Wireless Strain Measurement with Surface Acoustic Wave Sensors

THESIS

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By

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ABSTRACT

The instrumentation of rotating machinery with sensors typically requires wires and slip-rings for electrical connection to measurement equipment. These wired connections are subject to noise and have a high failure rate. Conventional wireless sensors are bulky, require delicate electronics, and their batteries need to be replaced periodically. Passive wireless sensors based on surface acoustic wave (SAW) devices, which are miniature, rugged, and require no electronics have been developed. These sensors have been shown to measure strain wirelessly in a multipath, time-variant environment. Sensors were fabricated on langasite using electron beam lithography. Detailed characterization of these SAW sensors is performed. An RF interrogation system has been developed to record the strain measurements and convert the data into usable information. Measurements from SAW sensors have been correlated with conventional strain measurements at high frequencies. A rugged package has been designed to encapsulate the delicate sensor and protect it from harsh environments. A miniature patch antenna was integrated into this package. Finally, conclusions and future work are discussed.
DEDICATION

This thesis is dedicated to my family.
I would like to thank my research advisers, Dr. Yakup Bayram and Dr. Eric Walton for their guidance, wisdom, and patience. They have been invaluable teachers and mentors throughout this research. I would also like to thank Dr. Roberto Rojas-Teran for presiding over my thesis examination committee.

Michael Wood, Alexander Reuge, Justin Burr, Zheyu Wang, and Elias Alwan; my friends and colleagues at the ElectroScience Lab have helped me get through my courses and research.

I would like to recognize Aimee Price at Nanotech West for her assistance in the painstaking fabrication of SAW devices and insight into the considerable experimentation that it required.

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Eric Belknap and Dr. Mark Walter in the Mechanical Engineering department have been instrumental in my understanding of the mechanical and material aspects of strain measurement and have helped broaden my research.
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Fields of Study

Major Field: Electrical and Computer Engineering
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CHAPTER 1: INTRODUCTION

Previous Research
Tuncay [1] studied the wireless measurement of static strain and temperature via surface acoustic wave (SAW) devices. A detailed literature review of SAW sensor fabrication was conducted and preliminary fabrication research was undertaken. The purpose of Tuncay’s work was to better understand the challenges of wireless strain and temperature measurement in multipath, high-temperature environments. His work, conducted with the same resources as this research, served as a very valuable springboard for this thesis material.

Problem Statement
In the operation of moving or rotating assemblies; strain, temperature, and other measurements must be taken to understand the operation of the system. Normally, conventional sensors (e.g. resistive strain gauges) are fixed to the rotating bodies or shafts. Wires are routed through the rotating equipment to slip rings on a shaft. A sophisticated measurement configuration can consist of hundreds of delicate slip rings. A simpler, more robust method for in-situ measurements is desired. By designing passive, wireless sensors, much of this complexity can be eliminated. The number of wireless sensors in a given system is limited by time, frequency, and space; not numbers of wires or slip rings. Because they require no electronics or power supply,
SAW devices lend themselves to this application. Further, the ability of some SAW devices to operate at temperatures of up to 600°C and their tolerance for shock and vibration make them rugged enough for the harsh environment inside rotating machinery [2].

Passive wireless sensors consist of a SAW device, an antenna, and protective packaging. These sensors must conform to the contours of the measurement surfaces as closely as possible, to avoid interfering with any fluid-dynamics and impacting other parts. An antenna must be collocated with one or more sensors to facilitate the wireless link. Further, a rugged package must be designed to protect the fragile sensor from a harsh, high-temperature environment.

**Proposed Solution**

A wireless strain and temperature measurement system has been proposed for the measurement of vibration moments and temperature change on rotating machines. The system consists of a wireless interrogator, similar to a coherent radar system, specially designed to measure the phase-difference between two discrete reflections of a SAW sensor. Data processing software was developed to interpret this phase difference and correlate it to strain or temperature data measured simultaneously by another method. Specialized SAW sensors have been designed and fabricated in a cleanroom. Using time and frequency-domain spacing facilitates the measurement from multiple sensors simultaneously. A miniature patch antenna, to be integrated into a rugged package, has been designed for the sensor’s communication link.
Dynamic Strain

In the operation of rotating equipment, of particular interest are the strain and vibration modes on objects like gears, fan blades, and turbines. Currently, resistive strain gauges are used for the measurement of these modes, but they are prone to temperature drift due to thermal expansion and other effects. Further, as previously discussed, the instrumentation of such sensors on a rotating machine is complex and problematic. Therefore, the primary focus of this research was to design a SAW device which could serve as a wireless strain gauge. Tuncay [1] utilized a SAW device to measure static strain on a cantilever, but of more interest is the vibration of a dynamic system, in which the strain oscillates at high frequencies. In this thesis, dynamic strain is measured with a SAW device on a mechanically oscillated cantilever and compared to a resistive strain gauge for validation of results.

Wireless Strain Sensors

Tuncay [1] discusses the fabrication of SAW devices on lithium niobate (LiNbO$_3$). In this thesis, langasite (La$_3$Ga$_5$SiO$_{14}$) is used as a substrate for its retention of piezoelectric properties at high-temperatures. Since Tuncay’s work was performed, the facilities available at The Ohio State University’s Nanotech West Laboratory have improved, streamlining the processes and improving the quality of SAW sensor fabrication. Additionally, thin-film deposition materials specified for high-temperature sensors had since become available. In this thesis fabrication processes, challenges, and results are explained thoroughly, leaving out key details due to their proprietary nature.
CHAPTER 2: SAW SENSORS

Introduction

As the complexity of modern mechanical systems increases, so do the challenges associated with the instrumentation and measurement of these systems. For rotating or fast moving structures, the use of conventional wired sensors can be problematic or unfeasible. The wiring of a rotating object (e.g. motor rotor windings) necessitates slip rings or commutators which require regular maintenance and generally introduce noise into delicate electronic channels. While precision slip rings do exist, they are immensely complex, requiring a great deal of skill and care in installation; they are also prone to failure.

Active wireless sensors, consisting of transponder circuitry and a power supply, are used in some applications where wiring is impractical. Some examples include soil moisture sensors for agriculture and home alarm systems. While these devices are able to be moved freely without being tethered by wires, they are usually powered by batteries which require periodic replacement. New technologies are enabling the design of passive sensors which require no power supply or electronics.

SAW Devices

An acoustic wave is generated by any impulse disruption of an elastic material. An intuitive example of this would be striking a block of gelatin swiftly with a rod. In this
case, shear and longitudinal vibrations are produced. Shear vibrational modes propagate isotropically into the bulk of the substrate and decay exponentially, they are called bulk acoustic waves (BAWs). Longitudinal waves propagate along the surface of the substrate, away from the source. The coupling of the vertically-polarized shear (SV-type) and longitudinal (L-type) components compose an eigenmode called a Rayleigh-type surface acoustic wave [3]. These waves propagate mainly along the surface with trough-depths of approximately one acoustic wavelength. The intensity of the SAW decays longitudinally at a linear rate.

The critical angle of decay into the bulk is given by

$$\theta_c = \cos^{-1}\left(\frac{v_s}{v_b}\right)$$

where $v_s$ is the velocity of SAW propagation and $v_b$ is the velocity of BAW propagation. Because this critical angle is relatively small, the energy of the SAW is mainly concentrated along the surface of the substrate as it decays with distance [3].

A SAW device consists of a piezoelectric substrate with conductive features deposited on the surface via nano-scale lithography methods, typically used in semiconductor and integrated optics fabrication. These metal features are typically arranged as interdigital transducers (IDT), periodic structures which resemble interlocked fingers. These fingers and the spaces between them are designed to resonate at one or more acoustic wavelengths. An incident RF signal coupled to the IDT results in an electric field across opposing fingers. Through the piezoelectric effect this electric field actually deforms the surface of the SAW device. As the electric field is oscillating at RF frequencies, an ultrasonic perturbation is generated and propagates from the IDT along the surface of the substrate, hence a surface acoustic wave [3].
Figure 1: The relevant dimensions of a single-electrode IDT, 4 pairs are shown here.

The IDT can make use of both the piezoelectric and inverse piezoelectric effects to both generate and detect SAWs respectively. Often used in telecommunications systems, SAW filters have a pair of IDTs, one to transmit and one to receive the acoustic signal.

An incident RF signal is converted to an acoustic one and then re-converted to RF by the second IDT. The transfer function of this acoustic translation can be tailored to design an inexpensive, high-Q filter for mobile phones, where size and power consumption are critical.

Other metallic designs on the piezoelectric surface can be used to create reflectors or absorbers. Like IDTs, reflectors are periodic fingers, but instead of being interlocked, they are separate. Those metallic elements form a series of periodically-spaced boundaries with mismatched acoustic impedance from that of the substrate. This
discontinuous impedance results in Bragg reflection of the SAW. [3] One-port SAW devices utilize the same IDT to transmit and detect the SAW after bouncing off of one or more of these reflectors.

![Diagram of Interdigital Transducer and Reflectors](image)

Figure 2: A one-port SAW device, here shown connected to an antenna, consists of an IDT and one or more reflectors.

By arranging these reflectors in a series, one incident pulse can produce a train of reflections from a one-port SAW device. SAW RFID tags have used this method to reflect a pulse code unique to each device without the need for other circuitry or batteries [4].

Pohl [4] proposed the idea of using a SAW device as a wireless sensor. His paper discussed various automotive applications for wireless SAW sensors including brake disc temperature measurement, tire pressure monitoring, and drive-shaft torque measurement. Because SAW devices require no power supply, capacitive charge structure, or semiconductor electronics, they readily lend themselves to applications in harsh environments that would destroy conventional electronics [4].
SAW Sensor Principles

A wireless sensor is expected to move about in its environment. This means that the propagation delay between a measurement system and the sensor is time-variant. As we will see, the delay of SAW device reflections is of interest; thus any change in sensor position will create errors. To solve this problem, each sensor needs a reference to which the reflection can be compared. This is accomplished by using two reflectors, a known distance apart on each sensor. The two reflectors are placed on either side of the IDT to make use of the acoustic wave traveling in both directions.

![Figure 3: An illustration of a one-port SAW device showing IDT, two reflectors, and wire bonding pads.](image)

Calculation of Strain

As the surface of the SAW device is elongated via thermal expansion or bending stress, its two reflectors are separated. The distance $L_{ik}$ between two reflectors $i$ and $k$, in this case the difference between $L_1$ and $L_2$, on a SAW device can be related to a time-delay $\tau_{ik}$ by

$$\tau_{ik} = \frac{L_{ik}}{v_s}$$
where $v_s$ is the velocity of propagation of the SAW for a specific cut of the piezoelectric substrate. The wavelength of the acoustic wave is naturally given by

$$\lambda_s = \frac{v_s}{f_o}$$

where $f_o$ is the frequency of both the RF and acoustic waves [4].

As the reflectors move apart, $\tau_{ik}$ increases such that the change in time-delay is given by

$$\Delta \tau_{ik} = \tau'_{ik} - \tau_{ik}$$

where $\tau'_{ik}$ is the time delay for an elongated SAW device. This time-delay can be easily measured and correlated with elongation. However, in time-varying environments (i.e. rotating machinery) time delays can be distorted due to multipath and range ambiguity.

Each reflection has a shifted phase proportional to the time-delay; this phase-difference is given by

$$\Delta \phi = \omega_o \cdot \Delta \tau_{ik}$$

[4]

where $\omega_o$ is the angular frequency of $f_o$. Through coherent I/Q demodulation, this phase-difference can be measured.

Strain is related to phase-difference by

$$\Delta \phi = \omega_o \frac{2 \varepsilon L_{ik}}{v_p} = \omega_o \varepsilon \tau_o$$

[5]

where $\varepsilon$ is mechanical strain $\left(\frac{\Delta L}{L}\right)$ and $\tau_o$ is the time-delay between the reflectors on an unstrained SAW device. Thus strain is linearly proportional to phase-difference as

$$\varepsilon = \frac{\Delta \phi}{\omega_o \tau_o}.$$ 

Elongation by strain and thermal expansion can be differentiated by frequency; the SAW device elongates slowly due to thermal expansion while strain oscillates much faster.
These parameters can be separated by frequency in post-processing. SAW device measurements can be compared to data measured with a resistive strain gauge, extensometer, or thermocouple for validation of results.

**SAW Sensor Design**

In designing a SAW device, the most critical choice is the substrate. There are many piezoelectric substrates which lend themselves to the transmission of acoustic waves along their surfaces. Typically, piezoelectric crystals are grown in a laboratory and precisely rotated and cut into polished wafers. The particular rotation of the cut is chosen based on crystalline structure and piezoelectricity characteristics which affect the SAW propagation. The SAW substrates discussed in this thesis include ST-X cut $\alpha$-quartz, 128° YX lithium niobate (LiNbO₃), and 48.5° Y langasite ($\text{La}_3\text{Ga}_5\text{SiO}_{14}$) [6]. The latter two substrates are more advanced and lend themselves more readily to SAW sensors than quartz. A few other modern SAW device substrates include lithium tantalate ($\text{LiTaO}_3$) and gallium orthophosphate ($\text{GaPO}_4$), but those are not discussed here.

Lithium niobate has a relatively high electro-acoustic coupling factor (4.5%) and will perform at temperatures up to 300°C [2] [6]. While 4.5% may seem poor, it is one of the highest among common substrates, and the highest listed in Table 1 by an order of magnitude. Further, a commercially available SAW filter has a typical input-output return loss of 22dB [7]. At our frequency of interest, 2.45GHz, lithium niobate has a specific acoustic attenuation of 3.9 dB/$\mu$s of delay [2]. While more than halving the power for a microsecond of delay sounds undesirable, in this category lithium niobate also comes out
ahead. This is a reminder of the significant attenuation that is common for all SAW devices. This means that a robust link budget is necessary for the implementation of any SAW sensor.

For extreme temperatures, langasite is more desirable as its attenuation does not significantly increase below 600°C. However its coupling factor is 0.44%, an order of magnitude worse than lithium niobate. Langasite’s acoustic attenuation is 17 dB/$\mu$s [2], so reflectors must be placed close to the IDT to minimize loss. Also, with attenuation of at least 30dB [6] the power of the incident pulse must also be very large to ensure a measurable reflection. Also considering fabrication challenges discussed in Chapter 3, langasite SAW devices are far more challenging to fabricate for wireless sensor applications. Langasite should only be chosen over lithium niobate as a substrate for high temperature applications.

<table>
<thead>
<tr>
<th>Substrate</th>
<th>Max Operating Temperature (°C)</th>
<th>Electro-Acoustic Coupling Factor (%)</th>
<th>SAW Propagation Velocity (m/s)</th>
<th>Acoustic Propagation Attenuation (dB/$\mu$s at 2.45GHz)</th>
</tr>
</thead>
</table>

Table 1: Lithium niobate has favorable properties when compared to common SAW device substrates.

Dr. Yakup Bayram has developed AccuSAW, a modeling system for SAW devices based on langasite and lithium niobate substrates. This simulation engine utilizes equivalent-circuit models from [3], [9], and [10]. Using this system, the design parameters of a SAW device can be easily varied for optimization. The parameters of AccuSAW
include: number of IDT or reflector fingers, aperture width, characteristic impedance, reflector location, and resonant frequency. AccuSAW produces time and frequency-domain plots of the simulated $S_{11}$ of the SAW device allowing one to parametrically optimize a device for its particular purpose. A screen shot of AccuSAW is included in Figure 4.

![AccuSAW GUI](image)

Figure 4: AccuSAW is a MATLAB-based GUI which simulates the time and frequency response of lithium niobate and langasite SAW devices.

AccuSAW assumes bi-directional single-electrode IDTs and open reflectors with finger spacing of $\lambda_s/4$ and finger width of $\lambda_m/4$ where

$$\lambda_m = \frac{v_m}{f}$$
where \( v_m \) is the velocity of acoustic propagation of the substrate with thin-film metallization atop. \( v_m \) is slightly slower than \( v_s \) due to the localized stiffening effect of the metallization and varies based on substrate characteristics, and metal composition and thickness [3]. For the purposes of electron-beam lithography

\[
\lambda_m = \lambda_s
\]

as the differences between the linewidths and gaps are slightly less than the resolution of the process.

Another parameter of the IDT is the aperture width (\( w_{ap} \)) which is typically a multiple of \( \lambda_s \). \( w_{ap} \) determines the input impedance of the IDT pair as well as the intensity and diffraction of an outgoing SAW. As previously mentioned, designing a SAW device on langasite is much more challenging than on lithium niobate. Not only is a langasite SAW device more lossy, but less research has been conducted on electron-beam lithography with a langasite substrate. Using the AccuSAW langasite model, it was determined that the aperture needed to remain narrow (\( w_{ap} \leq 20\lambda_s \)) to “focus” the SAW toward its target. For reflections, a narrower aperture requires less acoustic energy to generate an electric signal. It’s important to keep in mind, however, that too narrow an aperture will cause diffraction of the SAW away from the reflectors [3]. Assuming diffraction to be negligible for \( \frac{L_{dk}}{2} \leq 2\mu m \), the aperture width is the same for IDT and reflectors.

The number of finger pairs (\( N_f \)) affects the return loss of the SAW device as well as the dispersion and bandwidth characteristics. Too few IDT pairs, and the SAW device will be broadband but have poor efficiency. Conversely, too many IDT or reflector pairs will increase dispersion as the entire feature may be wider than the SAW pulse itself.
Increasing the number of IDT pairs increases the IDT capacitance and thus the input impedance of the SAW device, making matching more difficult. Conversely, the number of pairs of reflectors must be sufficient to reflect enough energy through the additive effect of Bragg reflection [3].

The first generation of our SAW sensors have been designed so that data can be collected from up to 16 of them simultaneously. Thus a time-domain and frequency-domain multiplexing regime was integral to their design. Table 2 displays the delay and frequency parameters for 16 SAW sensors. The maximum delay limits our sampling interval and risks higher attenuation, thusly was not to exceed 1600ns. It is advantageous in the development phase of this technology to use the unlicensed ISM band (2.4-2.5GHz) for simplicity; four channels were allocated in this band.

<table>
<thead>
<tr>
<th>Reflector Delays</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
</tr>
<tr>
<td>2.40 GHz</td>
</tr>
<tr>
<td>2.43 GHz</td>
</tr>
<tr>
<td>2.47 GHz</td>
</tr>
<tr>
<td>2.50 GHz</td>
</tr>
</tbody>
</table>

Table 2: 16 SAW devices can be measured independently by utilizing time and frequency multiplexing.

The SAW device parameters were optimized such that they would be consistent for all sensors. Minor adjustments to the geometries of the features are required for different frequencies and delays. Table 3 lists the feature dimensions as calculated by acoustic wavelength. Table 4 lists the reflector spacing from the IDT for each time delay, calculated by acoustic propagation velocity. These sensors were simulated in AccuSAW;
an example of the simulated time and frequency responses can be seen in Figure 5. Here SAW12 is simulated; we see good resonance at our target frequency and two distinct reflections in the time-domain. The reflections are dispersed in time because the reflectors necessary on langasite are many acoustic wavelengths in length.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Line/Gap Width ((\lambda_m/4 = \lambda_s/4))</th>
<th>Aperture Width ((w_{ap} = 20\lambda_s))</th>
<th>Number of IDT Pairs ((N_P))</th>
</tr>
</thead>
<tbody>
<tr>
<td>SAW_{1n}</td>
<td>2.40 GHz</td>
<td>285 nm</td>
<td>22.830 (\mu m)</td>
</tr>
<tr>
<td>SAW_{2n}</td>
<td>2.43 GHz</td>
<td>280 nm</td>
<td>22.550 (\mu m)</td>
</tr>
<tr>
<td>SAW_{3n}</td>
<td>2.47 GHz</td>
<td>275 nm</td>
<td>22.186 (\mu m)</td>
</tr>
<tr>
<td>SAW_{4n}</td>
<td>2.50 GHz</td>
<td>270 nm</td>
<td>21.920 (\mu m)</td>
</tr>
</tbody>
</table>

Table 3: The dimensions of the SAW device features are determined by the acoustic wavelength for a given frequency.

<table>
<thead>
<tr>
<th>Delays</th>
<th>Reflector 1 Spacing ((L_1))</th>
<th>Reflector 2 Spacing ((L_2))</th>
<th>Number of Reflector Elements ((N_R))</th>
</tr>
</thead>
<tbody>
<tr>
<td>SAW_{n1}</td>
<td>200, 400 ns</td>
<td>150 (\mu m)</td>
<td>420 (\mu m)</td>
</tr>
<tr>
<td>SAW_{n2}</td>
<td>600, 800 ns</td>
<td>700 (\mu m)</td>
<td>980 (\mu m)</td>
</tr>
<tr>
<td>SAW_{n3}</td>
<td>1000, 1200 ns</td>
<td>1250 (\mu m)</td>
<td>1520 (\mu m)</td>
</tr>
<tr>
<td>SAW_{n4}</td>
<td>1400, 1600 ns</td>
<td>1600 (\mu m)</td>
<td>2065 (\mu m)</td>
</tr>
</tbody>
</table>

Table 4: The reflector spacings of the SAW device are determined by the acoustic propagation velocity of the substrate.

In summary, 16 orthogonal SAW sensors have been designed and simulated in AccuSAW. The feature sizes of the relevant dimensions are calculated. Langasite is chosen as a substrate for its piezoelectric qualities at high temperatures, but it has a higher attenuation than some other substrates. These sensors have reflector delays of 1600ns or less to minimize this attenuation. Fabrication of these SAW devices will be discussed in Chapter 3.
Figure 5: AccuSAW simulation of time and frequency-domain responses for $\text{SAW}_{12}$. 
CHAPTER 3: SAW SENSOR FABRICATION

Introduction

SAW sensors were fabricated at The Ohio State University Nanotech West Laboratory. Due to the high frequency of operation of these SAW sensors, the features must be smaller than conventional optical lithography methods will allow. Though some advanced optical lithography techniques exist to make SAW devices at frequencies of 2.4 GHz, they are beyond our immediate capability. Also, it is planned to scale these sensors to 5 GHz for use in high-cutoff environments [1]. Finally, the ability to easily reconfigure the geometries of our patterns was desired. Therefore, we chose a two-cycle fabrication process based on electron-beam (e-beam) lithography.

Some details of the process are omitted for proprietary reasons, but the basic steps are outlined as well as challenges that were faced in fabrication. Electron micrographs of the metallic features are presented. Finally the testing of the finished SAW sensor is described as well as results.

Process

SAW sensor fabrication requires two cycles of the following steps outlined in Figure 6. The process largely occurs in an ISO 5 cleanroom at Nanotech West laboratory at The Ohio State University. Only the dicing of the wafer is preformed outside this controlled
environment, as the sample will be cleaned later. Before dicing, the wafer is cleaned with organic solvents and thoroughly dried. A thin-film polymer is baked onto the surface to protect it from scratches during the dicing process. During dicing, it is sometimes preferred to bond the back-side of the wafer to a glass plate to avoid damaging the delicate dicing blade. CrystalBond, a waxy epoxy, is used as a temporary adhesive. The wafer is then diced into smaller pieces or “samples” which can be processed individually. Using samples is advantageous for the experimental phase of SAW sensor development. Samples can be more easily handled than whole wafers, and various processes can be run concurrently while minimizing material consumption.

After dicing, the samples are each cleaned again, this time with more aggressive solvents to remove the protective polymer and CrystalBond; they are then dehydrated in an oven. Polymethyl methacrylate (PMMA) (\([C_5O_2H_8]_n\) ), also known as acrylic or Plexiglass, can be used as a high-resolution e-beam resist. It is applied to the sample in liquid form and then spun at several thousand RPM to create a uniformly thin, planarized layer of resist on the surface of the substrate. The sample is then baked allowing the liquid to solidify as the chemical chains of the polymer are thermally linked together. E-beam radiation will break these bonds, but the baked resist is not sensitive to light and is robust enough to endure other processes.

When piezoelectrics are bombarded with electrons, charging can occur due to their dielectric properties. This charging is undesirable for a proper exposure and can even destroy the substrate. Therefore, a conductive layer of aluminum is applied over the resist via plasma-vapor deposition or “sputtering”. Next, a precisely calibrated electron
beam exposes targeted areas of the resist through the aluminum, tracing the patterns of the metallic features. As the electrons collide with the resist, they break polymer bonds in certain areas which can then be dissolved.

Figure 6: The fabrication of a SAW device requires two cycles of eight steps to create the device features and wire-bonding pads respectively. The process for lithium niobate is shown here. After [1].

After electron-beam exposure, the aluminum is chemically etched away and the exposed resist is removed with development agents. The sample is still coated with resist, but with small voids or recesses in exposed areas that have been removed, revealing the surface of the substrate. During evaporation, an electron-beam heats a crucible of metal to its vaporization point in a high vacuum. Evaporation deposits two thin layers of metal on top of the sample; however the metal only adheres to the surface of the langasite in
areas where the resist was removed. The first of these metals is an adhesion layer, typically titanium (Ti), which is known for its tendency to stick to polished surfaces [11]. Next, a thicker layer of electrically conductive metal is deposited in the same manner. Langasite SAW sensors can operate at temperatures of up to 600°C, however the elements used for the metallic features (e.g. titanium and gold) tend to migrate together at these high temperatures, spoiling the performance of the device. Thiele et al. [11] suggest using platinum (Pt) and palladium (Pd), as these are robust conductors that can survive high temperatures for extended periods. For adhesion, zirconium (Zr) is specified for its strong adhesion performance on mono-crystalline surfaces and resistance to migration at high temperatures [11]. The electron-beam evaporation tools at Nanotech West had heretofore never used Zr, so a study was conducted by laboratory staff to ensure it would not interfere with their other processes.

The final and most critical process is lift-off, using a combination of solvents and ultrasonic baths the remaining resist is dissolved or “lifted-off”, and with it goes most of the deposited metal. Lift-off aggressiveness is increased by using stronger solvents and higher temperatures. The most aggressive chemical approaches use acetone near its boiling point, though the flammable vapors generated can be dangerous. A last-resort physical lift-off method involves removing excess metal with clear tape, but this will destroy integrity of the device. After successful lift-off, microscopic metallic elements have been adhered to the surface.

In order to electrically connect to this device, large conductive pads need to be placed atop via a repetition of the same process. The correct placement of the pads is critical, as
they may short-out the IDT if they are placed over the aperture. Alignment markers are written in the first e-beam exposure as a reference for the pad placement. These pads are vastly larger than the other elements, so the dose and resolution of the lithography process are not as critical. The above process is repeated, slightly modified for adhering pads overtop of the IDTs for electrical connection. The sample is then ready to be diced into individual SAW devices. After this step, the SAW device is ready for testing in a probe station or connected to an antenna.

Challenges

The above process specifies a single-layer polymer resist. These initial attempts on langasite consistently produced excellent results through the development and evaporation phases of the process. However, the critical lift-off step continued to exhibit conflicting signs. Metalization would fail to lift-off in some areas, indicating insufficient solvent strength at this step. However some features would be obliterated, implying lift-off was too aggressive, or that adhesion was insufficient in some areas.

To address the latter problem, an oxygen plasma “de-scum” etch was included in the process between development and evaporation. The purpose of plasma etching was to remove any residual resist that was not completely developed from the surface. Even a thin layer of residual resist would inhibit adhesion. The oxygen plasma should erode a small amount of residual resist, but it will also decrease the overall height of the resist profile, so it was used sparingly. This appeared to improve the adhesion problem, but the
lift-off was still incomplete in some areas despite using the most aggressive methods available.

One hypothesis to explain the lift-off problem can be described as “filming” where the deposited metal on the surface of the substrate remains connected to the metal atop the resist. The tensile strength of this film may prevent the lift-off from removing the metal in narrow spaces between features. The film may also be protecting the resist from lift-off solvents.

In cases where lift-off profile is to be improved, a bi-layer resist regime can be used [12]. In this regime, second layer of resist is deposited before the original PMMA resist by performing the spin-coating process twice. This new layer is a co-polymer of PMMA and methacrylic acid (MAA) (C₄H₆O₂). The co-polymer is more sensitive to electron-beam radiation, resulting in an undercut effect in development. The bi-layer process for langasite can be seen in Figure 7. In the bi-layer regime, the polymer forms an aperture and the co-polymer a cavity which facilitates the clean deposition of metals onto the surface of the substrate. This method eliminates the filming issue that was experienced with a single-layer resist regime.
Figure 7: A bi-layer resist regime improves lift-off results.

The bi-layer regime presents some new challenges however. As the co-polymer layer is more sensitive to electron-beam radiation, it can be completely eradicated when features are spaced too closely. The proximity effects of electron-beam radiation can compound the problem depending on feature density, dose, and beam current. Using positive resists like PMMA and its co-polymer, the limitation of e-beam resolution is often feature spacing, not line-width. The close proximity of two exposed features (in our case IDT or reflector fingers) can result in adjacent features bleeding together. As can be seen in Figure 8, this issue is more problematic with bi-layer regimes as the resist can collapse if the supporting layer is completely dissolved.
Figure 8: When the dose is too high or development-time too long, the resist structure collapses (a). Scanning electron micrographs post-lift-off illustrate such adverse effects (a) and correct dose and development time (b).

Development time, i.e. the time that a sample remains in a solvent bath, can affect the undercutting of the co-polymer layer. It was found that shortening development time decreased the minimum spacing for our bi-layer method, allowing us to achieve the feature dimensions as designed. The development time must be sufficient to fully dissolve the exposed PMMA top layer or the lines will appear jagged.

Results

After several iterations of the above process, the parameters were optimized such that a working 2.4GHz SAW device was achievable. Figure 9 shows some electron micrographs of a 100-pair IDT on langasite. The uniformity of the lines is critical for the
“inner” part of the IDT as this dictates its efficiency and bandwidth. Also, any touching of the lines results in the shorting-out of the IDT and a significant decrease in performance.

Figure 9: An electron micrograph of a 100-pair IDT on langasite, resonant at 2.4 GHz. Pads are not present in this micrograph.

A closer look at the metallic lines in Figure 10 shows some inconsistency inherent in the process. The precision of the process is ±5nm at best, however this may be improved with further fabrication refinements.
Figure 10: An electron micrograph showing the exact widths of IDT fingers and gaps.

Although steps were taken to minimize edge anomalies, further refinement is required to compensate for the e-beam proximity effect on the first and last finger-pairs. The proximity effect is more pronounced for the open-ended SAW device reflectors, shown in Figure 11. In the case of the reflectors, the uniformity and consistency of the lines is not as critical as any surface metallization will serve as a rudimentary reflector.
Figure 11: An electron micrograph of a 100-pair open-ended reflector grating. The under-dosed edges of the reflector are the result of the electron-beam proximity effect.

**SAW Sensor Analysis**

The above SAW device differs from previous designs in that $N_p = 100$ for the IDT and reflectors. This SAW device is discussed because it was the best specimen available at time of publication. An AccuSAW simulation of such a device is shown in Figure 12. Because the IDT and reflectors only consist of 100 pairs, the resonance is not as good to begin with. Further, the reflections are weaker and exhibit less time-dispersion. Once pads were placed on the above IDTs, the real SAW devices could be analyzed by electromagnetic probing. A coaxial SG probe with a characteristic impedance of 50Ω was placed on the pads and a vector network analyzer excited the device and measured the response. The return loss of the actual device is shown in Figure 13. We see that the resonance is shifted in frequency and is not as strong as simulated.
Figure 12: AccuSAW simulation results for SAW12 where $N_p = 100$.

Figure 13: The frequency response of the SAW device shows a weak resonance at 2.16GHz, a frequency somewhat lower than expected.
Taking the inverse Fourier transform of the frequency-domain data, we see the time-domain reflectometry of the SAW device in Figure 14. This plot, known as “range walk” in radar systems, shows the reflections after the expected delays of 600 and 800ns.

![SAW Device Time-Domain Reflectometry](image)

Figure 14: The time-domain reflectometry shows two reflections at 600 and 800ns respectively. The power of these reflections with respect to incident power is approximately -85dB.

Plotting the time and frequency-domain data together in a spectrogram, we can more clearly see that the reflections aren’t instantaneous impulses, but rather have some dispersion in the time-domain. This spectrogram, which can be seen in Figure 15, illustrates the available time-frequency space which could be occupied by orthogonal sensors in a measurement environment.
Figure 15: A spectrogram of the SAW device response shows good definition between the two reflectors but some dispersion in the time-domain. A faint echo can be seen at approximately 1600 ns and 2.15 GHz. The available space for other SAW sensors in the time-frequency plane is also visible. The colorbar is in units of dB.

The SAW device is highly capacitive, as expected. But the resistance is almost exactly 50 $\Omega$, as was designed. A Smith chart maps the complex impedance of the SAW device in Figure 16. This reactance can be tuned out with existing impedance-matching methods.
Figure 16: A Smith chart shows the SAW device impedance is nearly $50 - j140 \Omega$ at resonance. This impedance may be matched with conventional methods.
Introduction

A wireless interrogation system was designed to measure the phase-difference between two discrete reflections from a SAW sensor. The interrogator generates a burst of RF which is amplified as high as 35dBm. The pulse is transmitted to the sensor via an antenna or coaxial cable. Each reflected pulse from the sensor is received by a dedicated receiver circuit which amplifies and demodulates the signal via I/Q mixers. I and Q signals are acquired via a high-speed ADC, which is connected to a computer.

Two computers operate the interrogator, the first contains a precise timing control board which opens and closes the interrogator’s RF switches to generate pulses and receive reflections. The other computer is connected to data acquisition hardware to record I and Q data from the dual receiver circuits, it can also record data from other sources (e.g. a resistive strain gauge) simultaneously. A set of instrumentation amplifiers improve the dynamic range of the ADC and serve as analog low-pass filters. In MATLAB processing software, the I and Q channels are linearized, and digitally filtered. The phase-difference is then computed, and the data can be compared with that of the resistive strain gauge.
**Interrogator Design**

Figure 17: The SAW sensor interrogator features two receiver channels, one for each reflection.

The interrogator consists of a continuous-wave source, either a voltage-controlled oscillator or a signal generator. A signal generator is preferred for its higher power and precise frequency output. The continuous wave is split by power divider into transmit and reference signals. A high-isolation RF switch serves as a pulse gate for the transmit
path, switching on for short bursts at an adjustable pulse repetition frequency (PRF). A power amplifier boosts the transmit path by 15dB before it leaves the antenna. The CMOS-based RF switches must be placed so that they are not saturated by the amplified signal. In the many modifications to the Interrogator, the bistatic design shown in Figure 17 provides the best isolation between transmit and receive channels without limiting transmitter power by switch saturation.

A received pulse must be divided between two receive circuits by a power divider. A low-noise amplifier boosts the weak reflected pulses before they are divided. Each receive switch opens for a brief interval when its reflection is expected to arrive to minimize noise in the receiver. Separate I/Q demodulators compare the received signal with the continuous reference signal and produce in-phase and quadrature signals for the analog-to-digital converter (ADC). The I/Q channels are important for the measurement of $\Delta \phi$ as

$$\Delta \phi = \tan^{-1} \frac{Q_2}{I_2} - \tan^{-1} \frac{Q_1}{I_1}.$$ 

A computer connected to the ADC logs the I/Q data via customized Labview software. This software is capable of logging data from another source, such as a thermocouple or resistive strain gauge (RSG) simultaneously. The data is stored in a file which is fed to a Matlab routine that computes the phase-difference between the two pulses. The Matlab routine can also compare this phase-difference with data from another source.

While the interrogator is able to record 4-channels of I/Q data at 100kHz, the system must run at a sampling frequency of 50kHz when measuring from an RSG concurrently. Also, because there is no hardware synchronization between the ADC units, an arbitrary
delay is introduced between the I/Q and the RSG waveforms. This delay is caused by other Windows processes which compete for processor time between the initializations of both ADC modules. Because of its arbitrary nature, the delay is difficult to account for in post-processing, as we will see in Chapter 5.

**System Validation**

The interrogation system was validated by using a transmission line-based SAW device surrogate. A power divider and two different lengths of coaxial cables were assembled with open-circuited ends. The lengths of the cables needed to be selected to ensure the reflections would not impinge upon one another, so they could not be multiples of each other. The cable lengths were chosen to be 60 and 110 feet. Using

\[
\begin{align*}
    v_0 & \approx \frac{1 \text{ ft}}{\text{ns}} \quad \text{in free space} \\
    v_p & \approx \frac{2}{3} v_0 \quad \text{in coaxial cable} \\
    \therefore 2 \text{ ft} & = 3 \text{ ns}
\end{align*}
\]

the round-trip delays were calculated to be 180 and 330 nanoseconds, respectively. The impedance mismatch from the open-ended lengths of cable mimics the reflectors of a SAW device while the lengths of the cables represent the acoustic propagation time. The variable-length trombone line was attached to the longer of the cables, allowing the length of one line to be varied over several wavelengths. Reflections from the shorter line have a fixed phase-shift with respect to the variable line. This experiment was designed to mimic the performance of a SAW device, a schematic of the experiment can be found in Figure 18. I and Q data was collected from both I/Q demodulators while the length of
the trombone was modulated by hand. The quadrature of these signals was verified to confirm correct operation of the interrogator.

Figure 18: A simplified block-diagram of the interrogator evaluation test. Coaxial delay lines were used to simulate a SAW device.
CHAPTER 5: STRAIN MEASUREMENTS WITH INTERROGATOR

*Introduction*

Having validated the interrogator, we set out to collect strain data from a commercially available lithium niobate SAW device. This device was extracted from an RFID tag and contained 8 reflectors. The spacing between each reflector corresponded to the tag’s unique identification code. Before interrogation, time-domain reflectometry was measured via a vector network analyzer. Figure 19 shows the time-domain response of the SAW device.

![Graph showing time-domain reflectometry of the RFID SAW device](image)

*Figure 19: Time-domain reflectometry of the RFID SAW device shows 8 discrete reflections in 1250 ns.*
This SAW device was adhered to a printed circuit board with strain gauge adhesive. The board was a copper-clad Rogers Duroid substrate with both an antenna and coaxial feed etched onto it. The connector and antenna were selectable with an adjustable transmission line, allowing wired or wireless interrogation of the sensor. With one end fixed into a vice, the board acts as a cantilever. A conventional resistive strain gauge was mounted alongside the SAW device to act as a strain reference. The setup can be seen in Figure 20.

Figure 20: The strain cantilever can facilitate wireless or wired interrogation of a SAW sensor and has a resistive strain gauge for comparison.
The interrogation system was configured to collect data from both the SAW device and the resistive strain gauge simultaneously. The data would be compared and allow the correlation of calculated strain to measured strain.

**Dynamic Strain**

The above experiment was performed for various vibrational modes. Because of the stiffness of the cantilever and the peak force of the mechanical shaker, our maximum frequency of vibration was limited. Newton’s second law of motion states that

\[ \ddot{F} = m \cdot \ddot{a} \quad [13] \]

In this case, \( F \) is the peak force that can be generated by the mechanical shaker; \( m \) is the sum of the mass of the shaker; and \( a \) is the maximum acceleration of the assembly. Any friction or gravitational forces, and the stiffness of the assembly are neglected.

\[ a = \frac{dv}{dt} = \frac{d^2x}{dt^2} \]

where \( v \) is the velocity of the shaker mechanism and \( x \) is its displacement. The cantilever equation expresses strain as

\[ \varepsilon = -x \frac{3y_b(l - s)}{2l^3} \quad [5] \]

where \( y_b \) and \( l \) are the thickness and length of the cantilever, respectively. \( l - s \) is the distance from the free end of the beam to the SAW device. These values are all fixed, thus strain is proportionally dependent on displacement.

If sinusoidal displacement is given by

\[ x(t) = A\cos(\omega_s t) \]

where \( \omega_s \) is the angular frequency of shaker excitation, acceleration is derived as
\[
a(t) = \frac{d^2x(t)}{dt^2} = -A\omega_s^2 \cos(\omega_s t)
\]

where the coefficient \( A \) is the peak displacement of the assembly.

For the peak displacement, we can assume
\[
\cos(\omega_s t) = 1
\]

thus
\[
a|_{\text{max}} = A\omega_s^2.
\]

So our maximum displacement is
\[
A = \frac{a|_{\text{max}}}{\omega_s^2}
\]

and decreases with the square of frequency.

The shaker used for these experiments is rated for a peak acceleration of 31.5 m/s\(^2\). Due to the stiffness of the cantilever and the mass of the magnetic armature, this shaker was unable to achieve measurable displacement at frequencies greater than 200Hz. More powerful shakers were considered for this test; a shaker rated for 15 times more force has a maximum acceleration of only 98 m/s\(^2\) due to a much heavier armature. By the same derivation, it could operate no faster than 350Hz to impart measurable strain on the same setup. It is evident that generating significant strain on this setup at high frequencies can be very difficult.
Dynamic Strain Measurement

To validate the linearity of the SAW device at various levels of dynamic strain, the excitation amplitude was stepped up over several seconds so that the absolute measurements between the strain gauge and the SAW device could be linearly correlated. The time-domain plots of this experiment can be seen in Figures 21 and 22.

Figure 21: The SAW device's phase-difference can be correlated with absolute strain measurements for a conversion ratio.

Figure 22: The resistive strain gauge recorded absolute strain values for an amplitude-stepped shaker excitation.
A closer examination of one of these steps can be seen in Figure 23.

![SAW Device vs. Resistive Strain Gauge](image)

Figure 23: SAW Device phase difference detail. The time has been manually aligned using the stepped nature of the waveform.

Analysis of this data characterizes a phase-difference/strain ratio of approximately $8.75^\circ/\mu\varepsilon$ for this SAW device.

In another test, an amplitude-modulated signal was fed to the shaker. This time the modulation waveform was a continuous wave, more closely approximating a 2nd-order...
vibrational mode of a complex mechanical system. The time-domain plots are shown in Figures 24 and 25.

Figure 24: SAW device phase-difference for amplitude-modulated excitation.

Figure 25: Corresponding resistive strain gauge reading for amplitude-modulated excitation.
In this case, the resistive strain gauge seems to be biased against the SAW device. This discrepancy likely represents a zero-strain imbalance between the two sensors. This effect could be caused by deformation in the cantilever (e.g. torsion), or drift in the strain-gauge adhesive of either sensor.

![SAW Device vs. Resistive Strain Gauge](image)

Figure 26: A closer examination of the amplitude-modulated waveforms measured with SAW device and resistive strain gauge. Note that the times are not exactly aligned due to an arbitrary delay in the measurement system.

Analysis of this data for the same device provides a ratio of $7.47^\circ/\mu\varepsilon$. The discrepancy between these results can be accounted for as different reflectors were used for each measurement.
CHAPTER 6: SAW SENSOR PACKAGING

Introduction

In order to survive a harsh, high-temperature environment, the SAW sensor will require a robust package for protection. Because the sensor needs to be collocated with an antenna, the package should contain an integrated antenna. This package must have a small footprint and be as thin as possible. It should be able conform to curved surfaces and must not cause localized stiffening of non-rigid bodies (e.g. vibrating cantilevers).

A miniature patch antenna has been designed using Ansoft HFSS, an electromagnetic Finite-Element-based analysis tool. A characteristic bowtie shape, Seirpinski fractals, and a high-permittivity dielectric are used as miniaturization techniques.

Sensor Antenna

A miniaturized patch antenna has been designed for the SAW sensor. The fractalized bowtie patch has been discussed in [14]. Fractals are space-filling curves which can be used to symmetrically remove material from the conductive patch surface. By removing regions of conductor, the current must travel farther before encountering the edge of the patch. This longer current path effectively lowers the resonant frequency of the patch.

In order to make the sensor as conformal as possible, a dielectric substrate of thickness 0.5mm and relative permittivity ($\varepsilon_r$) of 9.9 was selected.
The high-dielectric constant helps to further reduce the antenna size by increasing the refractive index \((n)\), given by

\[ n = \sqrt{\varepsilon_r \mu_o} = \frac{c}{v_p} \]  

where \(\mu_o\) is the permeability of free space and \(c\) is the speed of light. As \(n\) is inversely proportional to the phase velocity \((v_p)\), the resonant frequency is further decreased.

Typically, thin, high-permittivity dielectrics create inefficient patch antennas due to their high capacitance. When excited at resonance, they tend to store more energy than they radiate. The reactance of such a resonator is compounded by dielectric attenuation in the substrate, such that radiating fields are further reduced. Because of the high-temperature requirements of the sensor, polymer dielectrics (e.g. Teflon) are unsuitable. A ceramic dielectric composed of 99.6% alumina \((\text{Al}_2\text{O}_3)\) was chosen for its tolerance of high-temperatures and its extremely low loss tangent of 0.0001.

An un-fractalized bowtie was first designed in HFSS as a baseline. The antenna’s dimensions were scaled for a resonance of 2.45 GHz. Next, a first-order Sierpinski fractal regime was removed from the patch. The resonant frequency of that patch was considerably lower than the un-fractalized antenna, so the dimensions could be scaled down. The process was repeated up to a third-order fractal. A rendering of these designs can be seen in Figure 27.
Figure 27: Actual size of 2.4GHz (a) first-order Sierpinski bowtie patch and (b) second-order Sierpinski bowtie patch. Shown to-scale with US postage stamp.

A third-order bowtie was simulated, but the miniaturization achieved was negligible. Further, the smaller fractal elements complicated the geometry of the patch pattern and increased antenna impedance. Therefore, a second-order fractal was utilized for further simulations.

Originally, a point-source feed was used to simulate excitation from a SAW device. The small size of this point required a very fine mesh for accurate simulation, this resulted in computationally intensive simulations which could take hours to complete. Further, results between simulations were sometimes inconsistent. A realistic feed for antenna testing is a 50Ω coaxial cable. Because of the antenna’s size, a semi-rigid RG-405 cable with outer-diameter of 2.20mm was constructed in the simulation. Once resonance was achieved at the desired frequency, feed position was parametrically changed in the antenna’s x-y plane to match impedance and achieve a desirable radiation pattern.
Figure 28: A 3D rendering of the sensor patch antenna. In this case it's being excited with an RG-405 coaxial cable feed.

At resonance, when the impedance is matched, the reflection coefficient ($\Gamma$) approaches zero. $S_{11}(f)$ represents the antenna reflection with respect to frequency, a decrease in $S_{11}$ indicates that energy is entering the antenna at a given frequency. A plot of $S_{11}$ vs. frequency for a particular feedpoint of our antenna is shown in Figure 29.
Figure 29: The $S_{11}$ vs. frequency plot shows a narrow resonance at 2.43GHz. The measured 3dB bandwidth is slightly less than 8MHz.

As can be seen in Figure 29, the fractalization of the antenna creates a very narrowband resonance. This is advantageous because we desire the sensor to be frequency-selective for multiplexing purposes. The resonant frequency is expected to shift down as the substrate and patch expand with rising temperature. The SAW device’s resonant frequency is also expected to shift down, but at a different rate [16] [17]. Thus, the system must be designed for the bands to align at the normal operating temperature, meaning that the sensor may not work at room temperature. A broadband antenna may solve this problem.
By rounding out the edges of the patch, the antenna bandwidth can be increased, this is called a rounded-edge bowtie antennas (REBA) [18].

Figure 30: A rounded-edge bowtie antenna (REBA) with fractals can achieve broad bandwidth and miniaturization.

A REBA antenna has been simulated in HFSS, however the removal of patch surface area increased the resonant frequency. The antenna dimensions had to be scaled up significantly to resonate at 2.45 GHz. While increasing the antenna bandwidth by a factor of 3, the size was increased by nearly 50%. Figure 31 shows the S11 response of the fractal REBA antenna.
Figure 31: The bandwidth of the fractal REBA is approximately 23MHz, an almost three-fold increase from the original bowtie. Note that the frequency has shifted up, requiring the antenna to be enlarged for 2.45GHz resonance.

It was found that moving the feed even a fraction of a millimeter could affect the antenna impedance significantly. In practice, it may be impractical to drill a through-hole precisely enough for a perfect match. Regardless, it is not difficult to find reasonable resonance while feeding from many locations on the patch.

The location of the feedpoint also affects the antenna radiation pattern. Feeding from the center of the patch results in symmetric fringing fields at the edges which generate a null at broadside ($\theta = 0^\circ$). This creates an inefficient radiation mode for a patch antenna. The location of the coaxial feed in Figure 28 (feedpoint A in Figure 32) is approximately the
ideal feedpoint for balancing impedance matching and radiation pattern. The radiation patterns for several feedpoints are shown in Figure 32.

![Figure 32: 3D gain pattern renderings all scaled to maximum antenna gain for feedpoint A (3.7dBi). Center feed (b) creates a null in the broadside direction and radiates poorly. Feeding the antenna in the corner (c) results in weak radiation skewed towards the feedpoint. Finally, feeding the antenna near the center of the edge (d) also results in a reasonable radiation pattern. These plots assume perfect impedance matching at all feedpoints.](image-url)
**Sensor Design and Fabrication**

The packaging will need to be strong enough to protect the delicate SAW device from impinging fluids (including gases), sudden temperature variations, and foreign material; yet it must be flexible enough not to inhibit the normal vibrations and stresses of a mechanical system. As the antenna substrate is a high-temperature ceramic, it could also serve as the enclosure for the SAW device. A small cavity can be machined out from the bottom of such a material. However ceramics like alumina are very stiff and brittle and may cause localized stiffening, obscuring strain measurements. Worse, they may crack while enduring moderate amounts of strain. A more flexible high-temperature material, such as silicone or a thermoplastic, would still protect the SAW device without adversely affecting the stiffness of the mechanical system. Such materials have the advantage of being molded, rather than machined. Molding allows for more complex packaging geometries while allowing other components of the sensor to be cast in-place.

The antenna needs to electrically connect to one of the SAW device pads at the feedpoint while another pad must connect to a ground plane. A ground plane is integrated onto the back-side of the antenna while the larger the metallic surface that the sensor is mounted to would also be contacted. By grounding the small antenna ground plane to a larger conductive body, floating-ground transients can be avoided. These conductors must not create an impedance mismatch or excessive losses, but because the thickness of the package is much less than an electrical wavelength, it is unlikely that this will occur. Because of the high-temperatures encountered by the sensor, the antenna’s conductive surfaces cannot be adhered to the substrate by conventional copper-cladding adhesives.
Further, typical soldering bonds will not survive a high-temperature environment. Solid conductors may be cast in-place during manufacturing of such a package. The conductor may also be press-fit through the antenna substrate via a pre-drilled hole. The conductive surfaces of the patch and ground-plane can be applied via high-temperature screen printing techniques.

*Passive Wireless Sensor Concept*

Three-dimensional renderings of a preliminary package concept in Figures 33 and 34 show the locations of the various components in a working sensor.

Figure 33: Isometric view of passive wireless sensor concept showing electrical connection detail.
The illustrations show the antenna substrate in white and the flexible packaging material in blue. The packaging material has been cast to allow room the SAW device without interfering with strain measurement. This can best be seen in Figure 34. The ground pin can also be seen protruding from the base of the sensor. Ideally, the bottom face of the sensor (packaging and SAW device) would be carefully coated in a high-temperature strain gauge adhesive and bonded directly to the object to be measured.

Figure 34: A side-view and underside of the sensor showing approximate thickness. Note that the SAW device is not directly touching the packaging and that its underside is exposed.
CHAPTER 7: CONCLUSIONS AND FUTURE WORK

Wireless measurement of the physical characteristics of rotating machines has been desired for unobtrusive monitoring of these systems. Existing measurement techniques require wired sensors, which are complex and prone to failure; or the use of active wireless sensors, which are bulky and high-maintenance. The presence of impinging fluids or debris and a high-temperature, time-varying propagation environment further complicate the problem of wireless sensing. Passive wireless sensors based on SAW devices were evaluated and designed. A practical, low-maintenance measurement system for the instrumentation of moving machines at high temperatures is proposed.

Wireless Sensing with SAW Devices

The physics of SAW devices have been explained. Design considerations for various substrates have been outlined. An interrogation system has been designed and validated to measure one or more SAW devices nearly simultaneously. Several orthogonal SAW devices for sensing in high-cutoff environments were designed and simulated using AccuSAW. Wireless measurement of dynamic strain of a cantilever with a SAW was conducted and correlated with a resistive strain gauge. SAW devices have been shown able to precisely measure vibrations of this cantilever simulating complex vibrational modes of mechanical systems.
Fabrication of SAW Devices

SAW devices specifically designed for high-temperature sensing have been designed, fabricated and tested. Fabrication of SAW devices on langasite was conducted in an ISO 5 cleanroom using electron-beam lithography processes. Considerable experimentation and error was necessary to perfect the process and fabricate a working sensor. With a perfected recipe, devices can now be mass-produced by-the-wafer using resources at The Ohio State University.

Packaging of Wireless Sensors

The delicate SAW device requires a robust packaging to protect it from harsh environments. Preliminary designs for the packaging of wireless sensors were discussed. Integral to this sensor package is a miniaturized patch antenna. This antenna has been designed using Ansoft HFSS; its size reduced using Seirpinski fractals and a characteristic bowtie shape. Using a very thin, high-permittivity dielectric, the antenna and package can conform to the surface of most machines. Metallic interconnections have been shown to be integrated into the packaging, which could likely be a moldable thermoplastic.

Future Work

This work demonstrated the efficacy of a wireless sensing system and proved the concept of such wireless sensors. The next phase in the development of this technology is the fabrication of the antenna and package around a SAW device, and subsequently testing it
in a high-temperature environment. Next, the sensor could be mounted in a real machine and wireless strain measurement can be made. With the development of a more advanced interrogation system, multiple sensors could be interrogated simultaneously over multiple time and frequency slots [19]. For example, a broadband interrogation signal combined with spectral analysis could measure from sensors over multiple frequencies instantaneously, increasing data throughput in multi-sensor environments. Though this thesis focuses on strain measurement with SAW sensors, there are methods to measure temperature, pressure, or the presence of some chemicals using specially designed sensors [4]. The ability to interrogate multiple measurands from a single sensor could be accomplished with specially-designed sensors and more complex data processing (e.g. wavelet transform) [20].
LIST OF REFERENCES


APPENDIX A: IMPEDANCE-MATCHING METHODS FOR SAW DEVICES

Note: this report was written separately in fulfillment of ECE793 independent study requirements.

Introduction

The characteristic impedance of surface acoustic wave (SAW) devices is often not naturally matched to standard impedances (e.g. 50Ω) and usually has a reactive component. The interdigital transducer (IDT) of a SAW device is generally a highly capacitive resonator. This impedance must be matched with surrounding circuit elements to ensure an efficient flow of RF power through a system and the minimization of loss. For SAW sensors, the loss of power between a SAW device and antenna is especially critical, as reflected power is often several decades below incident power.

There are several methods of matching the impedance of a SAW device. The simplest method is to calculate the impedance of the IDT based on its geometry. [1] states that IDT conductance ($G$) and susceptance ($B$) are given by:

\[
G = 8N_p^2K^2f_oC_s \left| \frac{\sin[N_p\pi(f - f_o)/f_o]}{N_p\pi(f - f_o)/f_o} \right|^2
\]

\[
B = 8N_p^2K^2f_oC_s \left| (\sin[N_p\pi(f - f_o)/f_o] - [N_p\pi(f - f_o)/f_o]) \left( [N_p\pi(f - f_o)/f_o]^2 \right)^{-1} \right|
\]

where $N_p$ is the number of IDT finger pairs, $K^2$ is the electro-acoustic coupling factor, and $C_s$ is the capacitance of a single IDT finger pair which depends on the particulars of
the device [1]. The conductance and susceptance of the IDT can be combined to form the admittance \((Y)\), and thus the impedance \((Z)\) of the IDT.

\[
Y = \frac{1}{Z} = G + jB
\]

\[
Y = \frac{1}{Z} = G + jB
\]

Design Approaches

In the design of a SAW device, this admittance can be optimized for an impedance-match by changing the substrate material, number of IDT fingers, factors of capacitance, etc. AccuSAW, a MATLAB application by Dr. Yakup Bayram, can make parametrically changing these parameters easier and more intuitive. It’s important to note that all these parameters must be carefully optimized. Engineering tradeoffs will be necessary to balance a good impedance-match with SAW device performance. This method does not allow the matching of all SAW configurations and cannot compensate for the capacitance intrinsic to the IDT.

In telecommunications applications, SAW devices are installed among various circuit elements. Impedance can be matched with typical methods in such circuits. These methods propose an external matching circuit using microstrip impedance transformers or lumped elements (e.g. inductors) [2].
Figure 35: The conventional method for matching the impedance of an IDT is with a lumped-element or microstrip matching circuit external to the SAW device. From [3]

For SAW sensors, such lumped elements cannot survive the high temperature constraints of operation. Further, electronics and microstrip lines increase the size requirements of a miniaturized package.

For the purposes of this research SAW devices were fabricated and tested. The impedance of one device was found to be 50-j140Ω at resonance. This impedance indicates a high capacitance that is intrinsic of the IDT. While the device was optimized for a 50Ω match as explained above, the capacitance was not compensated for.
A natural method for compensating for a capacitive load is to introduce an equivalent inductance into the circuit. As was already explained, the size constraints of the packaging make a typical microstrip impedance transformer impractical. However, there is ample space on the surface of the SAW device for such a microstrip, assuming that a groundplane will be beneath the piezoelectric substrate.

Effective Permittivity of Piezoelectric Substrate

For the design of a microstrip network, the permittivity of the dielectric substrate must be known. In the case of SAW device materials lithium niobate (LNB) and langasite (LGS), the permittivity is a second-order tensor, due to the crystal structure of the material. The permittivity of LNB and LGS has been studied at RF frequencies by [3] and [4], respectively. The 2nd order permittivity tensor is given by

$$\varepsilon = \begin{bmatrix} \varepsilon_{11} & 0 & 0 \\ 0 & \varepsilon_{11} & 0 \\ 0 & 0 & \varepsilon_{33} \end{bmatrix}$$

for an un-rotated trigonal crystal with hexagon symmetry. $\varepsilon_{11} = \varepsilon_{22}$ for these crystals due to their symmetry about the z-axis [3] [5].

The relative permittivity for LNB is given by:

$$\varepsilon_{LNB} = \begin{bmatrix} 44.7 & 0 & 0 \\ 0 & 44.7 & 0 \\ 0 & 0 & 27.0 \end{bmatrix} [3].$$

These values are for the original crystal orientation of lithium niobate. For the fabrication of SAW devices the wafers are cut from this crystal after it is rotated by 128°.
about the x-axis. Using the direction of cosines method, the permittivity tensor can be estimated for the rotated crystal [6]. The resulting rotated matrix is:

$$
\varepsilon_{LNB}^{128x} = \begin{bmatrix}
44.7 & 0 & 0 \\
0 & 33.7 & 8.59 \\
0 & 8.59 & 38.0
\end{bmatrix}.
$$

The number in bold indicates the effective permittivity for the z-direction of the wafer. This is the value of interest because it is in the direction of the electric field of a microstrip line.

The relative permittivity of langasite is given by:

$$
\varepsilon_{LGS} = \begin{bmatrix}
49.4 & 0 & 0 \\
0 & 49.4 & 0 \\
0 & 0 & 19.6
\end{bmatrix} [5].
$$

A langasite SAW device wafer has been rotated about two axes; first 48.5° about the x-axis, then 26.6° about the y-axis. Similarly, the direction of cosines for these rotations results in:

$$
\varepsilon_{LGS}^{48x,26y} = \begin{bmatrix}
46.8 & 6.62 & 5.24 \\
6.62 & 32.7 & 13.2 \\
5.24 & 13.2 & 39.0
\end{bmatrix}.
$$

Fortunately, the effective permittivity of interest for both substrates is found to be nearly the same, meaning that a single impedance-matching network should be effective for both devices with minimal modification. This value is relatively high, indicating a slow propagation of the electric field.

**Design of an Impedance-Matching Circuit**

Pozar [7] gives the procedure for a single-stub impedance-matching circuit. With knowledge of the dielectric permittivity and thickness, the circuit can be implemented on
the surface of a SAW device using a microstrip. The microstrip can be fabricated via the same lithography techniques that are used to create the SAW device itself. Figure 36 shows a circuit diagram of the inductive stub impedance-matching circuit.

For a $Z_o = 50\Omega$ microstrip transmission line, [7] gives the microstrip width as:

$$W = h \left\{ 8e^{-\frac{Z_o}{50}} \left[ \frac{\varepsilon_r + 1}{2} \cdot \varepsilon_r^{-1} \left( \frac{0.23 + 0.11}{\varepsilon_r} \right) \right] \right\}$$

where $h$ is the thickness of the wafer, usually 0.5mm, and $\varepsilon_r$ is the relative permittivity in the z-direction which was found above. Because the permittivity is relatively high, the microstrip must be very narrow. $W$ was found to be 81μm for LNB and 77μm for LGS.

For our operating frequency of 2.45 GHz, the capacitance of the load is given by:

$$C_L = -\frac{1}{2\pi f_0 X_L}$$ [7]

where $X_L$ is the reactance of the IDT, found to be -140Ω. The phase-shift of the real impedance transformer is known as:

Figure 36: The single-stub impedance transformer is designed to negate the capacitance of the IDT load with a shunt inductor. After [7]
\[ t = \tan \beta d \quad [7] \]

where the wavenumber, \( \beta \) is given by:

\[ \beta = \frac{2\pi}{\lambda}. \]

Because the real components of the impedances match, we only need to be concerned with the susceptance of the stub, which is a function of its length. In this case, [7] gives the length of the transformer as:

\[ t = -\frac{X_L}{2Z_o} \]

thus

\[ d = \frac{\lambda \tan^{-1} t}{2\pi} = 3.05\text{mm} \quad [7]. \]

The width of the stub is fixed by its 50\( \Omega \) resistance and was calculated above. The stub susceptance is given by:

\[ B_s = \frac{K^2 t - (Z_0 - X_L t)(X_L + Z_0 t)}{Z_0[K^2 + (X_L + Z_0 t)^2]} \quad [7]. \]

The length of the shorted stub is then found by

\[ l = -\frac{\lambda}{2\pi} \tan^{-1} \left( \frac{B_s}{Y_o} \right) = 1.11\text{mm} \quad [7] \]

where \( Y_o \) is the admittance of a 50\( \Omega \) line (0.02S). These values are almost exactly the same for each substrate due to their relatively similar dielectric constants. Note that the length of the stub, \( l \), is almost 14 times its width. This means that the stub might act more as an antenna than a resonator. Simulation software like ADS can confirm the operation of such a circuit. Furthermore, actually fabricating and testing this transmission line on a piezoelectric substrate would completely validate this solution.

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[8] shows that the construction of a microstrip-based circuit is realizable on lithium niobate. They designed a bandpass filter using coupled-transmission lines on such a substrate. The dimensions of the features are much smaller due to the higher operating frequencies of their particular filter.

Additional Approaches

If a microstrip approach is not realizable, there are other inductive circuit elements that can be created on the surface of a SAW device. [9] discusses the fabrication of inductive spirals to cancel the capacitance of the IDT. Figure 37 illustrates some of these concepts. Because of the proprietary nature of a patent, details about design are not included. A great deal of experimentation and computer modeling would need to be preformed to optimize this solution.

![Illustrations of surface inductors for the cancellation of IDT reactance. From [9]](image)

Figure 37: Illustrations of surface inductors for the cancellation of IDT reactance. From [9]

Works Cited


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