Saw Reflective Transducers And Antennas For Orthogonal Frequency Coded Saw Sensors

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SAW Reflective Transducers and Antennas for Orthogonal Frequency Coded SAW Sensors

by

Bianca Maria Cabalfin Santos
B.S. University of Central Florida, 2007

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in the School of Electrical Engineering and Computer Science in the College of Engineering and Computer Science at the University of Central Florida Orlando, Florida

Spring Term
2009

Major Professor: Donald C. Malocha
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ABSTRACT

Passive sensors that vary its impedance per measured parameter may be used with surface acoustic wave (SAW) reflective transducers (SRT) for wireless acquisition of the measurand. The device is composed of two transducers, where one, which may be attached to an antenna, is used to launch the wave within the device substrate, and the other is where the sensor load is attached to. The latter is able to reflect the incident wave. How much power is reflected is determined by the attached sensor load. Amplitude variations as well as peak frequency variations of the SRT reflectivity response are explored in this thesis.

SAW passive temperature sensors with an orthogonal frequency coded (OFC) time response were previously investigated and prove to be ideal for use in harsh environments. Each sensor is distinguishable from the other due to the OFC code embedded within its time response. However, this coding technique poses a difficulty in designing antennas for the sensor due to its inherently wide bandwidth, and capacitive, non-uniform input impedance. This work covers antenna design and testing for the 250MHz wireless temperature acquisition prototype with a 28% fractional bandwidth, and for the 912MHz system which has 10% fractional bandwidth. Apart from the tag, antennas for the transmitter and receiver were designed for 50Ω matching with the required bandwidth maintained. Wireless temperature acquisition runs for the 250MHz prototype were successfully performed and show good
agreement with measurements made by a thermocouple. Since a transceiver for the 912MHz system is not complete, the performance of the antennas was gauged by observing the signal transmitted wirelessly by the SAW tag and by comparing this with the sensor time response measured directly by a vector network analyzer.
To my mother and father
ACKNOWLEDGMENTS

I would like to thank my mother and father for all the support and for believing in me most especially in times when I lost faith in myself in finishing this endeavor. I am eternally grateful for all the advice, sacrifice, guidance and care throughout my 22 years in life.

Thank you, Dr. Donald C. Malocha, for being my advisor. I have grown tremendously as a person and as an engineer under your tutelage. Thank you for all the help and for the trust when you accepted me as your student even if you did not know me prior to hiring me. I would like to thank Dr. Parveen Wahid and Dr. Samuel Richie for serving as my committee member. In addition, I am grateful to Dr. Robert Youngquist and to NASA for the financial support.

To my colleagues at the Consortium for Applied Acoustoelectronic Technology - Mark and Daniel Gallagher, Nikolai Kozlovski, Matt Pavlina, Mike Roller, Brian Fisher, Nancy Saldanha, Yusuke Sato and Luis Rodriguez - I would like to thank you for selflessly willing to lend a hand whenever I needed it. You have made working at the CAAT a very fun and memorable experience. And to Rajesh Paryani, Yazid Yusuf and Ajay Subramanian, I would like thank you for the advice and for the anechoic chamber accommodations.
# TABLE OF CONTENTS

**LIST OF FIGURES** ................................................................. xi

**LIST OF TABLES** ................................................................. xx

**LIST OF ACRONYMS** ............................................................ xxi

## I SAW Reflective Transducers

### CHAPTER 1: INTRODUCTION ................................................... 2

### CHAPTER 2: P-MATRIX DERIVATION ......................................... 4

2.1 SRT Reflection Coefficient Derivation .................................... 4

2.2 Device Model ........................................................................ 6

### CHAPTER 3: AMPLITUDE VARIATIONS OF SRT FREQUENCY RESPONSE

3.1 Simple Model for SRT ......................................................... 9

3.1.1 Launching Transducer ..................................................... 10
3.1.2 SAW Reflective Transducer ........................................... 11
3.2 Device Modeling using P-Matrix Formula ............................... 14
  3.2.1 Open and Shorted Load ............................................... 15
  3.2.2 Resistive Load .......................................................... 16
  3.2.3 Inductive Load .......................................................... 17
  3.2.4 Capacitive Load ....................................................... 18
3.3 Experimental Result ........................................................ 19
  3.3.1 SRT Reflection Coefficient Extraction ............................. 20
  3.3.2 Open and Shorted Load ............................................... 21
  3.3.3 Inductive Load .......................................................... 22

CHAPTER 4: FREQUENCY PEAK DEVIATIONS IN THE SRT FREQUENCY RESPONSE ................................................................. 24
  4.1 Overview ........................................................................... 25
  4.2 Frequency Peak Deviations due to Susceptive Loads .............. 27
    4.2.1 Capacitive Load ........................................................ 28
    4.2.2 Inductive Load .......................................................... 30

II Antennas for Orthogonal Frequency Coded SAW Sensors 35
CHAPTER 5: INTRODUCTION .................................................. 36

CHAPTER 6: ANTENNAS FOR THE 250MHZ WIRELESS TEMPERATURE ACQUISITION SYSTEM PROTOTYPE ................................. 38

6.1 250MHz Transceiver Antennas ........................................... 38

   6.1.1 Lumped Element Wideband Matching to Match the Antenna to 50Ω 41

6.2 Antennas for the 250MHz UWB Orthogonal Frequency Coded SAW Temperature Sensor ........................................................... 44

   6.2.1 250MHz OFC SAW Temperature Sensor Device Overview ............................................ 45

   6.2.2 Lumped Element Wideband Matching to Match Device 1301 to the Antenna ............................................ 46

6.3 Antenna Performance .......................................................... 51

   6.3.1 Antenna Radiation Pattern and Gain .................................. 51

   6.3.2 Antenna Performance as Part of the 250MHz OFC Sensor Temperature Interrogation System .......................... 54

CHAPTER 7: ANTENNAS FOR THE 912MHZ TEMPERATURE ACQUISITION PROTOTYPE ............................................................... 58

7.1 912MHz Antenna Design ................................................... 58

   7.1.1 Open-Sleeve Dipole Antenna ........................................... 59

   7.1.2 Planar Open-Sleeve Dipole Antenna ................................... 60
7.2 Matching Network for the 912MHz OFC Time Division Multiplexed Temperature Sensor

7.2.1 912MHz OFC SAW Temperature Sensor Device Overview

7.2.2 Matching Network for the 912MHz OFC SAW Temperature Sensor

7.3 Antenna Gain

7.4 Antenna Radiation Pattern

7.5 Antenna Performance

7.6 Antenna Range Prediction

CHAPTER 8: CONCLUSION

LIST OF REFERENCES
LIST OF FIGURES

Figure 1.1 SAW reflective transducer (SRT), connected to a load sensor is used in conjunction with a SAW transducer. .............................................. 2

Figure 2.1 Block Diagram of SAW transducer. where a, b, V, I, refer to the forward, reverse waves, voltage and current applied to the SAW device respectively. ............. 4

Figure 2.2 Signal flow diagram for the device shown in Fig. 2.1 where each path is labeled with its respective gain. ................................................................. 7

Figure 3.1 Launching transducer plots ................................................................. 10

Figure 3.2 Reflection coefficient ($P_{11}$) at center frequency of the SRT for resistive, inductive or capacitive loads ................................................................. 11

Figure 3.3 Reflection coefficient of SRT with changes in resistive load. ............... 12

Figure 3.4 Reflection coefficient of SRT with changes in inductive load. ............. 13

Figure 3.5 Reflection coefficient of SRT with changes in capacitive load. ........... 14

Figure 3.6 Device input admittance plots ............................................................ 15

Figure 3.7 Device reflection coefficient for shorted and open load ................. 16
Figure 3.8  Device reflection coefficient for 50Ω, 500Ω and 5kΩ loads.  . . . . . . . 17
Figure 3.9  Device reflection coefficient for 0.637µH, 0.253µH and 63.7nH loads.  . 18
Figure 3.10 Device reflection coefficient for 1pF, 5pF, 10pF and 15pF loads.  . . . 18
Figure 3.11 Reflection coefficient of the DUT with an open (thick) and short (thin) load
attached  . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 21
Figure 3.12 Measured reflection coefficient at center frequency (140MHz) of the SRT for
different values of inductive load  . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 22
Figure 3.13 Measured reflection coefficient of the SRT (within the device whose char-
acteristics are found in Table 3.1) at center frequency (140MHz) for different values of
inductive load, and for open and short loads, in polar format  . . . . . . . . . . . . . 23
Figure 4.1  Input admittance of resonant SRT with a length of 25λ  . . . . . . . 24
Figure 4.2  Resonant SRT reflection coefficients for resistive, capacitive and inductive
loads in polar form taken at 253.2MHz, at the resonant frequency of 252.93MHz.  . 26
Figure 4.3  SRT susceptance and reflection coefficient simulations upon attaching dif-
f erent capacitive load values (in pF) to the SRT of length 25λ and W₀= 50λ.  . . . 28
Figure 4.4  Resonant frequency change versus capacitive load change for N₀ = 25, 35
and 45. The second plot is a magnified version of the first plot.  . . . . . . . . . . . 29
Figure 4.5  SRT susceptance and reflection coefficient simulations upon attaching dif-
f erent inductive load values (in nH) to the SRT of length 25λ and W₀= 50λ.  . . . 31
Figure 4.6  SRT susceptance and reflection coefficient simulations upon attaching different inductive load values (in nH) to the SRT of length $45\lambda$ and $W_a=50\lambda$. .......................... 32

Figure 4.7  Resonant frequency change versus inductive load change for $N_P = 25$, 35 and 35. The second plot is a magnified version of the first plot. .............................. 33

Figure 5.1  Diagram of the OFC tag system used for conducting wireless temperature measurements. ................................................................. 36

Figure 6.1  The antenna constructed for transmitter, receiver and the SAW sensor tag. 39

Figure 6.2  Antenna input impedance from 200MHz to 300MHz in Smith chart ($Z_0 = 50\Omega$) form prior to any matching done for the transmitter and receiver ends. .... 40

Figure 6.3  Antenna input reflection coefficient from 200MHz to 300MHz in dB prior to any matching done for the transmitter and receiver ends. ............................ 41

Figure 6.4  Lumped element matching network for transmitter and receiver antennas, where $Z_{in}$ refers to the input impedance looking into the transmitter or receiver circuit, which is $50\Omega$ .................................................... 41

Figure 6.5  Antenna input impedance from 200MHz to 300MHz in Smith chart ($Z_0 = 50\Omega$) form after connecting a series capacitor (21.2pF) and a shunt inductor (57.1nH) to the antenna. .................................................. 42

Figure 6.6  Input reflection coefficient measurement after matching for both transmitter and receiver antennas. .................................................. 44
Figure 6.7  Device layout of OFC SAW temperature sensor composed of a wideband bidirectional transducer (middle) with reflector banks on each side, both encoded with the same OFC code and both having seven time chips but with a different delay between the bidirectional transducer and each reflector bank. ................................. 45

Figure 6.8  Input impedance of the SAW sensor and the antenna in Smith chart form based on $Z_0 = 50\Omega$ prior to any matching. .................................................. 46

Figure 6.9  Resistance and reactance ($\Omega$) of the SAW sensor .............................. 47

Figure 6.10  Lumped element matching network for conjugate matching between the antenna and SAW device. ................................................................. 48

Figure 6.11  Impedance curves for $Z_{\text{saw}}$ (solid) and $Z_{\text{ant}}$ (dashed) after applying matching networks on both the antenna and Device 1301 based on the references indicated in Fig. 6.10. .................................................. 49

Figure 6.12  VSWR ADS simulation result (blue and dashed) and measurement (red and solid) between the antenna and the SAW device after using the matching network in Fig. 6.10 and eq. 6.2 ............................ 50

Figure 6.13  2D radiation pattern, in the E-plane (broken line) and the H-plane (solid line), in polar form for the transmitter and receiver antennas at a center frequency of 250MHz. .................................................. 52

Figure 6.14  3D radiation pattern for the transmitter and receiver antennas at a center frequency of 250MHz. .................................................. 53

xiv
Figure 6.15 Experimental setup used to determine the minimum instantaneous power required for a good correlation peak after matched filtering the response of the SAW sensor due to wireless transmission against an ideal OFC signal. .......................... 54

Figure 6.16 Correlation peak results due to matched filtering the ideal OFC signal with the wireless transmitted response due to the SAW sensor (using one reflector bank) for different attenuation values, as labeled, used in varying the instantaneous power fed to the transmitting antenna, as in Fig. 6.15. .......................... 55

Figure 6.17 Temperature measurement setup using the 250MHz transceiver prototype and the antennas designed herein with a hotplate used to vary ambient temperature 56

Figure 6.18 Temperature measurement results obtained by using the 250MHz transceiver prototype and the antennas designed herein versus the measurements made by the thermocouple. .......................... 57

Figure 7.1 Schematic of an open-sleeve dipole composed of an excited dipole with two sleeves or parasitics on both of its sides. .......................... 59

Figure 7.2 912MHz Antenna with 10% fractional bandwidth composed of a radiating structure and a balun integrated within the antenna feed. .......................... 60

Figure 7.3 912MHz antenna with a bandwidth of 800MHz to 1GHz and is composed of the excited dipole and a parasitic sleeve. The lengths, width (both in mm) and the angle are as labeled. .......................... 62

Figure 7.4 Antenna current distribution normalized by a value of 18.823A/m .......................... 63
Figure 7.5  Antenna reflection coefficient simulation results for the dipole without a sleeve and for the dipole with a sleeve, as obtained by using IE3D.  ................................ 64

Figure 7.6  Initial circuit design for the balun, which also serves as a matching network. The 1:1 transformer represents the coupling of the CPS structure to the microstrip line.  ........................................ 66

Figure 7.7  Zin  ........................................................................................................ 66

Figure 7.8  Characteristic impedance and effective permittivity of a CPS line with a strip width of 10mm and a gap of 0.5mm between the strips. ........................................ 67

Figure 7.9  Reflection coefficient simulation and experimental results of the antenna shown in Fig. 7.2  ................................................................. 70

Figure 7.10  912MHz OFC encoded SAW temperature sensor composed of two identical reflector banks but with the time delay between the wideband transducer (middle) and the reflectors (left and right) different for each bank.  ......................... 71

Figure 7.11  Measured 912MHz OFC encoded SAW temperature sensor time response which shows five chips for each reflector bank. Chip no. 2 for each bank is heavily attenuated, as highlighted above.  ........................................ 72

Figure 7.12  Network used in matching the 912MHz SAW device to the 50Ω antenna, while maintaining at least 10% fractional bandwidth.  ......................... 73

Figure 7.13  Measured 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form.  ........................................ 74
Figure 7.14 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting an inductor for resonance at center frequency.

Figure 7.15 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting a $\lambda/4$ impedance transformer with $Z_0 = 12.367\Omega$ to transform the center frequency impedance of Fig. 7.14 to $Z = 36.274\Omega$.

Figure 7.16 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting a $\lambda/4$ line with $Z_0 = 36.274\Omega$.

Figure 7.17 Input impedance of the parallel inductor-capacitor resonator (thicker line) only along with the 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after completion of the matching circuit.

Figure 7.18 Modification of the network shown in Fig. 7.12, wherein the inductor is replaced with a series short stub.

Figure 7.19 Kuroda’s identity used in transforming a series shorted stub ($Z_1$) into a shunt open stub. The transmission line elements are both $\lambda/8$ long.

Figure 7.20 Modification of the network shown in Fig. 7.18, wherein the series short stub has been replaced with an open shunt stub after Kuroda’s transformation shown in Fig. 7.19.

Figure 7.21 Final matching network that serves to transform the device to 50Ω over its 10% fractional bandwidth.
Figure 7.22 The calculated reflection coefficient of the SAW device after attaching the matching network looking into the reference shown in Fig. 7.21.

Figure 7.23 Final optimized matching network, with its dimensions (in mm) obtained using Momentum Optimizer, used in matching the 912MHz SAW device to the 50Ω antenna.

Figure 7.24 The impedance references used in calculating the reflection coefficient as a result of conjugate matching the SAW device to the 50Ω antenna.

Figure 7.25 Measured reflection coefficient due to conjugate matching between the SAW sensor and the 50Ω antenna, as illustrated in Fig. 7.24.

Figure 7.26 Measured antenna gain frequency sweep for antenna separation values of 0.7m and 0.85m.

Figure 7.27 Orientation of the antenna, within the Cartesian 3D space.

Figure 7.28 Measured normalized radiation pattern (dB) of the E-Plane for 866MHz, 912MHz and 957MHz.

Figure 7.29 Measured normalized radiation pattern (dB) of the H-Plane for 866MHz, 912MHz and 957MHz.

Figure 7.30 Test setup used in observing the wirelessly transmitted time response of the SAW device. It is composed of the transmitter, SAW tag and receiver antennas. They are arranged collinearly with a distance, d, separating any two adjacent antennas.

Figure 7.31 Wireless runs
Figure 7.32 Prediction of range of operation as a function of device loss.
LIST OF TABLES

Table 2.1 Definition of P-Parameters ................................................. 5
Table 2.2 Signal Flow Diagram Path for $P_{33}$ ............................. 7
Table 3.1 Part #856070 SAWTEK Lithium Niobate Filter ................. 19
Table 4.1 Summary of the Effects of Increasing SRT Length ............. 34
Table 7.1 Dimensions of the components used in the antenna feed, based on Figs. 7.2, 7.6 .......................................................... 69
<table>
<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>Ampere</td>
</tr>
<tr>
<td>B</td>
<td>Fractional bandwidth</td>
</tr>
<tr>
<td>c</td>
<td>Speed of light</td>
</tr>
<tr>
<td>C</td>
<td>Capacitance</td>
</tr>
<tr>
<td>cm</td>
<td>Centimeter</td>
</tr>
<tr>
<td>COM</td>
<td>Coupling of modes</td>
</tr>
<tr>
<td>CPS</td>
<td>Coplanar strip line</td>
</tr>
<tr>
<td>CPW</td>
<td>Coplanar waveguide</td>
</tr>
<tr>
<td>dB</td>
<td>Decibel</td>
</tr>
<tr>
<td>dBm</td>
<td>Decibel relative to 1 milliwatt</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite impulse response</td>
</tr>
<tr>
<td>FR4</td>
<td>Flame retardant 4 substrate used for printed circuit boards</td>
</tr>
<tr>
<td>G</td>
<td>Absolute gain</td>
</tr>
<tr>
<td>GHz</td>
<td>Gigahertz, 1,000,000,000 cycles per second</td>
</tr>
<tr>
<td>Hz</td>
<td>Hz, 1 cycle per second</td>
</tr>
<tr>
<td>l</td>
<td>Length of transmission line</td>
</tr>
<tr>
<td>L</td>
<td>Inductance</td>
</tr>
<tr>
<td>LiNbO$_3$</td>
<td>Lithium Niobate</td>
</tr>
<tr>
<td>$\text{Loss}_{\text{SAW}}$</td>
<td>SAW device loss</td>
</tr>
<tr>
<td>MHz</td>
<td>Megahertz, 1,000,000 cycles per second</td>
</tr>
<tr>
<td>m</td>
<td>Meter</td>
</tr>
<tr>
<td>mm</td>
<td>Millimeter</td>
</tr>
<tr>
<td>nH</td>
<td>nano-Henry</td>
</tr>
<tr>
<td>OFC</td>
<td>Orthogonal frequency coding</td>
</tr>
<tr>
<td>pF</td>
<td>pico-Farad</td>
</tr>
<tr>
<td>$P_r$</td>
<td>Power received</td>
</tr>
<tr>
<td>$P_t$</td>
<td>Power transmitted</td>
</tr>
<tr>
<td>PG</td>
<td>Processing gain</td>
</tr>
<tr>
<td>Q</td>
<td>Quality factor</td>
</tr>
<tr>
<td>$R_{EB}$</td>
<td>Center frequency input resistance to transform to for matching</td>
</tr>
<tr>
<td>$R_G$</td>
<td>Generator input resistance</td>
</tr>
<tr>
<td>$R_{SAW}$</td>
<td>Center frequency input resistance of SAW device</td>
</tr>
<tr>
<td>RX</td>
<td>Receiver</td>
</tr>
<tr>
<td>Symbol</td>
<td>Definition</td>
</tr>
<tr>
<td>--------</td>
<td>--------------------------------</td>
</tr>
<tr>
<td>s</td>
<td>Second</td>
</tr>
<tr>
<td>SAW</td>
<td>Surface acoustic wave</td>
</tr>
<tr>
<td>SMA</td>
<td>SubMiniature version A</td>
</tr>
<tr>
<td>SRT</td>
<td>SAW reflective transducer</td>
</tr>
<tr>
<td>SWR</td>
<td>Standing wave ratio</td>
</tr>
<tr>
<td>TCF</td>
<td>Temperature coefficient of frequency</td>
</tr>
<tr>
<td>TDM</td>
<td>Time Division Multiplexing</td>
</tr>
<tr>
<td>TX</td>
<td>Transmitter</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Transducer beamwidth</td>
</tr>
<tr>
<td>$Y_L$</td>
<td>Load admittance</td>
</tr>
<tr>
<td>$Z_0$</td>
<td>Transmission line characteristic impedance</td>
</tr>
<tr>
<td>$Z_{in}$</td>
<td>Input impedance</td>
</tr>
<tr>
<td>$Z_L$</td>
<td>Load impedance</td>
</tr>
<tr>
<td>$Z_{SAW}$</td>
<td>SAW device input impedance</td>
</tr>
</tbody>
</table>
Part I

SAW Reflective Transducers
Figure 1.1: SAW reflective transducer (SRT), connected to a load sensor is used in conjunction with a SAW transducer.

Fig. 1.1 shows a device, composed of two transducers. On the left, the launching transducer is connected to an antenna which receives an interrogation signal. The transducer cascaded and is on the right is connected to a sensor whose impedance varies. This will be referred to as the surface acoustic wave (SAW) reflective transducer (SRT) as it has the ability to reflect the incident wave coming from the launching transducer. The amplitude of the reflected wave is dependent upon the impedance of the load attached to it. The time response of the launching transducer is convolved with the time response of the SRT and is eventually transmitted by the antenna. The final response can be predicted by using the P-parameters.
Chapter 2 covers the theory and derivation necessary for subsequent chapters.

Chapter 3 covers the amplitude changes in the SRT response, as previously shown by [1]; theoretical and experimental results will be shown.

As noise is inherent in communication systems, it is necessary to find an alternative for measuring the sensor load attached to the SRT. Chapter 4 covers such alternative, where the resonant frequency changes as long as the SRT is designed for resonance to occur upon viewing its admittance response. The theory will be presented.

For this thesis, simulations are performed by using the finite impulse response (FIR) method. Coupling of modes (COM) simulation was not performed for any of the devices.
CHAPTER 2
P-MATRIX DERIVATION

2.1 SRT Reflection Coefficient Derivation

Figure 2.1: Block Diagram of SAW transducer, where a, b, V, I, refer to the forward, reverse waves, voltage and current applied to the SAW device respectively.

The P-matrix is shown in eq. 2.1, with the forward and reverse waves as shown in Fig. 2.1.

\[ P_{11} \], which is also the reflection coefficient of the SRT \( S_{11} \), is to be derived.
\[
\begin{bmatrix}
 b_1 \\
 b_2 \\
 I
\end{bmatrix} =
\begin{bmatrix}
 P_{11} & P_{12} & P_{13} \\
 P_{21} & P_{22} & P_{23} \\
 P_{31} & P_{32} & P_{33}
\end{bmatrix}
\begin{bmatrix}
 a_1 \\
 a_2 \\
 V
\end{bmatrix}
\]  

(2.1)

Table 2.1: Definition of P-Parameters

<table>
<thead>
<tr>
<th>P_{11}, P_{22}</th>
<th>Acoustic port reflection coefficient</th>
</tr>
</thead>
<tbody>
<tr>
<td>P_{12}, P_{21}</td>
<td>Acoustic port transmission coefficient</td>
</tr>
<tr>
<td>P_{13}, P_{23}</td>
<td>Voltage to SAW transfer elements (Ω^{-0.5})</td>
</tr>
<tr>
<td>P_{31}, P_{32}</td>
<td>SAW to voltage transfer elements (Ω^{-0.5})</td>
</tr>
<tr>
<td>P_{33}</td>
<td>Transducer admittance</td>
</tr>
</tbody>
</table>

Assuming that the mechanical reflectivity per electrode is zero (no reflections), the SRT transducer reflectivity, S_{11}, is solved with the following boundary conditions.

- Reverse wave at port 1, b_1 = 0, as there are no reflections due to the electrode

- Forward wave at port 2, a_2 = 0, as there is only one other transducer, the device launching transducer, which is located to the left of the SRT.

Thus, the wave incoming the SRT is only a_1, which is due to the launching transducer located to the left hand side; the impedance of the load is \( z_L = V/I = y_L^{-1} \), according to Fig. 2.1. This results in the following two equations, where \( P_{31} = -2P_{13} \).

\[
b_1 = P_{11}a_1 + P_{13}V = 0
\]  

(2.2)
\[ I = -Vy_L = P_{31}a_1 + P_{33}V \quad (2.3) \]

Using eq. 2.2 and eq. 2.3, \( S_{11} \) is given as follows [2], [1].

\[ S_{11} = \frac{2P_{13}^2}{P_{33} + Y_L} + S_{11,sc} \quad (2.4) \]

\( S_{11,sc} \) gives the reflection coefficient of the transducer if the load is a short circuit and is a non-zero value under the assumption that each electrode is able to reflect. For simplicity, since \( S_{11,sc} \) is a small value, let \( S_{11,sc} = 0 \).

According to eq. 2.4, the reflection coefficient for a short circuit load \( (Y_L = \infty) \) attached to the SRT is equal to zero. For an open load \( (Y_L = 0) \), the reflection coefficient is equal to \( \frac{2P_{13}^2}{P_{33}} \).

### 2.2 Device Model

By cascading the launching transducer with the SRT, as shown in Fig. 2.1, the overall device admittance response is derived using Mason’s Rule with the aid of the signal flow chart shown in Fig. 2.2.
Figure 2.2: Signal flow diagram for the device shown in Fig. 2.1 where each path is labeled with its respective gain.

There are two paths for $P_{33} = I_A/V_A$, the transfer function to be solved, which is also the overall device input admittance. These are listed in Table 2.2, [2].

<table>
<thead>
<tr>
<th>Path no.</th>
<th>Equation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$P_1 = P_{33A}$</td>
</tr>
<tr>
<td>2</td>
<td>$P_2 = P_{23A}e^{-jKL}P_{11B}e^{-jKL}P_{32A}$</td>
</tr>
</tbody>
</table>

For the overall system, there is only one loop which does not touch $P_1$ but touches the $P_2$. Thus, the determinant is given in eq. 2.5.

$$\Delta = 1 - P_{22A}P_{11B}e^{-2jKL}$$ (2.5)

Thus, as the loop does not touch $P_1$ but touches $P_2$, the cofactors are $\Delta_1 = \Delta$ and $\Delta_2 = 1$. The transfer function, $P_{33} = I_A/V_A$ is then solved using the following equation, as
dictated by Mason’s Rule.

\[ P_{33} = P_1 \frac{\Delta_1}{\Delta} + P_2 \frac{\Delta_2}{\Delta} \]  

(2.6)

Finally, from eq. 2.6, the device input admittance \[2\] is as follows after simplification as

\[ P_{32A} = -2P_{23A}. \]

\[ Y_{33} = P_{33A} + \frac{-2P_{23}^2 P_{11B} e^{-2jkL}}{1 - P_{11B} P_{22A} e^{-2jkL}} \]  

(2.7)
CHAPTER 3
AMPLITUDE VARIATIONS OF SRT FREQUENCY RESPONSE

This chapter discusses the variations in the amplitude response of the SRT with changes in impedance load attached to it. The impedance load variations are limited to purely resistive, capacitive or inductive loads. The changes will be observed both in theory and experiment. This work was previously done in [1].

3.1 Simple Model for SRT

The reflection coefficient of the SRT, by attaching varying loads and by using the impulse response method, is used for initial analysis.

For the calculations, let the substrate to be used be YZ Lithium Niobate, where both of the transducers are 10λ long. Let the beam width be equal to 25λ long, the reflectivity per electrode is equal zero and the center frequency of both transducers is 250MHz.

First, let the launching transducer and the SRT be analyzed separately.
3.1.1 Launching Transducer

Figure 3.1: (a) Reflection coefficient $S_{33}$ and (b) input admittance $P_{33}$ of the device without the presence of the SRT.

For the launching transducer, Figs. 3.1(a) and 3.1(b) show the $S_{33}$ (reflection coefficient) and $P_{33}$ (input admittance) plots of the device without the presence of the SRT. At center frequency, the input admittance of the device and the reflection coefficient are $2 + j2.52 \text{mS}$ and $-1.716 \text{dB}$, respectively.
3.1.2 SAW Reflective Transducer

The reflection coefficient ($S_{11}$) of the device at center frequency is first plotted to observe the changes in amplitude for different loads attached. The impedance load is varied as follows - purely resistive, capacitive and inductive load.

![Acoustic Reflectivity](image)

Figure 3.2: Reflection coefficient ($P_{11}$) at center frequency of the SRT for resistive, inductive or capacitive loads

Eq. 2.4 and Fig. 3.2 indicate where the short and open are located. A short load indicates no reflections for the ideal case, while for the open load; the reflection coefficient is equal to $2P_{13}^2/P_{33}$. In addition, Fig. 3.2 shows the paths the reflection coefficient curves take with
an increase in resistance, inductance or capacitance. These cases are presented individually below, at center frequency.

Resistive Load

![Reflection Coefficient Graph](image)

Figure 3.3: Reflection coefficient of SRT with changes in resistive load.

For a resistive load, Fig. 3.3 shows that a short load indicates no reflections occurring due to the SRT at its port 1. The reflection coefficient increases rapidly for lower values of resistance and saturates to a constant value as resistance increases further. This value is equal to the -4.57dB, the reflection coefficient for an open load.
Inductive Load

Figure 3.4: Reflection coefficient of SRT with changes in inductive load.

Fig. 3.4 shows the variation in reflection coefficient with an increase in inductive load. As the inductance increases, it saturates to reflection coefficient value of -4.57dB, equivalent to that of an open load. The peak of the curve in Fig. 3.4 should correspond to a point where the load is equal to but in opposite sign of the $P_{33}$ reactance. Resonance occurs at the peak, where $Im(P_{33} + Y_L) = 0$. Since the device at 250MHz is capacitive with a reactance of 2.52mS, an inductive load of -2.52mS or 0.253$\mu$H is needed so that resonance at center frequency results.
Capacitive Load

Figure 3.5: Reflection coefficient of SRT with changes in capacitive load.

The changes in center frequency amplitude of the reflection coefficient due to changes in the capacitance value are shown in Fig. 3.5. For low values of capacitance, the reflection coefficient is equal to -4.57dB, equivalent to an open load. As it increases in capacitance, the reflection coefficient decreases to low values as the capacitor resembles a short load.

3.2 Device Modeling using P-Matrix Formula

The input admittance of the device, where the launching transducer is cascaded with the SRT, was already derived in Chapter 2 and is referred to in eq. 2.7. Let L, the center to center distance between the transducers, be 0.872mm, which is equivalent to a one way time delay of 0.25µs.
3.2.1 Open and Shorted Load

Figure 3.6: Device input admittance plots due to (a) short load and (b) an open load.

Fig. 3.6(a), for the shorted load case, indicates that there is no reflection occurring as the input admittance of the device is the same as that of a single transducer only. Fig. 3.6(b), for the open load case, indicates that there are reflections occurring as the curves are
no longer smooth. As previously mentioned, the SRT reflection coefficient, $S_{11}$, is no longer equal to zero given an open load from eq. 2.4.

![S33 (Reflection Coefficient)](image)

Figure 3.7: Device reflection coefficient for shorted and open load

Fig. 3.7 shows the corresponding reflection coefficient of the device for short and open loads. If the load is a short, the reflection coefficient response shows no ringing, which is otherwise for the open load.

### 3.2.2 Resistive Load

Fig. 3.8 shows $S_{33}$ for the device given 50Ω, 500Ω and 5kΩ loads. From Fig. 3.3, it can be recalled that as the resistance increases the reflection coefficient of the SRT increases and saturates to the reflection coefficient of an open load. Fig. 3.8 shows the enlargement of ripples of the $S_{33}$ response with an increase in resistance and in the same way, the size of the
ripple should approach to that of the open load case, seen in Fig. 3.7 for further increases in resistive load.

![Image](image.png)

Figure 3.8: Device reflection coefficient for 50Ω, 500Ω and 5kΩ loads.

### 3.2.3 Inductive Load

For this case, the center frequency input admittance of the transducer is $2 + 2.52jmS$. To maximize the reflection coefficient value $S_{11}$ of the SRT, let the load be $Y_L = -2.52jmS$ ($0.253\mu H$) for resonance, which is equivalent to attaching an inductive load $|3|$. Due to resonance, the ripples are the largest for this amount of load inductance. As the inductance deviates from this value, the ripples are minimized, as observed in Fig. 3.9.
3.2.4 Capacitive Load

Figure 3.9: Device reflection coefficient for 0.637\,\mu H, 0.253\,\mu H and 63.7\,nH loads.

Figure 3.10: Device reflection coefficient for 1pF, 5pF, 10pF and 15pF loads.
Fig. 3.10 shows that with an increase of capacitive load, the ripples become smaller as the $S_{11}$ of the SRT becomes less with an increase of capacitance, as previously shown in Fig. 3.5.

### 3.3 Experimental Result

A lithium niobate filter composed of two transducers and with the following characteristics is used for this experiment, [4].

<table>
<thead>
<tr>
<th>Table 3.1: Part #856070 SAWTEK Lithium Niobate Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Die size</td>
</tr>
<tr>
<td>Substrate</td>
</tr>
<tr>
<td>TCF temperature coefficient of frequency</td>
</tr>
<tr>
<td>$K^2$ (%) coupling constant</td>
</tr>
<tr>
<td>$V_{free}$ surface</td>
</tr>
<tr>
<td>Package Dimension</td>
</tr>
<tr>
<td>Center Frequency</td>
</tr>
<tr>
<td>Bandwidth</td>
</tr>
</tbody>
</table>

The filter consists of input and output apodized transducers [4], which results in a filter that has an almost rectangular shape in the frequency domain.
3.3.1 SRT Reflection Coefficient Extraction

The SRT device response was extracted from the measured reflection coefficients of the device; the procedure is described below.

Eq. 2.7 shows the device $S_{33}$ or reflection coefficient response. Realistically, $S_{22A}$, the reflection coefficient of the launching transducer at port 2, could not be exactly equal to zero due to the ability of each electrode to reflect a small amount of energy. However, for simplicity, let $S_{22A} = 0$.

$P_{33A}$, the input admittance of the launching transducer, was attained experimentally by attaching a short load to the device and solving for admittance from the acquired reflection coefficient measurement and by using eqs. 2.4 and 2.7. Similar to $S_{22A}$, it was also assumed that $P_{11SC} = 0$ in eq. 2.4 for the SRT. It simplifies eq. 2.4 to $P_{11B} = S_{11} \approx 0$. Given these constraints and for as a short load is attached to the SRT, eq. 2.7 simplifies to $Y_{33} = P_{33A}$, the input admittance of the device.

$$Y_{33} = \left( Z_0 \frac{1 + \Gamma}{1 - \Gamma} \right)^{-1}$$ (3.1)

The reflection coefficient measurement, $\Gamma$ obtained by the vector network analyzer (VNA) can be easily converted to admittance through eq. 3.1, where $Z_0 = 50\Omega$ is the characteristic impedance of the VNA.
At center frequency, \(2 |P_{23}|^2\) is approximated as the center frequency conductance (\(\text{Re}(P_{33A})\)). Given \(P_{33A}\) and \(2 |P_{23}|^2\), \(P_{11B}\), the reflection coefficient due to the SRT alone, was extracted given \(Y_{33}\), which is the measured admittance of the device with any arbitrary load. Eq. 2.7 was used to obtain SRT reflectivity with the given parameters.

### 3.3.2 Open and Shorted Load

![Reflection Coefficient of DUT](image)

Figure 3.11: Reflection coefficient of the DUT with an open (thick) and short (thin) load attached

As seen in previous theoretical examples, a short load results in \(P_{33SC} = P_{33}\), where \(P_{33}\) is the transducer admittance without the presence of the SRT, which results in a smooth DUT response (Fig. 3.7). On the other hand, an open load results in an SRT reflection coefficient of \(P_{11} = 2P_{13}^2/P_{33}\), resulting in a DUT response with ringing (Fig. 3.7). By attaching a
short and an open load, Fig. 3.11 shows the measured frequency response. It clearly shows correlation with theoretical results as the ripple appears for an open load and a smooth curve results for a short load attached.

### 3.3.3 Inductive Load

![Reflection Coefficient Graph](image)

Figure 3.12: Measured reflection coefficient at center frequency (140MHz) of the SRT for different values of inductive load

Resonance can occur if the inductive load impedance is equivalent to $(-Im(P_{33}))$. Using the same device, 12 different inductors were used in this experiment.

Fig. 3.12 shows a similar trend to Fig. 3.4 where at the lower range for inductance, the reflection coefficient increases with an increase in inductive load until it reaches resonance point. From there, the reflection coefficient decreases with further increase in inductive load until it saturates to the same reflection coefficient value due to an open load.
The SRT reflectivity measurements from Figs. 3.11 and 3.12 were compiled and plotted in polar format in Fig. 3.13. These results show similarity to Fig. 3.2, which is the polar plot using the ideal case for a short, open and inductive load. Differences between ideal and experimental plots occur due to different device design. Nevertheless, this section proves that realistically, the SRT reflectivity should change accordingly as the impedance of the load attached is also changed. As a consequence, the overall device response, which is dependent upon SRT reflectivity values, should change as well. Finally, extracting the SRT reflectivity from the measured device response was previously presented.

Figure 3.13: Measured reflection coefficient of the SRT (within the device whose characteristics are found in Table 3.1) at center frequency (140MHz) for different values of inductive load, and for open and short loads, in polar format
(a) Input conductance $Re(P_{33})$ of resonant SRT

(b) Input susceptance $Im(P_{33})$ of resonant SRT

Figure 4.1: Input admittance ($P_{33}$) of the SRT, without a load. Boxed portion of the $P_{33}$ imaginary curve refers to the inductive portion of the plot.
Previously, discussions on using the SRT as an interface for measuring devices rely only on amplitude change as the load attached to it changes. As noise is inherent in systems and in wireless transmission of signals, an alternative to measuring amplitude was being sought. The peak frequency location of the reflected wave (\( P_{11} \) or \( S_{11} \)) of the SRT changes with load attached, as observed computationally and will be discussed further in this chapter. For the analysis herein, the finite impulse response method is used for simulation.

4.1 Overview

For demonstration purposes, a resonant SRT shown in Figs. 4.1(a) and 4.1(b), which has an overall length of \( 25\lambda \), is going to be used. This requires 25 pairs of electrodes. It was assumed for this analysis that each pair has an equivalent capacitance of \( 4.6\frac{pF}{cm\text{-pair}} \). For simplicity, non-reflective transducers are used. The reflectivity that results in the SRT is solely due to the load attached to it. The beamwidth is \( 50\lambda \) and the substrate used is Y-Z lithium niobate \( LiNbO_3 \). A longer time rectangular function results in a narrower bandwidth, and with the number of electrodes increasing, the conductance is increased. As shown in Fig. 4.1(b), as a consequence of larger conductance, the minimum of \( Im(P_{33}) \) curve is lower, resulting in an inductive \( P_{33} \) in frequencies slightly larger than center frequency. For this example, resonance (\( Im(P_{33}) = 0 \)) occurs at two frequency points, which are approximately, 252.93MHz and 256.42MHz. Resonant frequency points can be changed by adding a susceptance load, whether inductive or capacitive.
It is also interesting to note how a resonant SRT changes the amplitude of the reflection coefficient as a load is varied from a short to an open. This is best illustrated in Fig. 4.2, where the reflection coefficient changes are taken at the resonant frequency point of 252.93MHz.

Figure 4.2: Resonant SRT reflection coefficients for resistive, capacitive and inductive loads in polar form taken at 253.2MHz, at the resonant frequency of 252.93MHz.

Fig. 4.2 shows a difference in phase because the SRT has a linear phase response. At center frequency, the phase is zero but at the frequency point chosen, the phase deviates from zero linearly.

For inductive and resistive loads, as \( L \to \infty \) and \( R \to \infty \), an open load results. As expected, they converge at the same point. However, unlike Fig. 3.2, the point of convergence
is also the maximum achievable reflection coefficient, which is due to the fact that as an open load is attached \((Y_L = 0)\), resonance occurs and the denominator of eq. 2.4 is minimized.

For capacitive loads, a short load is equivalent to having a capacitance value of \(C \rightarrow \infty\). By further lowering values, an open load will eventually result. Hence, the reflection coefficient curve due to capacitive load begins with the maximum attainable reflection coefficient and eventually decreases in magnitude with an increase in capacitance, as indicated in Fig. 4.2.

### 4.2 Frequency Peak Deviations due to Susceptive Loads

Change mechanisms in frequency peak deviations of the SRT reflection coefficient are discussed in this section using capacitive and inductive loads. For this section, the SRT beamwidth is kept constant at 50\(\lambda\) while its width is varied to 25\(\lambda\), 35\(\lambda\) and 50\(\lambda\) to observe how the transducer length affects the range of loads that causes considerable changes in the frequency point of where the peak amplitude reflection coefficient of the device could occur.
4.2.1 Capacitive Load

Figure 4.3: SRT susceptance and reflection coefficient simulations upon attaching different capacitive load values (in pF) to the SRT of length $25\lambda$ and $W_a = 50\lambda$.

The peak resonant frequency has changed for each termination, as observed in Fig. 4.3. Based on eq. 2.4, as an increase in capacitance value causes an increase in $Y_L$, the $S_{11}$ SRT peak reflection coefficient amplitude is lowered. Although the peak conductance occurs at center frequency, shifts in the peak are due to a particular load applied that leads to $Im(P_{33} + Y_L) \approx 0$. Once this condition is achieved, resonance occurs as the denominator
in the first term of eq. 2.4 is only composed of real components at that specific frequency. Thus, increasing the capacitance in the manner described in Fig. 4.3 causes an increase then a decrease in resonant frequency. Starting from a smaller value of capacitance, increasing its value causes the resonant frequency to increase. Once the response is $\text{Im}(Y_L + P_{33}) > 0$ due to a higher capacitance value, the resonant frequency decreases back to the center frequency slowly, as seen in Fig. 4.3. This trend can be more clearly seen in Fig. 4.4.

Figure 4.4: Resonant frequency change versus capacitive load change for $N_P = 25, 35$ and 45. The second plot is a magnified version of the first plot.
Fig. 4.4 shows the change in resonant frequency due to changes in capacitive load for an SRT with $W_a = 50\lambda$ and lengths of $25\lambda$, $35\lambda$ and $45\lambda$. Increasing the number of electrode pairs to the SRT increases its central frequency conductance and narrows its bandwidth. The former effect allows resonant frequency changes for a wider range of capacitive load values as the minimum of the $Im(P_{33} + Y_L)$ curve becomes more negative, leaving a wider range of capacitor values that can cause $Im(P_{33} + Y_L)$ curve to equal zero, while noting that an increase in capacitance causes an upward shift in the $Im(P_{33} + Y_L)$ curve (Fig. 4.3). However, the latter effect of narrower bandwidth limits the range by which the resonant frequency could still change once the SRT length is increased.

### 4.2.2 Inductive Load

The inductive load is $Y_L = (j\omega L)^{-1}$. Using a shorted load or small values for inductive loads, the $Im(P_{33} + Y_L)$ curve is negative. As inductance is increased an upward shift in the $Im(P_{33} + Y_L)$ plot results since $Y_L$ becomes smaller. The $Im(P_{33} + Y_L)$ plot found in Fig. 4.5 shows that the frequency at which it is closest to $Im(P_{33} + Y_L) = 0$ changes. For an SRT length of $25\lambda$ and at a lower inductance value of $10\text{nH}$, the $Im(P_{33} + Y_L)$ curve is negative and far from $Im(P_{33} + Y_L) = 0$; the susceptance has nearly no effect on the resonant frequency. For this case, the dominating factor that decides where the frequency peak is located is the center frequency of the device (250MHz), as shown in the $S_{11}$ plot in Fig. 4.5. But as the susceptance is drawn closer to the $Im(P_{33} + Y_L) = 0$ line by increasing the inductive load,
the frequency point in which it is equal to $Im(P_{33} + Y_L) = 0$ is where or near where the peak $S_{11}$ frequency occurs. Further increasing inductance causes lesser amounts of upward shifts of the $Im(P_{33} + Y_L)$ curve until a saturation point is reached. At saturation, it is equivalent to an open load, resulting in the maximum peak amplitude in $S_{11}$. The frequency shift of the peak of the $S_{11}$ response becomes less noticeable until a shift no longer occurs due to saturation.

Figure 4.5: SRT susceptance and reflection coefficient simulations upon attaching different inductive load values (in nH) to the SRT of length $25\lambda$ and $W_a = 50\lambda$. 
As mentioned previously, increasing the number of electrodes to the SRT increases its central frequency conductance and narrows its bandwidth. Due to narrower bandwidth the range where the peak frequency could shift to becomes more limited. As shown in Fig. 4.6, which has a length of 45\(\lambda\), because of an increase in central frequency conductance, the peak of the \(Im(P_{33} + Y_L)\) curve becomes larger for a longer SRT. For a 10nH load, resonance already occurs causing shifts in frequency due to resonance unlike in Fig. 4.5 where the more dominating factor that affects where the peak is located is the center frequency of the
device given the same 10nH load. As a result, by increasing the length of the SRT, the range where more noticeable shifts in $S_{11}$ peak frequency is more limited.

![Resonant Frequency Change due to Inductive Load](image)

**Figure 4.7**: Resonant frequency change versus inductive load change for $N_p = 25, 35$ and 35. The second plot is a magnified version of the first plot.

In order to show these effects due to changes in the length of the SRT more clearly, Fig. 4.7 was plotted, which shows the frequency point of the peak of the $S_{11}$ response given a range of inductive loads.
Table 4.1: Summary of the Effects of Increasing SRT Length

<table>
<thead>
<tr>
<th>SRT Length</th>
<th>longer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Shift Range</td>
<td>shorter</td>
</tr>
<tr>
<td>Capacitance Value Range</td>
<td>longer</td>
</tr>
<tr>
<td>Inductance Value Range</td>
<td>shorter</td>
</tr>
</tbody>
</table>

Finally, Table 4.1 summarizes the effects in the range of frequencies where the peak could occur, and the capacitive and inductive load range that could still incur peak frequency shifts of the SRT $S_{11}$ response due to adding more electrodes to the SRT, thereby, increasing its time response length.

If an impedance sensor is used, which changes its response with its measured parameter, the SRT interface may be customized to work at the impedance range the sensor works by simply lengthening or shortening the SRT time response length. However, one must keep in mind of the frequency range the peak frequency could shift to in designing the SRT sensor to antenna interface.
Part II

Antennas for Orthogonal Frequency

Coded SAW Sensors
CHAPTER 5
INTRODUCTION

Figure 5.1: Diagram of the OFC tag system used for conducting wireless temperature measurements.

An existing 250MHz orthogonal frequency coded (OFC) tag system prototype consists of a transmitter, receiver and OFC SAW temperature sensors, where the sensors are compact and passive. The sensors take advantage of the non-zero temperature coefficient of lithium niobate ($LiNbO_3$), which is the substrate used for the device that causes the material to contract and expand depending upon the sensor ambient temperature [2]. As shown in Fig. 5.1, the transmitter emits through an antenna a wideband interrogation chirp. Upon its reception, the sensor, which is a one-port reflective device, convolves the chirp response with the SAW device response in time. The signal is relaunched towards the receiver for data
processing to determine the temperature reading. A multi-sensor system, with tags placed at different locations and with each sensor having a unique wideband OFC code for tag identification, allows for simultaneous temperature acquisition at different areas while utilizing the same allocated bandwidth. In order for this system to be plausible, a need for antennas for the SAW tags arises. It is essential to meet bandwidth requirements for each SAW tag to improve range measurements so that it could still be identified and discriminated from other tags, and for a more accurate reading as the receiver uses adaptive matched filtering to determine temperature. A good correlation peak can be achieved if bandwidth requirements are met. By default, the transmitter and receiver ends have 50Ω input impedance while the input impedance of the 250MHz OFC SAW device is capacitive. Variations in input impedance occur across its 28% fractional bandwidth. A brief overview of the device response and antenna designs for both the transceiver and SAW device are further discussed in Chapter 6.

The transceiver system for a 912MHz prototype works similarly as the 250MHz prototype. In the same way, both the transmitter and receiver ends have 50Ω input impedance. The 912MHz SAW device has 10% fractional bandwidth and is also capacitive with a non-uniform input impedance response. Due to higher frequency of operation and narrower device bandwidth, it is easier to devise ways on how to miniaturize a 912MHz tag antenna. Device overview and antenna designs with an attempt in antenna miniaturization for the transceiver and SAW device antennas, respectively, are further discussed in Chapter 7.
CHAPTER 6
ANTENNAS FOR THE 250MHZ WIRELESS TEMPERATURE ACQUISITION SYSTEM PROTOTYPE

As discussed in Chapter 5, a wireless temperature interrogation system employing OFC coded SAW devices would require antennas properly matched to both the transceiver system and the SAW sensor to allow successful ambient temperature acquisitions and to maximize operational distance between the transceiver system and the sensor tags. Techniques to address the challenges posed by the inherently wide bandwidth requirement (28% fractional bandwidth) of the (1) transceiver system and the (2) temperature sensor tags will be discussed in their own respective sections.

6.1 250MHz Transceiver Antennas

The transmitter and receiver are both 50Ω systems and require a fractional bandwidth of 28% centered at 250MHz (215MHz - 285MHz). Initial design procedures outlined in [5] were followed. The diameter of the disk was initially designed to be λ/4 long at 215MHz, which is the lowest frequency point of the band of interest. Thus, its diameter is 350mm and it is cropped at a 20° angle to reduce capacitance of the input impedance of the antenna. The
diameter and the angle of the disk cut were tuned with the aid of a parametric sweep done in Ansoft HFSS. The ground plane dimensions are 18in x 12in, which alters the radiation pattern due to its smaller size. Bandwidth requirements were met in simulation and the antenna was constructed using the said dimensions, as shown in Fig. 6.1.

Figure 6.1: The antenna constructed for transmitter, receiver and the SAW sensor tag
Its input impedance in Smith chart form is shown in Fig. 6.2, which indicates that matching performance is not optimal between 200MHz to 300MHz due to the curve not lying near the center of the Smith chart. Fig. 6.3 shows reflection coefficient of the antenna. Although wide bandwidth was achieved, it is desired to lower the frequency for good matching to 215MHz, the lower band edge of the system, as the current reflection coefficient at this frequency is -6.2dB that has a corresponding standing wave ratio (SWR) of 2.94. It is possible to have the SWR decreased to a value of 2 or less by performing lumped element matching for good antenna performance.
Figure 6.3: Antenna input reflection coefficient from 200MHz to 300MHz in dB prior to any matching done for the transmitter and receiver ends.

6.1.1 Lumped Element Wideband Matching to Match the Antenna to 50Ω

Figure 6.4: Lumped element matching network for transmitter and receiver antennas, where $Z_{in}$ refers to the input impedance looking into the transmitter or receiver circuit, which is 50Ω

Lumped element matching for wider bandwidth as outlined in [6] was employed where inductor, capacitor and inductor capacitor parallel resonator structures were used, as shown in Fig. 6.4.
Prior to adding the resonator, a series capacitor and a shunt inductor were used to provide more degrees of freedom in moving the impedance curve in Fig. 6.2. A series capacitor with a value of $21.2\text{pF}$ was used to reduce the antenna inductance and make it more capacitive. A shunt inductor with a value of $57.1\text{nH}$ was used to compensate for the added capacitance such that at center frequency, the antenna input impedance is resonant. As a result, the impedance curve shown in Fig. 6.5 is located near $Z_0 = 50\Omega$ point within the Smith chart for the entire band.

![Smith Chart](image)

Figure 6.5: Antenna input impedance from 200MHz to 300MHz in Smith chart ($Z_0 = 50\Omega$) form after connecting a series capacitor ($21.2\text{pF}$) and a shunt inductor ($57.1\text{nH}$) to the antenna.

For the parallel inductor capacitor resonator, the value of the shunt inductor was chosen to be $L = \frac{1}{(-11.383\text{jmS})(j2\pi215\text{MHz})} = 65.0\text{nH}$, where the susceptance based on Fig. 6.5 ($-11.383\text{jmS}$) at 215MHz was used. The shunt capacitor has a value of $6.2\text{pF}$ and was
chosen such that it fulfills the resonance condition, shown in eq. 6.1, where $\omega_0$ is the center frequency.

$$j\omega_0 C + \frac{1}{j\omega_0 L} = 0 \quad \text{(6.1)}$$

In building the matching circuit using variable capacitors and inductors, each circuit component from Fig. 6.4 was first measured and set to the theoretical values as close as possible. They were mounted on an FR4 64mil coplanar waveguide (CPW) structure whose line characteristic impedance is 75Ω, which provided ample space between the CPW positive and ground conductors and adjacent ground connections for all shunt elements. The matching circuits for the transmitter and the receiver were further manually tuned to compensate for any second-order effects such as the board fringe capacitance. The SWR measurements after matching are shown in Fig. 6.6. Finally, the bandwidth requirement (215-285MHz) was met and an SWR of less than 2 was achieved for the entire band.
6.2 Antennas for the 250MHz UWB Orthogonal Frequency Coded SAW Temperature Sensor

It is desired to design an UWB antenna that works for a 250MHz SAW temperature sensor. Unlike the transceiver system where 50Ω impedance matching is required, the device requires conjugate matching to be performed for maximum power to be delivered from the antenna to the device and vice versa. A brief overview of the device design and discussions on antenna design, simulation and testing will be presented in this section.
6.2.1 250MHz OFC SAW Temperature Sensor Device Overview

Figure 6.7: Device layout of OFC SAW temperature sensor composed of a wideband bidirectional transducer (middle) with reflector banks on each side, both encoded with the same OFC code and both having seven time chips but with a different delay between the bidirectional transducer and each reflector bank.

Fig. 6.7 shows the device layout; it has 28% fractional bandwidth with a center frequency of 250MHz. Its input impedance is non-uniform and capacitive. For each reflector bank, its time response is composed of seven contiguous chips, where each chip has a time duration of $\tau_c$ and has uniform length but with different carrier frequencies chosen to support orthogonality with each other; thus, it contributes to wider bandwidth. Fig. 6.7 shows that each chip in a reflector bank has a specific center frequency, as noted in the number of electrodes per chip and the width of the electrode for each chip. The material used is lithium niobate $LiNbO_3$, which has a nonzero temperature coefficient [2]. The material expands and contracts with changes in temperature, which makes the distance between electrodes change causing variations in the center frequency for each chip. Details of the device design and operation were previously presented [2]. SAW OFC system operation and benefits were tackled in more detail [7].
6.2.2 Lumped Element Wideband Matching to Match Device 1301 to the Antenna

Figure 6.8: Input impedance of the SAW sensor and the antenna in Smith chart form based on $Z_0 = 50\,\Omega$ prior to any matching.

The antenna needs to be conjugately matched to the SAW sensor (Device 1301) with a maximum SWR of 3 throughout its 28% fractional bandwidth. The input impedance measurements for both are shown in Fig. 6.8. Variations in resistance and the reactance of the SAW temperature sensor are more clearly shown in Fig. 6.9.
As indicated in Figs. 6.8 and 6.9, the center frequency resistance of Device 1301 is a low value whereas the antenna could be easily matched to 50Ω as previously shown in Sec. 6.1.1. To increase the device center frequency resistance to a higher value by using broadband tapered line impedance transformers such as the Binomial or Chebyshev transformer [8] is not an option as the low frequency of operation results in a microstrip or CPW circuit that is too large and impractical to use in conjunction with a large antenna. Thus, lumped element matching is used; however, it is noted that it is impossible to attain a very low reflection coefficient value over the entire bandwidth and adding more lumped elements saturates to a certain performance limit [6]. Instead of matching the antenna and Device 1301 to 50Ω, conjugate matching was done with the aid of the Smith chart, by using eq. 6.2 for reflection coefficient calculations [9], where $Z_{saw}$ and $Z_{ant}$ are the input impedance looking into Device
1301 and the antenna, respectively, and by using ADS tuning tools to achieve best possible performance.

\[
\Gamma = \frac{Z_{saw} - Z_{ant}}{Z_{saw} + Z_{ant}} \tag{6.2}
\]

For maximum power transfer, matching networks for Device 1301 and the antenna were made separately such that the reflection coefficient value from eq. 6.2 is minimized as much as possible while maintaining 28% fractional bandwidth.

![Lumped element matching network for conjugate matching between the antenna and SAW device.](image)

Figure 6.10: Lumped element matching network for conjugate matching between the antenna and SAW device.

Using steps indicated in [6], the device was first made resonant at the device center frequency (250MHz), which originally has an input impedance of \( Z_{in} = 15.5 - j106.3\Omega \). For resonance to occur, an inductor is connected to the SAW device with \( L = \frac{j106.3\Omega}{j2\pi250MHz} = 67.7nH \).
Thereafter, a parallel inductor and capacitor resonant network was used such that it brings impedance points at the band edges closer to where the center frequency impedance is located in the Smith chart without affecting the impedance points near the center frequency. After tuning using ADS, the resonant circuit components were determined to be $145pF$ and $2.8nH$. The red solid curve shown in Fig. 6.11 is the impedance curve for $Z_{saw}$ in the Smith chart based on $Z_0 = 50\Omega$. To conjugate match the antenna with Device 1301, the blue dashed impedance curve for $Z_{ant}$ prior to matching shown in Fig. 6.8 is to be brought closer to the $Z_{saw}$ impedance curve shown in Fig. 6.11. This was achieved by first using a shunt capacitor to bring the curve downwards. Adding a series inductor adds a degree of freedom as to how the curve will be moved on the Smith chart. Tuning for minimum reflection coefficient, the values used for the inductor and capacitor are $12.7nH$ and $19.3pF$, respectively.
respectively. The theoretical SWR that results from this network is shown in Fig. 6.12. The network was built using variable inductors and capacitors, where each lumped element value was first measured and adjusted to be as close as the theoretical values. These elements were mounted on a 64mil FR4 CPW structure whose characteristic impedance was chosen as 75Ω for a convenient gap between the CPW positive and ground planes. The CPW lines provide adjacent ground locations for shunt elements. Due to fringe capacitance, it was required to manually tune the matching network until the SWR was minimized while the bandwidth was maximized to cover the 28% fractional bandwidth as much as possible. The SWR measurement between the antenna and Device 1301 is shown in Fig. 6.12, which indicates that only six out of the seven chips of the OFC code of the device could be successfully transmitted.

![SWR Measurement](image)

Figure 6.12: VSWR ADS simulation result (blue and dashed) and measurement (red and solid) between the antenna and the SAW device after using the matching network in Fig. 6.10 and eq. 6.2
6.3 Antenna Performance

Antenna performance was gauged by evaluating its gain and radiation pattern along with performing measurements for actual wireless temperature interrogations after antenna installation. The system measurements include correlating the wirelessly transmitted sensor response with an ideal OFC response. If the antennas are efficient, correlation peaks could be seen after application of the adaptive matched filtering built within the system. Finally, a comparison of the wireless temperature measurements made by the temperature acquisition prototype compared to those obtained from a thermocouple will be presented.

6.3.1 Antenna Radiation Pattern and Gain

Due to the absence of an anechoic chamber that is suitable for measurement of an antenna designed to work from 215MHz to 285MHz, radiation pattern results from Ansoft HFSS antenna simulations are presented herein. Simulations were only done for 250MHz, the center frequency, as a frequency sweep for the gain is unavailable.
Figure 6.13: 2D radiation pattern, in the E-plane (broken line) and the H-plane (solid line), in polar form for the transmitter and receiver antennas at a center frequency of 250MHz.

Upon simulation, the same dimensions of the actual antenna built and described in Sec. 6.1 were used. At 250MHz, the radiation pattern is presented in Fig. 6.13, which shows the E-Plane ($\phi = 0^\circ$; $\theta$ is varied) and H-Plane ($\theta = 90^\circ$; $\phi$ is varied) radiation patterns of the disk monopole antenna. Although no simulations were done for the SAW antenna, it can be assumed that it has the same radiation pattern, approximately, as the transceiver antenna since good matching occurs at 250MHz, as seen in Fig. 6.12. With a reflection coefficient of -10.7dB, the power transmission coefficient, from [9], is $10\log(1 - |\Gamma|^2) = -0.387\, dB$, which is close to 0dB, for perfect matching.
Fig. 6.14 shows a 3D plot of the radiation pattern at 250MHz, with a schematic of the antenna whose orientation corresponds to the radiation pattern shown. It shows the variation of the gain in more detail. The maximum gain is 2.8dB. The shape of the radiation pattern resembles that of an apple compared to a doughnut shaped radiation pattern a dipole produces. This could be due to the smaller ground plane used for the antenna. Following the image theory [10], the smaller ground plane size does not provide as good of an image of the monopole compared to using a very large ground plane and hence, altering the pattern.
6.3.2 Antenna Performance as Part of the 250MHz OFC Sensor Temperature Interrogation System

Figure 6.15: Experimental setup used to determine the minimum instantaneous power required for a good correlation peak after matched filtering the response of the SAW sensor due to wireless transmission against an ideal OFC signal.

To gauge antenna performance for the system, a network analyzer was used to measure the response of the SAW sensor once implemented in the wireless system. The response due to the sensor was matched filtered adaptively, as in [11]. If the output has a good correlation peak, a meaningful temperature measurement can be made. Since the distance between electrodes in the SAW sensor expands or contracts with changes in temperature, each time chip, $\tau_c$, as discussed in Sec. 6.2.1, changes in frequency. The frequency shifts help determine the ambient temperature after adaptive matched filtering. The minimum instantaneous power fed unto the antenna that is required for a good temperature measurement is to be determined. Transmitter and receiver antennas are placed on each port of the VNA, which are 6m apart. The antenna for the SAW sensor is placed 3m away from either antenna; all antennas were arranged collinearly. An amplifier with 34.1dB gain and a variable attenuator were placed in series with the VNA, where the transmitter port output
power was set to -15dBm/Hz. Without any additional attenuation, the amount of power fed to the transmitting antenna is 19.1dBm/Hz. The setup for the experiment is shown in Fig. 6.15.

![Graph showing correlation peak results due to matched filtering.](image)

Figure 6.16: Correlation peak results due to matched filtering the ideal OFC signal with the wireless transmitted response due to the SAW sensor (using one reflector bank) for different attenuation values, as labeled, used in varying the instantaneous power fed to the transmitting antenna, as in Fig. 6.15.

Fig. 6.16 shows the adaptive matched filtering results, where measurements were taken at room temperature. It shows that the maximum attenuation that can be applied is 40dB for a good correlation peak, which equals -20.9dBm/Hz of power input to the transmitting antenna. While wave propagation loss in free space is a major contributor to loss, the device itself inherently contributes to loss due to the bidirectionality of the transducer (6dB loss),
electrode sheet resistance and OFC coded reflector losses. Details of loss mechanisms can be found in [2].

Figure 6.17: Temperature measurement setup using the 250MHz transceiver prototype and the antennas designed herein with a hotplate used to vary ambient temperature

Finally, Fig. 6.17 [12] shows the the setup used in acquiring the measurements shown in Fig. 6.18, which shows temperature measurements made by the 250MHz prototype after installing the antennas designed for the 250MHz transceiver system, along with the measurements read from the thermocouple, [12]. They correlate very well and prove that by using the existing transceiver prototype and the antennas discussed herein, wireless measurements for temperature can be successfully made.
Figure 6.18: Temperature measurement results obtained by using the 250MHz transceiver prototype and the antennas designed herein versus the measurements made by the thermocouple.
A 912MHz OFC coded and time division multiplexed (TDM) device was already fabricated. Antennas to allow wireless communication between the transceiver and the sensor are required, as previously done in Chapter 6. Along with a higher frequency of operation comes a smaller wavelength; therefore, it can reduce the size of the antennas in comparison to those used for the 250MHz system prototype. In addition, there is a less stringent fractional bandwidth requirement of 10% (866MHz to 957MHz). Discussions on antenna design for the transmitter, receiver and the SAW sensor, and an overview of the 912MHz device will be presented. Although a transceiver system for the 912MHz sensor is non-existent at this time, antenna performance can still be gauged by viewing the SAW tag response after antenna installation, aside from antenna gain and radiation pattern measurements.

7.1 912MHz Antenna Design

As in Chapter 6, both the transmitter and receiver antennas must have 50Ω input impedance. An open-sleeve planar dipole antenna is used and must be matched to the device and system
bandwidth, from 866MHz to 957MHz. Since a dipole requires a symmetrical or a balanced feed while a coaxial line is an unbalanced structure since the inner (positive) and outer (ground) conductors are not symmetric to each other, balun integration is necessary [10]. This section covers (1) the radiator, and (2) balun and matching network design. Simulation results on antenna gain, radiation pattern and return loss will be presented.

7.1.1 Open-Sleeve Dipole Antenna

Figure 7.1: Schematic of an open-sleeve dipole composed of an excited dipole with two sleeves or parasitics on both of its sides.

It was shown in [13] that an antenna with 1.8:1 bandwidth can be achieved by using an open-sleeve dipole antenna. The structure, as shown in Fig. 7.1, is composed of a cylindrical
dipole with two 'open-sleeves' or parasitic elements, placed on both sides that serve to broaden bandwidth. It is noted that the cylindrical dipole has a tapered feed to reduce antenna capacitance. Parametric studies conducted by King and Wong [13] by varying the distance between the excited dipole and the parasitic elements, the conductor diameter and length of the sleeves were done but with the antenna in front of a reflector. It was also stated in [13] that Barkley conducted similar studies to open-sleeve antennas but without the reflector adjacent to it and with sleeve length approximately half of that of the excited dipole. Finally, since a dipole must have a balanced feed whereas the coaxial feed is an unbalanced structure, a balun was used in [13].

### 7.1.2 Planar Open-Sleeve Dipole Antenna

![Figure 7.2: 912MHz Antenna with 10% fractional bandwidth composed of a radiating structure and a balun integrated within the antenna feed.](image-url)
Fig. 7.2 shows that the antenna is composed of a radiating structure, with a dipole and a parasitic sleeve, and a feed that integrates a balun, as highlighted above. The balun successfully transitions from a balanced coplanar strip line (CPS) to an unbalanced microstrip feed line. These parts are labeled in Fig. 7.2. It was built on a 32mil FR4 substrate with $\epsilon_r = 4.4$ and $\tan(\delta) = 0.02$. It is a two layer structure, where the radiating structure and the CPS line belong to one layer and the microstrip series open stub and connecting transmission line belong to another layer.

7.1.2.1 Radiating Element Design

Design for the radiator was initially performed. To ensure coverage of the entire band, let the antenna have a wider bandwidth of 800MHz to 1GHz. The width of the lines used for this part was chosen to be 5mm.

Based on Fig. 7.2, the excited dipole has a larger length, and thus, it covers the lower part of the band of interest. It was bent in an attempt to miniaturize the antenna and was initially assigned an overall length of $\lambda/2$ at 800MHz. A tapered feed was implemented to reduce capacitance. By using IE3D, its dimensions, including the angle used to taper the feed, were optimized for low VSWR values from 800MHz to 900MHz.

After which, the sleeve or parasitic element was added to the excited dipole to enhance the bandwidth. Initially, its length, which was shorter than that of the dipole, was equivalent to $\lambda/2$ at 1GHz. As a result, the reflection coefficient of the higher frequency part of the band
is lowered after optimization of its length and its distance from the excited dipole performed using IE3D. The optimized structure and its dimensions are shown in Fig. 7.3.

Figure 7.3: 912MHz antenna with a bandwidth of 800MHz to 1GHz and is composed of the excited dipole and a parasitic sleeve. The lengths, width (both in mm) and the angle are as labeled.

However, it is worth noting in Fig. 7.3 that the optimization results produce a dipole that is not symmetric as the lengths $EL_1$ and $EL_2$ are not equal, wherein the former is 4.04mm shorter than the latter. A current distribution simulation was performed to examine if the non-symmetrical structure affects this parameter greatly. As seen in Fig. 7.4, the current distribution is almost symmetrical between these two arms despite this anomaly. Furthermore, it is observed that it is non-uniform as the maximum current magnitude occurs near the feed and decreases in value in sinusoidal fashion towards the outermost parts of the radiator. This is typical for finite length dipoles [10].
Figure 7.4: Antenna current distribution normalized by a value of 18.823A/m

It is also observed in Fig. 7.4(a) that at 0.8GHz, the current is more concentrated within the excited dipole structure as it is designed with a longer length to account for the lower
frequency band. At 0.9GHz, it is shown in Fig. 7.4(b) that there is a slight increase in current within the parasitic sleeve compared to the previous frequency. Finally, at 1GHz, Fig. 7.4(c) shows that there is a higher concentration of current within the sleeve. Therefore, the antenna is capable of operating at the frequency band of 800MHz to 1GHz and that at higher frequencies, the structure still functions mainly due to the parasitic sleeve, whose length was designed to be shorter for this reason.

![Antenna Reflection Coefficient](image)

**Figure 7.5:** Antenna reflection coefficient simulation results for the dipole without a sleeve and for the dipole with a sleeve, as obtained by using IE3D.

The reflection coefficient simulation result comparison between the antenna that does not include a sleeve and one that includes a sleeve is presented in Fig. 7.5. It is shown that the one without a sleeve has narrower bandwidth compared to the one that uses a
parasitic element. This has $S_{11}$ values below -9dB from 800MHz to 900MHz. The antenna that includes a sleeve has larger bandwidth. Its highest $S_{11}$ value is -7.35dB, located at 800MHz, and it maintains a low value even up to 1GHz. Now that it has been proven that bandwidth requirement for operation has been met, the antenna feed is then designed.

7.1.2.2 Balun and Matching Network Design

Printed Marchand and double y baluns in [14] that provide a wideband transition between balanced and unbalanced transmission lines were presented. The balun was simplified and implemented as a matching network in [15], wherein it was integrated within a dual-band antenna that was successfully matched to 50Ω. It provided an excellent transition between the balanced CPS line, which was directly connected to the dipole, to a microstrip transmission line, which was unbalanced and can be readily connected to a coaxial SMA connector. The design guidelines in [15] were followed for a good transition between CPS and microstrip lines while maintaining the 10% fractional bandwidth requirement.
Figure 7.6: Initial circuit design for the balun, which also serves as a matching network. The 1:1 transformer represents the coupling of the CPS structure to the microstrip line.

(a) Without any transmission line (b) After the CPS line and before CPS to microstrip line transition (c) After the series open microstrip shunt stub and before microstrip connecting transmission line

Figure 7.7: Antenna input impedance in Smith chart form at the indicated references, as seen in the network shown in Fig. 7.6

Fig. 7.6 shows the initial circuit, which functions as a balun and a matching network. Prior to balun implementation, the input impedance of the antenna is shown in Fig. 7.7(a).
It shows that after the antenna, a balanced CPS line is added, and is labeled in Fig. 7.2. The width of each strip and the gap between the strips were arbitrarily chosen to be 10mm and 0.5mm, respectively. The wider width was chosen as it also served as a ground plane for the microstrip lines to be used later.

![Characteristics of CPS line](image.png)

**Figure 7.8**: Characteristic impedance and effective permittivity of a CPS line with a strip width of 10mm and a gap of 0.5mm between the strips.

In [16], CPS lines are dispersive. To investigate the changes in its effective permittivity ($\epsilon_{re}$) and characteristic impedance ($Z_0$) over the desired bandwidth, an IE3D simulation was performed. Fig. 7.8 shows the variations of these two parameters wherein $\epsilon_{re}$ increased and $Z_0$ decreased but not drastically. For initial design, let $\epsilon_{re} = 1.78$ and $Z_0 = 91.2\Omega$. The CPS line length used is $l = 25mm$. Since the quasi-TEM (transverse electromagnetic) mode is excited [17], the propagation constant is shown in eq. 7.1 [8], where $c = 3 \times 10^8m/s$ which is the speed of light in free space and $f$ is the frequency. To simplify analysis, let there be no attenuation, $\alpha$.
\[ \beta = \frac{2\pi f\sqrt{\epsilon_{re}}}{c} \quad (7.1) \]

As a result, the impedance curve found in Fig. 7.7(a), which serves as the load, \( Z_L \), is transformed and is moved clockwise, as shown in Fig. 7.7(b) by using the transmission line impedance equation [8] in eq. 7.2.

\[ Z_{in} = Z_0 \frac{Z_L + jZ_0\tan(\beta l)}{Z_0 + jZ_L\tan(\beta l)} \quad (7.2) \]

After which, the gap between the CPS line is coupled to the microstrip line structure, as indicated in Fig. 7.2. This is represented as a 1:1 transformer shown in Fig. 7.6. To lower the impedance curve shown in Fig. 7.7(b), a series open stub is added whose \( Z_0 = 50\Omega \). This is equivalent to adding an impedance of \( Z = -jZ_0\cot(\beta l) \), where the \( l = 25.898\, mm \) and \( \epsilon_{re} = 3.33 \), wherein \( \epsilon_{re} \) was derived from equations found in [8]. Eq. 7.1 was used to determine \( \beta \). As a result, the impedance was transformed from Fig. 7.7(b) to the one shown in Fig. 7.7(c). The microstrip line connected after the series open shunt stub, as shown in Fig. 7.6, by which an SMA connector is attached thereafter, has \( Z_0 = 50\Omega \). The VSWR after this feed line is maintained whatever the length of this transmission line may be.

The feed was optimized using IE3D to improve the reflection coefficient within the band. As a result, Table 7.1 lists the relevant component values of the antenna feed prior to and after optimization.
Table 7.1: Dimensions of the components used in the antenna feed, based on Figs. 7.2, 7.6

<table>
<thead>
<tr>
<th>Component</th>
<th>CPS</th>
<th>Microstrip Series Open Stub</th>
<th>Microstrip Feed</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial $Z_0(\Omega)$</td>
<td>91.2</td>
<td>50</td>
<td>50</td>
</tr>
<tr>
<td>Initial Width (mm)</td>
<td>10 per strip</td>
<td>1.544</td>
<td>1.544</td>
</tr>
<tr>
<td>Initial Gap (mm)</td>
<td>0.5</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Initial Length (mm)</td>
<td>25</td>
<td>25.898</td>
<td>any</td>
</tr>
<tr>
<td>Optimized $Z_0(\Omega)$</td>
<td>same as above</td>
<td>54.54</td>
<td>60.94</td>
</tr>
<tr>
<td>Optimized Width (mm)</td>
<td>same as above</td>
<td>1.335</td>
<td>1.092</td>
</tr>
<tr>
<td>Optimized Gap (mm)</td>
<td>same as above</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Optimized Length (mm)</td>
<td>same as above</td>
<td>27.52</td>
<td>30.55</td>
</tr>
</tbody>
</table>

As shown in Fig. 7.9, the simulation results after optimization show that the antenna has met the bandwidth requirement. Using the dimensions obtained, the structure was fabricated and its reflection coefficient was measured and is shown in Fig. 7.9. It also met the bandwidth requirement, with the highest reflection coefficient occurring at 0.866GHz, with a corresponding value of -9.35dB.
7.2 Matching Network for the 912MHz OFC Time Division Multiplexed Temperature Sensor

The 912MHz SAW sensor has a small resistance. However, the higher frequency of operation allows for a smaller and a more practical size for a matching network once implemented using microstrip lines. Thus, for the tag antenna, the same design for the transmitter and receiver antenna may be used, as long as a corresponding matching circuit is attached as an interface for the sensor. Simple stub matching is insufficient as a wider bandwidth should be maintained. In this section, an overview of the 912MHz SAW sensor, as well as techniques
used to match the entire 10% fractional bandwidth of the device to a 50Ω antenna will be presented.

### 7.2.1 912MHz OFC SAW Temperature Sensor Device Overview

![Figure 7.10: 912MHz OFC encoded SAW temperature sensor composed of two identical reflector banks but with the time delay between the wideband transducer (middle) and the reflectors (left and right) different for each bank.](image)

Fig. 7.10 shows the device layout for the temperature sensor. It has 10% fractional bandwidth centered at 912MHz. Similar to the 250MHz sensor, its input impedance is capacitive and non-uniform. Each reflector bank is placed to the left or to the right of a wideband transducer and is composed of five chips which have a uniform time length, $\tau_c$. The distance between the launching transducer and each bank is different, and as a result, a different time delay results for each bank. OFC is used for device identification where each chip response is orthogonal to the response of the other chips used. Unlike the 250MHz sensor, the chips are no longer contiguous in time due to TDM employed on the device that adds another degree of freedom for producing more codes [12]. Although TDM is employed, the magnitude response in frequency of the sensor should be the same as that of a sensor with the same
code but without TDM implementation as a time delay results only in a frequency phase change. As in the 250MHz sensor, the material used is lithium niobate \((\text{LiNbO}_3)\), which has a nonzero temperature coefficient \[2\] that causes it to expand and contract with changes in temperature, which makes the distance between electrodes change causing variations in the center frequency for each chip.

Figure 7.11: Measured 912MHz OFC encoded SAW temperature sensor time response which shows five chips for each reflector bank. Chip no. 2 for each bank is heavily attenuated, as highlighted above.

After device fabrication and packaging, the sensor time response was measured and the result is shown in Fig. 7.11. Although each bank was designed to contain five chips, it is noted that the second chip is heavily attenuated for both cases.
7.2.2 Matching Network for the 912MHz OFC SAW Temperature Sensor

Lumped and transmission line elements were both initially used to simplify the derivation of the matching network. After its completion, all lumped elements are converted to their respective microstrip line equivalent for easier circuit fabrication. For this section, all distances expressed in terms of wavelength, \( \lambda \), are based on the center frequency, \( 912 MHz \).

7.2.2.1 Matching Network using Lumped and Microstrip Line Elements

![Network diagram](image)

Figure 7.12: Network used in matching the 912MHz SAW device to the 50Ω antenna, while maintaining at least 10% fractional bandwidth.

A combination of lumped elements and microstrip lines was initially used to design a matching network suited to conjugate match the device within the band of interest with an antenna that was designed for 50Ω input impedance, which has its design details presented in Sec. 7.1.2. The steps below were made to achieve the network shown in Fig. 7.12. It is built on Rogers RO4003C substrate, which has \( \epsilon_r = 3.55 \).
The input impedance of the 912MHz SAW device was measured and plotted in Smith chart form as seen in Fig. 7.13. It indicates that it is capacitive and has a very low input resistance throughout its band.

Figure 7.14: 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting an inductor for resonance at center frequency.
An inductor was connected in series with the SAW device for center frequency resonance. Its value was calculated and shown below whereas its input impedance after this modification is presented in Fig. 7.14.

\[ L = \frac{-j\text{Im}(Z_{\text{SAW}}(f = 912 \text{MHz}))}{j2\pi 912 \text{MHz}} = 1.96 \text{nH} \] (7.3)

Figure 7.15: 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting a \(\lambda/4\) impedance transformer with \(Z_0 = 12.367 \Omega\) to transform the center frequency impedance of Fig. 7.14 to \(Z = 36.274 \Omega\)

As stated in Wheeler’s Method for wideband matching in [6], it is impossible to perfectly match over the entire band of interest. Instead of directly matching the device to the generator resistance, which is the antenna input impedance of \(R_G = 50 \Omega\), the center frequency impedance, \(R_{\text{SAW}}\), is transformed instead to \(R_{\text{EB}}\) that is calculated from the required SWR and bandwidth values. Let matching occur within \(B = 15\%\) fractional bandwidth based on an SWR of 2.5. The quality factor is then calculated by using eq. 7.4.
\[ Q = \frac{\text{SWR} - 1}{\sqrt{\text{SWR}}} \frac{1}{B} \]  

(7.4)

The ratio between the generator resistance, \( R_G \), and \( R_{EB} \) is calculated using eq. 7.5.

\[ \frac{R_G}{R_{EB}} = \sqrt{1 + Q^2 B^2} \]  

(7.5)

Based on the requirements, the center frequency SAW input impedance needs to be transformed to \( R_{EB} = 36.274\Omega \). A \( \lambda/4 \) impedance transformer with a characteristic impedance of \( Z_0 = \sqrt{R_{EB} R_{SAW}} = 12.433\Omega \), [8], is used to achieve the result shown in Fig. 7.15.

Figure 7.16: 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after connecting a \( \lambda/4 \) line with \( Z_0 = 36.274\Omega \)

In order to increase the bandwidth by bringing the rest of the band closer to the center of the Smith chart, a resonator may be attached. Upon transformation of a series inductor-capacitor resonant circuit to its microstrip transmission line equivalent, a series shorted and
series open stub with the appropriate characteristic impedance results in resonance that is achieved at center frequency. However, these components are rather difficult to implement and are avoided. It is desired to rotate the response in the Fig. 7.15 by 180° while maintaining the same center frequency impedance value. This prepares the curve such that a parallel inductor-capacitor resonant circuit may be used. It follows that its equivalent, the shunt shorted and open stubs that are easy to realize are also preferred. The rotated curve is shown in Fig. 7.16 and a λ/4 transmission line with $Z_0 = 36.274\Omega$ is used to obtain this response.

![Reflection Coefficient](Image)

Figure 7.17: Input impedance of the parallel inductor-capacitor resonator (thicker line) only along with the 912MHz OFC encoded SAW temperature sensor input impedance or reflection coefficient in Smith chart form after completion of the matching circuit.

The value of the capacitor was determined by first calculating the value of $Q_2$ using eq. 7.6, [6].
\[ Q_2 = \frac{Q}{\sqrt{1 + Q^2 B^2}} \]  

(7.6)

\[ Q_2 \] is the ratio of the susceptance, \( 2\pi 912 \text{MHz} C_1 \), to the conductance, \( 1/R_G \). The capacitance required for the resonator is as follows.

\[ C_1 = \frac{Q_2}{R_G 2\pi 912 \text{MHz}} = 16.014 \text{pF} \]  

(7.7)

To complete the resonator, the inductance must satisfy the following equation.

\[ j2\pi f C_1 + \frac{1}{j2\pi f L_1} = 0 \]  

(7.8)

Thus, \( L_1 = 1.902 \text{nH} \). The response of the parallel resonant circuit is shown in Fig. 7.17. Its inclusion, as shown in Fig. 7.12, successfully matches the SAW device to the 50Ω antenna, as seen in Fig. 7.17.

### 7.2.2.2 Matching Network using Microstrip Elements

Transformation of lumped elements to their respective microstrip line counterpart is necessary for easier fabrication. The elements that need to be transformed are the series 1.96nH and the parallel resonant circuit composed of 1.90nH and 16.01pF.
Figure 7.18: Modification of the network shown in Fig. 7.12, wherein the inductor is replaced with a series short stub.

To convert the inductor with a value of 1.96nH, as seen in Fig. 7.12, Richard’s Transformation [8] is used. An inductor in transmission line form is equivalent to a shorted stub with an impedance of $j\omega L = jZ_0\tan(\beta l)$. As $\beta = 2\pi/\lambda$ and by setting $l = \lambda/8$, $\tan(\beta l) = \tan(\pi/4) = 1$. Thus, the characteristic impedance required is $Z_0 = \omega L = 2\pi (912\, MH\)z) (1.96nH) = 11.219\Omega$. The circuit is modified, as shown in Fig. 7.18.

Figure 7.19: Kuroda’s identity used in transforming a series shorted stub ($Z_1$) into a shunt open stub. The transmission line elements are both $\lambda/8$ long.

However, it is desired to convert the series short stub into an open shunt stub in order to implement the circuit by using Kuroda’s identity [8]. The series short stub with a characteristic impedance of $Z_1$ and length $\lambda/8$, followed by a transmission line section with
a characteristic impedance of $Z_2$ and length $\lambda/8$ are shown in Fig. 7.19. Its conversion is shown in Fig. 7.19, where $n$ is as follows.

$$n = \sqrt{1 + \frac{Z_2}{Z_1}} \quad (7.9)$$

Figure 7.20: Modification of the network shown in Fig. 7.18, wherein the series short stub has been replaced with an open shunt stub after Kuroda’s transformation shown in Fig. 7.19.

The impedance transformer with $Z_0 = 12.367\Omega$ is divided into two. The circuit is transformed into the one shown in Fig. 7.20 by using Kuroda’s transformation discussed above. It is desired to modify the open shunt stub to make it narrower since it has a resulting width of 4.599mm. This is done by replacing its characteristic impedance into $Z_0 = 33\Omega$. Its length is $0.144\lambda = 27.663\, \text{mm}$.

Finally, the parallel resonator is converted, where the shunt capacitor is equivalent to an impedance of $1/j\omega C = -jZ_0 \cot(\beta l)$. Initially setting $l = \lambda/8$, the characteristic impedance for both shunt open and short stubs is $10.897\Omega$, which gives a microstrip line width of 12.976mm. It is desired to narrow the width of the line to 3.813mm, wherein $Z_0 = 30\Omega$. 80
This results in lengths of \(0.055\lambda = 10.611\, \text{mm}\) and \(0.195\lambda = 37.228\, \text{mm}\) for the shunt short and open stubs respectively. The final circuit is shown in Fig. 7.21.

![Diagram of matching network](image.png)

Figure 7.21: Final matching network that serves to transform the device to 50\(\Omega\) over its 10\% fractional bandwidth.

![Reflection Coefficient Graph](image.png)

Figure 7.22: The calculated reflection coefficient of the SAW device after attaching the matching network looking into the reference shown in Fig. 7.21.

The theoretical model of the matching circuit has met the requirements as it shows good matching performance over the desired band, as seen Fig. 7.22. Prior to circuit fabrication, its lengths and widths were optimized using Momentum ADS. The layout used in fabricating the circuit, along with its dimensions, are shown in Fig. 7.23.
After fabricating the antenna, measurements of $Z_{\text{ant}}$ and $Z_{\text{saw}}$ were made. The references used are shown in Fig. 7.24. Based on the measured data, the reflection coefficient was
calculated using eq. 6.2, [9]. The reflection coefficient plot is shown in Fig. 7.25, which indicates successful impedance matching over the band desired.

![Reflection Coefficient](image)

Figure 7.25: Measured reflection coefficient due to conjugate matching between the SAW sensor and the 50Ω antenna, as illustrated in Fig. 7.24.

7.3 Antenna Gain

![Antenna Gain](image)

Figure 7.26: Measured antenna gain frequency sweep for antenna separation values of 0.7m and 0.85m.
The gain measurement was performed in the laboratory without anechoic foam, resulting in a less reliable gain measurement due to reflections. They were not performed in the anechoic chamber in the absence of a needed reference antenna that was fully characterized at the frequency band desired. Instead, two identical antennas were mounted at bore sight wherein the sheets lay parallel to the ground. To determine the largest dimension of the antenna structure, the largest x and y dimensions were recorded. The largest x and y values were equivalent to the length of the sleeve (116.31mm) and the distance between the outermost edge of the sleeve to the feed location (50.75mm), respectively. It was calculated as \( \sqrt{50.75\text{mm}^2 + 116.31\text{mm}^2} = 126.9\text{mm} \) and is equivalent to 0.405\( \lambda \) at 957MHz. The distance, d, between the antennas is 0.7m for the first trial and 0.85m for the second trial. They are taken at the far-field since both are greater than 2\( \lambda \). The antenna sleeve is directly opposite to the other antenna for the maximum gain measurement as the sleeve is likened to a director in a Yagi-Uda array [10] where its direction points to where the maximum directivity is. The absolute gain, G, was determined using the two-antenna method in [10] and the Friis’ Equation, in eq. 7.10, wherein \( P_r/P_t \) is the measured ratio between the power received and the power transmitted.

\[
\frac{P_r}{P_t} = \left( \frac{\lambda}{4\pi d} \right)^2 G^2 \tag{7.10}
\]

The results are shown in Fig. 7.26, wherein the gain varies between 1dB to 3dB across the band of interest.
7.4 Antenna Radiation Pattern

The fabricated antenna has its orientation in 3D Cartesian space shown in Fig. 7.27, where the $\theta$ and $\phi$ axes are as labeled. The anechoic chamber facility available is suited to work at the frequency of operation of the antenna and radiation pattern measurements were made there at the frequency of 866MHz, 912MHz and 957MHz. They are all normalized to the highest value for gain.
Figure 7.28: Measured normalized radiation pattern (dB) of the E-Plane for 866MHz, 912MHz and 957MHz.

Referring to Fig. 7.27, the E-Plane measurement was taken where $\phi = 0^0$ and $\theta$ was varied. Based on the angle labeling in Fig. 7.28, it can be noticed that the gain is higher for $\theta = 0^0$ compared to $\theta = 180^0$ by approximately 4dB for all frequencies, whereas for a typical dipole, these values should have been the same [10]. This is due to the parasitic sleeve used to enhance the bandwidth. This case can be likened to a Yagi-Uda antenna array wherein parasitic directors are placed in front of an excited dipole and a few parasitic reflectors are located at its back. The directors are shorter in length compared to the excited dipole and are placed in the direction of high directivity [10]. As shown in the current distribution in Fig. 7.4, current parallel to the y and z axes exist (axes based on the one shown in Fig. 7.27); polarization purity was investigated by measuring antenna cross-polarization levels for all $\theta$. Ideally, since a dipole has linear polarization [10], the cross-polarization level should be equal to zero. Measurements in Fig. 7.28 indicate that cross-polarization is at least
approximately 33dB below co-polarization levels. Hence, the antennas fabricated indeed have linear polarization.

Figure 7.29: Measured normalized radiation pattern (dB) of the H-Plane for 866MHz, 912MHz and 957MHz.

The H-Plane measurement was taken where $\theta = 90^0$ and $\phi$ was varied, based on Fig. 7.27. Ideally, a dipole has uniform gain for all values of $\phi$ [10]. However, this is not the case based on the measurements shown in Fig. 7.29, where it indicates that at $\phi = 90^0$, the maximum gain occurs. For the worst case scenario, the lowest level of co-polarization for the 957MHz case is approximately 7.44dB below the maximum. This non-uniformity can be attributed to the parasitic sleeve, as explained in the E-plane case above. Similar to the E-plane measurement, the cross-polarization level is at least 34.2dB lower than the co-polarization levels, indicating that the antenna is linearly polarized.
To confirm the performance of the antennas designed above for the system, the wirelessly transmitted response of the SAW sensor was recorded and compared to the time response obtained by directly connecting the device to the VNA. The first port of the VNA was placed in series with an amplifier with its gain and input 1dB compression point characterized as 10dB and 11dBm, respectively. 10dBm/Hz of power was fed unto the amplifier, resulting in 20dBm/Hz of power input to the transmitting antenna. The SAW device was connected to another antenna with its appropriate matching network. The wideband chirp that the SAW tag antenna captures from the transmitter enables the device transducer to launch waves that are then reflected and encoded by the reflector bank response. The encoded signal is relaunched from the SAW tag antenna and is obtained by the receiver connected to the second port of the VNA. The time response wirelessly transmitted can be observed by
measuring the transmission parameter, \( S_{21} \), with an averaging performed over 30 iterations. This setup is shown in Fig. 7.30. They were arranged collinearly, with a separation of distance, \( d \), between any two adjacent antennas. Three trials were made, wherein \( d = 0.5\text{m}, 0.7\text{m} \) and 1m.

At a distance of \( d = 0.5\text{m} \), Fig. 7.31(a) shows that the entire time response was successfully transmitted. Propagation loss is increased for \( d = 0.7\text{m} \). According to Fig. 7.31(b), a few of the chips are missing from bank 2 due to its higher loss. For the wired response, compared to bank 1, its corresponding chip on bank 2 has lower magnitude, indicating that the latter is more lossy. Finally, for \( d = 1\text{m} \), some of the chips from both banks are gone due to higher propagation loss that comes as a result of increased distance between antennas. This problem may be alleviated by increasing the power emitted by the transmitter. Unfortunately, amplifiers with higher gain and 1dB compression point were not available at the time of experiment.
Figure 7.31: Measured time domain response when the SAW device was directly connected to the VNA, compared to the SAW device response time domain response obtained wirelessly using the setup found in Fig. 7.30, where the distance between any two adjacent antennas, d, was varied.
7.6 Antenna Range Prediction

\[
R_{\text{MAX}} = \left[\frac{G^4 \lambda^4 G_{\text{amp}} \text{Loss}_{\text{SAW}} P_t P_G}{(4\pi)^4 P_r}\right]^{1/4}
\]  

(7.11)

The distance required for wireless transmission of signals using the antennas at a frequency of 912MHz can be obtained using eq. 7.11, which was derived using Friis’ transmission equation [10]. It is suited for use with the setup shown in Fig. 7.30 where the transmitting antenna emits the chirp to be received by the SAW tag. At the transmitter, an amplifier of \(G_{\text{amp}} = 20\)dB gain is attached. Due to the SAW sensor reflective response, power is relaunched with the sensor time domain response encoded. The signal is then captured by the receiving antenna. For operation, it was assumed that the receiver requires 10\(\mu\)W of power and that the transmitter emits 1W of power at 912MHz. All antennas were successfully matched and have a measured gain of \(G = 2.73dB\) at center frequency. The device is lossy as it is not capable of reflecting all incident power, as indicated by the parameter \(\text{Loss}_{\text{SAW}}\). It cannot be extracted currently, so the distance required for power transmission is dependent on \(\text{Loss}_{\text{SAW}}\), as indicated in Fig. 7.32. Assuming an ideal and lossless device, the distance for operation is 2.76m. A distance of 0.5m is required for -30dB device loss. All of these assume no processing gain is used. However, due to OFC coding, the device has a processing gain of \(P_G = 25\) [12]. This enhances the range as a result, as shown in Fig. 7.32.
Figure 7.32: Prediction of range of operation as a function of device loss.
CHAPTER 8
CONCLUSION

The thesis includes discussions on SAW devices that use reflective transducers. It is composed of a launching transducer cascaded with a SAW reflective transducer (SRT). The magnitude of the reflectivity of the latter is dependent upon the load attached to the SRT, which can be inductive, capacitive or resistive. Amplitude variations of the SRT reflectivity were reported previously in [1] but variations in the frequency wherein peak reflectivity occurs due to the load attached were not explored previously. This is only made possible by lengthening the SRT such that its susceptance at some frequencies is equal to zero or below zero. By varying the inductive or capacitive load attached, the resonant frequency is changed. Trends as to how the peak frequency changes according to the load attached were presented and were based on the FIR model.

The scope of this thesis includes antennas built for the 250MHz OFC temperature acquisition system, which requires 28% fractional bandwidth. At the transmitter, a wideband chirp is emitted. The SAW sensor time response is transmitted after reception of the wideband chirp and is dependent upon the ambient temperature. Finally, as the signal is relaunched from the SAW tag antenna and arrives at the receiver, an adaptive matched filter is used to acquire temperature based on the received SAW tag time response.
The system utilizes the disk monopole to satisfy the wide bandwidth requirement and matching to the 50\(\Omega\) input impedance for the transmitter and receiver circuits. After constructing the antennas, it was shown that the matching requirement was not satisfied. Thus, Wheeler's technique to acquire wideband matching was used and presented. As the input impedance of the OFC SAW sensor varies with frequency, has low input resistance and is capacitive, similar techniques to the previous case were also used such that conjugate matching between the device and the antenna was achieved. Lumped elements were used for the matching circuit. Finally, after antenna installation, the 250MHz OFC temperature acquisition prototype is capable of measuring ambient temperature wirelessly and these results were presented.

For the 912MHz temperature acquisition system, the SAW sensor employs both OFC and TDM coding techniques. It requires the same specifications with the exception of a lower 10\% fractional bandwidth. For this case, in an attempt to miniaturize and make the antennas compact, a folded dipole with a balun was used, which also utilizes a sleeve for bandwidth enhancement. The same techniques previously presented for conjugately matching the SAW sensor to the antenna were employed. Methods used in transforming lumped elements to transmission lines, and to transform series elements to shunt elements were used. To test antenna performance, measurements of the gain, radiation pattern and time domain response of the wirelessly transmitted signal due to the SAW tag were recorded.
LIST OF REFERENCES


