H-SiC
active devices could be operated up to 1103°C, if high leakage currents can be tolerated. This
is what is targeted here. In order to meet turbine requirements for components exposed
to combustion products, diamond electronics must be investigated, as will be suggested
further in section 5.3. Nevertheless, SiC shows to be a good candidate for turbine parts
that are not exposed to combustion products, and for other gas turbine modules, like the
compressor section.

Sensor performance parameters like accuracy, resolution, repeatability, and hysteresis
still need to be validated at high temperatures. The macroscopic validation circuit only
reached 184°C, which is sufficient to validate the sensor concept, but not enough to draw
conclusions about sensor performance in a gas turbine environment.

In terms of integration inside engine environment, no major problem is foreseen, except
for fixation inside parts exposed to combustion products. Again, active device temperature
must remain lower than the SiC limit.

Real time data acquisition is possible with the new sensor array, at an expected rate
of one survey per second. This means that the reading sequence of table 4.3 would be
performed once every second, for all sensors installed in a turbine. As mentioned in sec-
tion 4.1.2, the use of a dedicated processor for frequency analysis will make this achievable.
In terms of durability, the electronic oscillators and connections need first to be protected by proper packaging to limit oxidation damage. Secondly, thermal fatigue and vibration could represent a considerable problem for both the electronic circuitry and the packaging; e.g., integrated circuit layers might separate from each other under low thermal cycles, and connections might be damaged under sustained vibration. Fixation cement might also break under fatigue. Those durability issues must be addressed as the integrated circuit will be developed further.

Finally, cost and manufacturability issues still need to be addressed. At this early stage of the research project, it is almost impossible to predict the cost of the sensing system, as it will mainly be labour dependent. Manufacturing and assembly times are difficult to assess at this point.

4.3 Microfabrication Process

The microfabrication process that is needed to build the integrated oscillator described in section 3.3 is presented in table 4.5. Only the high level steps are enumerated in this table. Low level detailed procedures still need to be elaborated, accounting for the availability of resources. Nonetheless, some information is provided in appendix B on the parameters of detailed microfabrication steps, with references.
Table 4.4: Temperature sensor array performance versus high pressure turbine sensing specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Predicted performance</th>
<th>Fully meets requirements</th>
<th>Partially meets requirements</th>
<th>Uncertain</th>
<th>Not acceptable</th>
</tr>
</thead>
<tbody>
<tr>
<td>Temperature sensing</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Temperature range</td>
<td>0-1100°C (30-2010°F) using 4H-SiC</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Accuracy at 1000°C</td>
<td>±6°C (±10°F), validated at low temperatures only</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Spatial resolution</td>
<td>&lt; 5 mm (0.200 in), sensitive wire with 1 x 1 mm integrated oscillators</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Temperature resolution at 1000°C</td>
<td>6°C (10°F), validated at low temperatures only</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Response time</td>
<td>&lt; 1 s</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Integration in engine environment</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Ambient temperature</td>
<td>0-1100°C (30-2010°F)</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Ambient pressure</td>
<td>Insensitive to pressure</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Resistance to contaminants</td>
<td>Insensitive to contaminants</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Excitation frequencies</td>
<td>Insensitive to mechanical vibrations</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Fixation surface temperature</td>
<td>0-1100°C (30-2010°F), corresponding to measured metal temperature</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Fixation surface material</td>
<td>Any metal</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Volume of one sensor</td>
<td>&lt; 1 mm x 1 mm x 1 mm</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Weight</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Data acquisition</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Data acquisition frequency</td>
<td>&lt; 1 Hz</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Data storage capacity</td>
<td>Real time storage possible, but not developed in this thesis</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Data transmission</td>
<td>One pair of leads for at least ten sensors</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Durability</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Oxidation life</td>
<td>To be determined, packaging issue</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Low cycle fatigue (LCF) life</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• High cycle fatigue (HCF) life</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Creep life</td>
<td>Not an issue with the current concept</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Cost</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Acquisition cost</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Operation cost</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Manufacturability</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>• Time for assembly</td>
<td>To be determined</td>
<td>✓</td>
<td>✓</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

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Table 4.5: Integrated oscillator microfabrication process

<table>
<thead>
<tr>
<th>Microfabrication step</th>
<th>Processes</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 Substrate 4H-SiC</td>
<td>Commercially available at Cree Inc. [33]</td>
</tr>
<tr>
<td>1 Semiconductor components</td>
<td>SiC CMOS process</td>
</tr>
<tr>
<td>Transistors, diodes</td>
<td>Ion implantation</td>
</tr>
<tr>
<td>Resistors</td>
<td>Ion implantation, oxide growth, electrode PVD$^a$</td>
</tr>
<tr>
<td>Capacitors (1st layer)</td>
<td></td>
</tr>
<tr>
<td>2 Intermediate layer</td>
<td>Oxide CVD$^b$, CMP$^c$</td>
</tr>
<tr>
<td>Insulating oxide</td>
<td></td>
</tr>
<tr>
<td>3 Metallization</td>
<td>Oxide etching, contact metal PVD, photoresist lift-off</td>
</tr>
<tr>
<td>Contacts</td>
<td>Platinum PVD, photoresist lift-off</td>
</tr>
<tr>
<td>Platinum resistors, connections</td>
<td>Contact metal PVD</td>
</tr>
<tr>
<td>Backside contact</td>
<td></td>
</tr>
<tr>
<td>4 Capacitors</td>
<td>Oxide CVD, electrode PVD</td>
</tr>
<tr>
<td>2nd and subsequent layers</td>
<td></td>
</tr>
<tr>
<td>5 Packaging</td>
<td>To be determined</td>
</tr>
<tr>
<td>Wire welding</td>
<td>To be determined</td>
</tr>
<tr>
<td>Protective coating</td>
<td></td>
</tr>
</tbody>
</table>

---

a. Physical Vapor Deposition  
b. Chemical Vapor Deposition  
c. Chemical-Mechanical Polishing
Chapter 5

Future Trends

5.1 Extension to Pressure and Strain Sensing

This entire master's thesis focused on temperature sensing. In order to measure temperature, temperature dependent resistors were introduced in the feedback network of oscillators to modulate the frequency of those oscillators.

Instead of those platinum resistors, however, piezoresistors or strain gauges could be placed in the feedback network. The value of those resistors would be influenced by stresses and strains respectively, rather than by temperature. Piezoresistors can be embedded in membranes for pressure sensing as illustrated in figure 5.1, whereas strain gauges can be fixed onto any metal surface to measure deformations.

Consequently, the sensor array concept developed in this thesis could be applied to the measurement of many parameters, including pressure and strain. A sensitive wire like the one of figure 3.1 could as well be developed to measure both pressure and temperature in a turbine cooling passage, for example, for fluid dynamics and heat transfer validation.
5.2 Active Control

Ideally, to obtain the best possible compromise between performance and durability, in a gas turbine, every part must be able to actively adapt itself to changing conditions, using multiple micro-devices, including actuators. At least, in a near and realistic future, the aircraft operator should get the possibility to monitor the health of multiple parts, and to adjust the engine running conditions accordingly, to maximize performance while maintaining an acceptable safety level. The temperature sensor array presented in this thesis is a first step toward many sophisticated active control applications, including

- Local control of turbine cooling, using macroscopic external valves, or local micro-valves;

- Enhanced active turbine tip clearance control, by locally controlling the temperature of the casing that supports the shroud segments;

- Local control of fuel injection based on local turbine temperature measurements.
The two first ideas could be applied to improve high pressure turbine efficiency, whereas the third one would help to ensure acceptable turbine durability. Other applications could be envisioned in gas turbine engines, outside the turbine environment, from combustor film cooling control to nacelle defrosting control.

5.3 Diamond Electronics

Developments are currently going on with diamond electronics. Calculating the intrinsic carrier concentration using parameters from Ruff et al. [11], and limiting this concentration to $10^{15}$ cm$^{-3}$, as was done in section 3.1.3 for Si and SiC, diamond would be operable up to temperatures of about 1700°C (1100°C for 4H-SiC). This semiconductor material offers good long-term promises for temperature monitoring of turbine parts that are directly exposed to combustion products.

5.4 Wireless Transmission

In order to eliminate all cumbersome leads from the turbine environment, electromagnetic waves can be envisioned for data transmission, in a long-term future. Obviously, no sensor could communicate its readings directly from the core of an engine to the external world by using electromagnetic waves. The metallic components and casings of the engine would prevent such direct transmission, acting as electromagnetic shields. Rather, a peer-to-peer communication protocol must be developed, so that less accessible sensors would communicate their readings to other more accessible sensors that are “visible” to them. Those latter sensors would do the same toward even more accessible sensors, and so on until the information reaches the external casing of the engine. Then, a pair of leads would be used to transmit the data across the casing, to reach the external world.

This futuristic idea would become more realistic if there existed a way to locally supply power to micro-devices, inside an engine, without the necessity of an external power source.
In this regard, a miniaturized heat conversion power source like a thermoelectric battery could represent a good solution, since heat transfer rates are very large through the walls of turbine mechanical components.
Chapter 6

Closing Remarks

This master’s project was initiated with the aim of developing a novel temperature sensor for gas turbine engines that would minimize the cumbersome leads associated with thermocouple readings. The needs for a high number of local sensors was first established, for the turbine section of an aircraft engine, and clear temperature sensor specifications were defined for this application. The entire project has been conducted keeping those background specifications in mind.

An innovative sensor array technology was studied and experimented throughout the master’s process. A frequency modulation concept was developed using electronic oscillators that will be embedded in a temperature-sensitive wire. The silicon carbide (SiC) semiconductor technology has been explored through high temperature macroscopic validation testing and microcircuit design. Ultimately, the developed sensitive wire could be used at temperatures exceeding 700°C, with 4H-SiC, but major issues still need to be faced in terms of electronic stability at high temperature. Electronic contacts represent one of the major hurdles to overcome.

For turbine blade temperature sensing, the developed temperature sensor array might not be competitive against optical pyrometry, if the lens fouling problems of the existing pyrometers are solved. For static components, however, like turbine vanes and shrouds, it
could become the preferred option. Eventually, the concept could even be extended to local pressure and strain distribution sensing, in which field the competition is weaker.

Despite all efforts, SiC electronics might never evolve fast enough to reach the operating temperatures of the turbine components that are directly exposed to combustion products, simply because the operating temperatures of the turbine materials are also growing very fast. Peripheral components like casings and cooled shrouds might be the most extreme applications that SiC could find in turbines. Nevertheless, applications in compressors are not out of the discussion, and could be sought as a first technology transfer opportunity in a the short-term future. As for turbines, research work must be stressed toward the development of higher temperature electronics, with new semiconductor materials like diamond.
Bibliography


Appendix A

Product Definition and Requirements Specification

Here follows a preliminary Product Definition and Requirements Specification (PDRS) document that was elaborated to describe the need for a temperature sensing system intended for local monitoring in high pressure turbines.
Product Definition and Requirements Specification (PDRS)

Temperature Sensing System for Gas Turbine Components

Master's Thesis Project
Lacy Label
Graduate Student
University of Sherbrooke
Mechanical Engineering Department

April 5, 2004
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1. Introduction

The current document describes the need for a new gas turbine temperature sensing technology. The detailed requirements for such a technology were established based on the author's experience in the Turbine Cooling & Static Structures department of Pratt & Whitney Canada, from January 2002 to September 2003, and upon discussions with colleagues in this same department. This document is subject to be revised as the technology development progresses, and therefore should be seen as a flexible working document rather than a rigid limiting frame, from a research perspective.

2. Product Definition

The proposed sensor should be capable of real time simultaneous measurements of metal temperature at multiple locations, within high pressure turbine components. Those components include turbine airfoils (vanes and blades), shrouds, discs, casings, and supports. The sensor must alleviate the problem of lead routing specifically encountered in small business or regional aircraft engines, and be operable in the turbine environment as specified herein.

3. Specifications

Every specification is given a letter to help prioritizing.

<table>
<thead>
<tr>
<th>Letter</th>
<th>Type</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>B</td>
<td>Basic</td>
<td>The customer does not even ask for this type of specification. He assumes that it will be met.</td>
</tr>
<tr>
<td>P</td>
<td>Performance</td>
<td>The customer shows an interest for this type of specification. It is met at different levels by the competition, but the customer wants to improve the existing technology towards a better satisfaction of this type of specification.</td>
</tr>
<tr>
<td>I</td>
<td>Innovation</td>
<td>This type of specification is not met by any rival product. It represents a new capability.</td>
</tr>
<tr>
<td>C</td>
<td>Constraint</td>
<td>If this type of specification is not met, it is simply impossible to use the product.</td>
</tr>
</tbody>
</table>

Levels related to turbine parameters are specified for a typical hot day block test take-off (HDBT-TO) condition, and hence describe the most exigent conditions encountered in a turbine. Some levels, e.g. temperatures and pressures, are specified as a range, because different levels can be encountered in different engine models or in different turbine parts of the same engine, at this same HDBT-TO condition.
S1. Temperature sensing

<table>
<thead>
<tr>
<th><strong>S1.1. Temperature range</strong></th>
<th><strong>B</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Range of metal temperatures measurable by the sensor.</td>
</tr>
<tr>
<td>Level</td>
<td>820 – 1260 °C (1500 – 2300 °F) for parts exposed to combustion products; 480 – 820 °C (900 – 1500 °F) for parts not exposed to combustion products.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be validated within the entire range of temperature specified.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th><strong>S1.2. Accuracy</strong></th>
<th><strong>P</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Accuracy of temperature measurement with the range of the sensor.</td>
</tr>
<tr>
<td>Level</td>
<td>± 6 °C (± 10 °F)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be tested for repeatability, hysteresis and deterioration with time.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th><strong>S1.3. Spatial resolution</strong></th>
<th><strong>P</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Smallest distance between two possible neighbour measurements on the same metal surface.</td>
</tr>
<tr>
<td>Level</td>
<td>5 mm (~0.200 in)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>Capability of the system must be demonstrated experimentally.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th><strong>S1.4. Temperature resolution</strong></th>
<th><strong>P</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Smallest temperature increase or decrease that can be felt by the sensor.</td>
</tr>
<tr>
<td>Level</td>
<td>6 °C (10 °F)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be validated within the entire range of temperature specified.</td>
</tr>
<tr>
<td><strong>S1.5. Response time</strong></td>
<td><strong>P</strong></td>
</tr>
<tr>
<td>------------------------</td>
<td>-------</td>
</tr>
<tr>
<td><strong>Definition</strong></td>
<td>Time required for the system to report any temperature change higher than the resolution of the sensor, within the specified sensor accuracy.</td>
</tr>
<tr>
<td><strong>Level</strong></td>
<td>1 x</td>
</tr>
<tr>
<td><strong>Flexibility</strong></td>
<td>maximum</td>
</tr>
<tr>
<td><strong>Validation</strong></td>
<td>Time required to get a stable signal after applying an instantaneous temperature change must be measured.</td>
</tr>
</tbody>
</table>
### S2. Integration in engine environment

#### S2.1. Ambient temperature

<table>
<thead>
<tr>
<th>Definition</th>
<th>Temperature of cooling air surrounding the system, under which the system must stay operational.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>480 – 760 °C (900 – 1400 °F)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be tested within the specified conditions of operation.</td>
</tr>
</tbody>
</table>

#### S2.2. Ambient pressure

<table>
<thead>
<tr>
<th>Definition</th>
<th>Pressure of cooling air surrounding the system, under which the system must stay operational.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>120 – 450 psi</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be tested within the specified conditions of operation.</td>
</tr>
</tbody>
</table>

#### S2.3. Resistance to contaminants

<table>
<thead>
<tr>
<th>Definition</th>
<th>Level of contaminants present in surrounding cooling air, under which the system must stay operational.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>Humidity: TBD</td>
</tr>
<tr>
<td></td>
<td>Dust: TBD</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be tested within the specified conditions of operation.</td>
</tr>
</tbody>
</table>

#### S2.4. Excitation frequencies

<table>
<thead>
<tr>
<th>Definition</th>
<th>Mechanical vibration frequencies, generated by the engine, to which the system is subjected.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>250 – 500 Hz (engine spool rotation speeds: 15 000 – 30 000 RPM); 10 000 – 20 000 Hz (~40 blades per rotor).</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The sensor must be tested within the specified conditions of operation or must be proven to have natural frequencies higher than the frequencies generated by the engine.</td>
</tr>
</tbody>
</table>
### S2.5. Fixation surface temperature

<table>
<thead>
<tr>
<th>Definition</th>
<th>Temperature of the part on which the sensors have to be attached.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>Depend on the concept. TBD.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>TBD</td>
</tr>
<tr>
<td>Validation</td>
<td>The mechanical fixation of the sensor on the host surface must be proven to satisfy durability specifications.</td>
</tr>
</tbody>
</table>

### S2.6. Fixation surface material

<table>
<thead>
<tr>
<th>Definition</th>
<th>Material on which the sensors have to be attached.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>Nickel-iron alloy (with or without coating, depending on concept and on specific component)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>N/A</td>
</tr>
<tr>
<td>Validation</td>
<td>The mechanical fixation of the sensor on the host surface must be proven to satisfy durability specifications.</td>
</tr>
</tbody>
</table>

### S2.7. Volume

<table>
<thead>
<tr>
<th>Definition</th>
<th>Volume occupied by the system (sensors and controller).</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>Sensors: 1 mm x 1 mm x 1 mm within airfoils, on the internal walls</td>
</tr>
<tr>
<td></td>
<td>Controller: 1 cm x 2 cm x 2 cm available over every airfoil of a stator</td>
</tr>
<tr>
<td></td>
<td>Subject to change depending on the concept chosen. Refer to description of engine components.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>Functions of other engine components must remain unaltered. Engine secondary flows must also remain unaltered.</td>
</tr>
</tbody>
</table>

### S2.8. Weight

<table>
<thead>
<tr>
<th>Definition</th>
<th>Weight of the complete system, including the sensors and the control boxes packaged inside the engine</th>
</tr>
</thead>
<tbody>
<tr>
<td>Level</td>
<td>500 g (~1 lb)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>N/A</td>
</tr>
</tbody>
</table>
S3. Data acquisition

<table>
<thead>
<tr>
<th>S3.1. Data acquisition frequency</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Frequency at which are made complete surveys of all the sensors managed by the system.</td>
</tr>
<tr>
<td>Level</td>
<td>1 Hz</td>
</tr>
<tr>
<td>Flexibility</td>
<td>minimum</td>
</tr>
<tr>
<td>Validation</td>
<td>N/A</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>S3.2. Data storage capacity</th>
<th>P</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Memory of the system, for storage of temperature measurements.</td>
</tr>
<tr>
<td>Level</td>
<td>4 MB per flight mission (onboard storage)</td>
</tr>
<tr>
<td>Flexibility</td>
<td>nice-to-have</td>
</tr>
<tr>
<td>Validation</td>
<td>N/A</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>S3.3. Data transmission</th>
<th>I</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Data transmission from the sensors to the inboard control computer.</td>
</tr>
<tr>
<td>Level</td>
<td>One pair of leads per airfoil cooling scheme feed passage</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>N/A</td>
</tr>
</tbody>
</table>

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S4. Durability

<table>
<thead>
<tr>
<th>S4.1. Oxidation life</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Number of hours of operation required for the product to lose enough material by oxidation to become unusable.</td>
</tr>
<tr>
<td>Level</td>
<td>10,000 – 20,000 h in flight conditions; 50 – 100 h in ground (block) test conditions.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>minimum</td>
</tr>
<tr>
<td>Validation</td>
<td>The product must stay operational until the overhaul time or until the end of the ground test.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>S4.2. LCF life</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Number of mission cycles required to make the product unusable due to Low Cycle Fatigue.</td>
</tr>
<tr>
<td>Level</td>
<td>10,000 – 20,000 cycles in flight conditions; 50 – 100 cycles in ground (block) test conditions.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>minimum</td>
</tr>
<tr>
<td>Validation</td>
<td>The product must stay operational until the overhaul time or until the end of the ground test.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>S4.3. HCF life</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Number of hours of operation required to make the product unusable due to high frequency fatigue.</td>
</tr>
<tr>
<td>Level</td>
<td>10,000 – 20,000 h in flight conditions; 50 – 100 h in ground (block) test conditions.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>minimum</td>
</tr>
<tr>
<td>Validation</td>
<td>The product must stay operational until the overhaul time or until the end of the ground test.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>S4.4. Creep life</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Number of hours of operation required for the product to become unusable due to creep.</td>
</tr>
<tr>
<td>Level</td>
<td>10,000 – 20,000 h in flight conditions; 50 – 100 h in ground (block) test conditions.</td>
</tr>
<tr>
<td>Flexibility</td>
<td>minimum</td>
</tr>
<tr>
<td>Validation</td>
<td>The product must stay operational until the overhaul time or until the end of the ground test.</td>
</tr>
</tbody>
</table>
## 5. Cost

<table>
<thead>
<tr>
<th><strong>5.1. Acquisition cost</strong></th>
<th><strong>P</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Definition</strong></td>
<td>Initial cost to buy and install the system in one engine.</td>
</tr>
<tr>
<td><strong>Level</strong></td>
<td>40 $ (Can.) per instrumented airfoil</td>
</tr>
<tr>
<td><strong>Flexibility</strong></td>
<td>maximum</td>
</tr>
<tr>
<td><strong>Validation</strong></td>
<td>N/A</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th><strong>5.2. Operation cost</strong></th>
<th><strong>P</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Definition</strong></td>
<td>Maintenance cost (replacement of parts at overhaul) required to keep the system operative.</td>
</tr>
<tr>
<td><strong>Level</strong></td>
<td>10 $ (Can.) per instrumented airfoil at every engine rebuilt</td>
</tr>
<tr>
<td><strong>Flexibility</strong></td>
<td>maximum</td>
</tr>
<tr>
<td><strong>Validation</strong></td>
<td>N/A</td>
</tr>
</tbody>
</table>
S6. Manufacturability

<table>
<thead>
<tr>
<th>S6.1. Time for assembly</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Definition</td>
<td>Time required on the assembly line to assemble a complete set of instrumented airfoils, to make the required settings and to plug cables.</td>
</tr>
<tr>
<td>Level</td>
<td>2 man-hours</td>
</tr>
<tr>
<td>Flexibility</td>
<td>maximum</td>
</tr>
<tr>
<td>Validation</td>
<td>N/A</td>
</tr>
</tbody>
</table>

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Appendix B

Details on the Microfabrication Process

Table B.1 gives various details about the microfabrication process of integrated temperature sensitive oscillators, with literature references. This table summarizes the literature survey that has been conducted so far, and is still to be refined and completed.
<table>
<thead>
<tr>
<th>Microfabrication step</th>
<th>Process</th>
<th>Equipment</th>
<th>Parameters</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 Substrate</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0.1 SiC wafer</td>
<td></td>
<td>N/A</td>
<td>4H-SiC, p-type, with grown n-type epitaxial layer</td>
<td>[33]</td>
</tr>
<tr>
<td>0.2 Substrate prep</td>
<td></td>
<td>Wet bench</td>
<td>To be determined</td>
<td></td>
</tr>
<tr>
<td>1 Semiconductor components</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1.1 Field oxide growth</td>
<td>Thermal oxidation, followed by CVD</td>
<td>Furnace, CVD apparatus</td>
<td>10 nm of thermal oxide, before CVD</td>
<td>[34]</td>
</tr>
<tr>
<td>1.2 Transistor windows opening</td>
<td>Lithography</td>
<td>Ion implanter</td>
<td>Nitrogen, high temperature process</td>
<td></td>
</tr>
<tr>
<td>1.3 n-type channel doping</td>
<td>Implantation</td>
<td>Ion implanter</td>
<td>For depletion mode transistors, no doping, use n-type epitaxial layer as the channel (1-2 ( \times 10^{10} \text{ cm}^{-2} ))</td>
<td></td>
</tr>
<tr>
<td>1.4 p-type channel doping</td>
<td>Implantation</td>
<td>Ion implanter</td>
<td>2.4 ( \times 10^{17} \text{ cm}^{-2} )</td>
<td>[33]</td>
</tr>
<tr>
<td>1.5 Gate oxide growth</td>
<td>Dry thermal oxidation</td>
<td>Furnace</td>
<td>Aluminum</td>
<td>[35]</td>
</tr>
<tr>
<td>1.6 Oxide annealing</td>
<td>Furnace bake</td>
<td>Furnace</td>
<td>40-50 nm thick, 1200°C</td>
<td>[35]</td>
</tr>
<tr>
<td>1.7 Polysilicon gate deposition</td>
<td>LPCVD</td>
<td>CVD apparatus</td>
<td>Low temperature</td>
<td>[35]</td>
</tr>
<tr>
<td>1.8 Polysilicon gate doping</td>
<td>Implantation</td>
<td>Ion implanter</td>
<td>500 nm thick, 620°C</td>
<td></td>
</tr>
<tr>
<td>1.9 Gate etching</td>
<td>Wet etching</td>
<td>Wet bench</td>
<td>n-type, phosphorus</td>
<td>[35]</td>
</tr>
<tr>
<td>1.10 Source and drain doping</td>
<td>Implantation</td>
<td>Ion implanter</td>
<td>Diffusion from Phosphorus Oxide Glass</td>
<td></td>
</tr>
<tr>
<td>1.11 Dopant activation anneal</td>
<td>Furnace bake</td>
<td>Furnace</td>
<td>n-type, dual implantation, nitrogen, 500°C, 5.0 ( \times 10^{14} \text{ cm}^{-2} ) at 70 keV, 3.35 ( \times 10^{14} \text{ cm}^{-2} ) at 40 keV</td>
<td>[35]</td>
</tr>
<tr>
<td>2 Intermediate layer</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2.1 Insulating oxide</td>
<td>Oxide CVD</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2.2 Surface polishing</td>
<td>CMP</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Microfabrication step</td>
<td>Process</td>
<td>Equipment</td>
<td>Parameters</td>
<td>Reference</td>
</tr>
<tr>
<td>-----------------------</td>
<td>---------</td>
<td>-----------</td>
<td>------------</td>
<td>-----------</td>
</tr>
<tr>
<td>3 Metallization</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3.1 Drain and source contact definition</td>
<td>Wet etching</td>
<td>Wet bench</td>
<td>Buffered HF, 20 × 20 μm contacts</td>
<td>[35]</td>
</tr>
<tr>
<td>3.2 Contact deposition</td>
<td>PVD, lift off, photoresist</td>
<td>Electron beam evaporator</td>
<td>Nickel, 50 nm thick</td>
<td>[34]</td>
</tr>
<tr>
<td>3.3 Deposition of interconnections and temperature sensitive resistors</td>
<td>PVD</td>
<td>Sputtering apparatus</td>
<td>Platinum, 100 nm thick</td>
<td></td>
</tr>
<tr>
<td>3.4 Backside contact deposition</td>
<td>PVD</td>
<td>Electron beam evaporator</td>
<td>Nickel</td>
<td>[26]</td>
</tr>
<tr>
<td>3.5 Contact anneal</td>
<td>Radiative bake</td>
<td>RTA</td>
<td>1000°C, 2 min for all contacts, in Ar atmosphere</td>
<td>[34]</td>
</tr>
<tr>
<td>3.6 Resistor anneal</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4 Capacitors</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4.1 Dielectric deposition</td>
<td>Oxide CVD</td>
<td>CVD apparatus</td>
<td>50 nm thick</td>
<td></td>
</tr>
<tr>
<td>4.2 Electrode deposition</td>
<td>Polysilicon LPCVD</td>
<td>CVD apparatus</td>
<td>500 nm thick</td>
<td></td>
</tr>
<tr>
<td>4.3 Electrode doping</td>
<td>See step 1.8</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5 Packaging</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5.1 Wire welding</td>
<td>To be determined</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5.2 Protective coating deposition</td>
<td>To be determined</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>