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Design, Fabrication, and Interrogation of Integrated Wireless SAW Temperature Sensors

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DESIGN, FABRICATION, AND INTERROGATION OF INTEGRATED WIRELESS SAW TEMPERATURE SENSORS

by

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A dissertation submitted in partial fulfilment of the requirements for the degree of Doctor of Philosophy in the Department of Electrical Engineering and Computer Science in the College of Engineering and Computer Science at the University of Central Florida Orlando, Florida

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Major Professor: Donald C. Malocha
ABSTRACT

Wireless surface acoustic wave (SAW) sensors offer unique advantages over other sensor technologies because of their inherent ability to operate in harsh environments and completely passive operation, providing a reliable, maintenance-free life cycle. For certain SAW sensor applications the challenge is building a wirelessly interrogatable device with the same lifetime as the SAW substrate. The design of these application intensive sensors is complicated by the degradation of device bond wires, die adhesive, and antenna substrate. In an effort to maximize the benefits of the platform, this dissertation demonstrates wafer-level integrated SAW sensors that directly connect the thin film SAW to a thick film on-wafer antenna. Fully integrated device embodiments are presented that operate over a wide range of temperatures using different fabrication techniques, substrates, and coding principles.

The design of orthogonal frequency coded, OFC, RFID sensors is presented on the lithium niobate substrate at 915 MHz. The OFC concept was developed previously at UCF, and uses both frequency and time modulation to provide signal diversity and processing gain for the simultaneous operation of multiple sensors in a noisy environment. In a multi-sensor system, discrete time multiplexing of individual OFC sensors, which share bandwidth and independently change temperature, is ideal for accurate sensor detection. Coding techniques are demonstrated that shorten the overall SAW response length while preserving code diversity and bandwidth by utilizing a multi-track SAW configuration. The extraction of temperature and identification information from the wireless sensors is investigated. A coherence correlator was developed that processes sensor information in the frequency domain, and is compared to techniques previously published. The design of a meander dipole antenna is presented for use on the lithium niobate substrate. Various antenna
fabrication techniques were used and wireless results are demonstrated over temperature.

In an effort to extend the operational temperature range of the OFC sensor SAW characterization was performed on the langatate substrate, which maintains piezoelectric operation up to its melting point of 1450°C. Fabrication techniques are investigated to extend SAW metallization lifetime at temperatures up to 1000°C. SAW coupling, substrate capacitance, velocity, and reflectivity are extracted at frequencies up to 915 MHz for various fabrication methods. The design of OFC sensors on langatate is presented, and results are demonstrated at 650 MHz. The design of a 650 MHz on-wafer dipole antenna is outlined and preliminary results are demonstrated over temperature. These integrated devices eliminate several locations of high temperature failure such as wire bonds and packages.
To my grandfather for his dedication to my education
and my wife for her dedication to our marriage.
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CHAPTER 1
INTRODUCTION

Wireless surface acoustic wave (SAW) sensors are an attractive platform for use in harsh environments because of their completely passive operation, inherent radiation hardness, and ability to operate at temperatures from cryogenic to 1000°C [1]. Several embodiments exist for SAW sensor architectures including resonators and reflective delay lines [2]. SAW sensors operate on a sensitivity of phase, delay, or amplitude to the environmental change of interest. Various implementations exist for the sensing of gas concentrations, strain, pressure, and other measurands [3–5]. In the simplest embodiment, the SAW temperature sensor capitalizes on the inherent temperature sensitivity of most substrate cuts to track delay changes.

NASA has made a strong push for the use of a distributed SAW sensor network for the monitoring of ground and space flight operations using the orthogonal frequency coded (OFC) platform. OFC was originally published by Malocha and Puccio in [6] and has been covered extensively in various embodiments [7–9]. The OFC SAW system offers several unique advantages over other published multi-sensor SAW systems like single frequency CDMA in its use of frequency and time diversity for increased noise immunity. The goal of this dissertation is to expand the operational temperature range of the OFC SAW multi-sensor system to 700°C, with a push toward 900°C. There is considerable interest at NASA in high temperature SAW sensors for increasing the safety and efficiency of turbine engines and other structural health monitoring applications with large temperature gradients. The work presented in this dissertation is in conjunction with the industry partner Krystal Engineering, who is growing boules of langatate used for the SAW sensor substrate of this work.
This dissertation will begin with an overview of wireless SAW sensors in Chapter 2. A short background of the OFC SAW concept will be given with a review of the results previously demonstrated. A survey of work on high temperature SAW substrates and sensors will be presented.

Chapter 3 will explore several developments that have been made during this dissertation for the interrogation and coding of OFC sensors. The extraction of sensor ID and temperature information will be discussed, and a new approach will be demonstrated for coherently extracting the sensor information in the frequency domain. Also a simple multi-track code manipulation technique is presented that reduces the impulse response length of OFC codes. This technique allows for decreased device lengths as the bandwidth of the device shrinks to meet loss or regulatory requirements.

Proposed in this dissertation is the concept of integrating the SAW sensor and its antenna directly onto the piezoelectric substrate, taking advantage of the inherent ruggedness of the SAW platform. As a proof of concept, initial devices were built on YZ-LiNbO₃ for use at temperatures below 400°C. Results are presented in Chapter 4 on the design and fabrication of 915 MHz on-wafer integrated OFC temperature sensors.

The remainder of the dissertation focuses on extending the operational temperature range of the OFC sensor platform beyond that demonstrated in previous research efforts. Chapter 5 will explore the lifetime of SAW electrodes on the langatate substrate at temperature up to 1000°C. Chapter 6 will investigate SAW operation on LGT at frequencies up to 1 GHz. Previously published research efforts have focused on low frequency resonant devices.

Based on these characterization results, OFC SAW devices are built on the langatate substrate in Chapter 7 for operation at 650 MHz. An on-wafer antenna is designed for operation at this new center frequency and substrate, and preliminary fabricated results are demonstrated using electroplated gold.

Finally, Chapter 8 will discuss the results of this dissertation and give a road map for future high temperature OFC sensor research.
CHAPTER 2
SURFACE ACOUSTIC WAVE SENSOR BACKGROUND

Surface acoustic wave (SAW) devices are common electronic components that use the piezoelectric effect to convert electrical signals into a mechanical wave propagating along an elastic surface. The benefit of a SAW device is in the decrease in velocity compared to an electromagnetic wave, allowing for long time delays in a short physical propagation distance. The focus of this dissertation is on the use of SAW technology for wireless temperature sensor applications. Wireless SAW sensors are RF devices that act as reflectors of the interrogation signal. Based on the measurand of interest the reflected interrogation signal will be altered in frequency, phase, delay, or amplitude. There are fundamentally two types of SAW temperature sensors: reflective delay lines and resonators.

Several groups have investigated multiplexing a network of sensors by encoding each sensor with a unique radio-frequency identification, RFID, code. Early commercial applications for SAW RFID included the tagging of rail cars [10]. Spread spectrum techniques for multi-access SAW sensors have been demonstrated by various groups [11, 12].

2.1 Orthogonal Frequency Coded SAW Sensors

In this dissertation orthogonal frequency coding, OFC, is used to encode the individual SAW sensors with an RFID. OFC is a spread spectrum technique that has the inherent advantage of processing gain and noise immunity. The basic OFC communication technique was outlined by Malocha and Puccio in [6], and briefly reviewed here. OFC is similar to M-ary frequency shift
keying in that they both use a set of orthogonal waveforms to spread the signal bandwidth. The OFC basis set assumes one of two forms given by

\[
    h_1(t) = \sum_{n=0}^{N} a_n \cos \left( \frac{2n\pi t}{\tau} \right) \text{rect} \left( \frac{t}{\tau} \right)
\]

or

\[
    h_2(t) = \sum_{m=0}^{M} b_m \cos \left( \frac{(2m+1)\pi t}{\tau} \right) \text{rect} \left( \frac{t}{\tau} \right)
\]

where

\[
    \text{rect}(x) = \begin{cases} 
    1 & |x| \leq 0.5 \\
    0 & \text{otherwise}. 
\end{cases}
\]

Each cosine term in the summation represents a time gated sinusoid whose local center frequencies are given by

\[
    f_n = \frac{n}{\tau} \quad \text{and} \quad f_m = \frac{(2m+1)}{2\tau}.
\]

For an orthogonal basis set over the time length \(\tau\), \(f_n \cdot \tau\) must be an integer, or an integer number of carrier cycles at frequency \(f_n\), and similarly an integer number of half cycles at \(f_m\). In the frequency domain this basis set is the sum of sampling functions with individual center frequencies given by (2.4), and null bandwidths of \(2\tau^{-1}\). The overall frequency response is dictated by the choice of even or odd transfer function, the set of basis frequencies, the individual weights, and either the bandwidth or time length.

The basic OFC definition can be used to define a desired signal with various levels of time and frequency diversity, and a number of multi-sensor OFC embodiments have been explored. A block coding technique where a set of orthogonal frequency chips of length \(\tau_c\) occur sequentially in time is the most basic and well developed. The basic technique places each code into a fixed bit
length, $\tau_B$ such that for an OFC code of $J$ chips

$$\tau_B = J \cdot \tau_{chip}. \quad (2.5)$$

This gives a time function for the total bit of

$$g_{bit}(t) = \sum_{j=1}^{J} \omega_j \cdot h_c \cdot (t - j \tau_c) \quad (2.6)$$

where $\omega_j$ is the individual chip weights and $h_c$ is one of the two orthogonal basis functions. Based on this definition the total bit null bandwidth is equal to $2J \cdot \tau_c^{-1}$. Although the chips are contiguous in frequency, individual frequencies can be shuffled in time to give a level of coding. Chip weight can also be manipulated $\pm 1$ for the realization of a PN code. The processing gain of the coded signal is the ratio of the bit length to that of the individual chip for equal energy signals.

An example transfer function is shown in Figure 2.1 using the definition given by (2.1) for a 5 frequency set. The time domain response demonstrates the well known linear stepped chirp. In the frequency domain the orthogonality condition places the peak of a carrier at the null of all others.
Figure 2.1: An example ideal transfer function response of an orthogonal frequency coded signal in the (a) time and (b) frequency domain.

The orthogonal frequency code is implemented on a SAW delay line as shown in Figure 2.2 and outlined in [7]. The platform consists of wideband input transducer and a series of Bragg reflector gratings. The transducer launches a surface wave based on the interrogation signal input on an antenna. The interrogation signal is convolved with the serially coded reflector array, reflected back to the transducer, and re-radiated out the antenna. The orthogonality condition is realized in the series of SAW Bragg reflectors, or chips, of constant length $\tau_{\text{chip}}$. Each of these chips has an individual center frequency $f_{\text{chip}}$ which satisfies the condition

$$N_g = f_{\text{chip}} \cdot \tau_{\text{chip}}$$

(2.7)

with $N_g$ being an integer number of half carrier cycles. $N_g$ maps to the individual Bragg gratings by metallized, grooved, or dielectric discontinuities. Each grating maintains the length $\tau_{\text{chip}}$ by varying the number of carrier cycles. Each chip is spaced $\tau_{\text{chip}}^{-1}$ apart in frequency, with the nulls of each adjacent Bragg reflector aligned with the peak of an individual reflector. The SAW travels
through the reflector array with minimal attenuation at orthogonal chip center frequencies, until the frequency of the interrogation signal coincides with a matching reflector frequency. Coding is accomplished by shuffling the chips in time/distance from the transducer. Sensor ID and temperature information can be extracted from the reflected sensor signal through the use of an adaptive matched filter or equivalent process at a receiver.

![Piezoelectric Substrate](image)

**Figure 2.2:** Schematic of a five chip SAW OFC RFID tag which can be used as the platform for a sensor. The figure depicts a chirp input signal and the returned coded signal which is the convolution of the OFC code and chirp.

An example OFC reflection coefficient is shown in Figure 2.3. In the frequency domain the S11 response of the SAW sensor has the wideband response of the transducer superimposed with the ripple from the distributed frequency code. Converting that response to the time domain shows the roll-off of the transducer, followed by the acoustic delay, and finally a series of orthogonal grating responses. For clarity time domain plots only show the envelope of each OFC grating. The plots shown are compared to a coupling of modes model developed in [13]. The model is used extensively throughout this thesis to design sensors, compare them to fabricated results, and extract parameters. As shown there is always a difference between fabrication and device layout. These discrepancies are calibrated out by individually correlating each grating and finding the actual frequency and delay.
Figure 2.3: An example RF probed reflection coefficient measurement of a OFC SAW sensor in (a) the frequency and (b) time domains operating at a center frequency of 915 MHz.

2.1.1 SAW Substrates for Sensor Applications

A goal of this dissertation is to extend the temperature range of the OFC sensor platform beyond that demonstrated in previous research efforts [9, 14]. Previous device implementations utilized aluminum electrodes on YZ-LiNbO$_3$ because of its large electromechanical coupling factor ($k^2$), low attenuation at high frequencies, and extensive characterization. The large coupling provides low-loss wideband operation and greater temperature sensitivity. However, lithium niobate and other extensively researched single crystal SAW substrates (quartz and lithium tantalate) are limited in operation to approximately 400°C [15]. Above this temperature lithium niobate will de-pole, losing piezoelectricity, and quartz twins.

Several alternative substrates exist for high temperature SAW sensor development including thin film on substrate and single crystal varieties [16, 17]. The most well researched of available materials is the langasite (LGX) family initially developed in Russia in the early 1980’s, and includes the materials langatate (LGT), langanite (LGN), and langasite (LGS) [18–20]. These
single crystal substrates are of class 32m symmetry, grown by the Czochralski method, and have similar piezoelectric properties to that of quartz. No phase transitions occur and piezoelectricity is maintained up to the melting point of 1450 °C. Extensive work has been conducted on material characterization first at low temperatures by [21] and more recently at temperatures up to 900 °C by [22]. Table 2.1 compares the basic SAW properties of langasite substrates to various common SAW wafer cuts.

Table 2.1: A comparison of common piezoelectric substrates and the performance of SAW cuts.

<table>
<thead>
<tr>
<th>Substrate</th>
<th>Euler Angle</th>
<th>Free Surface Velocity (m/s)</th>
<th>k² (%)</th>
<th>TCD (ppm/°C)</th>
<th>Maximum Temperature (°C)</th>
<th>Ref.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Y-Z LiNbO₃</td>
<td>&lt;0,-90,-90&gt;</td>
<td>3488</td>
<td>4.8</td>
<td>94</td>
<td>~450</td>
<td>[23]</td>
</tr>
<tr>
<td>128° Y-X LiNbO₃</td>
<td>&lt;0, 38, 0&gt;</td>
<td>3979</td>
<td>5.4</td>
<td>75</td>
<td>~450</td>
<td>[23]</td>
</tr>
<tr>
<td>ST-X Quartz</td>
<td>&lt;0,-47.25,0&gt;</td>
<td>3159</td>
<td>0.12</td>
<td>0</td>
<td>~575</td>
<td>[23]</td>
</tr>
<tr>
<td>Y-X LGT</td>
<td>&lt;0,-90,0&gt;</td>
<td>2210</td>
<td>0.44</td>
<td>-64.5</td>
<td>1450</td>
<td>[24]</td>
</tr>
<tr>
<td>LGS</td>
<td>&lt;0,138.5,26.6&gt;</td>
<td>2730</td>
<td>0.32</td>
<td>0</td>
<td>1450</td>
<td>[25]</td>
</tr>
</tbody>
</table>

Previous SAW device development has primarily focused on low frequency LGS implementations [26]. For this dissertation the langatate substrate has been investigated for temperature sensor applicability. High quality LGT boules were grown by Krystal Engineering in Titusville, Florida, and Y-cut wafer sizes of up to 3 inches in diameter were provided. High frequency SAW characteristics of the langatate substrate will be discussed. For wireless sensors, high frequency operation is ideal for minimizing the required antenna size, but other trade offs are necessary in
2.2 Antennas for SAW Sensors

Various traditional packaging and antenna configurations have been demonstrated for SAW sensors [27, 28]. Typical packaging for temperature sensors involves standard techniques demonstrated in the microelectronics industry, with application specific packaging implemented based on sensor environment. An example of a packaged SAW OFC temperature sensor is shown in Figure 2.4.

![Example of a packaged OFC temperature sensor](image)

Figure 2.4: An example of a packaged OFC temperature sensor consisting of a SAW die wire bonded into a ceramic package and soldered to a printed dipole.

2.2.1 Electrically Small Antenna Limitations

Ideally the footprint of an RFID tag or sensor occupies will be minimized. However, the overall volume occupied by a small antenna dictates its bandwidth and radiation efficiency. Wheeler and Chu defined an electrically small antenna as having a maximum dimension that occupies a sphere with radius \( r \) and

\[
k \cdot r < 1,
\]

(2.8)

where \( k = \frac{2\pi}{\lambda} \) [29]. As an antenna becomes smaller than this volume more energy is stored in evanescent fields and less is radiated into the far field.
The radiation quality factor, Q-factor, of an electrically small antenna is inversely proportional to the volume in which it occupies and has been theoretically derived in various publications. For a linearly polarized antenna, independent of geometry [30] defined the Q as

\[ Q = \frac{1 + 2(kr)^2}{(kr)^3[1 + (kr)^2]} . \]  

(2.9)

Fractional bandwidth is inversely related to Q by

\[ BW = \frac{S - 1}{Q\sqrt{S}} . \]  

(2.10)

where S is the voltage standing wave ratio of the matched antenna. Therefore, as the physical size of the antenna is decreased the bandwidth of the antenna will also decrease as more reactive energy is stored. Accordingly, a tradeoff must be made in the design of the antenna based on the gain, bandwidth, and size.
CHAPTER 3
CORRELATOR RECEIVER AND CODING DEVELOPMENTS

The orthogonal frequency coded (OFC) SAW concept was introduced theoretically in Chapter 2. However, there exists a large foundation of research that has been conducted beyond the initial concept to show feasibility of a multi-sensor platform [1,31,32]. This chapter will investigate several developments related to the SAW OFC system.

First a novel coherence correlator receiver routine will be presented that processes sensor information in the frequency domain. The updated correlator receiver system will be utilized throughout this dissertation for data extraction and is a general SAW receiver architecture. The algorithm demonstrates an improved noise immunity by coherently integrating the signal over the bandwidth of interest rather than utilizing a correlation peak detector. Based on this interrogation routine a discussion will be presented on inter-code interference for non-time orthogonal codes.

Finally an alternative device layout is demonstrated. A mixed orthogonal frequency coding (MOFC) technique is proposed that uses a multi-track SAW design for chip waveform manipulation. These devices provide flexibility in the realizable chip waveform and the ability to decrease the overall time response length for a constant bandwidth. These improvements provide greater flexibility when designing the bandwidth of the sensor system while controlling the length of the individual device, the inter-sensor interference, and total number of operable sensors.
3.1 SAW Sensor Matched Filter Receiver

Extensive research has been conducted on approaches for extracting sensor information from various wireless SAW embodiments. Initial work was conducted by Buff et al., based on tracking the shift in SAW resonator center frequency over temperature [33]. Resonate devices have been refined over the years and successfully applied in industrial applications [34]. There has been extensive work on SAW sensor interrogation using a frequency modulated continuous wave (FMCW) radar architecture [35]. By assuming the SAW delay line is a radar target, temperature is extracted from the change in reflected SAW pulse time delay. This technique has been extended to multi-pulse SAW delay extraction for greater accuracy using the frequency response phase change [36]. There has been extensive investigation into phase ambiguity caused by long delays requiring phase unwrapping and multi-step extraction techniques [37].

A signal waveform may not have a well-defined peak or even have constant group delay because of distortion from noise sources, inter-code interference, or inherent device frequency and phase dispersion. Pohl and his colleagues introduced the concept of spread spectrum coded multi-sensor systems interrogated by a correlator transceiver [11]. Their system interrogated the individual SAW tags first with the ideal sensor auto-correlation response, and then subsequently interrogated the sensor with a time scaled cross-correlation version. By comparing the level of the cross-correlation to the ideal response the sensor time delay could be extracted.

While the correlation process is a product integration of the expected ideal signal with the received signal, the matched filter process is a convolution of the received signal by an ideal version [38]. The matched filter process provides the highest signal to random-noise ratio, yielding the optimized detection condition. This is a general receiver architecture and can be utilized for any type of amplitude, phase, or frequency modulated signal because of the linear phase, band-limited frequency response. The perfect matched filter waveform is always a symmetric time domain compressed pulse, and dispersion is removed in the frequency domain. If the received signal time
delay is ideally recreated, the output of the matched filter is purely real in both domains and at \( t = 0 \) the output is equal to that of the auto-correlation.

An adaptive matched filter technique for use with SAW OFC temperature sensors has been published previously in [39, 40]. The receiver exploits the inherent SAW velocity change that occurs as the sensor moves over temperature to scale the time domain matched filter. The scaling is a function of the SAW substrates temperature coefficient of delay (TCD). TCD is defined as

\[
TCD = \frac{1}{\tau} \frac{d\tau}{dT},
\]  

where \( \tau \) and \( T \) are the reference SAW delay and temperature, respectively. The temperature coefficient of frequency can be defined as \( TCF = -TCD \). The parameter is assumed constant and linear over temperature for the following analysis.

A temperature dependent time scaling factor can then be defined as

\[
\alpha(T) = (1 + TCD \cdot \Delta T),
\]  

where \( \Delta T \) is the change in temperature relative to a reference temperature. At the reference temperature \( \alpha \) is 1. For YZ-LiNbO\(_3\) \( \alpha \) has an inverse relationship to temperature (TCD is negative), and heating the substrate increases SAW delay.

An ideal SAW response is used to generate a series of matched filters corresponding to discrete temperature steps. The transfer function of the ideal SAW response is synthesized as

\[
h(\alpha, t) = \sum_{i=1}^{N_c} a_i \cdot \cos \left[ 2\pi f_{chip} \left( \alpha t - \tau_{Di} \right) \right] \text{rect} \left( \frac{\alpha t - \tau_{Di}}{\tau_{chip}} \right),
\]

where \( N_c \) is the number of chips, \( \tau_D \) is the delay of each chip from the transducer, and \( a_i \) is a chip weighting coefficient. The individual chip waveforms can be modified beyond a simple rect function to include non-ideal waveform shapes. The factor \( \alpha \) scales the overall time domain response based
on the temperature of the sensor.

The matched filter correlation is then performed for each sensor, \( k \), in the frequency domain by

\[
G_k(\alpha, f) = H_R(f) \cdot \frac{1}{|\alpha|} H_k \left( \frac{f}{\alpha} \right)^*,
\]

(3.4)

where \( H_R \) is the signal at the input to the receiver ADC and \( H_k \) is the fourier transform of (3.3). \( H_R \) includes any sources of external noise and is assumed to be the summation of all signals in range,

\[
H_R(f) = \sum_{k=1}^{N} \left[ H_k(f) + E_k(f) \right] + H_{SN}(f) + H_{CN}(f).
\]

(3.5)

\( E_k \) is an error signal associated with each device due to device implementation and system effects. \( H_{SN} \) is random stationary noise and includes additive white gaussian noise, quantization error, and other noise terms. \( H_{CN} \) includes all constant noise and interference sources such as external jammers which are constant with time.

The output of the matched filter correlator gives a series of compressed pulses corresponding to a range of \( \alpha \) values. Device processing gain gives a compression of \( \tau_o/N_c \). Time gating is applied to isolate each compressed pulse and the peak value is plotted versus alpha. The maximum correlation energy corresponds to the actual temperature of the interrogated signal. All of the error and noise terms will set the minimal detectable signal within the receiver. This algorithm is illustrated further in Figure 3.1.
Figure 3.1: A demonstration of the adaptive matched filter process utilized for extracting the sensor temperature and RFID. A series of scaled matched filter functions, $H_{MF}$, are generated for a range of temperature values. Each matched filter is correlated to a time gated received signal, $H_R$, and the peak of the correlation response is found in the time domain. Plotting the maximum correlation signal versus scaling factor $\alpha$, the actual sensor temperature can be found.
3.1.1 Frequency Domain Coherence Correlator Receiver

The peak detection algorithm used previously is susceptible to high levels of noise that can cause correlation peak skewing or false peaks in the time domain. Correlation waveform shape is also dictated by sampling frequency, requiring curve fitting and zero padding. Because the convolution is typically performed in the frequency domain as in (3.4) and then transformed back to the time domain, this algorithm also requires multiple slow Fourier transforms. However, due to Parseval’s theorem the same signal information is contained in both domains, and the following section will outline a technique for extracting sensor information in the frequency domain.

In the optical domain a technique called coherence-domain reflectometry is extensively used to extract the location of a device under test by sweeping the location (delay) of a reference mirror and measuring the coherence of a reflected wideband white source [41]. This technique can be expanded to a matched filter correlator by assuming the received signal has a transfer function of the form

\[ H_R(f) = H_k(f) \exp(-j2\pi f \tau_D), \]  

(3.6)

where \( \tau_D \) is the unknown delay of the received signal, \( k \). For simplicity the explicit frequency dependence will be dropped from further notation, unless required.

Assuming an estimate of the received signal delay of \( (\tau_D + \Delta \tau_E) \), the necessary matched filter is defined as

\[ H_{MF} = H_k^* \exp(j2\pi f(\tau_D + \Delta \tau_E)). \]  

(3.7)

With \( H_k^* \) containing any code dispersion.

The output of the matched filter would then have the form

\[ G = |H_k|^2 \exp(j2\pi f \Delta \tau_E) \]  

\[ = |H_k|^2 \cdot [\cos(2\pi f \Delta \tau_E) + j \sin(2\pi f \Delta \tau_E)]. \]  

(3.8)
Therefore, $G$ will be maximized and purely real when the estimate error, $\Delta \tau_E$, is minimized, independent of frequency.

Based on sampling and error terms, $\Delta \tau_E$ will rarely be zero. By defining a delay error factor as

$$DF(\Delta \tau_E) = \sum_f \text{Re}[G(\Delta \tau_E)]$$

$$= \sum_f |H_k|^2 \cdot \cos(2\pi f \Delta \tau_E)$$

(3.9)

an estimate of the delay $\tau_D$ can be obtained by integrating over the band of interest. Sweeping $\Delta \tau_E$ over a range of values and determining the maximum $DF$ will yield a best estimate of the actual delay $\tau_D$. The delay step is independent of sampling and set by software.

### 3.1.1.1 Modifications for Random Sensor Ranging

The process above is easily applied to a multi-sensor SAW system by using the time scaling property in (3.2). By scaling $H_{MF}$ for a given sensor $k$ and sweeping $\Delta \alpha$ the correct temperature can be extracted by maximizing (3.9). However, if there exists a random temperature independent electromagnetic delay, $\tau_{EM}$, the form of the solution must be modified slightly. Looking at the frequency scaled version of the matched filter

$$H_{MF} = \frac{1}{|\alpha + \Delta \alpha|} \cdot H_{code_k} \left( \frac{f}{\alpha + \Delta \alpha} \right) \cdot \exp \left[ j2\pi f \left( \frac{\tau_D}{\alpha + \Delta \alpha} + (\tau_{EM} + \Delta \tau_{EM}) \right) \right]$$

(3.10)

there exists a large phase ambiguity caused by the difference in electromagnetic delay, $\Delta \tau_{EM}$. The transfer function of each sensor, $H_{code}$, includes the distributed code of the OFC RFID;

$$h_{code} = \sum_{i=1}^{N_c} A_i \cdot \text{sinc} \left[ \frac{\pi}{\alpha + \Delta \alpha} \cdot f_{ci} \right] \cdot \tau_c \cdot \exp \left[ \frac{2\pi f}{\alpha + \Delta \alpha} + \Delta \tau_{Di} \right]$$

(3.11)
ignoring the negative portion of the frequency spectrum. $H_{code}$ is a distributed frequency response with non-linear phase given by the chip separation $\Delta \tau_{Di}$, and this dispersion must be accurately removed to estimate the temperature. The device matched filter is calibrated at a reference temperature, setting $\alpha = 1$.

If the sensor is stationary relative to the transceiver $\tau_{EM}$ can also be fully calibrated. Otherwise when $\Delta \alpha = 0$, the output of the matched filter will become

$$G_k = \frac{1}{\alpha^2} \left| H_{codek} \left( \frac{f}{\alpha} \right) \right|^2 \exp(j2\pi f \Delta \tau_{EM})$$

$$= \frac{1}{\alpha^2} \left| H_{codek} \left( \frac{f}{\alpha} \right) \right|^2 \cdot \left[ \cos(2\pi f \Delta \tau_{EM}) + j \sin(2\pi f \Delta \tau_{EM}) \right].$$

If $\Delta \tau_{EM}$ is zero the output will be equivalent to (3.8); otherwise the real part of $G_k$ may no longer be maximized when the correct $\Delta \alpha$ is found.

However, the magnitude of $G_k$ remains an independent measure of the device correlation. Therefore, defining the alpha error factor as

$$AF_k (\Delta \alpha) = \sum_{f} |G(f, \Delta \alpha)|^2$$

$$= \sum_{f} \left[ \left( \frac{1}{\alpha} \cdot \frac{1}{\alpha + \Delta \alpha} \right) \left( H_{codek} \left( \frac{f}{\alpha} \right) \cdot H_{codek} \left( \frac{f}{\alpha + \Delta \alpha} \right) \right) \right]^2,$$

removes the dependence on phase minimization. A series of error factors can be found for a range of temperatures with a software defined step of $\Delta \alpha$. By properly applying a time gate to the expected device window, this technique removes the device orthogonal frequency code and extracts temperature simultaneously. By taking $|G(f, \Delta \alpha)|^2$ we get the total energy of the correlation signal by Parsavel’s theorem. The received signal is also bandlimited in frequency, and sets the noise floor and peak width of $AF$.

Once the correct device temperature is extracted the actual alpha value can be utilized
to create a temperature scaled matched filter for the device. The delay error factor (3.9) can be maximized to extract any residual delay due to device range. For a fixed range, 2 pulse configuration, (3.13) can be used to find the temperature of the reference pulse, and knowing the correct $\alpha$ a temperature compensated matched filter can be created for the second code. The difference in pulse delay caused by a second stimulus (strain, pressure, etc.) can then be extracted once again by (3.9).

### 3.1.2 Comparison of Extraction Methods

The benefit of the frequency domain coherence technique may not be readily apparent when compared to work previously demonstrated in [39]. For the ideal case both receiver types should contain the same information. However, the benefit occurs when trying to extract information from the peak correlation response.

Consider the received signal given by (3.5), if we want to extract the information for the first sensor in the system the output of the matched filter will become

$$G_1 = H_R \cdot H_1^*$$

$$= (|H_1|^2 + E_1) \exp(j2\pi f \Delta \tau_E) +$$

$$\sum_{k=2}^{N} (H_k + E_k) \cdot H_1^* \exp(-j2\pi f (\tau_{D_k} - \tau_{D_1} - \Delta \tau_E)) +$$

$$(H_{CN} + HSN) \cdot H_1^* \exp(j2\pi f (\tau_{D_1} + \Delta \tau_E)).$$

(3.14)

In the time domain this signal is compressed, but could be distorted by the error term and inter-sensor interference. By time gating and finding the change in signal energy by (3.13), the solution will have less bias toward a single erroneous peak.

Time domain skewing is exacerbated by low time resolution, $\delta t$. Accuracy in the time domain is set by the sampling rate with $dt = 1/f_s$. Typically the sampling rate is set by the Nyquist
rate as twice the bandwidth of interest, but for a distorted, compressed signal this accuracy typically is not enough. Correlation signals are usually zero padded in frequency, slowing data processing. As an example consider an OFC code consisting of 5, 50 ns long chips. Figure 3.2 compares the correlation of such an OFC device when sampled at the Nyquist rate and also one padded by zeros to 10x the Nyquist rate, which decreases $\delta t$ from 2.5 ns to 0.25 ns, respectively. For this example comparing these results to the output of (3.9) illustrates the benefit of the coherence technique. $DF$ calculation can be performed quickly with an arbitrarily small $\delta \tau$, 0.1 ns, by simple matrix multiplication.

![Normalized Magnitude](image)

(a) Nyquist Sampled  
(b) Zero Padded

Figure 3.2: A comparison of the output of a matched filter sampled (a) at the Nyquist rate of twice the bandwidth of interest and (b) zero padded to 10x the bandwidth of interest. The OFC code utilized consisted of 5, 50 ns long chips.

### 3.2 Time Division Multiplexed OFC

The issue of coding and RF identification is crucial for determining the number of simultaneous sensors that can be successfully identified at the receiver. In an effort to increase the number of simultaneously operating devices, some research efforts have focused on determining ideal code-
sets for delay line type SAW tags based on large bit sequences [42] [43]. Most orthogonal code sets assume high levels of synchronization, but multiple SAW sensors generally work asynchronously due to environmental effects and random electromagnetic time delays destroying the ensemble code optimization. In addition, simulations indicate that the overall transfer function of actual implemented SAW sensors convolved with transceiver system components, as well as nonlinear effects and fabrication tolerances, make overlapping codes exceptionally difficult to differentiate. Some previous publications ignored these effects or failed to account for composite, uncompressed sensor interaction [44]. In some cases the theoretical predictions claimed 10’s of sensors, but measured results reduce the number to 1-3 [45].

To operate a multi-sensor system with a large number of RFID sensors, it is necessary to optimize various diversity techniques such as time, frequency, spatial, etc. Discrete time multiplexing of individual sensors, which share bandwidth and independently change temperature, is ideal for accurate sensor detection because of minimized cross correlation levels. However, time orthogonality requires a large time window and long device lengths. A drawback of a passive SAW sensor system is devices cannot be turned off, and the total energy at the receiver will always be a superposition of each reflected signal. Even if the codes are time orthogonal, any time spreading (EM, temperature changes, non-linearities, etc.) will cause time interaction. The following section will investigate the effect of inter-sensor interference in an overlapping multi-sensor system utilizing the device parameter extraction presented in Section 3.1.

3.2.1 Signal to Interference Derivation

Consider an OFC sensor system as shown in Figure 3.3 where a stationary code is time gated and processed for extraction of time delay $\tau_k$ or other sensor parameter. Within that sensor window a number of undesired interference codes overlap the desired code with a time overlap of
\( \Delta \tau_{jk} = \tau_k - \tau_j \). The frequency domain transfer function of the desired OFC device is given by

\[
H_{S_k} \left( \frac{f}{\alpha} \right) = \frac{1}{|\alpha|} H_{\text{code}k} \left( \frac{f}{\alpha} \right) \exp \left( -j 2\pi \left( \frac{f}{\alpha} \right) \tau_{Dk} \right),
\]

(3.15)

and for the remainder of this derivation all temperature shifts will be assumed zeros, \( \alpha = 1 \). The matched filter for the desired code is assumed to be ideal and the complex conjugate of the reference transfer function.

![Diagram](image)

Figure 3.3: Illustration of an OFC multi-sensor system with arbitrary code overlap \( \Delta \tau_{jk} \). In this example the code centered at \( \tau_k \) is the desired device, and the code at \( \tau_j \) represents an overlapping, interfering code.

The total response of all sensors in range is

\[
H_{ST} = H_{S_k} + \sum_{j \neq k} H_{\text{code}j} \cdot \exp \left( -j 2\pi f \tau_{Dj} \right).
\]

(3.16)

This response can include an arbitrary device overlap either by design or temperature scaling. Ignoring temperature effects for now and concentrating on ideal overlap the output of the matched
filter function will give a response of

$$H_T = H_{SYS} \cdot [H_{ST} \cdot H_{S_k}^*]$$

$$= H_{SYS} \cdot \left[|H_{code_k}|^2 + \sum_{j \neq k} H_{code_j} \cdot H_{code_k}^* \exp \left( -j2\pi f \delta \tau_{Dj} \right) \right].$$  \tag{3.17}

From (3.13), the energy of the matched filter correlation is found by maximizing the summation over frequency. The relative signal to inter-sensor interference level can be found by comparing the delay error factor for the desired code to the cross correlation value of the other signals, or

$$SNR = \frac{DF_k}{DF_{jk}}$$

$$= \frac{\sum_f \left[ |H_{SYS}| \cdot |H_{code_k}|^2 \right]^2}{\sum_f \left[ |H_{SYS}| \cdot H_{code_{jk}} \cdot \exp \left( -j2\pi f \Delta \tau_{jk} \right) \right]^2},$$  \tag{3.18}

where $H_{code_{jk}} = H_{code_k}^* \cdot H_{code_j}$.

From (3.18), the system transfer function has no affect on the SNR calculation, assuming a near-linear passband response. The signal degradation can be estimated for arbitrary overlap in a multi-sensor system.

### 3.2.2 OFC Sensor Overlap Simulations

The following are ideal simulations of multi-sensor system overlap, excluding gaussian white and other noise terms. The simulations overlap several OFC codes as shown in figure 3.4, where a stationary sensor remains at delay $\tau_{Dk}$ and other sensors have staggered delays. The delays of the non-stationary sensor signals are defined by

$$\tau_{Dj} = \tau_{Dk} - (k - j) \cdot \delta t,$$ \tag{3.19}
where $\delta t$ is the simulation time step. An overlap of 1 chip between adjacent sensors is equivalent to 1/2 chip overlap at the lower and upper ends of the desired response time window. When $\Delta \tau_{jk} = 0$, there is a complete overlap of all sensors in the simulation.

Figure 3.4: Example OFC layout utilized for signal to interference simulations. One desired code remains stationary within a given time window, and undesired codes are time shifted to create an overlapping code set. Complete overlap occurs when $\Delta \tau_{jk} = 0$.

An ideal, 915 MHz, SAW OFC sensor system was simulated with 6, 50 ns long reflectors per sensor. The OFC codes have been randomized, but absolute SNR levels will change dependent on code sequencing. The overlap is controlled by $\delta t$, and for each step the SNR and extracted delay error was found. Figure 3.5 shows the calculated SNR when 2 codes are superimposed on top of the stationary code (3 code overlap) and the dashed lined shows the average SNR over a single chip. The random fluctuations in SNR are expected due to carrier interaction and random coding.
Figure 3.5: Signal to interference ratio of an OFC 3 sensor system. Each sensor is comprised of 6 chips. As $\delta t$ decreases the overlap of 2 interfering codes are increased. Dashed line represents the average SNR per chip.

Figure 3.6 demonstrates the average SNR over each chip for 2, 4, 6, and 8 overlapping code sets. As expected the greater the signal overlap the poorer the average SNR. As the number of interacting sensor signals increases, the worse the SNR and the poorer the extraction of device parameters, such as delay. As the number of simultaneously operating sensors in a system increases, the system and device parameters need to be optimized.
3.3 Mixed Orthogonal Frequency Coding

From Section 3.2, the ideal sensor system maintains orthogonality of device responses within a given time window. However, as device bandwidth decreases and the number of simultaneous non-overlapping codes increase, code lengths become prohibitively long and code diversity becomes more difficult. Although since time and frequency are orthogonal domains, device implementations can use this flexibility for optimizing the various system parameters. This section outlines a mixed orthogonal frequency code (MOFC) concept that maintains chip frequency orthogonality, but overlaps chip time responses within a given device. The overlapped chip time responses are implemented in two or more parallel SAW tracks, which are designed to simultaneously minimize intra-chip reflections, provide high reflectivity, decrease a sensor’s time length, and allow arbitrary desired bandwidth. The MOFC SAW embodiment allows serial-parallel chip placement to obtain more diverse time and frequency coding by adding a third dimension, spatial variation, to the embodiment.
Previously reported OFC designs placed each chip in a discrete sequential time cell, as shown in Figure 2.2, though this is not implicit in the OFC definition. OFC devices can also utilize non-orthogonal time diversity while maintaining the frequency orthogonality condition. If individual chip delays, $\tau_D$ are relaxed by allowing

$$\tau_{Di} \leq \tau_{Di+1} < \tau_{Di} + \tau_{chip},$$

an MOFC chip will be produced. Physically, the overlapping chips are spatially implemented through a multi-track SAW layout in which multiple SAW propagation paths recombine electrically at a parallel/series transducer configuration. As shown in [46], the transducer configuration is also further optimized for matching to the sensor antenna.

### 3.3.1 Illustrative MOFC Waveform Creation

As an example, for a special two chip MOFC case where

$$\tau_{Dm} = \tau_{Dn} = 0,$$  \hspace{1cm} (3.21)

and with m and n being adjacent orthogonal frequencies, (3.3) can be expanded to produce the transfer function of a cosine-envelope waveform composed of the sum of orthogonal sinc functions in frequency. In the time domain this transfer function is given by

$$h_{\cos}(t) = \cos \left( \frac{\pi t}{\tau_{chip}} \right) \cos \left( \frac{2\pi t(f_{chip_m} + f_{chip_n})}{2} \right),$$

(3.22)

due to orthogonality setting

$$\tau_{chip} = \left( \frac{1}{f_{chip_m} - f_{chip_n}} \right).$$

(3.23)
In this form the average of the two chip frequencies becomes the signal carrier frequency [47]. The cosine envelope chip is illustrated in Figure 3.7 where the waveform function is derived from the sum of the two individual chips in the frequency domain. The chip response in the time domain exhibits the expected cosine modulation and illustrates a method where parallel tracks can be summed to produce a desired amplitude weighted waveform from multiple uniform time domain waveforms. Expanding upon the result in (3.22), if the two chip frequencies are not adjacent than the MOFC chip will instead be modulated with a frequency equal to a multiple of \(2 \cdot (\tau_{chip}^{-1})\).

![Figure 3.7: (a) A demonstration of the summation of two synchronous orthogonal chips (--- and \(-\cdot-\)) yielding a cosine-envelope waveform response in the frequency domain. The signal carrier frequency, \(f_0\), is the mean of the individual orthogonal frequencies and the bandwidth of the overall chip is \(2 \cdot (\tau_{chip}^{-1})\). (b) The equivalent time domain signal is shown, verifying the cosine weighted chip response.](image)

Physically this device is implemented by a two track SAW configuration as shown in Figure 3.8, where the Bragg reflector in the second track is offset a quarter wavelength from that in the first. By properly offsetting one track from the other the peak of the cosine will occur at the midpoint of \(\tau_{chip}\) and the response in time will resemble an ideal truncated cosine. The device was designed for operation at 915 MHz, with two parallel tracks, and each reflector has 50 shorted
electrodes. Figure 3.9 compares the results of a probed device to a coupling of modes (COM) simulation, as outlined in [13].

Figure 3.8: An example of a fabricated 915 MHz cosine envelope reflector structure on YZ-LiNbO$_3$. The device has two identical SAW transducers connected in parallel (left) and two overlapping orthogonal frequency chips (right).

Figure 3.9: Experimental results of an RF probed cosine envelope reflector structure. (a) S11 time domain results of the device comparing the device to the coupling of modes (COM) simulation. The device shows minimum distortion due to the non-ideal stored energy of the reflector. (b) The time domain response is gated and the fourier transform shows the frequency domain response.

As a second illustrative example, as more chips with identical delays are added in parallel, the dispersion of a typical OFC code is removed and the time and frequency responses resemble that of a non-dispersive filter with a carrier frequency equal to the average of the individual
chip frequencies. This technique is well known for the design of filters and can be modified for OFC coding applications [48]. Figure 3.10 shows the ideal result for a five chip synthesis, with equal amplitude, and all frequencies occurring simultaneously in time. As shown, the frequency magnitude response resembles a typical wideband SAW filter, with constant group delay, and the time response resembles a truncated sinc function corresponding to that bandwidth. The approach can be implemented using five parallel tracks each having a chip utilizing reflectors, transducers, or both.

![Frequency Response](image1)

![Time Response](image2)

Figure 3.10: An example of the theoretical response of a 5 chip OFC code with all chips occupying the same time slot. In the frequency domain (b) the individual chip frequencies are shown (— — —) along with the overall synthesized frequency response. The dispersion of a typical OFC response is removed and the time domain plot (b) resembles a typical non-dispersive filter with a carrier frequency equal to the average chip frequencies.

The device was also built on LiNbO$_3$, as shown in Figure 3.11, with four chip frequencies centered about 915 MHz. The device has two transducers in parallel, each with a cosine modulated chip as shown previously. Probed experimental S11 responses are shown in Figure 3.12. Time domain results in Figure 3.12a highlight skewing in the impulse response due to non-ideality in the reflector response. Gating this time response and taking the fourier transform gives the frequency

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response of the composite chip, shown in Figure 3.12b. Optimization would yield a more ideal frequency response.

Figure 3.11: A fabricated multi-frequency non-dispersive reflector structure for operation at 915 MHz on YZ-LiNbO$_3$ viewed under an optical microscope. The structure consists of 2 parallel transducers (left) with 4 overlapping orthogonal Bragg reflectors (right).

Figure 3.12: Experimental results of an RF probed multi-frequency non-dispersive reflector structure. (a) S11 time domain results of the device comparing the device to the coupling of modes (COM) simulation. The device shows the skewing toward the beginning of the impulse response due to the non-ideal response of a SAW reflector. (b) Frequency domain measurement of the gated reflector impulse response showing the filter like characteristic of the device.

In general, the SAW MOFC implementation can use serial/parallel chips in multiple tracks with varying time overlaps to achieve optimum auto- and cross-correlation properties. However, for simplicity, all of the devices discussed and fabricated here will have a time overlap of a maximum
of two chips using two parallel tracks, although theoretically the overlap can be made arbitrary for an increased and optimized codeset.

As a final example, an MOFC and OFC coded signal format comparison using ideal transfer functions is presented. Assuming a center frequency of 915 MHz with five frequencies and a 7% fractional bandwidth, a serial five chip OFC signal and a five chip MOFC signal having only three discrete time cells were synthesized. The MOFC signal is 40% shorter in time than the OFC signal, which will aid in minimizing code collisions from concurrently received signals and also optimize overall die size. Figure 3.13 illustrates the autocorrelation properties of the OFC and MOFC signals, with peak to side lobe levels for both devices greater than 15 dB. As expected, the width of the compressed pulses are equal due to the equivalent bandwidth for both code types. A 2.2 dB decrease in processing gain for MOFC compared to OFC signals is obtained as a result of the shortened MOFC response time length.

![Figure 3.13: A comparison of the autocorrelation of a five chip OFC device to that of a similar device that contains mixed frequency cells. The correlation peak widths are identical, and MOFC has nearly 15 dB correlation sidelobes over three chips in time.](image)

The MOFC device simulated in Figure 3.13 did not have any two adjacent frequencies
occupying the same cell. When this restriction is removed, as in Figure 3.14, the autocorrelation shows an increase in sidelobe level to greater than 10 dB, illustrating that code selection and design can be optimized within a system. Also of note, chip overlaps can be variable in time, with partial overlaps across or within multiple cells, to achieve optimum correlation properties within a code or codeset.

![Autocorrelation Diagram](image)

Figure 3.14: A comparison between the autocorrelation of an MOFC device with adjacent frequencies, e.g. $f_1/f_2$, in the same cell and without, highlighting the decrease in peak to sidelobe level when adjacent chip frequencies overlap in time.

### 3.3.2 Experimental Device Design and Results

For a proper comparison on the performance of a MOFC to a traditional OFC device, as utilized in [9], both types of devices have been similarly designed. These comparative OFC and MOFC devices were fabricated on YZ-LiNbO$_3$ for fundamental operation, with a minimum feature size of 0.9 µm. Each reflector was designed with 68 shorted electrodes, such that a five chip sensor bandwidth occupies a 70 MHz interrogation bandwidth at room temperature. For these grating lengths, intra-chip reflectivity and interaction in a serial embodiment is no longer negligible, and
must be considered to maintain good correlation properties. In order to minimize this adjacent frequency chip interaction, a two track approach was implemented with the restriction that all even frequency chips would be in a separate track from odd frequency chips. An example of a fabricated MOFC sensor is shown in Figure 3.15. As shown, the device uses five chip frequencies within three cells in two adjacent parallel tracks; each IDT is wideband, covering the required bandwidth.

![Figure 3.15: A fabricated 915 MHz MOFC device on YZ-LiNbO$_3$ viewed under an optical microscope, transducer (left) and Bragg reflectors (right). The device utilizes two parallel tracks with five chip frequencies in three cells. Two wideband transducers are connected in parallel by the thicker dark traces used for RF device probing and for wire bonding.](image)

Figure 3.16 compares the probed results of a two track OFC device to a coupling of modes (COM) simulation. Inter-chip roll off has been minimized to 3 dB, and a slight gap has been added between chips to eliminate any stored energy interaction. The two track approach yields a 3 dB increase in insertion loss due to the multiple transduction paths, but the parallel transducer configuration was designed for a close conjugate match to the antenna. The devices have a $9\lambda_0$ long interdigital transducer with an approximate 2 dB mismatch loss between the sensor and planar dipole antenna.
Figure 3.16: A comparison of RF probed, experimental data to that of a coupling of modes simulation for a five cell OFC device. The relative chip frequencies are shown in the time domain plot.

Similar constraints were used to design a four device set of MOFC devices with five frequencies in three cells. Each MOFC cell contains two chips with a complete spatial overlap and one cell per device with only a single chip, similar to the example shown in Figure 3.15. The codes were designed for minimal cross-correlation with the OFC devices, and with the restriction that no adjacent chip frequencies occupy the same spatial cell. The measured results of a fabricated MOFC device are shown in Figure 3.17 and exhibit strong agreement to COM simulations. As expected, the amplitude response of the two MOFC cells, first and third, exhibit a modulation of the typical chip response relative to the frequency separation of the Bragg reflectors.
Figure 3.17: A comparison of RF probed, experimental data to that of a coupling of modes simulation for a three cell MOFC device. The relative chip frequencies are shown in the time domain plot.

An ideal matched filter was calibrated to the experimental measurements using the transfer function in (3.3) which accounts for fabrication tolerances. Results of the correlation of this matched filter to the data shown in Figure 3.17 are presented in Figure 3.18, compared with the autocorrelation of the ideal matched filter. The peak to sidelobe level shown is greater than 13 dB throughout the time window of interest for post processing.
The designed and measured device results are consistent with expectations. The MOFC and OFC devices had equivalent bandwidths and yielded equivalent correlation coherence times for similar designs. The MOFC device yields a 40% reduction in code time length but at the expense of a 40% decrease in processing gain, consistent with the devices time-bandwidth product. The reduced MOFC code time length allows a greater number of sensors for a given inter-code interference level in a multi-sensor system and, the overall device size is also reduced, allowing more devices to fit within any length constraint. By designing with even and odd chip frequencies in parallel tracks, the SAW intra-symbol chip effects are minimized at the cost of a 3 dB increase in insertion loss when using broadband transducers.

3.3.3 Wireless Multi-Sensor System Tests

A total of eight devices were fabricated, for simultaneous use as temperature sensors, and each device was mounted to a bent dipole antenna on an FR4 substrate for wireless testing [28]. Figure 3.19 plots the individual wireless VNA measurements of the eight sensors on a single graph.
to show how the devices are designed spatially (time delay offset) in relation to each other. The sensors are multiplexed in time, alternating between OFC and MOFC devices, allowing each set to be used individually or as a complete, eight device set. The available time window for all sensors is set by the pulse repetition rate of the interrogator, with the window set at 5 $\mu$s for the system used. Within that window, each device occupies a discrete time slot that accounts for device drift with temperature, while limiting composite device interaction. These devices were fabricated on YZ-LiNbO$_3$, having a temperature coefficient of $-94$ ppm.

Figure 3.19: Time domain measured data of eight 915 MHz wireless sensors measured independently, at 10 cm, by a network analyzer. The set of devices contains four MOFC devices (light grey) that have been interlaced between four traditional OFC devices (black).

A chirped transceiver system was utilized for wireless testing of the devices. The interrogator produced a 700 ns chirp with 28 dBm of peak power available to the transmit antenna [49]. The system was configured with two identical dipole antennas for transmit and receive, each with 2 dBi of gain. The adaptive matched filter software routine developed in Section 3.1 was able to track realtime temperature for each sensor.

The set of eight devices were randomly placed about the interrogation antennas in a labora-
tory environment, at distances between 1 and 4 meters. Figure 3.20 shows the extracted temperature of the eight devices simultaneously received and processed, with four interrogation averages per sample reading. Device 1002 was cooled by liquid nitrogen and devices 1006 and 1005 were heated by a heat gun. The three devices being heated/cooled were fitted with a wired thermocouple and the two measurements track well, with variation due to mechanical attachment and differences in thermal conductivity. The expanded graph shows the temperature tracking for the additional five sensors present in the system that were kept at room temperature. Temperature variation from the nominal value for all devices is $\leq 2$ °C over a temperature range of $\pm 130$ °C.
Figure 3.20: Measurement results from a set of 8 concurrent SAW OFC temperature sensors being interrogated by a chirped transceiver system. Four OFC and four MOFC devices are arbitrarily placed at distances between 1 and 4 meters in a lab environment. The solid lines represent the extracted temperature from the SAW device and the · · · lines plot the temperature from a wired thermocouple. The sensor temperature variations were introduced by successive heating and cooling, allowing 'free running' variations.
CHAPTER 4
INTEGRATED WIRELESS SAW SENSORS ON LITHIUM NIOBATE

For certain SAW sensor applications the challenge is building a wirelessly interrogatable device with the same lifetime as the SAW substrate. The design of these application intensive sensors is complicated by the degradation of device bond wires, adhesive, and antenna substrate. The purpose of this chapter is to demonstrate the effectiveness of integrating the sensor and antenna directly onto the SAW substrate. These techniques directly connect the thin-film SAW device to a thick film antenna metallization, eliminating external interconnects. Results shown will highlight a comparable antenna performance to that of a traditionally packaged and interrogated device. Devices from Section 3.3 will be utilized as example temperature sensors integrated onto YZ-LiNbO$_3$.

4.1 On-Wafer Antenna Design

The main considerations when designing an antenna for a SAW sensor is matching the transducer bandwidth, a function of device coding and expected temperature variation, and maximizing the radiation efficiency. A SAW substrate is not ideal for meeting these requirements due to its high relative permittivity and also the uniaxial anisotropy between the crystal axes. Initial integrated sensors used lithium niobate because of the availability of large wafer sizes, large SAW coupling, and previous delay line SAW design experience. There is a 2:1 difference in permittivity between the X or Y ($\epsilon_{11} = 45.6$) and Z ($\epsilon_{33} = 26.3$) cuts [50]. These properties make the use of a patch antenna difficult due to the high permittivity and thin substrate giving a narrow bandwidth and
low efficiency [51]. A loop antenna would be a natural choice for its inductive input impedance; however, for a single loop the radiation efficiency is poor [52]. Therefore, for proof-of-concept a dipole was chosen for an antenna embodiment.

Standard LiNbO$_3$ wafers come in sizes up to 100 mm, which is too small for a half-wave dipole at the target center frequency of 915 MHz. As discussed in Chapter 2, the Q of an electrically small antenna is reciprocal to the volume the antenna occupies and sets the maximum bandwidth. The bandwidth can be expanded by detuning the antenna, but with the effect of decreasing the radiation efficiency. At 915 MHz the electrically small antenna limit is a sphere with a radius of 52 mm, and therefore the design of an antenna fitting on a substrate with radius 38.1 mm will require a tradeoff of efficiency and bandwidth. A meander-line dipole was chosen for miniaturization as the radiation efficiency is acceptable for a small antenna [53].

The realizable gain of the antenna sets the read range of sensor tag based on Friis transmission equation [52]. Ignoring polarization mismatch and dielectric losses, the gain of the antenna is defined as

$$G(\omega) = \eta_r(\omega) \left(1 - |\Gamma(\omega)|^2\right) D(\omega). \tag{4.1}$$

The maximum antenna gain is set by the antenna directivity, $D(\omega)$, and independent of geometry for a small antenna it approaches a value of 1.8 dB. However, the realizable gain is a function of the efficiency

$$\eta_0(\omega) = \eta_r(\omega) \left(1 - |\Gamma(\omega)|^2\right). \tag{4.2}$$

$(1 - |\Gamma(\omega)|^2)$ represents the impedance mismatch loss at the SAW/antenna interface. The radiation efficiency $\eta_r$ is defined as $R_r/(R_L + R_r)$, the ratio of power radiated to that lost to conductor or dielectric losses. Therefore it is important to design an antenna that minimizes the mismatch to the sensor and also does not introduce high resistive losses.

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4.1.1 SAW Sensor Impedance

Before discussing the design of the on-wafer antenna it is important to discuss the input impedance of the SAW sensor. Ignoring parasitics, the input admittance of a SAW transducer is comprised of a parallel network of three terms:

\[ Y(\omega) = G_a(\omega) + jB_a(\omega) + j\omega C_T \]  \hspace{1cm} (4.3)

where \( G_a \) is the acoustic conductance, \( C_T \) is the electrode capacitance, and \( B_a \) the symmetric Hilbert Transform susceptance \([23]\). The input impedance is typically dominated by the electrode capacitance which is a function of the substrate and device geometry. At center frequency the Hilbert transform susceptance can be ignored, and from the impulse response model the transducer capacitive susceptance is

\[ X(\omega_0) = j\omega_0 \tilde{C} \tilde{C}_s N_p W_a. \]  \hspace{1cm} (4.4)

The transducer electrical capacitance is constant over frequency and a function of the normalized capacitance based on transducer sampling \( \tilde{C} \), the length of the transducer in electrode pairs \( N_p \), the width of the transducer \( W_a \), and the substrate dependent static per finger capacitance \( C_s \). The acoustic conductance of a short transducer can be approximated as

\[ G_a(\omega_0) = \frac{1}{2} \omega_0 \tilde{G}_a k^2 C_s W_a N_p^2, \]  \hspace{1cm} (4.5)

with \( \tilde{G}_a \) representing the normalized conductance based on transducer sampling and harmonic of operation. The electrical quality factor of the transducer can thus be calculated from the above equations as:

\[ Q_T = \frac{X(\omega_0)}{G_a(\omega_0)} = \frac{2\tilde{Q}}{N_p k^2}. \]  \hspace{1cm} (4.6)
For YZ-LiNbO$_3$ $k^2$ is 4.8\% and $C_s$ is 4.6 pF cm$^{-1}$. From the impulse response length, the 4 dB acoustic bandwidth of the SAW transducer is given by $\Delta \omega/\omega_0 = N_p^{-1}$. From the electrical quality factor above, for a given acoustic bandwidth the impedance of the transducer is fixed along a constant Q-factor arc on the Smith chart. The necessary acoustic bandwidth of each transducer is a function of the OFC code layout and the two do not necessarily match. From the Bode-Fano criterion higher Q circuits are intrinsically more difficult to match over frequency [54].

The goal is to minimize the mismatch loss between the SAW sensor and antenna, without the use of a matching network. Since a SAW transducer is inherently a balanced impedance element there is no need for a balun. For a given sensor bandwidth, the impedance can be manipulated by modulating the individual transducer beamwidth. In [46], Kozlovski and Malocha demonstrated the of manipulation of impedance through the use of parallel/serial combinations of individual transducers. In order to minimize the reflection at the SAW/antenna interface, the transducer would lie at the point on the Q-arc 90° from $Z_0$, assuming the antenna is matched directly to $Z_0$, on the Smith chart. However, this would not maximize the power transferred from the antenna to the SAW transducer which can be calculated as

$$ P = \frac{1}{2} \left| V_g \right|^2 \frac{R_S}{(R_S + R_A)^2 + (X_S + X_A)^2}. $$

(4.7)

Assuming the SAW impedance is equal to $Z_S = R_S - jX_S$, maximum power is transferred when the antenna impedance is conjugately matched as $Z_A = R_S + jX_S$. This sets the reflection coefficient looking into the SAW sensor at $\Gamma_S$ and looking into the antenna at $\Gamma_A^*$. In [55] a power reflection coefficient is defined as

$$ \Gamma_p = \frac{Z_S - Z_A^*}{Z_S + Z_A^*}, $$

(4.8)

and this value will be utilized when comparing the mismatch loss, $(1 - |\Gamma_p^2|)$, of the antenna throughout this dissertation. At center frequency, conjugately matching minimizes mismatch loss.
and maximized power transferred to the SAW at \( P = V_g^2/8R_S \).

For a dipole by designing the antenna to be electrically long the inductive reactance of the antenna can conjugately match the capacitive reactance of the SAW sensor. The meander dipole allows more design freedom for creating an arbitrary antenna input impedance because there are more dimensions for tuning. If the Q of the antenna is less than the electrical Q of the transducer bandwidth skewing will be minimized. The Q of the antenna is a relation between the power stored in the reactive field and power radiated, and dependent on antenna geometry as discussed in Chapter 2.

### 4.1.2 Meander Dipole Design

The antenna layout used for these initial devices is shown schematically in Figure 4.1, and represents the extensively researched meander dipole antenna [53, 56, 57]. The design of these antennas start with a straight dipole of a given length that is electrically short and has a capacitive input impedance. The resonant frequency of the antenna is lowered by maintaining overall length while increasing total wire length through meander sections [58]. These sections do not contribute to the radiation of the antenna because the current is orthogonal to the main dipole current, but rather contribute mutual inductance to the antenna impedance. By increasing the height, \( h \), of each meander section the mutual inductance will increase at the cost of an increase in parasitic resistance, lowering the radiation efficiency. For a helix type antenna the current vector in each meander section is constructive. However, for a planar dipole the current in each section is opposite in polarity and decreasing the width, \( w \), will cancel the increasing mutual inductance. This effect is more pronounced on the high permittivity lithium niobate substrate because of the large capacitance between each meander section. By decreasing the width of the antenna traces more meanders can be placed within a given area for a lower resonant frequency, but at the cost of a lower radiation efficiency from an increased parasitic resistance.
Antenna designs were conducted using Ansoft HFSS® 3-D full wave electromagnetic simulator. HFSS allows uniaxial dielectric anisotropy to be incorporated into the simulation. For comparison two antennas were simulated on YZ-LiNbO$_3$, one slightly long electrically and one resonant at the SAW center frequency. Regular straight dipole antennas were simulated on an artificially large substrate to remove the effect of meander sections on the quality factor. Figure 4.2 compares the power reflection coefficient calculated by (4.8) at the interface of each antenna to that of a simulated SAW transducer with 9 electrode pairs.
Figure 4.2: (a) Simulated input impedance of a SAW transducer compared to an electrically long dipole antenna and resonant dipole on YZ-LiNbO$_3$. The substrate is allowed to arbitrarily sized and dipole antennas are regular straight 1 mm wide copper traces. 915 MHz is marked by •. (b) The calculated power reflection coefficient at the interface of the SAW and antenna. SAW center frequency is 915 MHz and has a 100 MHz bandwidth.

A final design was chosen that conjugately matched a 915 MHz SAW device with 7% fractional OFC bandwidth. Predicted radiation efficiency is 82%, and max directivity is 1.6 dB. The final antenna design has a d of 1 mm, h of 11 mm, w of 4 mm, and 10 total bends. The overall length of the antenna is 60 mm, which is short enough to fit 3 antennas onto a 3 inch wafer. A straight dipole of the same length would resonate at 1.4 GHz. Predicted current distribution is shown in Figure 4.3. As expected the dipole has a linearly decreasing current profile because of the length being less than $\lambda/4$. The strong coupling between meander sections is also evident. Simulated antenna impedance is shown on a Smith chart in Figure 4.8 and compared to an example fabricated device. The fabrication of the antenna will be discussed in the following sections.
4.2 Electroplated Gold On-Wafer Antenna

For demonstration purposes, initial device fabrication was performed with traditional physical deposition techniques. The OFC SAW was patterned first using aluminum deposited by electron beam evaporation, with a titanium adhesion layer. Depositing the SAW first allows for RF probing and also the use of contact photolithography for the necessary sub-micron SAW linewidths.

Various physical deposition techniques were tested for defining antenna patterns, with electroplated gold having the best adhesion and lowest resistivity. Figure 4.4 shows an example integrated antenna fabricated on a black YZ-LiNbO$_3$ wafer. A thick titanium adhesion layer and gold seed were patterned by contact mask aligner and deposited by electron beam evaporation. Gold electroplating was then performed to a final thickness of 2.5 µm after trial and error of determining the ideal plating settings. Results from these initial tests exposed possible improvements to the antenna design that were implemented in the following section.

Figure 4.3: Simulated current distribution of the designed meander dipole antenna operating at 915 MHz.
Figure 4.4: An example integrated SAW sensor fabricated on black YZ-LiNbO$_3$ with electroplated gold antenna pattern. Antenna size as shown is approximately 50mm x 14mm.

4.3 Direct Write Copper On-Wafer Antenna

The initial devices fabricated suffered from poor adhesion and high resistive losses due to the relatively thin metallization. In order to improve upon the electroplated devices, a direct write process was used to pattern a copper antenna on lithium niobate. Direct write fabrication has the advantage of rapid conformal fabrication without the need for a mask. Antenna traces can be arbitrarily thick, and simultaneously exhibit superior adhesion compared to other physical deposition methods.

4.3.1 Direct Write Background

Several techniques are available for the direct write of patterns onto materials, including ink jet, laser assisted, and plasma spray; each process has its own drawbacks and merits. For this dissertation work a partnership was formed with Mesoscribe Technologies for the patterning of copper on lithium niobate. Mesoscribe has developed a proprietary high-definition plasma spray process that creates traces by injecting material in powder form into a plasma flame and directing the molten material toward the substrate [59]. Dynamic collimation technology allows the line width to be controlled and other process parameters, gas flow and power level, control the density.
A strength of the Mesoscribe process is the ability to conform onto non-planar surfaces while maintaining high throughput production for research or commercial products. A variety of materials can be deposited including dielectrics, metals, or sensor alloys. Materials can be stacked for integrated packaging and electrical isolation, eliminating adhesives. Films are robust and high temperature tolerant. Traces are typically $\geq 20\mu m$ in thickness and trace widths are actively controlled at $\geq 250\mu m$. The width can also be laser scribed after deposition to a width of $25\mu m$. Deposition is fully automated and performed by a 6-axis robotic control spray head, as shown in Figure 4.5a. Traces are able to write up and over a variety of surfaces and at varying angles; Figure 4.5b shows a copper trace written from a circuit board, up the die adhesive, and connecting to a SAW delay line. These films adhere well to substantially thinner dissimilar metals, aluminum in this example.

Figure 4.5: (a) The Mesoscribe automated 6-axis direct write robot writing material to a part. (b) An example of copper directly written onto a SAW delay line on lithium niobate. The diced SAW device is attached to a printed circuit board by adhesive and connections are made by directly writing copper traces up and onto the thin-film aluminum device.

A process was developed for depositing high quality copper traces onto LiNbO$_3$. Fig-
Figure 4.6a shows an example of a 1 mm wide copper trace written onto a 3 inch diameter wafer. No surface preparation or trenching is required prior to deposition. Copper adheres well without an adhesion layer, passing a simple tape test. The copper is nominally 20 µm thick. Measurement of film thickness by contact profilometer yields an average surface roughness of 2.1 µm, shown in Figure 4.6b, which is also evident in the optical image of the copper. These thick copper lines successfully bond to the 80 nm electron beam evaporated aluminum pad. The copper trace has been laser trimmed to provide a taper to the 0.5 mm wide aluminum probe pads of a SAW transducer. Measured sheet resistance is approximately 1.2 mΩ/□.

Thick conductors are important in reducing resistive losses of microwave transmission lines because of skin depth effects [60]. A thin conductor, closer to the skin depth, will exhibit higher losses. For the frequency of 915 MHz, the skin depth is 2.2 µm or 10% of the total thickness. The thickness achieved is an 8x increase compared to electroplated devices in Section 4.2.

Figure 4.6: (a) Optical image of a 1 mm wide copper trace directly written onto a 3 inch diameter LiNbO₃ wafer. The line has been laser trimmed to interface with an 80 nm aluminum probe pad. (b) A surface scan of the Cu/Al junction on YZ-LiNbO₃ performed by contact profilometer showing the relative surface heights and roughness. Average step height is approximately 20 µm for the copper and 80 nm for the aluminum.
4.3.2 Direct Write Antenna Results

Several experimental devices were fabricated for testing as integrated sensors and also for experimental verification of the stand-alone antenna. SAW metallization is performed first as described in Section 4.2, with Ti/Al deposited by electron beam evaporation. A photoresist mask is applied to protect the SAW propagation path from debris generated when depositing and trimming the antenna. Antennas have been written at various thicknesses and levels of corner trimming. However, antennas shown are only trimmed to taper to the SAW pads and for any over spray on either end. This allows for higher throughput of integrated devices, but does not remove corner rounding or over spray of traces. All copper shown is approximately 20µm in thickness. An example diced SAW/Antenna device is shown in Figure 4.7, and compromises the middle third of a 3 inch diameter YZ-LiNbO₃ wafer.

![Figure 4.7: An example of a 915MHz SAW sensor and direct write antenna integrated onto a 3 inch diameter, Y-cut lithium niobate wafer.](image)

4.3.2.1 Antenna Characterization

After fabrication, the SAW transducer was removed for stand-alone antenna characterization. Impedance was measured at the input to the antenna, and Figure 4.8 shows the results. The figure also plots the response of the probed SAW transducer of interest and the designed antenna layout. The SAW device utilized has a 9 pair transducer, quarter wavelength electrodes, and 2-50λ beamwidth transducers in parallel. Differences in antenna impedance between simulation
and measurements can be attributed to a shift in the reference plane during measurement. For measurement the diced antenna was placed on an FR4 substrate and short leads made connection from the SMA connector to antenna, shifting the reference plane and adding additional parasitic resistance.

Figure 4.8: Impedance Smith chart comparison of the response from an RF probed SAW device, simulated meander dipole antenna, and measured on-wafer antenna.

The measured antenna impedance was used to calculate the mismatch loss at the SAW/Antenna interface. The SAW impedance is determined from RF probed results taken prior to antenna fabrication. Figure 4.9 compares the mismatch of the on-wafer antenna to that of a simple dipole fabricated on a traditional FR4 substrate. These results do not consider added parasitics from the device packaging required when mounting to the FR4 antenna. The results highlight a shift in device center frequency to the lower part of the sensor bandwidth. The following section will demonstrate the relative performance of the integrated antenna to that of the packaged sensor.
Figure 4.9: The predicted and extracted mismatch loss at the SAW/antenna interface found by HFSS simulation and measurement of both a traditionally packaged FR4 device and on-wafer direct write device.

4.3.2.2 Wireless Sensor Measurements

Figure 4.10 compares the wireless measurement of similar wideband reflective delay line temperature sensors connected to direct write and FR4 antennas. A vector network analyzer was used to wirelessly interrogate each device at a distance of 20 cm. Each device has a coded reflector bank with relative orthogonal reflector frequencies shown. Devices are designed to have 7% fractional bandwidth and a center frequency of 915 MHz [14]. The distributed frequency reflector bank of these devices illustrate the relative bandwidth of each SAW/Antenna combination. Device designs are identical and minimal SAW fabrication differences are assumed. Loss is comparable between the FR4 and integrated devices throughout the band of interest.
Figure 4.10: A time domain comparison of the separate measurement of two SAW sensor configurations, interrogated wirelessly at 20 cm by a vector network analyzer. Each device has approximately 7% fractional bandwidth and the relative center frequencies of each coded reflector is shown.

An ideal matched filter was generated for the two measured devices. Figure 4.11 compares the relative correlation of the matched filter to the data obtained wirelessly at 20 cm. As expected the width of the two correlation peaks are the same because the processing gain between the two devices should be similar. There is a 0.5 dB decrease in correlation amplitude between the FR4 and direct write because of slight differences in the reflected SAW energy. These results highlight the ability of the on-wafer antenna to operate in a coded multi-sensor environment without added attenuation.
Figure 4.11: Matched filter correlation of an integrated on-wafer device and a standard packaged device. Both responses are normalized to the integrated device for comparison. Measurements were taken wirelessly at 20 cm by a vector network analyzer.

The integrated sensors were tested over temperature as shown in Figure 4.12. The wireless tests consisted of three sensors interrogated simultaneously by a chirped interrogator and temperature extracted by the method described in Chapter 3. As a comparison an integrated and traditionally packaged device were placed in a closed container and cooled simultaneously with liquid nitrogen. A separate integrated device and wired thermocouple were left outside of the container and kept at room temperature. For these tests each reading takes approximately 2 seconds, and the total plot takes place over 90 minutes. After an hour the container was opened causing liquid to condense on the open, integrated device. The liquid caused a damping of the SAW and a temporary loss of returned signal, as evident in the loss of temperature lock around reading 2000.
Figure 4.12: The simultaneous interrogation of three wireless SAW sensors, including two sensors integrated on-wafer with an antenna. As a comparison a integrated and traditional FR4 antenna were cooled in a container with liquid nitrogen. A third sensor was left at room temperature with a wired thermocouple.
CHAPTER 5
HIGH TEMPERATURE SAW FABRICATION

The focus of this chapter is on extending the operational temperature range of the OFC sensor platform beyond that demonstrated in previous research efforts [9, 14]. Previous device implementations utilized aluminum electrodes because of its favorable resistivity, low density, and extensive characterization. Commercially, extensive work has been performed on extending the lifetime of SAW filters under high power conditions utilizing combinations of aluminum, copper, and other metals [61]. High power SAW filters require large temperature dissipation capabilities and resilience to stress migration. However, these materials are still limited to temperature ranges below approximately 200°C. Typical wired sensors such as thermocouples and platinum RTDs withstand high temperatures, but have thicknesses orders of magnitude greater than that required for SAW operation. The following chapter will investigate fabrication techniques that extend the operational lifetime of SAW sensors at temperature ranges up to 1000°C.

5.1 High Temperature SAW Metallization Background

Inherently the langatate substrate maintains its piezoelectric properties up to approximately 1450°C. However, the SAW metallization is typically the limiting factor on device lifetime. Therefore a main focus of this dissertation has been on determining a thin film metallization technique that exhibits an extended performance at temperatures above 900 °C and also exhibits an acceptable SAW performance, low loss and high reflectivity. Most previous efforts have utilized platinum for an electrode material due to its chemical inertness and high melting temperature [62].
However, the stability of a thin platinum film on a dielectric substrate can be poor due to platinum agglomeration caused by stresses in the deposited films and poor adhesion quality [63].

The chemical inertness of platinum can be a slight hindrance due to its extremely poor adhesion to oxide substrates like LGT. This problem is common in SAW device fabrication and an interstitial metal layer, such as titanium, is typically used for adhesion of the electrode film to the substrate. The adhesion layer is typically a more reactive metal than platinum causing the interface of the films to alloy [63]. Previous studies have shown the limitation of using titanium above temperatures of 600 °C. Titanium easily diffuses through the platinum film and TiO2 is formed causing a destruction of the adhesion layer and platinum film delamination. An adhesion layer of zirconium has been shown to be adequate up to approximately 750 °C, but also tends to migrate into the platinum film causing failure [64].

Other research efforts have shown that there is a significant increase in device lifetime when switching to a higher melting point refractory metal such as tantalum because it will not diffuse into the platinum [65]. However, the platinum films themselves still intrinsically fail at temperatures above 900 °C. The primary source of electrode failure appears to be the agglomeration of the platinum in an effort to minimize the films surface energy [66]. Agglomeraion occurs through a two-step process in which first the platinum film begins to recrystallize causing breaks in the electrodes and then these holes grow through surface diffusion to form islands. These effects are expedited by stresses in the film, such as from the large surface to volume ratio of the thin film, and also from oxidized and diffused adhesion layers [67]. Since SAW electrode thickness is on the order of hundreds of nanometers, film failure typically occurs at a temperature well below the melting point of the bulk material.

Figure 5.1 demonstrates the degradation of a Zr/Pt thin film on the langatate substrate after heating the device to 800 °C for 2 hours and allowing it to cool back to room temperature. Electrode breaks begin to appear along the SAW transducer causing an increase in insertion loss and eventual loss of signal.
Figure 5.1: Optical image of a SAW transducer with Zr adhesion layer and Pt electrodes on langatate substrate (a) before and (b) after heating in ambient air environment to 800 °C for 2 hours.

There has been promising results on the utilization of a non-conducting overlay film that protects the thin film from agglomeration [68]. An overlay layer mechanically fixes the platinum, reducing film stress by raising the binding energy at the platinum surface. These layers also passivate the thin film and slow adhesion layer oxidation. There are also reports of using an oxide coating between the SAW substrate and electrode metallization to protect against oxygen and gallium loss at the surface of langasite [69]. Highly oriented films grown epitaxially could also extend the lifetime of metal films because of lower intrinsic stress after deposition [70].

Promising results of high temperature SAW electrode metallization have been demonstrated in [64], with a lifetime exceeding 16 hours at 1000 °C and 5.5 months at 800 °C, on an LGS substrate. These devices utilized a film of Pt/10%Rh alloy simultaneously deposited with ZrO2 and passivated with SiAlON. The goal of this research effort will be to characterize other electrode metallizations to find a comparable lifetime with a simpler morphology.
5.2 In-Situ Resistance Test Fixture

In an effort to characterize the thin film versus thickness, morphology, and other factors, the resistance of various thin films have been monitored over temperature. A simple test fixture was fabricated for in-situ monitoring of resistance changes at temperatures up to 1000 °C. A two part test fixture was designed in the 3D modeling software SolidWorks for the in-situ resistance measurements. The fixture consists of a base, shown in figure 5.2, where the sample is inverted to make contact with platinum leads. A ceramic feedthrough is used to bring two platinum wires from outside the furnace to the lip where the test resistor structure will rest. This feedthrough is a separate component that was epoxied to the base after fabrication of the test fixture. The top of the test fixture compresses the resistor die onto the platinum wires under its own weight.

![Figure 5.2: 3D layout of the test fixture base design in SolidWorks. Dimensions shown are in mm. The circular cutout will accept a standard high temperature thermocouple feedthrough.](image)

A mold was created of the modeled part using a 3D printer. Significant draft, approximately 10°, was added to the parts to ensure the final cast could be extracted from the mold. Rescor 780 castable Al₂O₃ from Cotronics Corporation was used to create the test fixture. The ceramic powder and activator was mixed and then shaken into the 3D printed molds. The cast cures for 24 hours at
room temperature. The test fixture was extracted from the mold and residual moisture was baked out at 110°C for two hours. The final casted alumina pieces can be machined for modifications such as fastener holes, but it was found that the weight of the test fixture made sufficient contact pressure on resistor test structures.

Figure 5.3 shows the final opened test fixture after the addition of the platinum test leads. The highlighted area shows the location where the device is measured during testing. Final device size is a maximum of 15 mm × 5 mm. Platinum wire of 1 mil thickness is used for a contact to the sample. The platinum leads are brought from outside the furnace through an Omegatite 450 alumina insulator purchased from Omega Engineering Inc. An alumina adhesive, Cotronics Resbond 989, was used to fix the feedthrough to the test fixture.

![Figure 5.3: The completed test fixture showing the separated top and bottom halves. The test device is inverted, making contact with two 1 mil Pt test leads.](image)

5.3 Thin Film Resistor Measurements

Devices were fabricated on x-cut langatate for use in the ceramic test fixture. X-cut material was utilized because of an abundance of small square substrates that are difficult to use for sub-micron fabrication. These devices have a 3 mm pad on either end for test contacts. A set of 4 shorted reflector gratings were connected in series as resistor test patterns. The gratings had 61
electrodes of 4 μm width, 250 μm length, and metallization ratio of 50%. The test pattern is shown schematically in Figure 5.4. The goal is to approximate the linewidth requirements of the final SAW device and investigate the stability of the final application. Resistance is a function of thin film sheet resistance, dielectric overlay losses, and LGT conductivity.

![Figure 5.4: Example test structure with 4 resistor cells in series as shown in the image cut out.](image)

Resistors were fabricated with thin adhesion layers of Ti, Zr, or Ta. Results from Ti adhesion were abandoned early as device lifetime was greatly reduced. Difficulty in evaporating tantalum caused poor lift off profiles and was therefore not investigated further. Platinum and palladium were investigated for electrode materials at various thicknesses. Alumina overlays were used to protect the resistor. All depositions were performed using an electron beam evaporator system, with adhesion and electrode deposited sequentially and alumina deposited separately. A lift-off process was used to define all patterns.

Figure 5.5 shows an example resistor pattern compromised of a 5 nm Zr adhesion, 95 nm Pt electrode, and a 125 nm alumina overlay. Initial resistance was 70 Ω. The device was heated for several hours to a final temperature of 950 °C and allowed to soak for 30 minutes. As shown, no degradation is apparent in the resistor pattern that was protected by the alumina overlay. However, areas left unprotected for electrical connection were completely destroyed.
Figure 5.5: (a) Example resistor pattern fabricated on X-cut LGT after heating to 950 °C for 30 minutes. The device uses a 5 nm Zr adhesion, 95 nm Pt electrode, and 125 nm Alumina overlay. (b) The Pt pad used for connection to the resistor. The lower half of the image shows the deterioration of the area of the resistor that was left unprotected and the upper area shows the resilience of the area covered with alumina.

After initial testing it was determined that stability would increase by fixing the test leads to the casted fixture. Electro-Science Laboratories 9595-A platinum-silver epoxy was used to adhere the test leads and completely coat the ledge of the resistor test fixture. The epoxy was then cured to ensure a uniform film for contact to the resistor. The bare contact portion of each resistor die was also coated with a thick layer of the same conductive adhesive and cured at 350 °C prior to being placed in the test fixture. The epoxy guaranteed the lifetime of the test point extends beyond that of the thin film.

An in-situ temperature measurement of a similar device with epoxy connections is shown in Figure 5.6. This film consists of an approximately 100 nm metal film with a 125 nm dielectric overlay. The implemented preventative measures allowed the device to be able to ramp up to 950 °C and return to room temperature. The dashed line plots the expected resistance change based on the temperature coefficient of resistance for bulk platinum (3.729 mΩ/°C).
Figure 5.6: Resistance measurement of a 95 nm thick platinum resistor structure with a 5 nm zirconium adhesion layer and 125 nm alumina protective coating. The theoretical dashed line plots the expected resistance based on an ideal bulk platinum temperature coefficient of resistance.

A similar Zr/Pt resistor was cycled 3 times to a temperature of 900 °C and allowed to cool back to room temperature. Figure 5.7a shows the combined results of this resistor over the multi-step annealing process. Figure 5.7b shows the degradation of the thin film through SEM measurement.
Figure 5.7: (a) In-situ resistance measurement of a Zr/Pt/Al₂O₃ test structure being cycled to 900 °C and allowed to cool back to room temperature. (b) SEM image of the electrodes after the final temperature cycle.

In an effort to understand the role of the langatate substrate on the in-situ measurements the resistor test pattern was modified to an interdigitated pattern. A 7.9 mm wide transducer with 30 88.4 µm wide electrode pairs was placed across the test pads. Ideally these films would have
megohm resistance, depending on the resistance of the substrate combined with the alumina overlay. However, as shown in Figure 5.8 the resistance of the substrate slowly decreased as the film is heated above 500 °C.

Figure 5.8: Measurement of LGT resistance over temperature using an interdigitated Pt pattern. The film starts off highly resistive, but at 950 °C has a resistance in the kΩ range.

Palladium and palladium/platinum combination films were also investigated. The hope was that Pd would have lower high frequency losses compared to Pt because of the lower density of the films, 12.02 versus 21.45 g cm\(^{-3}\). However, all films using Pd failed above 700 °C, with an example measurement of a Zr/Pd/Al2O3 film shown in Figure 5.9. The failure mechanism of Pd at high temperatures was not investigated further and work focused on high frequency losses of Pt transducers and reflectors on LGT.
Figure 5.9: Resistance of a 65 nm Zr/Pd film with a 120 nm alumina overlay measured in-situ over temperature. The film failed to stabilize with the furnace at 700 °C.

Based on high frequency SAW measurements, outlined in the following section of this dissertation, the necessity for a thinner Pt metallization became apparent. However, the concern was that film resilience would decrease as the thickness decreased. A resistor pattern was therefore also fabricated with Zr/Pt at 63 nm and a 130 nm thick alumina overlay. Results of temperature cycling this film are shown in Figure 5.10a. Results show the thinner film began to degrade more rapidly above 900 °C. However, inspection by SEM, as shown in Figure 5.10b does not show substantial film degradation. Probing through the alumina layer shows similar resistance to original room temperature measurements.
Figure 5.10: (a) Resistance measurement of a 63 nm Zr/Pt film with a 130 nm alumina overlay. (b) An SEM image of the electrodes after temperature cycling.

Upon inspection the Pt/Ag paste appears to be the point of failure. The SEM image in Figure 5.11 shows a discontinuity in the film at the interface of the adhesive and thin film. This adhesive failure restricted the length of these in-situ resistance measurements. However, preliminary results indicate that alumina prevents the deterioration of thin platinum electrodes on
LGT for several hours at 900°C, and additional research is necessary to find a reasonable interface to the thin film.

Figure 5.11: SEM image of the interface a thin film Zr/Pt resistor and Pt/Ag paste used for electrical contact after heating to degC900.
CHAPTER 6
LANGATATE SUBSTRATE SAW CHARACTERIZATION

Langatate SAW development has been limited previously to frequencies around several hundred MHz, making a major focus of this dissertation the characterization of LGT at frequencies up to 915 MHz. Going to high frequencies allows the antenna size to decrease, but there are several inherent challenges with increasing the frequency when compared to previous efforts on lithium niobate. Since the velocity is slower the wavelength will decrease for a similar center frequency. Therefore the line width will decrease and make it difficult to fabricate devices, especially those with fundamental operation. Also metallization thickness will be more restricted as device performance is directly related to the height of the metal relative to the wavelength, \( h/\lambda \), of the SAW. These device restrictions also become more pronounced when the metal is switched from aluminum to platinum. The following chapter will investigate SAW operation on LGT with respect to the high temperature fabrication outlined in Chapter 5.

6.1 Delay Line Test Structure

The basic test structure utilized to characterize SAW operation on langatate is demonstrated in Figure 6.1. The device consists of a 2-port delay line with an inline reflector structure. As shown, the devices are probed by an RF probe station and full 2-port S-parameters are collected by a vector network analyzer. Phase velocity and insertion loss are extracted from the direct S21 response between the two transducers. Per strip reflectivity is extracted from the reflected grating S21 response by fitting to coupling-of-modes predictions. Coupling and electrode capacitance are
calculated from the S11/S22 response of the transducer. By building an identical device at a longer delay the propagation loss in dB/µs can also be determined. All parameters can be compared versus harmonic of operation, metallization type/thickness, and with/without protective overlays.

Figure 6.1: An example of the SAW test structure utilized for langatate characterization being probed by an RF probe station. The device consists of a 2-port SAW delay line with input/output transducer on the left and inline grating structure on the right.

Typical probed transmission S21 results from the structure are shown in Figure 6.2. The ripple in S21 passband represents the reflected grating response. The frequency domain response is converted to the time domain by performing an inverse FFT. An example time domain response is shown with the direct SAW response and grating response highlighted.
The initial goal was to simply extract enough device parameters for OFC sensor layout at 915 MHz. An initial test mask was made with $4 \cdot f_0$ sampled transducers having a 7.2 $\mu$m wavelength. This transducer periodicity allows for fundamental and third harmonic operation at 305 MHz and 915 MHz, respectively [23]. Devices were fabricated on Y-X LGT because of wafer availability. Aluminum and platinum metallizations were performed for comparison. Probed delay line responses were gated in the time domain and transformed back to frequency for comparison of insertion loss and bandwidth distortion.

An example device response is shown in Figure 6.3, operating at the fundamental frequency of 305 MHz and third harmonic of 915 MHz. Aluminum harmonic device performance is comparable at both frequencies, insertion loss decreases slightly. However, for a relatively thin Pt metallization thickness of 110 nm, but a relatively large relative $h/\lambda$ value of 4.5%, the SAW response on LGT is completely damped at 915 MHz. Devices were also fabricated with a thinner Pt metal thickness of 44 nm, matching the $h/\lambda$ from the fundamental operation at 305 MHz. Compar-
ing the results of the fundamental and harmonic device, for the same relative Pt thickness, shows that there is an 8 dB increase in insertion loss. Also, there is a 10 dB increase in insertion loss and a 10 MHz change in center frequency between devices with thin platinum compared to the thicker aluminum film.

Figure 6.3: (a) Frequency response of a SAW delay line at 305MHz on Y-cut LGT. (b) Frequency response of a SAW delay line at 915MHz on Y-cut LGT, third harmonic 7.2 µm device. Each plot corresponds to a different h/λ value for the metallization. Device wavelength is approximately 7.2 µm, giving a metallization thickness of 110 nm Al, 110 nm Pt, and 44 nm Pt, respectively, for the 3 plots shown.

These initial results demonstrated that operation at higher frequencies with a platinum film would require extensive device analysis. Final device implementation requires a trade-off in insertion loss, center frequency, and bandwidth. Tests masks and devices were fabricated with transducer sampling of 2, 3 and 4 · f₀, with various periodicity, and harmonic operation to fully characterize these parameters. The following sections will examine the transduction, reflectivity, and propagation loss of surface acoustic waves on Y-cut langatate.
6.2 SAW Transduction on Y-Cut LGT

In an effort to quantify the ability of the SAW transducer to couple energy into a surface acoustic wave, SAW coupling coefficient and static capacitance were extracted from various transducer configurations. The admittance of a SAW transducer was outlined in (4.3) as the parallel combination of three terms, a real and imaginary acoustic contribution and a capacitive electrical term. An example probed transducer admittance is outlined in Figure 6.4. The following section will outline the extraction of $k^2$ and $C_s$ from these probed measurements, and provide extracted results over a range of frequencies and device topologies.

![Figure 6.4: An example of the admittance of a SAW transducer on langatate used for the extraction of coupling coefficient and substrate capacitance.](image)

Frequency dependent admittance is a function of the electrostatic charge density, but at center frequency the extracted capacitance and coupling can easily be found by scaling values based on geometry. Therefore, transducer center frequency is found first by time gating the direct $S_{21}$ response, converting back to frequency, and minimizing the integral squared error of the experimental data and an ideal sampling function approximation throughout the passband. $S_{11}$
responses are then gated to minimize reflections and converted to impedance. Parasitic resistance is removed by minimizing the resistance at the null of the experimental transducer impedance. Parasitic capacitance is more difficult to determine and requires a similar probe pad structure on wafer.

The parasitic free SAW impedance is converted to admittance and the conductance is found at center frequency. The transducer capacitance is then extracted from the susceptance. Total device susceptibility is $B_a(\omega) + \omega C_T$, where $B_a$ is the Hilbert transform susceptance required for a causal system.

$$B_a(\omega) = G_a(\omega) \ast -1/(\pi \omega), \quad (6.1)$$

where $\ast$ indicates convolution. From Fourier transform properties, an approximation of $B_a$ is found from the extracted conductance as

$$B_a(t) = G_a(t) \cdot -j \text{sgn}(t) \quad (6.2)$$

and converting back to the frequency domain. Capacitance can then be found by again minimizing the integral squared error between the approximation of $B_a + \omega C_T$ and the actual extracted susceptance.

With center frequency capacitance and conductance found the substrate parameters can be extracted based on device geometry. Using the impulse response model of [23], substrate capacitance can be found from

$$C_s = \frac{C_T}{CN_p W_a}. \quad (6.3)$$

Coupling is extracted from the quality factor with

$$k^2 = \frac{Q_T N_p}{2Q}, \quad (6.4)$$
where
\[
Q_T = \frac{\omega_c C_T}{G_a(\omega_c)}.
\]  \hspace{1cm} (6.5)

Variations in fabricated metallization ratio are removed by estimating the scaling factors \( \bar{Q} \) and \( \bar{C} \), which are a function of the element factor for each transducer sampling and analytically given by [23]. Approximations to those analytical functions are shown in Figure 6.5, and by normalizing to this value the actual substrate coupling and capacitance can be determined irrespective of sampling. Metallization ratio is measured optically on each probed device. Variations due to harmonic, M, of operation are automatically removed by this method, as center frequency scales the conductance will also scale.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure6_5}
\caption{The relative scaling factors for transducer coupling and quality factor versus metallization ratio for 2, 3 and 4 \( \cdot f_0 \) sampled transducers. The relative coupling for a transducer is shown in (a) and the inverse Q is shown in (b).}
\end{figure}

Devices were initially fabricated with aluminum electrodes for substrate verification and comparison to published SAW characteristics in [24, 71]. In these previous publications, the extracted coupling coefficient was 0.423\% and the static capacitance was 2.08 \( \text{pF cm}^{-1} \) for Y-X SAW on LGT. Extracted results using 109 nm thick aluminum metallization are demonstrated in
Figure 5.3. Transducer wavelengths are 7.25 µm, 10.2 µm and 15.28 µm, with $4 \cdot f_0$ sampling for operation at the fundamental frequency and third harmonic. The average coupling and capacitance were 0.41% and 2.02 pF cm$^{-1}$, respectively.

![Figure 5.3](image)

Figure 6.6: (a) The extracted static capacitance of the langatate substrate at various center frequencies. (b) The extracted SAW coupling coefficient of the langatate substrate at various center frequencies. All devices have aluminum electrodes with a total thickness between 109 nm. Devices operate at both the fundamental and third harmonic as indicated.

Subsequent devices were fabricated with platinum electrodes for comparison of transducer performance, and results are shown in Figure 6.7. Various transducer wavelengths were utilized for operation from 100 MHz to 900 MHz using fundamental, second, and third harmonic transducers having 2, 3, and $4 \cdot f_0$ sampling, respectively. Typical metallization thickness was kept relatively thin between 40 nm and 70 nm. The average capacitance is 1.90 pF cm$^{-1}$ with a standard deviation of 0.16 pF cm$^{-1}$. The coupling coefficient at center frequencies below 450 MHz is consistent with theory and aluminum measurements with a mean of 0.43%. However, measurements made of devices operating at 650 MHz and 915 MHz slowly decrease in coupling strength, with 915 MHz devices having an average coupling of half at 0.23%. Varying the impulse response length had no correlation to the decrease in coupling strength.
Figure 6.7: (a) The extracted static capacitance of the langatate substrate at various center frequencies. (b) The extracted SAW coupling coefficient of the langatate substrate at various center frequencies. All devices have platinum electrodes with a total thickness between 40 nm and 70 nm. Harmonic operation of each transducer is indicated.

6.3 Electrode Reflectivity on Y-Cut LGT

As outlined in [72], reflectivity is of particular importance when designing OFC sensors because it determines operable bandwidth and insertion loss. Extracted reflectivity of LGT was published previously by [24], but only for shorted aluminum electrodes. Test devices were fabricated with open and short circuit gratings, several relative metal thicknesses, and a series of metallization ratios to better predict the reflectivity of LGT. Reflectivity is the combination of three effects, mass loading, electrical shorting, and stored energy. For a low coupling material, such as LGT, the mechanical reflections should dominate the reflectivity, and it is expected that the extracted reflectivity will be a maximum at 50% metallization ratio [73]. However, the reflectivity must be determined experimentally for a given metallization.

Reflectivity extraction follows the method outlined by [13], which uses the S21 response of the test structure. The direct response and grating response are separately gated in the time
domain and converted back to frequency. The total frequency domain reflection coefficient of the grating is extracted by normalizing to the direct transducer response. Per strip reflectivity is found by minimizing the error between the experimental reflection coefficient and the COM model approximation outlined in [13]. By comparing responses in the time domain the amplitude and grating loss can be determined from the initial amplitude and roll-off. An example frequency and time domain response are shown in Figure 6.8.

![Figure 6.8](image)

Figure 6.8: A comparison of the measured grating response to that of a COM prediction in (a) the frequency domain and (b) the time domain.

Initially a wafer was fabricated with 0.7% $h/\lambda$ Pt metallization, or 105 nm, using the test structure of Figure 6.1, but with metallization ratio stepped from 20, 35, 50, 65 and 80% for both open and shorted electrodes. The test structure had a wavelength of 15.28 µm, with 50 $\lambda/4$ strips. These initial test devices quantify the reflectivity trend at fundamental operation and results are presented in Figure 6.9. Additional measurements from several other test structures are shown at various relative thicknesses.
Figure 6.9: The extracted strip reflectivity versus measured metallization ratio for (a) short circuit grating and (b) open circuit grating.

The velocity shift of the grating relative to open circuit is an important parameter for designing the center frequency of OFC gratings. By extracting the center frequency of the grating, the velocity shift of the grating can be determined from the layout wavelength. Figure 6.10 plots the extracted shift in grating velocity compared to the free surface velocity for both open and short circuit gratings. The velocity decreases linearly as the SAW is shorted by grating electrode, and both types of gratings follow a similar trend.
Figure 6.10: The extracted grating velocity versus measured metallization ratio for (a) short circuit grating and (b) open circuit grating.

### 6.4 Propagation Loss of Alumina on LGT

The change in propagation loss of langatate before and after deposition of a passivation layer of alumina was extracted over a range of frequencies. Similarly to devices in Section 5.1, alumina was deposited by electron beam and patterned over the entire device interaction area using a lift off process. Devices were subsequently annealed in an oxygen environment at 500 °C for 2 hours. Figure 6.11 compares the change in insertion loss of an example device before and after deposition of the alumina film in dB/µs⁻¹.
6.4.1 Alumina Reflector Gratings

The loss associated with an electron beam deposited alumina overlay is quite high and also unpredictable over frequency. The loss could be a function of stoichiometry issues related to the ebeam and using a sputter or atomic layer deposition could lower these losses. But covering the entire device in aluminum oxide could still be unnecessary. Since the reflection mechanism of langatate is primarily a mechanical edge effect a single deposition could be performed passivating the transducer and patterning an alumina grating. Since the density of alumina (3.95 g cm\(^{-3}\)) is close to that of aluminum (2.7 g cm\(^{-3}\)), results are comparable because of the lower acoustic mismatch to langatate (6.15 g cm\(^{-3}\)) when compared to platinum (21.4 g cm\(^{-3}\)).

Gratings of alumina have been fabricated on Y-cut LGT for comparison to those of platinum. At an \(h/\lambda\) of 1.2% extracted reflectivity is 0.11%. An example alumina grating response operating at the fundamental frequency is shown in Figure 6.12. Each grating consists of 100 strips with a center frequency wavelength of 7.25 μm. Also shown is a similar platinum grating that is completely covered with an alumina passivation. These devices are not intended for a direct comparison, but
rather demonstrate the ability of the alumina grating. Although the reflectivity for alumina is lower, longer grating lengths will have less distortion in the waveform and matched filter synthesis should be simplified.

Figure 6.12: The comparison of an alumina grating and a platinum grating with an alumina overlay operating at 305 MHz.
CHAPTER 7
INTEGRATED WIRELESS SAW SENSORS ON LANGATATE

Based on the results shown in Chapter 6 it was determined a tradeoff was necessary between center frequency, metallization thickness, and fabrication limitations. For a proof of concept integrated design, 650 MHz was chosen as the device center frequency. This center frequency allows for fundamental device fabrication with 0.85 µm electrode width using a 2 · f₀ sampled transducer. This center frequency should keep the platinum metallization at a reasonable thickness when considering resistive losses and high temperature lifetime. Antenna bandwidth and loss will be sacrificed slightly for antenna integration onto a 3 inch wafer.

7.1 650 MHz Orthogonal Frequency Coded Sensor Design

A two track OFC design approach was used so that strongly coupled narrowband transducers could be used and still maintain a wide band OFC code. The basic device layout is shown in Figure 7.1. In the design the upper track contains the chip frequencies f₁ and f₂ and the lower track f₃ and f₄. The transducers are designed with a center frequency between the two orthogonal chips in the respective acoustic track. This design did not compensate for transducer beam width to ensure equal conductance in each track. The chip length is 138.5 ns, and the total sensor bandwidth is 29 MHz. The remaining design parameters are given in Table 7.1.
Figure 7.1: Example layout of an OFC device on LGT. The device has two tracks and 4 orthogonal chip frequencies.

Table 7.1: Design parameters for the OFC device on the LGT substrate.

<table>
<thead>
<tr>
<th>Track #</th>
<th>Transducer</th>
<th>Grating</th>
</tr>
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<tbody>
<tr>
<td></td>
<td>$\lambda_0$</td>
<td>$N_p$</td>
</tr>
<tr>
<td></td>
<td>(\mu m)</td>
<td>(2·$f_0$ Pairs)</td>
</tr>
<tr>
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<td>30</td>
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Results of OFC devices fabricated on Y-cut LGT are presented in Figure 7.2. The fabrication was performed with a 65 nm thick Zr/Pt film. Since adjacent chip frequencies are in the same track there is a significant roll off between the first and second response in each track. This roll off is due to each chip not being completely orthogonal and having significant energy overlap in frequency, which is exacerbated by the high reflectivity of the platinum grating. As described in Chapter 2, there is a tradeoff between insertion loss and waveform distortion for an OFC grating.
Figure 7.2: Initial OFC sensor response on LGT. Frequency domain return loss shown in (a) and time domain shown in (b).

A second design was performed using withdrawal weighted reflector gratings that removed every other electrode. These reflectors have a lower net reflectivity, but maintain their orthogonality as the grating impulse becomes longer. The layout parameters for a withdrawl weighted OFC device
are shown in Table 7.2. Although the number of strips in each grating is cut in half, the individual chip length, 138.5 ns, is maintained compared to the previous design. These updated devices also used wider bandwidth input transducers at 25 pair, and compensated the beamwidth for each track.

Table 7.2: Design parameters for an OFC device on the LGT substrate using withdrawal weighted gratings.

<table>
<thead>
<tr>
<th>Track #</th>
<th>Transducer</th>
<th>Grating</th>
</tr>
</thead>
<tbody>
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<td>( \lambda_0 ) (µm)</td>
<td>( N_p ) (2( \cdot ) f(_0) Pairs)</td>
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An example probed withdrawal weighted device response is shown in Figure 7.3. As shown the weighting of each grating removed the strong roll-off between the first and second grating in each track. Preliminary fabrication did not use an alumina overlay, and device layout was conducted without consideration of the velocity shift associated with an alumina overlay.
Figure 7.3: Withdrawal weighted OFC sensor response on LGT. Frequency domain return loss shown in (a) and time domain shown in (b).
7.2 650 MHz Meander Dipole Design

A meander dipole antenna was designed for operation on LGT similarly to the design completed for LiNbO$_3$ in Chapter 4. At 650 MHz the limit for an electrically small antenna is a sphere with a radius of 73.5 mm, which is again less than the size of a 3 inch wafer. Therefore in order to fit an antenna on a 3 inch substrate there must be a tradeoff between bandwidth and efficiency as described in Chapter 2. Assuming a final antenna metallization of platinum, there will also be larger conductor losses compared to previous iterations because of a resistivity increase to 105 nΩ m.

Y-cut LGT has a larger disparity in the permittivity between crystal axis, but with the same uniaxial symmetry as LiNbO$_3$. From [21] the relative permittivity is 18.27 for $\epsilon_{11}$ and 78.95 for $\epsilon_{33}$. For Y-cut X propagating LGT the rotation places the larger permittivity value perpendicular to the direction of propagation, and parallel to the dipole current vector, considering previous antenna layouts demonstrated on LiNbO$_3$. The high permittivity in this direction greatly increases the coupling strength between parallel meander sections. HFSS was once again used to simulate the antenna and factors in the anisotropic permittivity.

The design of the antenna has been modified slightly to more easily match to the capacitive SAW input impedance. A shunt T-match connection was utilized to connect the antenna impedance to the SAW sensor. The basic antenna structure is outlined in Figure 7.4, and consists of an input transmission line of radius $a'$ and length $L'$ shorted to a dipole of radius $a$ and length $L$. Effectively the shunt input network flips the input impedance of the antenna 180°, making it inductive for an electrically short antenna.
Figure 7.4: Antenna layout using a T-match connection, where a standard dipole of length $L$ is fed by a shunt transmission line of length $L'$.  

The theoretical calculation of input impedance for a T-match antenna is outlined in [52], which assumes the input impedance of the antenna is a combination of a transmission line and radiating antenna modes. The equivalent circuit is shown schematically in Figure 7.5. The total current input onto the feed is divided between the transmission line and radiative conductors based on the separation, $s$, and relative conductor sizes, $a$ and $a'$. A current division factor can be defined as 

$$\alpha \approx \frac{\ln\left( \frac{s}{a'} \right)}{\ln\left( \frac{s}{a} \right) - \ln\left( \frac{a}{a'} \right)}.$$  

(7.1) 

Total input impedance can then be expressed as 

$$Z_{in} = \frac{2Z_t \left[ (1 + \alpha)^2 Z_a \right]}{2Z_t + \left[ (1 + \alpha)^2 Z_a \right]}.$$  

(7.2)
$Z_t$ is the impedance of the short circuit 2-wire transmission line

$$Z_t = jZ_0 \tan \left( k \frac{L'}{2} \right),$$

(7.3)

with characteristic impedance $Z_0$ that depends on the conductor sizes and separation. For the simple case in which the conductor sizes are equal and $Z_t >> (1 + \alpha)^2 Z_a$

$$Z_{in} \approx 4Z_a,$$

(7.4)

The T match feed is useful for electrically short dipole antennas with small radiation resistances because the feed can step up the impedance of the antenna to more closely match that of the SAW sensor. Controlling the ratio of $a/a'$ controls the ratio of the input impedance to the radiation resistance.

![Figure 7.5: Equivalent circuit of the T-match fed dipole antenna where $Z_\alpha$ is the antenna impedance, $Z_t$ is the shunt transmission line impedance, and $\alpha$ is the current division factor between the two modes.](image)

A meander dipole antenna was designed utilizing a T-match feed on the LGT substrate. For simplicity the antenna conductor diameter was kept constant at 1.5 mm. Final antenna length was maintained at 65 mm. Figure 7.6 outlines the remaining antenna layout parameters.
Figure 7.6: Final simulated antenna layout with dimensions shown in mm.

A comparison of the designed antenna impedance and the probed SAW impedance is shown on a Smith chart in Figure 7.7 for both gold and platinum metallizations of 10 µm thickness. Simulated antenna radiation efficiency is 60% with platinum or 75% with gold metallization. Simulated surface current is shown in Figure 7.8 and shows the concentration of the current in the radiating element.

Figure 7.7: A comparison of the input reflection coefficient of the probed SAW sensor and simulated on-wafer antenna. The 30 MHz OFC bandwidth is highlighted, with ♦ representing center frequency.
7.3 Integrated LGT Results

An example antenna was fabricated on Y-Cut LGT using electroplated gold at 5.3 μm thickness, shown in Figure 7.9. Gold was utilized to ensure the antenna and device performed as expected and for initial temperature experiments below 600 °C. Preliminary fabrication did not use an alumina overlay because of high electron beam deposited alumina propagation loss. SAW metallization was Zr/Pt at 650 nm nominal thickness. The long interconnects between the SAW probe pads and antenna feed were created by the Ti/Au antenna seed layer at 240 nm.

The integrated device was interrogated at a distance of 15 cm using the vector network
analyzer and two whip digital television antennas. Power output on the VNA was set to 10 dBm and no averaging was performed. Figure 7.10 compares the time domain response of the probed signal and that extracted from the wireless measurement. A matched filter was also calibrated to the sensor using gated sine functions, and a comparison of the relative correlation of each device is shown.

![Figure 7.10](image)

Figure 7.10: (a) Comparison of the probed results to the wireless measurement of the integrated at 15 cm. (b) Matched filter correlation comparison of the probed and wireless data. The same matched filter is used for both signals, showing the relative time shift of EM and cable delays.

The experimental TCF for a SAW on Y-cut LGT was published in [24] as 64.5 ppm/°C. However, this value was for room temperature only, and it is anticipated that the coefficient will vary greatly with temperature. The expected temperature coefficient can be theoretically calculated by the well known matrix method developed by Adler in [74]. This system matrix approach was implemented in Matlab for the calculation of velocity, coupling, and temperature coefficient of the SAW mode. A complete set of temperature dependent material coefficients for langatate were provided in [75], with the thermal coefficients of expansion outlined in [76]. Recently an updated set of temperature coefficient of stiffness, permittivity, and piezoelectricity were demonstrated over
two temperature ranges by [22], room temperature to 500 °C and 500 °C to 900 °C. The theoretical temperature coefficient for Y-cut LGT using both sets of constants is plotted in Figure 7.11 from 0 °C to 400 °C. The integrated device was then placed into a furnace and interrogated through a ceramic door over the same temperature range. The algorithm demonstrated in Section 3.1 was then used to extract the TCF from the extracted frequency scaling factor. TCF is related to frequency scaling factor by

\[
TCF = \frac{\alpha - 1}{T_m - T_c},
\]

(7.5)

where \(T_m\) is the temperature measured by a thermocouple and \(T_c\) is the temperature of the sensor during calibration. Two separate runs were performed, initially below 200 °C and subsequently over the full range.

![Figure 7.11: A comparison of the measurement of temperature coefficient of frequency versus calculations using published material constants.](image)

After heating to 400 °C for 30 minutes the antenna was allowed to cool to room temperature. Upon inspection under microscope the area where the gold thick film/thin film connection had
seriously degraded. The image in Figure 7.12 shows a break forming at the connection between the antenna and interconnect. A noticeable decrease in signal level had occurred in the temperature extraction software at approximately 250°C.

Figure 7.12: Optical image of the interface of the Au antenna/probe pad interface after heating to 400°C.
CHAPTER 8
DISCUSSION AND CONCLUSION

The goal of this dissertation was to investigate the integration of an antenna and temperature sensor directly onto a surface acoustic wave (SAW) substrate. Integrated sensors remove die adhesives, bond wires, and traditional packaging, opening a range of sensor applications that leverage the unique advantages of the SAW substrate. As a proof of concept, sensors were designed and fabricated on the lithium niobate and langatate substrate using orthogonal frequency coding (OFC) for use in a multi-sensor system. The OFC concept has been outlined extensively in previous publications as a passive spread spectrum coding technique that uses both time and frequency diversity for improved inter-sensor interference. However, this dissertation built upon that foundational work by developing improvements to the coding, interrogation, and temperature range of the sensors.

Initial work presented in this dissertation highlighted several novel improvements to the wireless OFC sensor system, before focusing on the use of integrated devices. It was shown that for optimal signal to noise ratio each sensor should by discreetly time multiplexed, requiring long time windows for an increased code set. A coding technique was investigated that decreased overall sensor time response length by allowing arbitrary chip overlap through the use of multiple acoustic tracks. This mixed orthogonal frequency coding allows for another level of code diversity and increases the number of sensors that fit within a given interrogation repetition rate. Based on these principles a set of eight uniquely coded sensors were designed and fabricated to fit into a \(5\,\mu s\) window with a \(68\,\text{MHz}\) bandwidth. For interrogating these and other sensors in this dissertation a coherence correlator system was presented that increases speed and accuracy by processing data.
in the frequency domain. The system does not rely on resolution in the time domain, but rather integrates the signal energy over the bandwidth of interest. The system code was written in both Matlab and Python, and demonstrated millisecond processing time of a single sensor. Interrogation times are now primarily limited by the speed of the interface to the radio hardware. These implemented changes enabled the first set of eight simultaneously interrogated sensors measured in a 915 MHz system over a temperature range of ±130 °C.

Initial on-wafer antenna designs were performed on lithium niobate because of the strong foundation of research for designing low temperature OFC sensors using the substrate. A meander dipole antenna was designed for use on standard 3 inch Y-cut LiNbO$_3$ with a center frequency of 915 MHz. Matching of the antenna to the SAW sensor was performed without the use of external components and a mismatch loss of less than 2 dB was achieved over a 7% fractional bandwidth. For comparison, antennas were fabricated using thermal evaporation, electroplating, and a direct write system. The direct write technique demonstrated the ability to deposit between 10 to 20µm thick copper onto lithium niobate and make connection to the sub 100 nm aluminum SAW metallization. Measured integrated sensor performance was comparable to devices traditionally packaged and soldered onto an FR4 dipole antenna. Wireless results were demonstrated in a three sensor system over temperature. These devices have application in gas, strain, and high temperature sensors where traditional packaging is impractical.

In an effort to expand the operational temperature range of the OFC sensor platform an investigation was conducted on the langatate substrate. This single crystal substrate maintains piezoelectric operation up to its melting temperature of 1450 °C. The material properties have been well characterized and initial SAW operation has been demonstrated in previous publications. However, little work has been published on devices operating at frequencies above several hundred MHz. To exploit the substrates increased high temperature lifetime a change in SAW metallization, from the aluminum used in previous experiments, would be necessary and require characterization versus chosen electrode material. Therefore langatate characterization performed in this dissertation
consisted of two inter-related tasks, an electrode metallization study and a high frequency SAW characterization. The goal of this second portion of the dissertation was to develop a stable high temperature sensor with an extended lifetime above 700 °C, with a push toward 1000 °C.

Platinum was chosen for an electrode metallization due to its chemical inertness and high melting temperature. However, sub-micron thick platinum agglomerates well below the expected bulk melting temperature and has a poor adhesion to dielectric substrates. Additionally, adhesion layers typically employed in SAW fabrication cause platinum delamination through oxidation. In-situ resistance measurements were therefore conducted of various film morphologies up to 950 °C to determine the lifetime of a given metallization versus morphology. A short term stability was found for a Zr/Pt film of thicknesses above 60 nm at temperatures below 800 °C. The use of an aluminum oxide passivation layer increased the lifetime up to 950 °C by restricting film oxidation and mechanically fixing the platinum. However, film lifetime was not determined definitively because of issues at the connection point of the resistor sample and the high temperature adhesive used for connection to the test fixture. Investigation by SEM showed film migration at the site of the thin electrode and thick connecting paste on devices heated for extended periods at 900 °C. Future work will be necessary to determine the failure mechanism of thick/thin film materials and a potential solution.

Based on the results of electrode fabrication limits, a study was conducted on SAW operation on the langatate substrate. SAW coupling, static capacitance, and electrode reflectivity were extracted at frequencies up to 1 GHz, versus metallization thickness. Initial results fabricated using aluminum match values reported in literature across all frequencies of operation. However, there is a strong decline in coupling strength at high frequencies when the metallization is switched to platinum. The decrease in coupling makes the use of a SAW on langatate impractical at 915 MHz when considering reasonable platinum thicknesses greater than 60 nm. The mechanism for the decrease in SAW coupling is currently unknown, and a change in sensor center frequency was required.
Finally, the design of OFC sensors on langatate using platinum electrodes was demonstrated. A decrease in sensor center frequency was required from previous work to 650 MHz, where the device loss and metallization thickness are balanced. The decrease in langatate coupling, as compared to lithium niobate, requires narrowband transducers and reflectors for reasonable insertion loss. Devices were demonstrated with chip amplitude compensation over a 30 MHz bandwidth with 4 orthogonal frequencies. A meander dipole antenna was also designed for use on 3 inch langatate substrates. Matching improvements were shown compared to devices on lithium niobate by using a T-match shunt antenna feed. Proof of concept integrated devices were fabricated with electroplated gold for the on-wafer antenna metallization. These devices were demonstrated at temperatures up to 400 °C. Similar to platinum at high temperatures, a degradation of the thick/thin gold interface was evident after temperature cycling. Future work will be necessary to determine if this problem occurs with a thick platinum antenna connected directly to the SAW metallization. Capacitive coupling through the alumina passivation or using an inductive coupling instead of T-match feed could alleviate this problem in subsequent integrated sensors.
LIST OF REFERENCES


