Design And Simulation For Encoded Pn-ofc Saw Sensor Systems

John Pavlina

University of Central Florida

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DESIGN AND SIMULATION
FOR ENCODED PN-OFC SAW SENSOR SYSTEMS

by

JOHN M. PAVLINA
B.S. University of Central Florida, 2004
M.S. University of Central Florida, 2007

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Major Professor: Donald C. Malocha
ABSTRACT

Surface acoustic wave (SAW) sensors provide versatility in that they can offer wireless, passive operation in numerous environments. Various SAW device embodiments may also be employed for retrieval of the sensed data. Single sensor systems typically use a single carrier frequency and a simple device embodiment since tagging is not required. However, it is necessary in a multi-sensor environment to both identify the sensor and retrieve the information. Overlapping sensor data signals in time and frequency present problems when attempting to collect the sensed data at the receiver. This dissertation defines a system simulation environment exclusive to SAW sensors. The major parameters associated with a multi-device system include the transmitter, the channel, and the receiver characteristics. These characteristics are studied for implementation into the simulation environment. A coupling of modes (COM) model for SAW devices is utilized as an accurate software representation of the various SAW devices. Measured device results are presented and compared with COM model predictions to verify performance of devices and system. Several coding techniques to alleviate code collisions and detection errors were investigated and evaluated. These specialized techniques apply the use of time, frequency, and spatial diversity to the devices. Utilizing these multiple-access techniques a multi-device system is realized. An optimal system based on coding technique, frequency of operation, range, and related parameters is presented.

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CHAPTER 1: INTRODUCTION

Surface acoustic wave (SAW) sensors possess many unique advantages over possible competing technologies. Attractive characteristics include the following: passive, radiation hard, small, rugged, inexpensive, identifiable, and operable over a wide temperature range. Current embodiments include resonators, delay lines, differential delay lines, and devices with multiple reflective structures. SAW sensors are able to measure physical, chemical, and biological parameters [1, 2]. However, few SAW sensor embodiments are capable of operation in multi-sensor environments because the sensor must transmit identification and sensed information simultaneously [3-6]. Most examples show less than four being utilized simultaneously.

Orthogonal frequency coding (OFC) has been proven a viable method for coding SAW tags [7]. Orthogonality concepts have been in use for multiplexing applications for some time [8]. Generally, there is a required relationship between the basis set frequencies and their bandwidths that meets the orthogonality condition. For instance, if a time signal is broken into a number of finite-length serial chips, the local adjacent chip frequencies are contiguous and orthogonal, and the signal has a linear group delay, then a stepped chirp response is obtained [9]. The use of spread spectrum and chirp interrogation signals for SAW sensor applications has been previously reported [10, 11].

This dissertation is the study, design, and implementation of orthogonal frequency coded surface acoustic wave sensors for use in a multiple sensor environment. The challenges presented are in the form of designing of a SAW sensor simulator, developing a usable coding type, and describing an optimal system setup for SAW sensors.
Chapter 2 will discuss basic communication theory including basic signal concepts, spread spectrum signals, and orthogonality. The advantages inherent within a spread spectrum signal are presented as well as common uses. Orthogonal frequencies will be defined with a brief look into orthogonal frequency division multiplexing. A short introduction into SAW devices and the implementation of OFC SAW will also be discussed.

In chapter 3, a detailed look at a sensor-level simulator will be presented. The parameters of interest discussed are the transmitter, channel characteristics, target, and the receiver output. The simulations utilize a SAW device coupling of modes (COM) model and combined with RF system parameters. A simulation tool for the overall SAW system is utilized when performing analysis on important parameters, such as signal to noise ratio (SNR), SAW coding type, and range effects. System evaluation of a multi-frequency SAW sensor system utilizing multiple targets will be discussed.

Various coding approaches will be considered in chapter 4. The coding of a system with decreased collisions is of primary interest. The techniques presented are used in an attempt to create a working multiple sensor code set. Simple correlation techniques, spread spectrum techniques, and multiplexing schemes will be reviewed. The use of multiplexing in frequency and two types of time division multiplexing will be addressed.

The discussion and implementation of an optimized SAW sensor system is discussed in chapter 5. The design consideration affecting the range of the system will be addressed. The target will be considered as to how the antenna and device parameters affect the system. Different decoding techniques will be addressed for the extraction of the sensed information. Ultimately, a set of workable design parameters will be established.
CHAPTER 2: BACKGROUND

Early work on spread spectrum (SS) systems was done for military applications such as RADAR and a method for secure communication systems [12]. The design of communication systems has recently given rise to the use of spread spectrum technology in new ways [5, 13]. The current rise in the design of sensor systems has allowed a reexamination of the use of spread spectrum for wireless sensor devices [3, 7, 10, 14, 15]. The basic communication theory and spread spectrum signals pertaining to this thesis will be discussed in this chapter.

Communication Theory and Spread Spectrum Signals

The use of spread spectrum in its basic form has many inherent benefits but when used in conjunction with PSK, FSK or some other type of modulation scheme it can become even more powerful [9, 16]. The transmission of information by the use of spread spectrum has been utilized for many years. Spread spectrum systems encompass communications, data transmission, message privacy, signal hiding, and position location [17]. Their ability to reject unwanted signals including jamming and selectively addressing specific receivers has given them a broad range of uses.

Basic Signal Transmission Concepts

![Diagram](image)

Figure 2-1: A basic communication system block diagram consisting of an input source that is sent to a transmitter and then through the channel to the receiver to be output at its destination.

A generic block diagram as seen in Figure 2-1 shows the basic components of a communication system. The signal transmitted at the source is defined as $x(t)$. This signal can be defined many
different ways based on its type; periodic, non-periodic, energy, power. This allows further classification as well as how the signal is to be handled.

The spectral density of a signal characterizes the distribution of the signal energy in the frequency domain. Common forms are energy spectral density and power spectral density defined as

\[
E_x = \int_{-\infty}^{\infty} |x(t)|^2 \, dt \tag{2.1}
\]

\[
P_x = \frac{1}{T_0} \int_{-T_0/2}^{T_0/2} |x(t)|^2 \, dt \tag{2.2}
\]

where \(T_0\) is the period [18]. Correlation is a matching process commonly used in communications. Autocorrelation refers the matching of a signal to a delayed version of itself. The autocorrelation of \(x(t)\) is defined as

\[
R_x(\tau) = \int_{-\infty}^{\infty} x(t)x(t + \tau) \, dt. \tag{2.3}
\]

This correlation process is key in the creation of a matched filter or integrator. In the matched filter process a signal is created as a time-flipped version of itself to be correlated against in order to observe how closely the signals resemble each other [19]. An example of the creation of a matched filter (MF) is shown in Figure 2-2.
Frequency Transform Overview

When transforming from time to frequency and vice versa, the Fourier transform is the commonly used method. This employs the known formula

\[ H(f) = \int_{-\infty}^{\infty} h(t) e^{-j2\pi ft} \, dt \]  \hspace{1cm} (2.4)

and its corollary

\[ h(t) = \int_{-\infty}^{\infty} H(f) e^{j2\pi ft} \, df. \]  \hspace{1cm} (2.5)

When working with finite time only some finite number of frequencies can be computed for \( H(f) \). For this reason the relationship between the samples of \( h(t) \) and \( H(f) \) is determined. The samples of \( h(t) \) are taken at uniform intervals \( \Delta t \); the number of samples over the interval \( T_{\text{max}} \) is

\[ N0 = \frac{T_{\text{max}}}{\Delta t}. \]  \hspace{1cm} (2.6)

Now,
\[ H(f) = \int_0^{T_{\text{max}}} h(t)e^{-j2\pi ft} dt \]

\[ = \lim_{\Delta t \to 0} \sum_{n=0}^{N_0-1} h(n\Delta t)e^{-j2\pi fn\Delta t} \Delta t. \]  

(2.7)

Consider the samples of \( H(f) \) as uniform intervals of \( \Delta f \). If \( H_k \) is the kth sample, that is, 
\[ H_k = H(k\Delta f), \]

then Equation (2.7) becomes

\[ H_k = \sum_{n=0}^{N_0-1} h(n\Delta t)e^{-j2\pi kn\Delta t} \Delta t \]

(2.8)

\[ = \sum_{n=0}^{N_0-1} h_n e^{-j2\pi nk/N_0}, \]

where

\[ h_n = \Delta t \cdot h(n\Delta t) \text{ and } N_0 = \frac{1}{\Delta t \cdot \Delta f}. \]  

(2.9)

The final form of (2.8) is the familiar form of the discrete Fourier transform (DFT) [20]. The standard form for the inverse DFT (IDFT) is

\[ h_n = \frac{1}{N_0} \sum_{k=0}^{N_0-1} H_k e^{j2\pi nk/N_0}. \]

(2.10)

Both sequences \( h_n \) and \( H_k \) are periodic with a period of \( N_0 \) samples. This results in \( h_n \) repeating every \( T_{\text{max}} \) seconds and \( H_k \) repeating with a period of \( F_{\text{max}} = f_{\text{samp}} = 1/\Delta t \) (the sampling frequency). The sampling interval of \( h_n \) is \( \Delta t \) and of \( H_k \) is \( \Delta f = 1/T_{\text{max}} \).
The DFT as applied in most computer applications (i.e. MATLAB) assumes that when \( x_n \) is put into the DFT operator, it has already been scaled by the appropriate factor; mainly, \( \Delta t \) [21]. Without taking into account this scaling factor, the transform, while correct in relative amplitude and phase, will have a scaling factor offset. It would in effect be as if the sampling interval was disregarded or divided out, yielding

\[
DFT(h(n\Delta t)) = \frac{DFT(h_n)}{\Delta t} = \frac{H_k}{\Delta t} = H_{sk}.
\] (2.11)

This would clearly yield incorrect values if attempting to obtain an exact picture of the frequency spectrum of the signal \( h(t) \). The DFT, if used with unscaled time data will yield a scaled frequency response of \( H_{sk} \). Similarly with the IDFT, if the real values \( h(n\Delta t) \) are desired, Equation (2.10) must take into account the sampling period. This is shown as

\[
h_n = h(n\Delta t) \cdot \Delta t
\] (2.12)

\[
h(n\Delta t) = \frac{h_n}{\Delta t} = \Delta f \sum_{k=0}^{N0-1} H_k e^{-j2\pi nk/N0}
\] (2.13)

which, if the limit is taken of the right side, yields the integral form of Equation (2.5).

Now it is known that Parseval’s theorem in integral form states

\[
\int_{-\infty}^{\infty} |h(t)|^2 dt = \int_{-\infty}^{\infty} |H(f)|^2 df
\] (2.14)

where \( H(f) \) is the Fourier transform of \( h(t) \), or in the form of a summation
If only a discrete set of \( N \) numbers are evaluated, it takes the form

\[
\sum_{n=0}^{N} |h(n\Delta t)|^2 \Delta t = \sum_{k=0}^{N} |H(k\Delta f)|^2 \Delta f.
\] (2.16)

For discrete signals, according to [22], Parseval’s theorem takes on the form

\[
\sum_{n=0}^{N} |h(n\Delta t)|^2 = \frac{1}{N} \sum_{k=0}^{N} |H(k\Delta f)|^2.
\] (2.17)

Recall from Equation (2.11) when using the DFT,

\[
DFT\{h(n\Delta t)\} = H_s(k\Delta f) = \frac{H(k\Delta f)}{\Delta t}
\] (2.18)

so Equation (2.17) is a different form of Equation (2.16) because the Fourier transform of \( h(n\Delta t) \neq H(k\Delta f) \), as such a substitution is in need in Equation (2.17)

\[
\sum_{n=0}^{N} |h(n\Delta t)|^2 = \frac{1}{N} \sum_{k=0}^{N} \left| \frac{H(k\Delta f)}{\Delta t} \right|^2,
\] (2.19)

This leads to the form of Parseval’s theorem for discrete signals Equations (2.15) and (2.16).

\[
\sum_{n=0}^{N} |h(n\Delta t)|^2 = \frac{\Delta t \cdot \Delta f}{\Delta t^2} \sum_{k=0}^{N} |H(k\Delta f)|^2
\] (2.20)

or
\[ \sum_{n=0}^{N} |h(n\Delta t)|^2 \Delta t = \sum_{k=0}^{N} |H(k\Delta f)|^2 \Delta f. \] (2.21)

This shows the equations are interchangeable given certain constants are understood.

Therefore as long as the intended use of the equation is understood within the programming language there should arise no significant difference between the discrete and continuous forms of the equations.

**Basic Spread Spectrum**

Spread spectrum (SS) is a communication model for sending digital information through a channel. It was developed by the military as a way to combat the jamming of signals [23]. SS signals are referred to in this way because the transmission bandwidth required to send the signal is larger than the data bandwidth. A spread spectrum system is defined by the following two aspects: the signal is occupied in a bandwidth much larger than the minimum bandwidth required in sending the data, and the signal spreading is accomplished by a spreading signal [17]. This code signal is independent of the data being transmitted. Despreading of the signal is accomplished using a correlator and a synchronous replica of the code signal used in the spreading process.
Figure 2-3: Basic spread spectrum technique diagram illustrating the convolution of the spreading code with the data to spread the bandwidth and recover the data even with unwanted signal interference.

Advantage of Using Common Bandwidth

In some applications it is advantageous to utilize the bandwidth in such a way that it occupies a large number of users. The advantage is using the whole bandwidth all the time while still maintaining good interactions between the users. The benefits of generic SS techniques are interference suppression, fine time resolution, low power transmission, and multiple-access [24].

Interference Combating

White Gaussian noise is the probabilistic channel model used for most of communications. It has infinite average power spread uniformly over all frequencies. When transmitting information this source of interference is in general combating with the desired signal. This noise is however limited by the detector. For a narrowband signal that means only the noise within the bandwidth contributes. Since SS was developed for its anti jamming capabilities it is fitting to start there.
When a spreading signal set exists, the likelihood of the jammer knowing the set is small. Since a jammer has finite energy a decision must be made: will he jam the entire frequency bandwidth of interest placing small amounts of power in each location, or will he jam a smaller frequency band with a larger amount of power in attempts to disrupt communications?

Figure 2-4: Frequency spreading of a standard spread spectrum system with noise (N0) and jammer (J0) signals. The advantage over standard AWGN noise is minimal, while the advantage over a jammer signal can be significant.

Figure 2-4 compares the effect of spreading on white noise and a jammer signal of the two types described above. Since the Gaussian noise is of infinite average power, the effect of spreading is nothing. The jammer has a fixed finite power, $P_j$, and a power spectral density $\text{PSD}_j$. When the signal is spread the jammer can make choice one (seen in the figure distributing its power over the entire SS BW)
or choice two only (over part of the spectrum place the power). If a poor choice is made on the placement of the power in choice two, then the average effect will be less than if it was a good choice.

Jamming is not always a malicious act but can be due to natural phenomena. It is also sometimes a result of self-interference caused by multipath, where delayed versions of the same signal bounce off of buildings in an outdoor environment or walls indoors, so they return at a delayed time from the direct signal [25].

**Fine time resolution**

Spread spectrum signals are often used for ranging applications or determination of position. Distance can be measured by accurately determining the time delay through the channel. Uncertainty in delay measurement is inversely proportional to the bandwidth of the signal pulse. This uncertainty is proportional to the rise time of the pulse which is inversely proportional to the bandwidth of the pulse signal $\Delta t = 1/BW$. The larger the bandwidth the better the precision of the time delay measurement. The implementation of a long chain polarity change code signaling pulses instead of a single can then be correlated against at the receiver. The result of the correlation can then be used for more accurate measurements.

**Low Probability of Intercept**

If one were able to send a signal that could not be detected by anyone but the intended receiver, then this would be extremely advantageous. These systems are referred to as low probability of detection (LPD) or low probability of intercept (LPI). The goal of these systems is to create an optimum spreading in order to reduce the amount of energy sent and necessary to receive. These
systems can be thought of as buried in the noise only coming up when the intended recipient with the appropriate spreading signal receives it and despreads it.

Multiple-Access

SS lends itself very easily to use as a multiple-access technique. The technique involving a unique spreading signal for each simultaneous user is referred to as code division multiple-access (CDMA). CDMA has inherent privacy benefits since each user is unaware of any others code signal. An unauthorized user cannot easily monitor the communication of the intended user.

Common Uses

A few popular techniques have emerged from the idea of spreading the information signal. For signals with bandwidth B and duration T, the dimensionality of the signal space is approximately 2BT. It is possible therefore to increase dimensionality by either increasing B with spectrum spreading or T by time spreading. There are two common types of spectrum spreading, direct sequence (DS), and frequency hopping (FH). Time spreading is referred to as time hopping (TH) the third jamming rejection technique. It is also possible to combine many of the techniques to create hybrids: DS/FH, FH/TH, and DS/FH/TH, for example.

Direct Sequence Spread Spectrum (DS/SS)

Direct sequence is the name given to the spectrum spreading technique whereby a carrier signal is first modulated by a data stream followed by a wideband spreading signal. Consider a data modulated carrier having the waveform given by

\[ h_x(t) = A \cos(\omega_0 t + \theta_x(t)) \]  (2.22)
where the data phase modulation is $\theta_x(t)$, the radian frequency is $\omega_0$, and $A$ is the amplitude. The signal is then further modulated by the spreading signal $g(t)$, and the transmitted waveform can now be expressed as

$$h(t) = A \cos[\omega_0 t + \theta_x(t) + \theta_g(t)]$$  \hspace{1cm} (2.23)

The phase of the signal now has two components: one due to the data, and one due to the spreading signal.

If the signal is modulated by a binary phase shift keying (BPSK) modulation it is known that the modulation can be expressed by a multiplication of the carrier signal by a $\pm 1$. If both the data and the spreading sequence modulation are BPSK signals the waveform can be written as

$$h(t) = A x(t) g(t) \cos(\omega_0 t).$$  \hspace{1cm} (2.24)

An example of DS/BPSK modulation and demodulation is shown in Figure 2-5. The figure shows the binary waveform $x(t)$, where the 1 corresponds to negative pulse value and a 0 corresponds to a positive value. An example of a binary spreading sequence $g(t)$ is shown in Figure 2-5. A modulo-2 addition of the pulse waveform and the spreading sequence and the equivalent waveform of the $x(t)g(t)$ product is shown in Figure 2-5.

For BPSK modulation the phase of the carrier is $\pi$ when the waveform product is $-1$. Similarly the phase is zero when the value of $x(t)g(t)$ is $+1$. The carrier phase seen in Figure 2-5 (d) allows one to appreciate the way the signal “hides” when using this technique. It is difficult to distinguish the slowly changing data signal $x(t)$ within the rapidly changing signal $x(t)g(t)$.
For demodulation of the signal a two step process is employed. Firstly the disspreading of the signal is accomplished be correlating the received signal with a synchronized replica of the spreading signal. The second step is accomplished using a traditional demodulator.

Figure 2-5: Direct sequence is the spectrum spreading technique whereby a carrier signal is first modulated by a data stream followed by a wideband spreading signal. Shown here is a basic depiction of a data signal, spreading signal, and transmitted signal composed of the combination of the two. The signal phase shows just how difficult it would be to detect the data from the transmitted signal.

Processing Gain

An interesting interpretation of the performance characteristics for the DS/SS signal is expressing the signal energy per bit $E_b$ in terms of the average power. The signal energy per bit may be expressed as

$$ E_b = P_{av} T_b = \frac{P_{av}}{R} \quad (2.25) $$
where $R$ is the information rate in bits/s. The power spectral density for a broadband jamming signal may be expressed as

$$J_0 = \frac{J_{av}}{W} \quad (2.26)$$

where $W$ is the spreading bandwidth. Using the above relations, the ratio $E_b/J_0$ may be expressed as

$$\frac{E_b}{J_0} = \frac{P_{av}}{R} = \frac{W/R}{J_{av}/P_{av}}. \quad (2.27)$$

The ratio $J_{av}/P_{av}$ is the jamming to signal power ratio, which is generally larger than unity. The ratio $W/R = T_b/T_c$ is the bandwidth expansion factor or equivalently the number of chips per information bit. This ratio is usually called the processing gain of the DS/SS system. It represents the advantage over the jammer gained by implementing DS/SS [26].

**CDMA**

Code division multiple-access (CDMA) uses the enabling of processing gain and coding gain obtained from DS/SS systems for the use of the same channel by multiple users. CDMA was implemented for wireless cellular communications [16]. The use of the same channel bandwidth by multiple users is made possible provided each has a distinct pseudo noise (PN) sequence. In the demodulation of each PN signal, the signals from other users appear as additive interference. This interference varies depending on the number of users at any given time. In CDMA a major advantage relies on the users transmitting messages only for a short period of time [16, 27].
Each of the $N$ users is given its own code, $g_i(t)$ where $i = 1, 2, ..., N$. The user codes are approximately orthogonal so that the cross-correlation of two codes is near zero. The data modulated signal is multiplied by the spreading code designated for group 1, $g_1(t)$. Simultaneously, groups 2 through $N$ are multiplied by their spreading code. The signal present at the receiver is the linear combination of the production from each of the users. This linear combination can be represented as

$$g_1(t)h_1(t) + g_2(t)h_2(t) + \cdots + g_N(t)h_N(t).$$  \hspace{1cm} (2.28)

If the receiver is assumed to be configured to receive messages from group 1 and the multiplied spreading code is assumed to be synchronous with the received signal, the output of the multiplier will produce the desired result

$$g_1^2(t)h_1(t)$$  \hspace{1cm} (2.29)

plus a combination of all the undesired signals

$$g_1(t)g_2(t)h_2(t) + g_1(t)g_3(t)h_3(t) + \cdots + g_1(t)g_N(t)h_N(t).$$  \hspace{1cm} (2.30)

If the code functions are chosen with orthogonal properties, the desired signal can be extracted perfectly in the absence of noise and the undesired signals can be easily rejected. In practical application the code functions are not perfectly orthogonal, thus introducing limitations on both performance and the number of simultaneous users [16].

**Time Hopping Spread Spectrum**

Time hopping is a spread spectrum technique in which the carrier is turned on and off by a pseudorandom code sequence. The original data rate $R$ is increased to create a longer time than would normally be necessary to send the message. During the longer time interval the data is sent in bursts
According to the dictates of the code sequence. Each symbol that is to be transmitted is represented by a sequence of time shifted pulses.

**Time Division Multiplexing/Multiple-Access**

Time division multiplexing is a useful way to share a common spectral resource with multiple users. Each of the $M$ signals or users is given full use of the spectrum for a short duration called a time slot. The unused time between the slots are referred to as guard times, and allow for a small time uncertainty between signals in adjacent time slots to reduce interference. TDM is commonly used in satellite communications systems [28]. The user’s data is transmitted sequentially within a given period and repeated, while the commutator on the receiver end demultiplexes the signal. This allows for a greater number of users using the same frequency band without increasing the bandwidth. A customary TDM implementation requires synchronization or a signaling bit in order to determine the beginning and the end of the periods.

![Figure 2-6: A basic depiction of time division multiplexing. Time slots utilize the same frequency spectrum and maintain guard bands between each slot to prevent interference.](image)
Another common technique used in spread spectrum is frequency hopping. Frequency hopping spread spectrum (FH/SS) transmits the signal by rapidly switching its carrier signal among many frequency channels. The switching is performed utilizing a pseudorandom sequence which is known to the transmitter and receiver.

Most commonly used with M-ary frequency shift keying (MFSK) [29]. The position of the M-ary signal set is shifted pseudo randomly by a frequency synthesizer over a hopping bandwidth. A depiction of a single user hopping scheme is seen in Figure 2-7.

![Diagram of frequency hopping spread spectrum](image)

*Figure 2-7: Frequency hopping spread spectrum uses a pseudo noise code to jump quickly from frequency to frequency. Each user may jump to a different frequency at a different time to accommodate multiple-access within the same frequency and time bands.*
**FDMA Frequency Division Multiple-Access**

FHSS is a type of frequency division. Like all multiple-access techniques, the goal is to share a common spectral resource among a large number of users. Frequency division multiplexing (FDM) or multiple-access (FDMA) specifies sub bands of frequency to each user. The assignment in this way is long term or permanent. Like time division frequency division makes use of buffer zones to ensure the spectrum intervals don’t overlap as depicted in Figure 2-8.

![Figure 2-8: The basic description of a frequency division multiplexing scheme. Each potential user is given a frequency band to use for the duration of time. The use of guard bands prevents interference between the users.](image)

**Orthogonality**

It is often helpful to think of signal waveforms in the form of vectors. An N-dimensional orthogonal space is defined as a space characterized by a set of N linearly independent functions, $f_n(t)$, called basis functions. Any arbitrary function in the space can be generated using these functions. The basis functions must satisfy the condition

$$\int_{-\infty}^{\infty} f_n(t) f_m(t) \, dt = \begin{cases} 0 & (m \neq n) \\ 1 & (m = n) \end{cases}$$

(2.31)
The principle of orthogonality can be stated as follows. Each \( f_n(t) \) function of the set of basis functions must be independent of the other members of the set. Each \( f_n(t) \) must not interfere with any other members of the set in the detection process. One reason for the focus on orthogonal signal space is that any arbitrary signal can be expressed as a linear combination of \( N \) orthogonal waveforms.

In Figure 2-9 an example of an antipodal signal, which is an orthogonal set with \( N=2 \). This orthogonal signal set is made up of pulse waveforms described by

\[
h_1(t) = p(t) \quad 0 \leq t \leq T \tag{2.32}
\]

and

\[
h_2(t) = p\left(t - \frac{T}{2}\right) \quad 0 \leq t \leq T \tag{2.33}
\]

where \( p(t) \) is a pulse with a duration of \( \tau = T/2 \), and \( T \) is the symbol duration. In general a set of equal-energy signals \( h_i(t) \), where \( i = 1,2,\ldots,M \), constitutes a orthonormal (orthogonal, normalized to unity) set if and only if

\[
z_{ij} = \frac{1}{E} \int_0^T h_i(t)h_j(t) \, dt = \begin{cases} 1 & \text{for } i = j \\ 0 & \text{otherwise} \end{cases} \tag{2.34}
\]

where \( z_{ij} \) is called the cross-correlation coefficient, and where \( E \) is the signal energy expressed as

\[
E = \int_0^T s_i^2(t) \, dt. \tag{2.35}
\]
Figure 2-9: Antipodal signal representative of an orthogonal waveform. The two signals occupy different time durations within the period. This allows orthogonal definition of the cross-correlation coefficient to be achieved.

The transforming of a waveform set representing the message into an improved waveform set is referred to as waveform coding. The most popular of such waveform codes are referred to as orthogonal and bi-orthogonal codes. The encoding procedure endeavors to make the coded signals as unlike as possible. The goal is to make the cross-correlation coefficient among all pairs of signals as small as possible. The smallest value for the cross-correlation is $z_{ij} = -1$ which only occurs when the signals are antipodal. In general it is possible to make the cross-correlation coefficients equal to zero. The set is then said to be orthogonal.

When the orthogonality condition is applied to a waveform that consist of a sequence of pulses, where each pulse is designated with a level +1 or -1, which in turn represents a binary 1 or 0. When the set is expressed in this way, Equation (2.34) can be simplified to
Orthogonal Frequency Definitions

Frequency shift keying (FSK) is often implemented using orthogonal signaling, but not all FSK is orthogonal. Frequency tones can manifest orthogonality if, for a tone at \( f_1 \), the sampled envelop of the receiver output tuned to \( f_2 \) is zero. A property that ensures such orthogonality states that any pair of tones must have frequency separation that is a multiple \( 1/T \) Hertz. To show this consider a time limited, nonzero time function defined as

\[
h(t) = \sum_{n=0}^{N} a_n \psi_n(t), \quad |t| \leq \frac{T}{2}
\]

\[
= 0, \quad |t| \leq \frac{T}{2}
\]

where

\[
\psi_n(t) = \cos \left( n \frac{\pi t}{T} \right)
\]

The function, \( \psi_n(t) \), represents a complete orthogonal basis set with real coefficient \( a_n \). The members of the orthogonal are over the given time interval if

\[
\int_{-T/2}^{T/2} \psi_n(t) \psi_m(t) = K_n, \quad n = m
\]

where \( K_n \) is a constant.
Given the basis set and constraints, two functional descriptions are obtained which are of the form:

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

(2.40)

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

(2.41)

Each cosine term in Equation (2.40) and (2.41) represents a time-gated sinusoid whose local center frequencies are given by

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

(2.42)

The frequency domain representation of the basis functions are well known sampling functions with center frequencies given in Equation (2.42).
Orthogonal Frequency Coding Discussion

The previous section defines the functional descriptions which will be used to define the desired time function. The desired signal will have both frequency and time diversity which will provide a systematic way of implementing a code in a SAW device embodiment. From [30], given a time function $g_{\text{bit}}(t)$, having a time length of $\tau_B$ defined as the bit length, the bit will be divided into an integer number of chips such that

$$\tau_B = J \cdot \tau_c \quad \text{where } J = \# \text{of chips.}$$  \hspace{1cm} (2.43)

The chip interval $\tau_c$ is set as the time interval in Equation (2.42) for the basis set. Allowing a time delay of $\tau_D$, such that $t = (t - \tau_D)$ in Equations (2.37)-(2.42), and given a definition of each chip as $h_{cj}(t)$, then a bit is defined as the sum of $J$ chips as

![Orthogonal Frequencies](image-url)
\[ g_{\text{bit}}(t) = \sum_{j=1}^{J} w_j \cdot h_{cj}(t - j \cdot \tau_c). \]  

(2.44)

Each chip is contiguous without time overlap and \( w_j \) is the bit weight and the functional form for the chip definition \( h_{cj}(t - j \cdot \tau_c) \) is given in Equations (2.40) or (2.41). In general, multiple local carrier frequencies are possible in each chip depending on their weighting coefficient. Assuming a chip uses the basis set in Equation (2.40), similar results are obtained using Equation (2.41). Then, in general,

\[ h_{cj}(t - j \cdot \tau_c) = \sum_{m=1}^{M} b_{jm} \cos \left( \frac{(2m + 1)\pi(t - j \cdot \tau_c)}{\tau_c} \right) \text{rect} \left( \frac{t - j \cdot \tau_c}{\tau_c} \right). \]  

(2.45)

To generate the required signal, let \( b_{jm} = 0 \) for all \( m \), except \( m = C_j \) where \( 1 \leq C_j \leq M \). Then,

\[ h_{cj}(t - j \cdot \tau_c) = b_j \cos \left( \frac{(2C_j + 1)\pi(t - j \cdot \tau_c)}{\tau_c} \right) \text{rect} \left( \frac{t - j \cdot \tau_c}{\tau_c} \right). \]  

(2.46)

The form in (2.46) shows that each chip has a single local carrier frequency of \( f_{cj} = \frac{2C_j + 1}{2\tau_c} \) and \( b_j \) is the chip weight. In order to build the desired time function, the following design rules are used: 1) \( b_j = \pm 1 \) for all \( j \), 2) the bit null bandwidth is \( BW_{\text{bit}} = J \cdot 2 \cdot \tau_c^{-1} \), and 3) \( C_j \) is a sequence of unique integers which means that \( f_{cj} \) form a contiguous, non-repetitive set, similar to Figure 2-11. The rules, however, do not require that the local frequency of adjacent chips that are contiguous in frequency must be contiguous in time. In fact, the time function of a bit provides a level of frequency coding by allowing a shuffling of the chip frequencies in time.
Figure 2-11 shows an example seven chip sequence where \( f_{cm} \neq f_{cn} \) for all \( m \neq n \), and there is an integer number of half wavelengths in each chip. The seven local chip frequencies are contiguous in frequency but are not ordered sequentially in time, and the chip weights are all unity. If the local chip frequencies were ordered high to low or low to high, the time sequence would be a stepped down-chirp or up-chirp, respectively.

**Number of Codes**

Using the OFC definitions above there are a large number of possible permutations of codes possible due to the orthogonal nature of the coding concept. If only the frequencies are permutated in order to produce a code, then there would be \( N! \) possible code combinations where \( N \) is the number of orthogonal frequencies in the chip sequence. The given chip sequence represents the orthogonal
frequency code for the bit. For the single frequency case of a signal, N chips long, and $b_j = 1$, the signal is a simple gated RF burst $\tau_B$ long.

In addition to the OFC coding, each chip can be weighted as $\pm 1$, giving a PN code in addition to the OFC, namely PN-OFC [31]. This does not provide any additional processing gain since there is no increase in the time bandwidth product, but does provide additional code diversity for tagging. For conventional PN coding, the number of available codes is $2^N$. When using PN-OFC coding, the number of available codes is increased to $2^N \cdot N!$. A comparison of the number of codes possible using N chips for OFC and CDMA is shown in Figure 2-12. A seven chip OFC code has approximately the same numerical number of code possibilities as a nineteen chip conventional PN code. A benefit of OFC is clearly seen in this example.

Figure 2-12: A comparison of the number of possible codes given a certain number of chips when using either CDMA or PN-OFC. The number of CDMA type codes increases as $2^N$ shown in solid blue while the number of PN-OFC codes increases as $2^N \cdot N!$ shown in dotted red.
Orthogonal Frequency Division Multiplexing

OFDM is a technique that breaks a bit stream in several subcarriers to transmit data in parallel. The waveforms have the ability to cope with several channel conditions including narrowband interference and frequency-selective fading due to multipath. Using orthogonal subcarriers efficiently utilizes the spectrum without causing undue interference.

\[
\text{Figure 2-13: Block diagram of a simple OFDM transmitter with serial to parallel conversion of the bit stream before being modulated by a carrier frequency and summed to go through the channel. The plot on the right is a depiction of orthogonal subcarriers utilizing efficient spectrum bandwidth.}
\]

One key advantage in communication systems is the advantage of transmitting a number of low-rate data streams in parallel instead of a single high rate stream [17]. The subcarriers are separated by the bit rate \( r \) which is related to the bit length \( \tau \) as

\[
r = \frac{1}{\tau}
\]  

(2.47)

Using this equation and the definition for orthogonality established previously each subcarrier is allowed to overlap without causing interference between the channels. The complications arise as the demodulation of this coding scheme is susceptible to incorrect frequency synthesis.
SAW Device Introduction

SAW devices have been utilized for many years in varying applications. A typical SAW device is a solid state device that converts its electrical energy into a mechanical wave on a single crystal substrate. Its popularity and expanded use into sensing applications comes from its ability to provide very complex signal processing in very small packages. Small, rugged devices offer a low-cost, light-weight solution for operation in harsh environments [32]. It also provides a monolithic structure fabricated with current IC photolithography techniques. Generally a SAW device consists of one or more transducers on the substrate. The coupled energy is then transferred to another transducer in the case of filtering operations or to a reflector structure. The reflector structure is similar to that shown in Figure 2-14. It consist of metal on a piezoelectric substrate with a certain periodicity and metalized to none metalized region ratio typically referred to as the a to p ratio. The periodicity and a/p ratio determine the frequency of reflection or resonance. When the wave passes underneath the metalized region a certain portion of the energy is reflected. This creates a forward and reverse traveling wave shown in Figure 2-14.

The device can have multiple ports. An example of a device utilizing multiple ports is shown in Figure 2-15. These devices may be attached to external elements for obtaining sensory information.
Many current sensors allow the use of external stimuli affects device parameters (frequency, phase, amplitude, delay) [33, 34].

![Figure 2-15: A film type sensor with input and output transducers. The reactive film modifies the SAW wave yielding the sensory information.](image)

A chirp or weighted chirp maybe implemented for either of the transducers within the SAW device [9, 35, 36]. This allows for an amount of signal processing to take place at the device. The ease of which SAW devices create these complex signals is one of the technologies main advantages.

A one port device returns an altered interrogation signal [37]. The sensor acts as a cross sectional reflector as in a radar system. Similar to radar current SAW devices can provide wireless and passive operation [1, 3, 4, 31, 38-44]. These returned signal yield the desired sensed information.
Figure 2-16: Single frequency type device showing the reduction in reflected power as the number of single frequency reflectors increases.

The maximum range for a single frequency CDMA or PPM device is limited by the received power from the last reflector. Sensor range currently limited to a few meters [33, 38, 45]. Range of these devices may be increased using coherent integration of multiple responses [45]. Figure 2-17 plots the returned power against the number of reflectors. For CDMA each reflector would act as a chip, N chips would then return a signal that corresponds to the curve in Figure 2-17. Weighting the reflectors or other techniques may be applied to combat the severity of the difference between the first reflector and the last [46].
Figure 2-17: Plot of returned power for a typical delay line sensor utilizing N number of reflectors. The amount of energy quickly decreases with increased reflectors, leveling out as the number gets to be large.

Current devices offer capability of operating over a large frequency range \( \sim 10 \text{ MHz} - 3 \text{ GHz} \) \([2, 4, 34, 42, 47-49]\). SAW sensors sensitivity is quantifiable for different measurements on various substrates \([50-56]\). The issue of multiple simultaneously illuminated sensors still remains a prevalent research topic. Currently there are few embodiments that can account for multiple tags/sensors \([14]\). The limitation produced by multiple sensors may be circumvented by using interrogator proximity/antenna directionality to excite single device \([57]\).

**OFC SAW Implementation**

Given the functional description stated above, a SAW device can be applied. A device schematic is shown in Figure 2-18. Additional information about SAW OFC device implementation can be viewed here \([7, 30, 41, 58-61]\).
Figure 2-18: Schematic of a 7 chip SAW OFC tag. Each reflector contains a chip and has a reflector period corresponding to the desired frequency. Shown in red is a chirp interrogation signal and in blue the convoluted response from the chirp and reflector interactions.

For use in a sensing application the SAW PN-OFC device utilizes a differential mode of data modulation. The sensor is employed by creating identical reflector banks on either side of the device as depicted in Figure 2-19. The figure shows the use of a wideband transducer. Due to the large bandwidth occupied by the device the individual chip amplitude can become susceptible to variation. This amplitude variation can be combated by utilizing a weighted transducer [30, 59]. The resulting chips will correspond more closely to the ideal and compressed pulses appear sharper.

Figure 2-19: Sensor schematic for a differentially operated SAW PN-OFC device. The delays on either side of the device allow for simultaneous tagging and sensing operation. The difference in the delay will also set somewhat the precision of the data.
A sensor with a weighted transducer was tested using an RF probe station and a temperature controlled chuck. Figure 2-20 shows the results of the experiment. Additional results of the implementation are here [7, 30].

![Temperature Sensor Results](image)

Figure 2-20: Sensor temperature versus thermocouple temperature for a YZ LiNbO3 OFC sensor. Extracted temperatures (red circles) are shown, and expected measurements (blue line) are also indicated [30].

**Discussion**

Spread spectrum communications have been discussed and necessary formulae have been established. The remainder of the chapters will build on the fundamentals established in this chapter. The next chapter will focus on building up a system that will implement the SAW OFC devices presented here. The following chapter will discuss the coding implementations for OFC SAW, and followed by the multiple-access techniques that can be used.
CHAPTER 3: SYSTEM LEVEL SIMULATION

The evaluation of modern communication systems encompasses a diverse number of parameters. For a proper presentation of an actual signal environment for OFC SAW tags, the essential communication characteristics must be implemented into the simulation. As such this section will cover the basic communication aspects implemented within the system simulation. The leading characteristic of a SAW passive sensor system is an interrogation device that sends an RF signal and receives a modified response from a sensor tag. It is vital that the targets be differentiable when interrogated simultaneously.

System Components Introduction

The system can be broken into several distinct parts: the transmitter, the channel, the device, and the receiver. The transmitter and receiver are often combined into a single block and referred to as the transceiver; the two will remain separate here in order to easily distinguish between their functions. A block diagram illustration of the basic system parameters is show in Figure 3-1.
The transmitter delivers power. The input source into the transmitter can be varied, i.e. continuous wave, pulse, chirped, and the source may be categorized as either analog or digital in nature. Any power the tag receives will first be manipulated and transmitted from the transmitter. It is possible to transmit nearly unlimited power if one were in space, but FCC and health regulations must be maintained for terrestrial based communications. Given that terrestrial communications is assumed, there is in fact a limit to the amount of power one can emanate and therefore be received by the targets. Techniques exist that can lower the peak power transmitted while maintaining the same average power. One such technique is known as chirping and is commonly used in radar systems [62]. These systems often utilize SAW devices to implement these chirps because of their quick response and superb linearity. The figure of merit for the simulation used in this thesis is the average transmitted power regardless of implementation.
The channel is the propagation medium that connects the transmitter with the device. In general, the channel can consist of coaxial wires, fiber optic cables, waveguides, the atmosphere, or empty space. As with most terrestrial communications, the channel considered here is the atmosphere.

The equations for the OFC SAW device have been previously derived. An ideal device can be simulated using these equations. The ideal device can be interpreted as a bit of a digital signal that has a SS code placed upon it. As formerly described, the bit is broken into chips of equal time length, and each is given a distinct frequency allocation.

![Figure 3-2: Ideal time response for an OFC SAW device. The device can be thought of as a bit with a SS code embedded within it. Each bit is comprised of several chips of equal time length.](image)

A generic receiver contains a few key components: an antenna, a RF filter, an amplifier, a mixer, a LP filter, and an analog to digital converter (ADC). These basic components comprise a very straightforward receiver. The receiver shown in Figure 3-3 displays a generic receiver with filters, a mixer, power amplifiers, an ADC, and a summer. The ADC will usually output the information to a computer for the summing operation.
Figure 3-3: Basic receiver diagram consisting of the main components. The antenna receives the signal and passes it through the filter to remove unwanted signals before going to an amplifier, mixer, low-pass filter, and additional amplification. The signal then is passed to an analog-to-digital converter to allow post processing, such as a summer, in the digital domain.

Full System Signal Simulation

The previous section provided the background for a generic system level simulation. The following section will describe in more detail the exact methods used to obtain a system response from a sensor array. Modeling and simulation are keys to analyzing communication systems of any type before actual fabrication and physical testing begin. MathCAD and Matlab are common tools in communication simulation and will be utilized in this thesis. As discussed in the previous chapter, similar simulations have been done on SAW OFC devices [7, 63, 64] with only single devices implemented.
System Overview

Figure 3-4: A representation of a multiple sensor array field. Any sensor receiving energy from the transmitter responds with a signal. The transmitter will then act as the receiver collecting the signals to be passed through to the signal processing unit.

The system seen in Figure 3-4 is a representation of a multiple sensor environment. The system is stationary for the purposes of this evaluation. The transceiver is where the analysis is performed on the individual sensors of interest. When this is modeled in a computer simulation, it closely resembles the block diagram shown in Figure 3-5. Each signal is generated separately either using ideal equations or COM modeling. These signals are placed in a random fictional space by the input of a delay. The spacing is designated by the operating parameters. These signals are all summed as they pass through
the channel. This received signal is then sent to a bank of N matched filters for correlation. A decision is made based on this correlation data, and sensed information is extracted.

**System Simulation Block Diagram**

1. Target
2. Channel
3. Receiver

Figure 3-5: Block diagram of the computer modeled data present in the simulation of a system. The signals are created and a spatial delay with corresponding attenuation representative of the channel is then implemented. Noise is added at the receiver antenna before being passed to the matched filter and signal processing units.

**Transmitter Parameters**

The transmitter emits the interrogation signal to be sent to the targets. With the usage of an antenna on the transmitter, a new parameter can be defined. The effective radiated power, with respect to an isotropic radiator (EIRP), is defined by
\[ \text{EIRP} = P_t G_t \quad (3.1) \]

where \( G_t \) is the gain of the transmitting antenna. The gain is an antenna parameter that relates the power output (or input) to that of an isotropic radiator. The directive gain is

\[ G = \frac{\text{maximum power intensity}}{\text{average power intensity over } 4\pi \text{ steradians}}. \quad (3.2) \]

The transmitter is then the interrogator of the SAW target either generating or modifying an input signal for this purpose. The signal leaves the transmitting antenna, is passed into the channel, and no further manipulation performed.

The power radiated from the transmitter antenna is assumed to be uniform over the bandwidth of interest. This implies the use of an idealized impulse function in the time domain. The power spectral density is an effective way to observe the power over a certain spectrum. Figure 3-6 shows an example of a bandlimited PSD of the transmitter. The band limiting can be thought of as caused by the transmitting antenna or by some limiting filter in the transmitter. The average power transmitted is assumed unaffected by any impedance mismatch loss or dissipative losses. Therefore, all power generated is transmitted through the antenna. The antenna on the transmitter is specified a gain of unity. This allows for the EIRP to be equivalent to the power transmitted, \( P_t \). Recall the average power calculated in the time domain utilizes Equation (2.2). A flat PSD response allows for a simple evaluation of the average power utilizing the frequency response.
Figure 3-6: Band limited single sided PSD of the transmitter assuming uniform transmission over the bandwidth of interest.

Channel Characterization

Within the channel, the concept of free space is employed. Free space assumes that the channel is free of any hindrance to RF propagation such as absorption, reflection, or refraction [65]. The energy arriving at the receiver is assumed to be a function of the distance from the transmitter only. A free space channel characterizes an ideal RF propagation path; in practice, absorption, reflection, and refraction will play some role in the amount of received energy.

The governing equations for the channel are based on range equation derivations. An isotropic radiator creates a power density on a hypothetical sphere at a distance $r$ from the source and is related to the transmitted power $P_t$

$$p(r) = \frac{P_t}{4\pi r^2} \text{watts/m}^2.$$  

(3.3)

For a $r$ much greater than the propagation wavelength (known as the far field), the power extracted at a receiving antenna is
\[ P_r = p(r)A_{er} = \frac{P_t A_{er}}{4\pi r^2} \]  

(3.4)

where the parameter \( A_{er} \) is the absorption cross section (effective area) of the receiving antenna, defined by

\[
A_{er} = \frac{\text{total power extracted}}{\text{incident power flux density}}.
\]  

(3.5)

The relationship between the antenna gain \( G \) and antenna effective area \( A_e \) is

\[
G = \frac{4\pi A_e}{\lambda^2} \quad (for \ A_e \gg \lambda^2).
\]  

(3.6)

The power that passes through the channel and is received at the device can then be defined by

\[
P_r = EIRP \frac{G_r \lambda^2}{(4\pi r)^2}.
\]  

(3.7)

Since the antenna gains and transmitted power are not directly governed by the channel, it is common to group the remaining terms. These collection of terms are often called the path loss, or the free-space loss, and designated by \( L_s \), where

\[
L_s = \left(\frac{4\pi r}{\lambda}\right)^2.
\]  

(3.8)

The same equations apply when the device reradiates its energy and acts as the transmitter. Thus in the simulation the channel effects must be taken into account twice.

While passing through the channel to the device, the signal is affected by the path loss established in Equation (3.8). The equation is clearly affected by the frequency of operation. While this is not a vital concern in the simulation of the channel, it will affect the optimal system setup. Operating
range, antenna size, and ease of fabrication are just a few of the factors affected by the choice of frequency. The operating frequency employed by the initial simulations was selected to concur with previously fabricated and verified devices [63, 64]. With a large fractional bandwidth, it is possible that fading type effects will influence the response of the signal. A 250MHz device with a fractional bandwidth of 30% will have an amplitude variation at the band edges of approx 2.3dB. Therefore, the chip corresponding to the highest and lowest frequencies will vary in amplitude by 2.3dB. The consequences of the amplitude variation are possible to account for since they are known to exist. The simulation will therefore take the average path loss based upon the center frequency. Figure 3-7 shows a plot of the change in path loss due to a small frequency variation.

Figure 3-7: Path Loss calculation in dB normalized by the amplitude at 250 MHz. The path loss at the band edges of a 30% fractional bandwidth signal is approximately 2.3dB.

The power present at the antenna of the device when the previous assumptions are made is now governed by
\[ P_{rd} = \frac{P_t G_e}{L_s}. \]  

(3.9)

Target Parameters

The device when viewed on a substrate would resemble Figure 3-8.

Figure 3-8: OFC SAW device on a piezoelectric substrate. The differential device depicted is used for the extraction of information based on the different delays on each side of the transducer.

The SAW device can be modeled by utilizing the coupling of modes. This is a P-matrices cascading of parameters to accurately represent the SAW device. The use of the COM model allows for the entire system to be calculated quite accurately [66, 67]. The model of such a device can be seen in Figure 3-9.:
Figure 3-9: Coupling of modes simulation of an OFC SAW device on LiNOb3. The two reflector banks, seen in the time domain, represent the two responses from a differential device.

The device would also encompass an antenna and matching network for attachment to the substrate.

Antenna Considerations
The equations for an antenna would be similar to those based on range (see Equations (3.4)-(3.6)). If an electrically small antenna (ESA) is utilized, then there are theoretical limits to dimension with respect to bandwidth and gain [68].

For a linear ESA with lossless matching network, the Q of the ESA is defined by

$$Q_a = \frac{1}{ka} \left[ 1 + \frac{1}{(ka)^2} \right]$$  \hspace{1cm} (3.10)

where k is the wavenumber, and a is radius of a sphere that can enclose the antenna. The percentage bandwidth can then be given by
\[ \text{PercentBW} = \frac{S - 1}{Q_a \sqrt{S}} \]  

(3.11)

where \( S \) is the efficiency of antenna and matching, and \( S=1 \) VSWR.

\[ G = k \cdot a(2 + k \cdot a). \]  

(3.12)

Figure 3-10: Gain and percentage bandwidth attainable versus wavenumber-radius product for an ESA. The solid line is for gain and the dotted line is for fractional bandwidth, \( S=2 \).

Figure 3-10 plots ESA gain and fractional bandwidth versus the wavenumber-radius product \((k \cdot \alpha)\). The solid line represents gain and the dotted line fractional bandwidth; both curves are frequency independent. The antenna effects will be discussed further in the optimized system section.

For the remaining calculations, the antenna gains have dropped out as they are assumed isotropic radiators, leaving only the transmitted power and path loss. This energy is then manipulated by the device. Impedance mismatch between the antenna and device are assumed to contribute minimal loss and signal distortion. The loss based on [30] is approximately 0.168 dB, best case scenario for perfect matching. The signal distortion will not be modeled in this simulation.
Figure 3-11: SAW device model illustrating the major contributors to the overall system range. Modeling of the device must account for these gains and loss, allowing for accurate simulation.

Figure 3-11 illustrates the major mechanisms affecting the device power received and reradiated. For the ideal case, the substrate characteristics are not taken into account. The ideal, single signal response from theoretical equations without device parameter simulation can be seen in Figure 3-12. The transduction and reflector losses are unaccounted for in the ideal simulation. These must be factored in when calculating the signal retransmitted for a more accurate assessment.
The COM model does, in fact, model these parameters and allow for accurate device representation without additional manipulation. Figure 3-13 illustrates a COM simulation with the transducer and reflectors represented. Note the distinct difference in the envelope of the COM model data and that of the ideal case. The transducer impulse response directly affects these responses and research has previously been conducted to improve these profiles [30, 59].
Figure 3-13: Coupling of modes simulation of a 5 chip, 5 orthogonal frequency device. The amplitude profile is observed for the time domain response. The transducer electrical response and triple transit response present in a SAW device are clearly seen. The non-uniform amplitude response is due to the narrowband input transducer used in the simulation.

After the device receives and manipulates the power, it can be thought of generically as the transmitter. The passive sensor system is similar to a radar system. The device acts as the cross sectional area power reflector. One can now describe the power that is reradiated or transmitted, $P_{td}$, by the device as

$$P_{td} = \frac{P_{rd} G_d^2 P_d}{L_d} \quad (3.13)$$

where $G_d$ is the devices antenna gain which was assumed to be unity, $P_d$ is the average power of the device whether ideal or COM, and $L_d$ is the device loss. $L_d$ takes on two values, one if an ideal
simulation is performed and another for a COM simulation. This allows for the transduction loss and reflector loss to be considered more simply.

Receiver

The device’s response travels once more through the channel and propagates over the distance \( r \). The overall general equation for a single device’s received power at the receiver now closely resembles the radar equation shown as

\[
P_r = \frac{P_t G_r G_t^2 P_d}{L_s L_d}. \tag{3.14}
\]

Equation (3.14) states the power present at the receiver from an individual device. Each of the \( N \) devices present in the field of view of the antenna would utilize the same transmitted energy with which to modify and reradiate. Recall that each of the returned signals is added together as they reradiate their energy and pass through the channel. The returned signal will be

\[
r(t) = \sum_{m=1}^{N} A_m \cdot g_m (t - \tau_{D_m}) \tag{3.15}
\]

where each of the \( g_m \) signals are delayed by some random time \( \tau_{D_m} \). \( A_m \) and \( \tau_{D_m} \) are controlled by the corresponding spatial delay. If target \( m \) is at range \( r_m \), then the corresponding time delay is \( \tau_{D_m} = r_m / c \). A pictorial representation of device placement is seen in Figure 3-14. This figure contains devices scattered around a 5m, and 10m arch from the transceiver.
Figure 3-14: Pictorial representation of device placement in 2-D space. The devices are placed randomly around two arches at 5m and 10m. The device placement implies a random delay with respect to one another creating asynchronous reception of the signals.

For example, the maximum amount of displacement implemented within the system simulation is along a 5m arch and dithering around that within ±0.5m. The received signal would therefore be defined by Equation (3.15) with each $g_m$ having a different received power and corresponding amplitude $A_m$. The received signal for a thirty-two device OFC-PN ideal simulations is shown in Figure 3-15.
Figure 3-15: Time response for 32 devices utilizing an ideal simulation. Each device has a random delay with respect to one another. Notice the amplitude variation caused by the asynchronous nature of the overlapping chips within each device.

**Additive White Gaussian Noise**

When the signal is received after passing through the channel, noise is assumed to have been added. A white noise generation in the frequency domain was performed for implementation in the simulation environment. Additive white Gaussian noise (AWGN) is generally described as a signal with constant power spectral density (PSD) and a time signal with a zero mean value [69]. If the PSD is described as $\eta$ watts per Hz when measured over positive frequencies, then the power spectrum is

$$G_n(f) = \frac{\eta}{2} \text{ for all frequency.} \quad (3.16)$$

The factor of one-half is introduced due to the two sided power spectrum. Note that the spectrum is defined over a power basis. This requires a multiplication by $R$ to obtain the mean-square voltage. Recall that this does not describe any real process as its energy is infinite, but one is generally
not interested in all frequencies. For band limited white noise, the noise power is independent of the frequency of operation. The choice of bandwidth determines the noise power as

\[ P_n = \int_{-B}^{B} \frac{\eta}{2} df = \eta B. \]  

(3.17)

Assuming this is generated across a resistor R, the mean-square noise voltage is

\[ |n(t)|^2 = R P_n = \eta BR \text{ volts}^2. \]  

(3.18)

Thus, one must take into account bandwidth and resistor values when generating a signal in the frequency domain.

Noise is added into the system at the receiver’s antenna. Figure 3-16 shows an ideal signal that has been corrupted by AWGN.

![Figure 3-16: Ideal OFC device response corrupted by AWGN. The device contains 5 chips and 5 orthogonal frequencies and has an approximate 3dB signal to noise ratio.](image)

55
Figure 3-17: COM modeled OFC device response corrupted by AWGN with signal to noise ratio of 3 dB. The transducer electrical response and reflector response are still clearly seen but the triple transit response is now nearly buried within the noise.

A white noise generation in the frequency domain was performed for implementation in the simulation environment. Recall that the PSD can be defined as

$$G_n(f) = \lim_{T \to \infty} \frac{|F_T(f)|^2}{T}$$  \hspace{1cm} (3.19)

where $F_T(f)$ is the fourier transform of signal $f(t)$, and $T$ is the period of the truncated time signal [70]. Equation (3.19) is necessary when attempting to create a signal that conforms to Equation (3.16) while still being realizable in the frequency domain. For implementation, a random number generator is used to create a uniform random variable $\theta$ between $\pm\pi$ which is then passed to

$$noise = \cos \theta + j \cdot \sin \theta,$$  \hspace{1cm} (3.20)

with magnitude is one with a random phase. This signal is created for each data point. Since AWGN is considered a wide sense stationary random variable, no two data points are directly correlated. The
unity magnitude frequency vector must be scaled by $\sqrt{kT_0 T_{max}}$, where $k$ is Boltzmann’s constant, $T_0$ is the temperature, and $T_{max}$ is the signal period. Scaling in this way allows for the average power of any bandwidth chosen to be the familiar $kTB$. Below is a pictorial representation of the signal and corresponding histogram.

Figure 3-18: The band limited frequency response of the generated noise signal amplitude and angle of the frequency signal (top) and real and imaginary of the time signal (bottom).
The signal to noise ratio (SNR) is implemented by taking the average power of the generated signal and the average power of the noise implemented signal and simply dividing them. This yields the SNR at the receiver antenna

$$\text{SNR} = \frac{P_r}{P_n}. \quad (3.21)$$

For a desired SNR given in dB, consider

$$\frac{P_r}{P_n} = 10^{\frac{\text{SNR}}{10}}. \quad (3.22)$$

The noise at the receiver is fixed by the antenna bandwidth or the IF filter bandwidth. Simply, the power received must be changed if a desired SNR is to be achieved. In observing Equation (3.14), the only scalable parameter after the devices are placed in the environment spatially is the power transmitted. The scaling factor to be multiplied by $P_r$ is found to be
\[
\frac{A_{SNR} \cdot P_L}{P_n} = 10^{\frac{SNR}{10}} \quad \text{or} \quad A_{SNR} = \frac{10^{SNR} P_n}{P_r}
\] (3.23)

While it is not necessarily practical to change the power transmitted, certainly if it is to make the value extraordinarily high, it is possible in the simulation to imagine it varying as such. The PSD of the received signal after scaling and the noise are shown in Figure 3-20. The signals have been normalized to the maximum of the received signal.

![Power Spectral Density of COM Modeled S11 of SAW Device and AWGN](image)

Figure 3-20: The normalized signal and noise power spectral density functions for a 3dB SNR.

**Digital Receiver for AWGN**

A correlator type and matched filter type receivers are considered optimum receiver types for signals corrupted by AWGN [19]. As such, the matched filter type receiver is employed for use with SAW OFC sensors. The matched filters’ basic property can be described by the impulse response
\[
h(t) = \begin{cases} 
    s(T - t) & 0 \leq t \leq T \\
    0 & \text{elsewhere}
\end{cases}
\]  
\hfill (3.24)

where \( s(t) \) is the real valued signal. The correlation realization of the matched filter is described in the time domain as the convolution of the received signals' waveform \( r(t) \) with the impulse response of the filter:

\[
z(t) = r(t) * h(t) = \int_0^t r(\tau) \cdot h(t - \tau) d\tau.
\]  
\hfill (3.25)

Substituting (3.24) into (3.25)

\[
z(t) = \int_0^t r(\tau) \cdot s(T - t - \tau) d\tau.
\]  
\hfill (3.26)

When \( t = T \),

\[
z(T) = \int_0^T r(\tau) \cdot s(\tau) d\tau.
\]  
\hfill (3.27)

This product integration of the received signal with a replica over a period is known as the correlation of \( r(t) \) with \( s(t) \). Therefore, the received signal can be thought of as running through a bank of \( N \) correlators or matched filters. A noise-free comparison of the ideal correlation and the received signal correlation is shown in Figure 3-21.
Figure 3-21: Comparison of an ideal correlation peak seen in blue and the peak generated by passing the received signal through the matched filter in dashed red. The received signal is comprised of 32 individual device signals utilizing PN-OFC coding. Each device utilized 5 OFC chips and occupies approximately 28% fractional bandwidth. The peaks clearly do not match in this case.

With the SNR established, the signal is then input into the N matched filters for processing.

Figure 3-22 shows an example of a single device matched filter response. The received signal contains three device responses and is corrupted by AWGN with a signal to noise ratio of 10 dB. Compare the ideal response and the COM response of the same simulation variables and set of codes shown in Figure 3-22.
Figure 3-22: A representation of a COM model and an ideal signal’s matched filter responses. The COM model correlation shown on top tends to be more susceptible to high side lobes than the ideal on bottom. The received signal contains three device responses and is corrupted by AWGN with a signal-to-noise ratio of 10 dB. Each device contains 5 OFC chips each with a PN sequence placed upon it, and occupies 28% fractional bandwidth at 250 MHz.

The correlation peaks have expected, known locations. One advantage when utilizing a SAW device is the ability to distinguish the response signals from environmental echoes [36]. This feature is
beneficial in this case because the correlation peak will be more heavily dependent upon substrate placement than temperature or spatial variation. As the substrate placement is a designed variable, it will be known a priori. The search algorithm can then narrow the possible search window instead of attempting a max peak search. An example of windowing the resulting reflected time response is seen in Figure 3-23. In this case, the result has removed the consequential high sidelobe cross-correlation and left only the main auto-correlation peak.

![Ideal OFC MF Response with AWGN](image)

**Figure 3-23:** An example of the windowing algorithm used in peak detection. The resulting signal yields a smaller search area and improved peak detection.

After the algorithm is used to narrow the search area, the max signal point is found. For the purposes of the system simulation, this maximum signal point is compared with the known inputted delay. The delay that was placed due to spatial variation, temperature variation, and on substrate delay, should match this value closely. The closeness determines an error percentage. An arbitrary value was initially chosen for availability of comparison. A peak located within five sample points of its intended
value was considered correct. This allowed for observation of the effects of a system with a varying number of simultaneous devices simulated.

Figure 3-24: Percentage of erroneous device matched filter peak locations plotted against then number of devices simulated simultaneously in a system level simulation. The ideal simulation is compared with a COM modeled simulation of the same system parameters and identical code set.

After the matched filter process is performed and a peak location established, additional post-processing can be performed on the received signal. Techniques utilized to increase peak delay accuracy will be discussed in the optimal system design section in the next chapter.
COM System Simulation

Through the use of a reliable simulator, a large amount of data analysis can be done without having to manufacture a device. This is especially valuable in circumstances when cost or time is extremely prohibitive in the fabrication and implementation of devices. The coupling of modes (COM) technique has proved itself reliable for the prediction of SAW devices [67, 71-73]. The COM model for SAW reflective structures uses a p-matrices analysis of the SAW device to characterize its frequency response.

When simulating a system of sensors utilizing SAW devices, it is advantageous to simulate the potential interactions prior to fabricating and testing a set of codes. It is often difficult to obtain a proper testing environment for multiple sensors. In this instance, the COM model allows device interactions to be ascertained free from physical testing.

Reliability of COM model

The COM model is very dependable when simulating SAW devices. For means of a comparison devices have been fabricated and their data extracted using a probe station and a vector network analyzer (VNA). The data taken from the VNA was then compared to the COM model data in an attempt to verify predictions. The frequency domain response of a one port device, representative of the SAW sensors being discussed, is shown in Figure 3-25. The reflectivity of the response matches very closely and with proper accounting for the static capacitance and resistance provided by the probe and probe pads, the slope of the transducer matches as well.
Figure 3-25: Frequency domain comparison of a one port SAW device with input transducer and reflector bank. The device is fabricated on Y-Z lithium niobate at 250 MHz. This device is uses a wideband transducer and eight OFC reflectors. The simulation is shown in blue, and the experimental probed data is shown in red.

The time domain data from the previous device, shown in Figure 3-26, also has very good agreement between experimental and simulated response.
Figure 3-26: Time domain comparison of a one port SAW device with input transducer and reflector bank. The reflector bank is an eight frequency OFC device with eight chips. The device is fabricated on Y-Z lithium niobate at 250 MHz. The simulation is shown in blue, and the experimental probed data is shown in red.

Given this data set and many more like it, it is possible to ascertain that the simulation utilizing these equations will provide adequate results. It was necessary then to fully understand the COM model implementation if the delay was to be extracted correctly.

*Simulation Environment*

The COM model yields scattering parameters that are able to be used in the simulation. Unlike the ideal model each device is comprised of an input transducer and a set of reflectors. The transducer is uniform with five finger pair at the operating frequency. The transducer uses its fundamental mode of
operation with one-fourth wavelength electrodes. The reflectors also use their fundamental operation utilizing one-fourth wavelength electrodes.

![SAW device schematic](image)

**Figure 3-27. SAW device schematic, includes an input/output transducer and 5 reflectors of varying periodicities.**

The specifications of the transducer operating at 250 MHz are given in Table 3-1.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Number of Periods</strong></td>
<td>5</td>
</tr>
<tr>
<td><strong>Wavelength</strong></td>
<td>14.02 µm</td>
</tr>
<tr>
<td><strong>A/P Ratio</strong></td>
<td>0.5</td>
</tr>
<tr>
<td><strong>Metal Thickness</strong></td>
<td>1500 Å</td>
</tr>
<tr>
<td><strong>Beam Width</strong></td>
<td>20 λ</td>
</tr>
<tr>
<td><strong>Tap Width</strong></td>
<td>¼ λ</td>
</tr>
</tbody>
</table>

**Table 3-1: Wideband transducer specification entered into the COM model for proper modeling.**

![Wideband transducer model](image)

**Figure 3-28: Wideband transducer model used in COM model simulations.**
The specifications of the reflectors as defined within the COM model for a five frequency OFC device are given in Table 3-2.

Table 3-2: Reflector specifications corresponding to 5 different reflectors utilizing 5 differing frequencies. The number of periods and wavelength dictate the appropriate frequency when the device is modeled.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Number of Periods</strong></td>
<td>7.5</td>
<td>8</td>
<td>8.5</td>
<td>9</td>
<td>9.5</td>
</tr>
<tr>
<td><strong>Wavelength</strong></td>
<td>15.51 µm</td>
<td>14.54 µm</td>
<td>13.68 µm</td>
<td>12.92 µm</td>
<td>12.23 µm</td>
</tr>
<tr>
<td><strong>A/P Ratio</strong></td>
<td>0.5</td>
<td>0.5</td>
<td>0.5</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td><strong>Metal Thickness</strong></td>
<td>1500 Å</td>
<td>1500 Å</td>
<td>1500 Å</td>
<td>1500 Å</td>
<td>1500 Å</td>
</tr>
</tbody>
</table>

Figure 3-29: A reflector structure with 5 reflectors. The periodicities are not meant to be accurate representations but are for visualization only.

**Delay Implementation**

The time delay of the reflectors’ response corresponds to a spatial delay on the substrate. This delay is intentionally made to be larger than one microsecond to allow for any other responses reflected off environmental echoes to be disregarded. At a distance of five meters, environmental echoes will be received within twenty nanoseconds. While the delay is quite large because of the much slower velocity of the SAW, it can be implemented with a much smaller distance. For LiNOB3, a 1us delay corresponds to an on substrate delay of 0.0035m or 3.5mm.

When the matched filter operation is performed on a real device, ideal assumptions must be reworked. The exact on substrate delay must be known when attempting to find whether the
correlation peak is in error. The distance between the transducer and the reflectors, \( D_r \), corresponds to a time delay of

\[
\tau_r = \frac{D_r}{v_s}
\]  

(3.28)

where \( v_s = \text{substrate velocity} \)

When the transducer’s corollary on the delay is considered, the convolution effect must be taken into account. As pictured in Figure 3-31 a slope is created in the time response when the transducer is convolved with the reflectors. The beginning of the slope corresponds to the delay \( \tau_r \) but the corresponding first chip occurs at \( \tau_r + \tau_{idt} \). Because of this, the correlation peak will also start at the value \( \tau_r + \tau_{idt} \).
Figure 3-31: Convolution of two $\text{rect}(t)$ functions of different lengths showing the slope that is induced by the convolution. This is representative of the convolution incurred by the transducer convolving with the reflector structure in a SAW device.

Once this delay is specified, appropriate decision algorithms can be implemented in order to verify the accuracy of the model. This delay corresponds to the first reflector bank that is to be implemented from the code set. Each of the additional reflector banks will be offset from this starting point, the $N \cdot \tau_c$ specified in the design of the code set. In addition to the initial delays, each reflector bank uses varying OFC-PN codes.

**COM Simulation Example**

The frequency and time response for an example simulation can be seen in Figure 3-32. The COM model simulation yields the S11 response of the device. This response can then be manipulated. The FFT is performed on the S11 response yielding the time response. The response of three separate returned signals is plotted in Figure 3-32. The offset implemented between devices is two chip lengths and can be clearly seen.
Further manipulation can be implemented now that the COM data has been acquired. In order to more distinctly view the time response the reflectors only, the transducer response can be removed from the data. This can be done in one of two ways: (1) modeling the reflector on its own and subtracting that information or (2) smoothing the data and subtracting the smoothed response. The S11 response of the reflectors is shown in frequency, Figure 3-33a, and time, Figure 3-33b. The response is then conjugate flipped to create the negative frequency response. This technique is employed to ensure that the time response of the signal has no imaginary information associated with it. The two sided spectrum and time responses are shown in Figure 3-33c, and Figure 3-33d. The time domain data now has the same information sampled twice as frequently.
The COM data can then be passed to the same functions that are used for the ideal simulation; implementing a time delay due to spatial offset or temperature; the matched filter process; and finally the peak detection.

**Full Computer System Model and Simulation**

When a user runs the simulation, they go through a series of dialog boxes used to setup the simulation parameters. Initially the user must input some basic parameters such as how many devices to simulate simultaneously and the number of times they would like the simulation run, to create a better statistical average. This dialog is shown in Figure 3-34.
The user must then input the code set under test utilizing an excel file created beforehand, shown in Figure 3-35. This allows for ease of manipulation outside of the simulator to speed up simulation time.
A friendly dialog box then asks the user to wait while it continually updates its progress to show how much has been performed and how much is left.

![Progress bar indicating the time elapsed and the percentage complete for the simulation running.](image)

Figure 3-36: Progress bar indicating the time elapsed and the percentage complete for the simulation running.

Then based upon the initial inputs a percentage error figure is presented to the user. An example is shown in Figure 3-37.

![Example output of the system simulation stating the number of errors versus the number of tags simulated simultaneously.](image)

Figure 3-37: Example output of the system simulation stating the number of errors versus the number of tags simulated simultaneously. The error percentage allows a simplistic view of the merit of the design being tested. This simulation utilized OFC-PN coding with no offset; 3dB SNR; and a range variation of +1m from a 5m arch.
The simulation also outputs a data file containing an array of the observed distance away from the expected value. This data can be manipulated later in order to better understand the outcome of the simulation run.

**Discussion**

A system simulator was presented and the necessary equations established. When the simulation of sensor environment is combined with the performance of the COM model a very reliable simulator can be realized. This allows for more rigorous testing of code sets and susceptibility of errors due to varying other parameters, i.e. SNR. The next chapter investigates different coding techniques.
CHAPTER 4: CODING APPROACHES

Using the techniques developed in the previous chapter, simulations were performed for ideal SAW PN-OFC sensor tags. In the course of these simulations, cross-correlation effects were causing deviation in the correlation peaks. A method to initially decrease these effects needed to be implemented. Initial efforts explored changing the PN sequences because the signals were thought to be adding in-phase and out-of-phase. PN code sequences are commonly used in wireless telephone and other technology. These code sequences are also referred to as maximal sequences; Gold codes and Barker codes are considered as such [19]. These codes have become very common in the use of optimal PN coding. Although helpful in many situations, normal code sequence generation could not be used due to the length of codes and method of implementation. This presented the challenge of finding another way to lessen the sidelobes of the cross-correlations that were causing the peak distortion.

Code Finding Techniques

As there is only one PN code sequence that generates optimal sidelobes, its implementation to create code diversity was not beneficial. Rather, the shuffling of chip frequencies could be optimized to add code diversity. In an attempt to eliminate the cross talk between the different SAW tags, the discovery of different but complementary code sequences was imperative. The codes also must eliminate the erroneous bits that presented themselves. No known technique had been established for this type of code-finding.

Autocorrelation

Since the appearance of high sidelobes was the initial concern, the autocorrelation of the code set needed to have the lowest sidelobes possible. This did not ensure low collision properties or even
that the choice of code would necessarily be optimal. The process began with the use of the autocorrelation of each of the different chip sequences. Every sequence of the $N_c$ number of chips using $N_f$ different frequencies was created. For example, all 5040 chip sequences were created for $N_c = 7$, and $N_f = 7$. The autocorrelation was performed on each sequence, followed by the finding of the max sidelobe level.

A sidelobe level criterion of the users’ choosing was implemented; this allowed for different amounts of codes to be outputted. For instance, if the criterion for the levels was $\text{sidelobe} < A$ where $A = 0$, then all 5040 codes would be outputted. However, if the criteria is set to -15.21dB, there would be 1060 codes or if -17.21dB, 160. This program also gave outputs for different $N_c$’s as well as different $N_f$’s with the allowance of one or more $N_f$’s to be repeated. Figure 4-2 below demonstrates the output of the program where $N_c = 7$, and $N_f = 7$. The plot displays the code sequence which gives the lowest autocorrelation sidelobes. This initial step allowed for the narrowing of sequence possibilities. The code set created can now be manipulated, used as an initial set in the next phase, or used as a set in and of itself. Though this narrows the possibilities, it does not show how these codes would interact with each other which led to cross-correlation.
Cross-Correlation

The autocorrelation code set was applied to further narrow the results to create an \( M \) code set, where \( M \) is the number of targets that would be used simultaneously. Next, a set of codes was created that had low cross-correlation sidelobes. This can be done with any amount of codes desired. Due to time limitations, however, only codes created from the sets obtained from a set with no more than two interactions would be used. For example, consider the following two sequences: 1237654 and 4512367. This pair possesses a group of three sequentially in common during correlation and would therefore be discarded. The program obtains each code selected and correlates it with all other selected codes. It then checks the sidelobe levels versus the criteria created. When the sidelobe of the cross-correlation of the two codes reaches beyond the max, the corresponding crossed code is discarded. This step is repeated for each of the different codes until only one set per code exists. It is necessary then to
narrow the sets down even further in order to be confident of the result. All sets are compared to see which codes set $A_1$ had in common with set $A_2$. Any not in common are discarded from set $A_1$. Then set $A_1$ and set $A_3$ are compared and any not in common are discarded. This process is then repeated until the loop breaks when there are fewer than $C$ codes where $C$ is the desired number of codes. Once this is achieved, one can then be assured that the remaining set has below $X$ level sidelobes when correlated with code $A_1, A_2, ... A_N$ where $N$ is the set where the loop. If $N > C$, the codes fit the required criterion.
Figure 4-3: Block diagram of the program used to find the code sets with the lowest cross-correlation sidelobes.
**PN Coding**

After utilizing the cross-correlation code sets with no PN coding sequence, a way of implementing the PN logically was needed. As previously discussed, the optimal PN codes are well-known in forms such as Barker Codes and Gold Codes. Gold codes are not realizable in the system as they are utilized for continuously changing scrambling sequences. The Barker codes are an optimal code set with respect to cross-correlation sidelobes for short code sequences. However, the code set is extremely limited. Though it has been shown to work for CDMA, in 802.11 for example, multiple code sequences for a fixed chip length do not exist [19]. The application to OFC shows no distinct advantages. It was determined however that if a Barker code existed for the number of chips $N_c$ being utilized that the Barker code would be used. The Barker code is determined to provide the lowest number of interactions and therefore the lowest sidelobes during autocorrelation. When a $N_c$ is chosen that is not one of the lengths of a Barker code, the use of customary PN coding techniques is inefficient or nonexistent. Maximal length codes or m-codes are easily implementable but not optimal PN sequences. Although Gold, Walsh, and Kasami sequences derived from maximal length codes show increased performance, they are not suitable for use in OFC SAW devices because of the small number of chips, usually less than ten. Consequently, sidelobe reduction was the decided route of implementation.

The use of autocorrelation sidelobe levels was utilized in the PN code creation process. The user inputs the codes in the system and the program returns the PN sequence that provides the smallest autocorrelation sidelobes for each code.
Figure 4-4: Block diagram of the program used to find the PN code with the lowest auto-correlation sidelobes for a given input code set.

While the previous attempts created slightly better results, a clear distinction between the purely random case and these optimized codes was not apparent. Thus, further study was needed in addition to alternative coding approaches while maintaining the benefits of the PN-OFC.

**Spatial Division and Frequency Division Multiple-Access**

Spatial Division Multiple-Access

Spatial division multiple-access (SDMA), also called multiple-beam frequency reuse, allows the frequency allocation to be used fully between two users at the same time. For instance, a dual beam
antenna feeding two receivers allows the simultaneous access to the satellite from two different receivers on earth [17]. Or, a base station in a cellular network can utilize multiple beam formed antenna instead of a single dipole in order to increase gain in a particular direction as well as separate between different users and increase the potential capacity [16].

Figure 4-5: An example of the use of spatial division to account for multiple simultaneous users in a cell phone tower scenario. Each antenna covers only a certain swath of area around the base station.

A similar technique may be applied to SAW sensors. A receiver with multiple antennas or a phase array antenna allows the receiver to selectively view only certain sensors. This could also be done with an extremely narrow band antenna that either sweeps the area of desired observation or only views certain devices at certain times if the devices or receiver are in motion [57].

The appeal of this method is clear: if one is able to clearly distinguish x number of devices simultaneously, then a spatial division can be performed and the number of viewable devices doubles to 2x. Though clearly a simplistic case, it still proves noteworthy. The use of this type of multiple-access
depends on the application. A static application versus dynamic or a fixed versus an open ended system creates advantages to the use of this type of multiple-access.

Frequency Division Multiplexing

While SDMA has specific benefits, it does not provide universal application desired. This led to the investigation of other types of multiple-access. As previously shown, frequency division multiplexing or frequency division multiple-access (FDMA) allocates a user with a specific frequency band over its entire time of usage. The concept of breaking the frequency into parts for each user to use is not new. The use of frequency division combined with OFC required observation to determine potential anti-collision properties.

The investigation presented here involves the utilization of 30% fractional BW at 250 MHz with 9 OFC chips. The first four chip frequencies were allocated to half of the devices and the other five frequencies were allocated to the other half. The targets were also separated into two groups of different delay offsets. The start of the first sensor group’s return signal is referred to as reference zero time, and the second group of targets delay starts at \(11 \tau_e\) respectively. The frequency allocation of the two signals is illustrated in Figure 4-6. The red line indicates the bandwidth occupied by one group of devices and the blue line is the second group.
Figure 4-6: A plot showing the normalized frequency versus amplitude in dB. A different frequency bandwidth is assigned to two different code sets. The bandwidth is utilized by nine orthogonal frequencies. One half of the devices utilize the lowest four orthogonal frequencies and the remaining devices use the upper five.

The benefit of utilizing such a scheme rests in the reduction of collisions obtainable by dividing the frequency spectrum in half. This benefit, however, needs to be weighed against the reduction in time, bandwidth product, or processing gain of the potential device that utilizes the entire bandwidth simultaneously. The error percentage versus number of simultaneously simulated devices is shown in Figure 4-7.
Figure 4-7: Error rate simulation of the system utilizing 4 and 5 frequency chips which are in different frequency bandwidths. For a small number of devices this approach works well, however when an increasing number of devices begin to occupy the same time and spectrum the errors begin to rise.

The benefit for a small number of devices is clear, but as the overlap increases so do the errors. The use of frequency division is readily applied to current OFC technology. The design need only account for two different center frequencies when calculating the OFC chips. As such, a thorough investigation into experimental testing for verification of device versus ideal data is not needed.

In-Band Frequency Division Multiplexing

The proposed solution takes the center frequency of interest and the allocated bandwidth and uses them to create two separate orthogonal frequency sets. Recall that the basis functions require

\[ f_n = \frac{n}{\tau} \quad \text{and} \quad f_m = \frac{2m + 1}{2T} \] (4.1)
Accordingly, two totally different sets can be created from the same center frequency and bandwidth. For example, in Figure 4-8 the basis sets were used to create an orthogonal frequency set of five frequencies and another of four frequencies. These sets are not orthogonal to each other but are orthogonal within themselves. This creates two distinct sets that can be used on separate devices in order to utilize fully the orthogonality within the chips.

\[ \tau_B = J \cdot \tau_c \text{ where } J = \# of chips \]  (4.2)

Figure 4-8: Representation of two orthogonal frequency sets occupying approximately the same bandwidth around the same center frequency. One set is created with four frequencies and the other using five.
\[ g_{\text{bit}}(t) = \sum_{j=1}^{J} w_j \cdot h_{cj}(t - j \cdot \tau_c). \quad (4.3) \]

For set one the functional description of the chip is

\[ h_{1 cj}(t - j \cdot \tau_c) = b_j \cos \left( \frac{2C_j \pi(t - j \cdot \tau_c)}{\tau_c} \right) \text{rect} \left( \frac{t - j \cdot \tau_c}{\tau_c} \right) \quad (4.4) \]

with each chip having a single local carrier frequency of \( f_{1 cj} = \frac{C_j}{\tau_1 c} \). For set two the functional description of the chip is

\[ h_{2 cj}(t - j \cdot \tau_c) = b_j \cos \left( \frac{2(C_j + 1) \pi(t - j \cdot \tau_c)}{\tau_c} \right) \text{rect} \left( \frac{t - j \cdot \tau_c}{\tau_c} \right) \quad (4.5) \]

with each chip having a single local carrier frequency of \( f_{2 cj} = \frac{2C_j + 1}{2\cdot\tau_2 c} \). Recall generically the bandwidth of the bit is

\[ BW_{\text{bit}} = J \cdot 2 \cdot \tau_c^{-1}. \quad (4.6) \]

Since the sets have different taus and set one has a \( J \) of five while set two uses a \( J \) of four they may still occupy approximately the same bandwidth but utilize different orthogonal chips.

**Same BW Same Center Frequency**

A simple analysis was performed to determine what kind of cross-correlation sidelobes the proposed solution would create. The time and frequency response of the two separate signal sets can be seen in Figure 4-9 and Figure 4-10, respectively. The frequency plot shows that they occupy approximately the same bandwidth around the same frequency.
Figure 4-9: Time response of two separate signals one utilizing 5 orthogonal frequencies (blue) and one utilizing 4 (red) both centered around 250MHz.

Figure 4-10: Frequency response of two separate signals one utilizing 4 orthogonal frequencies (blue) and one utilizing 5 (red) both centered around 250MHz.
A comparison was done using cross-correlation technique. A single frequency from set one consisting of just four frequencies was crossed with set one and two to observe the correlation between the frequencies. The resulting cross-correlation of the single frequency with set two, the different bandwidth code set, is shown in red in Figure 4-11 while the correlation of the frequency with set one, its own set, is shown in blue.

![Cross-correlation Response](image)

Figure 4-11: Cross-correlation of one frequency from the five frequency set crossed with both signals shown above. The dotted blue shows the correlation of the chip with its own frequency and the others from its orthogonal set. The red shows the correlation of the chip with the four chips from the other orthogonal frequency set.

The largest peak shown in blue is the frequency crossed with itself, similar to an autocorrelation. The largest of the red peaks is similar in shape to that of the blue peak. This is the most similar frequency from the four OFC set creating a large cross-correlation peak. It is unclear from this analysis if the utilization of the in-band frequency diversity will yield better results.

A simulation was run using the new parameters of one-half of the targets utilizing five frequencies in their code and the other half using four. By design, the first 8 targets would use five frequencies and then targets 9-16 would use four frequencies and so forth to 32.
Figure 4-12: Error rate of coding using the same frequency spectrum with different chip bandwidth allotments corresponding to utilizing 5 frequencies in one set of codes and 4 in the other. The devices were split in half with every set of four utilizing a different code set, e.g. the first four devices use five frequencies in their code set and then the next four devices use the code set with only four frequencies.

The results seem to indicate limited benefit to this type of frequency division. There is some benefit when the number of targets is low but not as advantageous as desired. The overlapping of the signals in time and frequency appears to be critical in the elimination of peak distortion.

When the frequency is non-overlapping the results are superior to the overlapping case. There is still a limit to the amount of non-overlapping targets one can produce in FDMA. There will still be a certain amount of overlap once the number of targets exceeds the available bandwidth. These results lead to the investigation of limiting the overlap in time.

**Time Division Multiplexing and Time Division Duplexing**

As stated previously, time division multiplexing (TDM) is a useful way to share a common spectral resource with multiple users. Each of the $M$ signals or users is given full use of the spectrum for
a short duration. This technique allows for a greater number of users to operate on the same frequency band without increasing the bandwidth. A customary TDM implementation requires synchronization or a signaling bit in order to determine the beginning and end of the periods [17]. While the concepts for use in conjunction with OFC come from this characteristic TDM signaling, OFC TDM does not necessarily contain all of these components.

TDM Basics

OFC TDM makes use of multiple time slots each with identical time length $\tau_c$. For example, in one period $T$ there are $10 \tau_c$ length time slots. The period is determined by limitations either in device length or interrogation repetition time. Thus, the governing equations for orthogonality do not change from the original OFC. However, the equation that describes a chip would change to the form:

$$h_{chip}(t, f_{0j}, j) = h(t - (j + k_j)\tau_c, f_{0j})$$

(4.7)

with the new term $k_j$ encoding into which $\tau_c$ slot each chip is placed. Instead of being continuous as with OFC, the chips can be offset by an integer number of $\tau_c$. 
In the example seen in Figure 4-13, each user (or, in this system, each sensor) has been placed in alternating time slots. There are two sensors each with a different shuffling of their frequencies in time (OFC) and alternating time slots (TDM). Sensor 1 is allocated time slots 1, 3, 5, 7, 9 and is seen in solid red, while Sensor 2 is allocated slots 2, 4, 6, 8, 10 in dotted blue. This approach allows for frequency spreading using orthogonal frequencies and time spreading using time division to reduce collisions. This illustration shows how an OFC coding implementation can be overlaid with a TDM system. Note that while TDM spreads the chips in time, the frequency allocation remains the same. Also, because there are times where the signal is zero, there is no effective increase in power transmitted from OFC.
Multi-User Detection: A Comparison

Robust solutions to multi-user, passive, wireless RFID are difficult and often technology-dependent. Some multiple-access techniques have been applied to SAW RFID, such as CDMA, but have limitations with code collisions in a multi-tag environment.

The possible TDM combinations that could be formed will not be examined at length since the number of possibilities is not as important as the anti-collision properties. For a simple TDM example, observe 1_2_3_4_ and _1_2_3_4 where each number and each underscore represents a chip length. Arranged in this way, the same PN-OFC code would not overlap. This is essentially the basics of a TDM scheme where each RFID sensor is given specific time slots where its frequencies can be placed. This creates a desirable feature, namely minimal overlap.

Figure 4-14 displays a plot of the number of possible PN-OFC codes versus CDMA coding, assuming the chips are contiguous in time. Also shown is the number of codes for a sample TDM with N frequencies and 2N+1 time slots. While this shows the vast number of possible codes that can be introduced, it does not account for the usability of the code. Criterion for usability has not been established for PN-OFC or PN-OFC-TDM. Simple logic was employed to empirically create fixed code sets that would work well together.
A system simulation was performed for the OFC-PN and then OFC-PN TDM for ideal theoretical comparison. Five orthogonal frequencies with 11 TDM slots compared to the five frequencies with just 5 slots is shown in Figure 4-15. The code sets were considered as a fixed set for a static sensor system setup.

This provides a simple visual to view the performance possibilities. The performance was based on peak location criteria. When a peak was not found within ±5 samples of its prescribed location, it was determined to be in error. When the simulation is run multiple times each time varying the random time parameter, a percentage of errors can be established.
Figure 4-15: Comparison of the percentage of erroneous peaks observed during system level simulation. OFC is seen in red and OFC TDM seen in blue. The number of bits in the system is equivalent to the number of devices or sensors. An error is a variation greater than 2.5 ns.

OFC TDM Device Design

As with OFC SAW work previously reported, OFC TDM can be applied to communication and tagging applications, as well as sensors. The following describes a tag designed at 456 MHz composed of a PN coded 5 chip, 5 frequency, OFC TDM with 11 time slots. Each chip’s time length is 54.7 ns with a 3dB bandwidth of 18.2 MHz with a total device 3dB bandwidth of 91 MHz. The reflectors are assumed to have equal reflectivity and a rectangular time function response; the device schematic is shown in Figure 4-16.
The input transducer is assumed wideband and its effect generally negligible for the example. The ideal OFC tag responds to a pulsed interrogation signal with its impulse response shown in Figure 4-17. This response has uniform amplitude and is \( 11 \tau_c \) long. The signal response is a wideband spread spectrum signal with the inherent benefits of security and processing gain. Because of the orthogonal nature of the chips and time diversity, there is no inter-symbol interference within the device. At the receiver, the post-processor makes use of a matched filter to create the compressed pulse shown in Figure 4-17. The compressed pulse is \( 0.2 \cdot \tau_c \) long, yielding an OFC processing gain of 25.
Figure 4-17: (Top) Ideal PN-OFC-TDM impulse response signal ideally returned from the SAW device embodiment. (Bottom) Simulated matched filter autocorrelation of the signal. While increasing diversity, PN OFC-TDM does not add processing gain to the code.

OFC TDM Sensor Design

The OFC TDM concept is readily applied to SAW-sensing applications. The resulting system has the advantage of having simultaneous tagging and sensing. The sensor embodiment utilizes two identical reflector banks on either side of the transducer as shown in Figure 4-18. The two reflector banks have differing delays on either side of the device, $\tau_1$ and $\tau_2$. When applied in this way, two
compressed pulses will result from the matched filter operation. Using this differential time delay, sensed information can be extracted.

![Figure 4-18: OFC SAW sensor schematic. Relative grating periodicities are shown for each chip. The shuffling of these chips generates the OFC code sequence. The device operates in differential mode for sensing applications depicted as two different delays on either side of the transducer.]

While constructing the SAW systems coding scheme, the reduced collisions and improved performance are positive outcomes. Utilizing the SAW coupling of modes (COM) model, device performance can be accurately obtained which allows the devices to be characterized thoroughly before fabrication. The differential mode sensor schematic is given in Figure 4-16. Notice the separation between the different chips as representation of the TDM signal.

As a demonstration, a 456 MHz PN-OFC-TDM differential embodied sensor having five chips was fabricated on an YZ LiNbO3 substrate. The COM simulation and measured device time response is shown in Figure 4-19. The agreement is excellent; observed chip amplitude variations are due to the transducer bandwidth effects, as predicted.
The experimentally measured sensor S11 response was then entered into the simulation transceiver. The result shown in Figure 4-20 illustrates the expected multiple correlation peaks. Good correlation exists between the theoretical and expected result. These results provide sufficient evidence for the use of OFC TDM in SAW sensor applications.
Figure 4-20: This plot shows the compressed pulses from a differential mode device similar to the one shown in Fig. 11. The experimental data shown yields comparable correlation peaks to the ideal shown in Figure 4-17.

The addition of TDM OFC provides decreased code collision, which translates to more useful tags and an even larger number of code sets. The approach provides a novel coding technique, advantages associated with spread spectrum, and the ability to simultaneously tag and sense temperature in a multi-sensor environment. While this technique did provide additional accuracy from decreased peak distortion, one must use very long devices to reduce the errors with a large number of devices. Further work is needed in coding techniques to reduce collision errors and determine an optimal system configuration, where device size versus performance must be evaluated.

**Time Division Duplexing**

The application of time division within a SAW sensor system is slightly less restrictive than in other applications. The SAW sensor can be allowed some overlap in time without causing a total loss of information. This is helpful for SAW sensors because time division implemented in SAW is done in space on the substrate. To clarify, the more time blocks used, the larger the substrate needed to implement them.
The use of time division in this way allows for variations within the implementation. The sensors’ overlap can be varied based on end user requirements.

The mathematical description of the returned signal from a single OFC sensor has been derived previously and given as

\[
g(t) = \sum_{j=1}^{J} w_j \cdot h_{ej}(t - j \cdot \tau_c).
\] (4.8)

For time division duplexing (TDD), the signal time function would be subject to the placement in time with respect to a reference. Each signal time function would then look like

\[
p(t) = g(t - \tau_D)
\] (4.9)

where \(\tau_D\) is the delay of each signal. \(\tau_D\) can have any value but is generally confined to a time slot corresponding to an integer number of \(\tau_c\). This allows for a prescribed offset that can be easily varied in order to observe changes in the number of errors. Consider
\[ total(t) = \sum_{k=1}^{M} g_k(t - i(k - 1)\tau_c) \] (4.10)

where, \(i\), is the offset parameter, and \(M\) is the total number of sensors. This represents the total returned signal for the ideal case.

**TDD OFC SAW Simulation**

When this encoding scheme is used with assumed synchronous reception, the number of erroneous peaks that occur is nil. This would be an ideal case of multiple target responses returning orthogonally. The example shown below contains a repeated set of three codes offset two chips. Thirty-two targets are placed in the field of view of the antennae for this simulation. The correlation peaks shown in Figure 4-22 are unique from other implementations in that each peak is separated from the others. In all previous applications, the peaks would coincide if no spatial delay was present.
Figure 4-22: Matched filter correlation peaks of a repeated code set. The code set used five frequencies with three different OFC-PN codes repeated until the desired number of devices was met. The simulation shown was a synchronous idealized case study.

When asynchronous reception is implemented, errors begin to emerge. These errors, however, level off to some value regardless of the number of targets placed in the system simultaneously. This error level is directly dependent upon the number of overlapping chips, and thus the larger the overlap, the larger the percentage of errors. The example below includes the same repeated code set pictured in the example above but given random offsets using the +1m relationship developed earlier.
Figure 4-23: Matched filter correlation peaks of the same repeated code set mentioned previously. Two different runs are presented each with different random offsets placed on each individual device. The random offsets depicting sensor placement and temperature variation. Notice the amplitude does vary but the main peaks still remain apart from the remaining devices.
There is a plateau of percentage error despite increased number of devices being utilized in the system. The simulation was performed at 250 MHz utilizing 28% fractional bandwidth with 5 OFC chips. An overlap of 2 chips was allowed and the devices were placed at a 5 meter arch ±0.5 meter.

The results of this type of implementation clearly show the possibility of having as many sensors in the field of view of the transceiver as desired. The limitations concern device size. Additionally, significant on substrate propagation losses become a factor at higher frequencies. Clearly, the system designer determines the tradeoff of performance versus inhibiting factors. While zero-chip overlap would most assuredly create negligible errors, this design would create the largest spatial device set when a large set is implemented.
TDD SAW Device Considerations

If the number of chips that each device offset with respect to one another is $M$, then the final device in the set would be at time

$$\tau_{final\ device} = (N - 1) \cdot M \cdot \tau_c + P \cdot \tau_c$$

(4.11)

where $N$ is then number of devices, $P$ is the number of OFC chips, and $\tau_c$ is the chip length. This would be the time allotted for the last device in the set. A device of this maximum time would correspond to an on substrate spatial delay of

$$delay = v_s \cdot \frac{\tau_{final\ device}}{2}$$

(4.12)

where $v_s = substrate\ velocity$ which varies by substrate (3488 m/s on Y/Z LiNOb3 for example). If constrained to a certain packaging device, the system designer can easily assess the number of possible devices vs. offset. An example of device length versus number of devices for a $\tau_c$ of 100ns is shown in Figure 4-25.
Figure 4-25: A comparison showing the dependency of device length to number of devices in system for varying number of chip offsets when utilizing OFC-PN-TDD. Shown is device length corresponding to a set of devices utilizing 250 MHz center frequency and 20% fractional bandwidth. Each device has a 0.5 us initial delay, 5 OFC chips and is assumed fabricated on YZ-lithium niobate.

For instance, examine an N=20 sensor device set with an offset of 2, 5 OFC chips, and a chip length of 100ns. The result yields a time delay of 4.3us, which would be 0.75 cm on substrate. This value is for a single-sided device, meaning only one reflector bank. An in-line differential type device would be approximately twice as long.

Performance Comparison

The system level simulations were performed using MATLAB. The system is comprised of 32 SAW sensors. The devices are assumed to be 5m +1m from the TX/RX. A matched filter receiver type
structure is used and an example MF peak is shown in Figure 4-26. A peak search algorithm is used to obtain matched filter peaks for comparison with known delay values. This allows the designer to determine if the scenario presented to the sensor would yield false readings or even no reading at all.

Figure 4-26: A comparison of matched filter responses from the returned signal comprised of 32 devices and the ideal single device returned signal. The main peak still remains quite clear even when 32 devices are present, representing the distinct advantage of using time division when trying to distinguish between different sensors. Each of the devices utilizes five OFC frequencies and five chips. Center frequency of 250 MHz with 28% fractional bandwidth is used with a two TDD offset in place between the devices.

The returned signal decision matrix is performed just as it was previously to determine an error percentage. A similar result to previous examples is obtained and can be seen in Figure 4-27.
Figure 4-27: Error percentage of COM system simulation with varying number of targets placed in the FOV simultaneously. The simulation was performed at 250 MHz utilizing 28% fractional bandwidth with 5 OFC chips. An overlap of 2 chips was allowed and the devices were placed at a 5 meter arch ±0.5 meter.
CHAPTER 5: CONSIDERATIONS FOR OPTIMAL SYSTEM DESIGN

Potential range, frequency and bandwidth use, coding for multi-user reception, and decoding accuracy are all challenges present in designing an optimal OFC sensor system. The range is generally described using a radar equation [74]. Frequency and bandwidth allocation are functions of not only substrate and SAW parameters but also the transceiver operation. The coding type will directly affect the number of simultaneous sensors that are viewable. Decoding accuracy has not yet been addressed but will dictate to what extent the sensed information is deemed reliable.

Range of System

In determining the range within the system simulation, the power obtained at the receiver is dictated by Frii’s equation. Examining this simple equation provides a look into how a change in center frequency would affect the overall system. The basic form of Frii’s one-way transmission equation [75] takes the form of

\[ \frac{P_r}{P_t} = G_t G_r \left( \frac{\lambda}{4\pi R} \right)^2, \]  

which can then modified for a radar like sensor to

\[ \frac{P_r}{P_t} = G_t G_r G_{SAW} \left( G_{SAW,Ant} \left( \frac{\lambda}{4\pi R} \right)^2 \right)^2, \]  

in order to include the two-way transmission through the channel and the gain (or loss) of the SAW device, \( G_{SAW} \), and the SAW device antenna, \( G_{SAW,Ant} \).

Recall from Equation (3.1) that the EIRP is \( P_t G_t \), which is equivalent to the maximum power transmitted from the transceiver, now referred to as \( P_{t_{max}} \). The device components, \( G_{SAW} \) and \( (G_{SAW,Ant})^2 \), can be lumped into a single gain parameter, \( G_{target} \). As stated previously, it is
also common to collect the propagation loss terms together as in Equation (3.8), but now the term is squared for two-way propagation, \( G_{prop} = \left( \frac{1}{L_s} \right)^2 \). As such, Equation (5.2) may be rewritten as

\[
\frac{P_r}{P_{t,max}} = G_r G_{prop} G_{target} .
\] (5.3)

If the gains of the antennas are set to unity the power received is dictated primarily by the channel propagation loss. The transmitted power and SAW device gain can be set arbitrarily in order to view the effect the center frequency has on received power shown as

\[
P_r = C_{prop} = \left( \frac{\lambda}{4\pi R} \right)^4 = \left( \frac{v_{em}}{4\pi} \right)^4 \left( \frac{1}{f^4} \right) \left( \frac{1}{R^4} \right)
\] (5.4)

where \( v_{em} \approx 3 \cdot 10^8 \text{ m/s} \), or the velocity electromagnetic wave through a medium, in this case assuming air. This simply means that in dB there would be a 40dB loss per decade for a given range as the frequency increases. Also, there would be a 40dB loss per decade for a given frequency as the range increases as shown in Figure 5-1.
Figure 5-1: The power received normalized to the transmitter power at a transceiver from a target that is a specified distance away. The power decays steeply very close to the transceiver creating a large difference in the power received from a target a few meters away up to 20 meters away.

When observing not merely the power received but also trying to determine a maximum range based on transceiver characteristics, a more encompassing equation must be used. The maximum range, $R_{\text{max}}$, is dictated by the radar equation [62] with the cross-sectional reflector replaced by the target information. The altered radar equation is given as

$$R_{\text{max}} = \frac{\lambda}{4\pi} \sqrt{\frac{P_t G_t^2 G_{\text{SAW, Ant}}^2 P_G N_{\text{sum}}}{N F G_{\text{SAW}} N_{\text{tot}} SNR}}$$  \hspace{1cm} (5.5)$$

where the new parameters are: the processing gain, $P_G$, the noise figure, $N F$, the number of integrations if any, $N_{\text{sum}}$, the total noise, $N_{\text{tot}}$, the required signal to noise ratio for correct interpretation of the signal, $SNR$, established previously in Equation (3.21). As shown in Equation (5.5), the range directly varies according to the wavelength. One can then clearly see that the frequency will have the greatest impact on the potential range of the system. When calculating the potential range parameter, one can also examine the impact of the signal to noise ratio (SNR). The SNR as defined by the radar equation is equivalently called $SNR_{\text{min}}$, the minimum detectable SNR. The required SNR will
primarily depend upon the type of coding used. This required SNR will be established based on minimum requirements for discerning the signal and resulting correlation. The SNR will have an impact upon the maximum resolution of the sensed information.

Figure 5-2: The range is plotted as a function of frequency while varying the minimum acceptable SNR. The SNR as given in the legend is user defined based on detection criteria for acceptable errors.

Figure 5-2 illustrates that the potential range for a low frequency signal can be quite high; therefore, one might want to determine the dynamic range of the system. The dynamic range differs from the maximum range as it dictates the near and far of a discernable signal. Unlike cellular or other wireless communication systems [27], it is not possible to instruct a sensor to decrease the amount of reradiated power. Once in place, it acts as a dummy receiver, converter, and reradiator. Therefore, a maximum dynamic range that is discernable exists for any given system.

The receiver dynamic range is a function of the hardware. Typically, the ADC will determine the final system dynamic range, usually established at the detector. The dynamic range of an ADC is found by taking the peak signal power and comparing it to the quantization noise power. Since these are the theoretical largest and smallest values that can be observed by the ADC, the same procedure is
performed in order to obtain the theoretical best SNR. The quantization noise power is equal to $q^2/12$ with $q$ being the quantile level [17]. The peak power of the analog signal normalized to 1 $\Omega$ can be expressed as

$$V_p^2 = \left( \frac{q(2^N - 1)}{2} \right)^2 \approx \frac{q^2 2^{2N}}{4}. \quad (5.6)$$

The peak signal power to quantization noise power is then defined as

$$\frac{S}{N} = \frac{q^2 2^{2N}}{4} \approx \frac{q^2}{12} = (3)2^{2N}. \quad (5.7)$$

This expression is more commonly seen in the dB form of

$$SNR_{DR} = 6.02N + 1.76 \text{ dB}, \quad (5.8)$$

where 1.76dB rather than an expected 4.77dB is used to account for the use of a sinewave input instead of a simple peak power steady signal [20, 76, 77]. The dynamic range is then clearly a function of the number of bit resolution of the ADC without accounting for scale. This communicates to the receiver designer that a larger resolution ADC must be used if a higher dynamic range is desired. For an 8-bit ADC, the theoretical best SNR is 49.92 dB, which also corresponds to the best possible dynamic range of this ADC.

When the dynamic range is defined by choosing the detector, the range is spatially defined. When there is specified a maximum range desired, there also exists a minimum, distortionless observed distance. Any targets observed further than the maximum would be considered below the minimum detectable signal (MDS) and any targets closer would saturate the ADC. This creates the case where

$$MDS < P_6 < MDS + SNR_{DR} \quad (5.9)$$
where $P_0$ is defined as the power input into the detector shown previously in Figure 3-3. Illustrated in Figure 5-3, a point on the line corresponds to the near and far distance. For example, the closest the ADC can resolve is a target at 3m, for an eight bit ADC, when the range to the far target is 60 meters.

![Target Near Far Dynamic Range](image)

**Figure 5-3**: The dynamic range allowable by different resolution ADC’s detectors. The range to the closest target illuminated by the receiver dictates the maximum distance the receiver can see without saturation.

While the ADC determines the ultimate dynamic range of the receiver, it does not inevitably decide the dynamic range of the system. Because of the inability of the individual targets to adjust their output power judicially, a potential issue arises within the correlation peak detection. For example, two targets are placed in the FOV of the receiver, one at 5m and the other at 10m. Figure 5-4 displays the correlation peaks of the target at 10m. The blue line represents the autocorrelation peak of the target and red the cross-correlation. The chips of each of the devices are intentionally separated for readability. The autocorrelation peak of the target in this case does not compare to the cross-correlation sidelobes from the target at 5m.
Figure 5-4: The depiction of a cross-correlation and an auto-correlation with amplitude variation placed on the received signal while the matched filter remains constant. The cross-correlation of the device with a near target with greater received power would “drown out” the autocorrelation signal from the further device. The simulation was done utilizing two devices with TDM coding to space out the chips.

Given a prior knowledge of system implementation, it is possible to alleviate some of the dynamic range issues. One can always put an attenuator on the devices that are known to be close to the receiver in order to more easily view the further sensors. Of course, this would also limit the use of the proposed system since it would have to be stationary as well as known beforehand.

Analysis was performed upon the impact of the dithering of the sensors around an arch in space and the results shown in Figure 5-5. The targets were allowed to vary within a certain percentage of a center arch at ten meters from the transceiver. When the entire set of sensors coincided on the same ten meter arch, e.g. zero percent variation, the results were predictably quite good since this represents an ideal system, e.g. no random overlap. The targets were allowed further separation (10 %, 20 %, etc.) until it became apparent that the peaks were no longer distinctly visible as half of the targets or more performed poorly. The simulation design parameters are outlined in Table 5-1.
Table 5-1: The system simulation design parameters used when simulating a variation in range. Unless stated otherwise, the center frequency, fractional bandwidth, number of frequencies, and number of chips will remain unchanged for the remainder of the simulations.

<table>
<thead>
<tr>
<th>OFC System Simulation Design Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center Frequency</td>
</tr>
<tr>
<td>Fractional Bandwidth</td>
</tr>
<tr>
<td>Number of OFC frequencies</td>
</tr>
<tr>
<td>Number of chips</td>
</tr>
<tr>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>Coding type</td>
</tr>
<tr>
<td>Number of chips offset</td>
</tr>
</tbody>
</table>

Figure 5-5: Simulation analysis of the effect of the dynamic range on the number of errors for a given sensor scenario. The sensors were dithered a percentage around a fictional arch in space and the number of errors were observed for different numbers of sensors viewed simultaneously. For example, around a 10m arch in space, a range variation of 10% is equivalent to ±1m from the arch, or 9m to 11m. A device was considered erroneous if further than five sample points from the known location where a sample point is 0.5 ns.

**Target Considerations**

When decreasing the potential range due to the radar equation, an increase in frequency can provide benefits to the target. The most prevalent benefit would be a decrease in the footprint of the antenna needed to achieve the desired bandwidth. The use of an electrically small antenna (ESA) is of
considerable interest on the target. The target is desired to be diminutive for use as a sensor; therefore, the entire embodiment must be relatively small.

For an ESA, the $k \cdot a < 1$ criteria should be met. In many SAW OFC applications, $k \cdot a < 0.8$ will be adequate which yields a gain of approximately 4 dB and %BW=22%. At 1 GHz, a~3.8 cm while at 100 MHz a~38 cm. The equations reveal the existence of a minimum size at a given frequency to attain a desired fractional bandwidth. As the frequency increases, a larger fractional bandwidth is achievable for a smaller antenna size. As the effective size of the antenna increases, the gain and bandwidth both increase. From Equations (3.10)-(3.12), it is shown that a decrease in the frequency will allow for additional gain and bandwidth within the same area. Consequently, for antenna design and small devices, higher center frequency is better.

Another consideration pertaining to the target would be which coding technique to operate. It is apparent that OFC-PN offers many distinct benefits over CDMA when creating a sensor. Utilizing a YZ lithium niobate substrate, the SAW tag losses are all assumed to be the same except for the reflector loss. The OFC reflectors possess low loss because of the reflector’s orthogonal nature, making each chip nearly transparent to all other chips. The CDMA reflectors, however, are all at the same frequency and therefore must have low reflectivity in order to penetrate to the last reflector chip. Based on good design principles for the reflectors, the anticipated reflector loss for CDMA is ~20 to 60 dB depending on design and for OFC is ~3 dB [45, 78]. A summary of the design results is shown in Table 5-2. Based on analysis at 1 GHz, the SAW OFC design has approximately 17 dB less loss than the approximately equivalent CDMA design.
Table 5-2: SAW tag design consideration, comparing OFC to CDMA.

<table>
<thead>
<tr>
<th>SAW Parameters</th>
<th>CDMA</th>
<th>OFC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center frequency (GHz)</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>%Bandwidth</td>
<td>3.60%</td>
<td>25%</td>
</tr>
<tr>
<td># of chips</td>
<td>19</td>
<td>7</td>
</tr>
<tr>
<td># of codes</td>
<td>.52 M</td>
<td>.65 M</td>
</tr>
<tr>
<td># of frequencies</td>
<td>1</td>
<td>7</td>
</tr>
<tr>
<td>Processing Gain</td>
<td>19</td>
<td>49</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>SAW Tag Losses (dB)</th>
<th>CDMA</th>
<th>OFC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reflector loss</td>
<td>20</td>
<td>3</td>
</tr>
<tr>
<td>Propagation loss</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>Bidirectional loss</td>
<td>6</td>
<td>6</td>
</tr>
<tr>
<td>Parasitic loss</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td><strong>Total Loss (dB)</strong></td>
<td><strong>32</strong></td>
<td><strong>15</strong></td>
</tr>
</tbody>
</table>

OFC-PN with TDM and TDD provide several distinct advantages to the device design in addition to those inherit within OFC-PN. An additional criterion for comparison is needed to determine which is optimal. The number of viewable targets correctly received is a promising review mechanism. There must be a determinable lower end on the SNR in order to distinguish the targets. Also of consideration is the device size relative to its performance. For example, the device can be quite large when using TDD if little overlap is allowed. The precision of the any proposed technique is also a common concern. Though the target may be perceived correctly, continued accuracy under changing circumstances is still in question. These criteria will be evaluated in an attempt to determine an optimal system setup.

The various coding techniques were exposed to different criteria simulations. Each simulation utilized the same center frequency of 250MHz, the same approximate fractional bandwidth of 28%, and all other simulation variables remained the same unless stated otherwise. Figure 5-6 compares the different time multiplexing schemes as they were deemed the most effective in elimination of collision.
Table 5-3: Simulation design parameters of a fixed number of targets, fixed SNR, and with varied coding types.

<table>
<thead>
<tr>
<th>OFC System Simulation Design Parameters</th>
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<tbody>
<tr>
<td>Center Frequency</td>
</tr>
<tr>
<td>Fractional Bandwidth</td>
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<td>Number of OFC frequencies</td>
</tr>
<tr>
<td>Number of chips</td>
</tr>
<tr>
<td>Signal to Noise Ratio</td>
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<table>
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<tr>
<th>Coding Types</th>
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</thead>
<tbody>
<tr>
<td>OFC-PN</td>
</tr>
<tr>
<td>Standard</td>
</tr>
<tr>
<td>Time Division Duplexing</td>
</tr>
<tr>
<td>Number of chips offset</td>
</tr>
<tr>
<td>Time Division Multiplexing</td>
</tr>
<tr>
<td>Number of available chip slots</td>
</tr>
</tbody>
</table>

Figure 5-6: Comparison of different coding techniques under the same system simulation. A SNR of 3 dB and range of 5m ±10% was used to compare the error percentage. The simulation was run 160 times when 16 devices were in the system, to create a better average. A device was considered erroneous if further than five sample points from the known location where a sample point is 0.5 ns.

The most common noise sources presented when studying electronics are thermal, shot, and 1/f noise [79]. 1/f noise is commonly represented as the noise in the electrical resistance of a circuit but was not modeled within the simulation. Thermal and shot noise are white noise processes modeled
using the AWGN noise Equations (3.16)-(3.18) discussed in chapter 3. There are other sources of unwanted distortion that will also be combined into what will be simply referred to as noise. These include the quantization noise from the ADC, discussed previously, and the unwanted signals returning from other targets. The target returned signals will effectively act like noise when cross-correlated with the signal of interest similar to a CDMA system [16]. There is then more noise when more targets are simultaneously simulated. A simulation was performed with the fixed parameters stated in Table 5-4 and varying SNR. The noise effects of multiple devices remain constant while the white noise component is varied. The results are shown in Figure 5-7. The results indicate that there is a signal to noise ratio below which the performance starts to degrade, around minus twelve decibel.

Table 5-4: Simulation design parameters of a fixed number of targets with varied signal to noise ratio.

<table>
<thead>
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<tr>
<td>Fractional Bandwidth</td>
</tr>
<tr>
<td>Number of OFC frequencies</td>
</tr>
<tr>
<td>Number of chips</td>
</tr>
<tr>
<td>Number of devices simultaneously</td>
</tr>
<tr>
<td>simulated</td>
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<table>
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<tr>
<th>Coding Types</th>
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<tbody>
<tr>
<td>OFC-PN</td>
</tr>
<tr>
<td><strong>Standard</strong></td>
</tr>
<tr>
<td>Time Division Duplexing</td>
</tr>
<tr>
<td>Number of chips offset</td>
</tr>
<tr>
<td>Time Division Multiplexing</td>
</tr>
<tr>
<td>Number of available chip slots</td>
</tr>
</tbody>
</table>
Once it was fairly clear the OFC-PN utilizing TDD was allowing for the largest number of correctly received targets, other possible tradeoffs needed investigation. The comparison for device size using OFC-PN-TDD is depicted in Figure 5-8. The use of only five orthogonal frequencies and five chips creates less processing gain and more errors for the zero offset case. This example also clearly demonstrates that when given five frequencies, no great benefit to using more than a two-chip offset exists. That is, unless one is going to eliminate that overlap completely as is the case with greater than five chips. With TDD performing superior, the success criterion was altered slightly from the previous examples to allow distinguish ability.
Figure 5-8: Evaluation of time division duplexing utilizing different offsets. No offset corresponds to OFC-PN and 5 chip offset would mean no overlap barring any distance diversity. A device was considered erroneous if further than four sample points from the known location where a sample point is 0.5 ns.

When considering device length, one must also account for variation in frequency. For even though the device might be very long at a low frequency when using a 5 chip offset, at a high frequency the length might be acceptable. Also of consideration when examining the frequency and device length is on substrate propagation loss. The on substrate propagation loss is defined by

\[ L_p = a \cdot f + b \cdot f^2 \frac{dB}{\mu s} \]  

(5.10)

where \( a = 0.19 \) and \( b = 0.88 \) for Y-Z lithium niobate, and \( a = 0.47 \) and \( b = 2.62 \) for ST-X quartz [37, 80]. For lithium niobate, the on substrate loss as a function of frequency can be seen in Figure 5-9.
Recall from Equation (4.12) that the spatial delay on substrate is due to the velocity of the wave as well as the time required for the wave to reach the furthest device from the transducer. As TDD has given the most enhanced performance, it will be evaluated here. Given that the initial delay is equivalent on all devices, it is imperative then to look at the difference as the frequency changes. The bandwidth $BW$ will change with frequency while the fractional bandwidth, %$BW$, will remain constant at 25 percent to satisfy the ultra wideband criterion. If the number of chips and number of frequencies, $N_f$, is the same and set to 5, then

$$BW_{chip}(f) = \frac{%BW}{N_f} = \frac{0.25 \cdot f}{5} \quad (5.11)$$

and the time length of a single chip is

$$\tau_{chip}(f) = \frac{1}{BW_{chip}(f)} = \frac{1}{\frac{0.25 \cdot f}{5}} \quad (5.12)$$

Now the tau is varying with frequency and can be plugged in to Equation (4.11) and (4.12) to determine how the spatial delay varies with frequency given as
\[
delay(f) = v_s \cdot \frac{(N - 1) \cdot M \cdot \tau_{chip}(f) + P \cdot \tau_{chip}(f)}{2},
\]
(5.13)

where \( N \) is the number of devices, \( P \) is the number of OFC chips, \( M \) is the chip offset, and shown in Figure 5-10.

Figure 5-10: Illustration of the dependency of device length upon center frequency of operation. The TDD coding type is compared at different chip offsets showing a clear increase in device size at low frequency. The velocity, \( v_s \), for lithium niobate is 3488 m/s. The number of devices, \( N \), is 16 and the number of chips, \( P \), is 5.

These results can then be compared directly to help determine an optimal frequency range. Figure 5-11. In Figure 5-11 the TDD offset has been set at 2. With the delay between transducer and reflector now set by the TDD offset (2), number of chips (5), and number of devices (16), the comparison is done between propagation loss in dB and device length in millimeter.
Figure 5-11: A direct comparison of propagation loss shown in solid blue and device length in dotted green versus frequency. The comparison was done on a TDD device with a two-chip offset in place.

Decoding Technique/Peak Location Finder

Although a coding technique was discovered that greatly increased the number of simultaneously viewable targets, a small amount of unexplainable errors remained. To better understand the consequences of a potential error, consider the temperature coefficient of delay (TCD) of YZ-LiNbO$_3$ is approximately 93 ppm/°C.[37, 81] The arbitrary value used to denote an error previously of 5 sample points corresponds to a time of 2.5ns. This yields a potential accuracy for temperature based purely on peak detection of 17°C. With the utilization of zero padding in the post-processing application, it is possible to obtain any accuracy desired. The zero padding, however, adds significant refresh time when obtaining the information. Also, the zero padding will not yield a more
precise result. These potential limitations of the current decoding technique lead to other possible techniques to add precision and accuracy to the result.

Peak Location Estimation Type One

A generic system block diagram is depicted below in Figure 5-12. As previously illustrated, the signals are each individually generated and then summed with AWGN as they pass through the channel into the receiver front end.

![Simulation of System Block Diagram](image)

Figure 5-12: A generic block diagram for an OFC SAW sensor system, depicted with the addition of a possible jammer signal.

The SAW targets signals 1 through N are given generically as

$$h(t) = \sum_{m=1}^{N} g_m(t - \tau_{D_m})$$

(5.14)

where $g_m$ is the generic form of the OFC SAW defined previously, Equation (2.44), and $\tau_{D_m}$ is the delay due to spatial variation. This time delay causes the targets to become asynchronous. Through use of the Fourier transform, the signal can be equally described by its frequency domain representation $H(f)$. 
Consider the frequency representation of the signal just before the receiver front end. The signal can be represented by

\[ H_R(f) = \sum_{i=1}^{N} H_i(f) \cdot \exp(-j \cdot 2\pi f \cdot \tau_{D_i}) + H_{AWGN}(f) + H_{jam}(f) \]  

(5.15)

where \( H_{AWGN}(f) \) and \( H_{jam}(f) \) are the frequency domain representations of the noise and any jammer signal. For simplicity, the frequency variable will be dropped from the notation.

For a single matched filter output, the following would be the frequency domain signal of the filter matched to signal 1

\[ H_R \cdot H_1^* = H_1 \cdot H_1^* \exp(-j 2\pi f \cdot \tau_{D_1}) + \sum_{i=2}^{N} H_i \cdot H_i^* \cdot \exp(-j 2\pi f \cdot \tau_{D_i}) \]  

(5.16)

\[ + H_{AWGN} \cdot H_1^* + H_{jam} \cdot H_1^*. \]

The \([ \quad ]^*\) denotes the complex conjugate of the ideal signal. The delay, \( \tau_{D_1} \), is currently unknown and is estimated initially based on simple peak detection in the first decision block, shown in Figure 5-13. The estimated value is designated \( \tau_{1E} \). An error term is calculated as

\[ H_{Error} = H_R \cdot H_1^* - H_1 \cdot H_1^* \cdot \exp(-j 2\pi f \cdot \tau_{1E}). \]  

(5.17)
The value of $\tau_{1i}$ in Equation (5.17) is then dithered around the originally estimated delay to create a set of estimations $\tau_{1i}$. A matrix of $H_{\text{Error}}$ terms are created each with the different delay, as illustrated in Figure 5-13,

$$H_{\text{Error}_i} = H_R \cdot H_1^* - H_1 \cdot H_1^* \cdot \exp(-j2\pi f \cdot \tau_{1i}) \text{ for } i = 1 \ldots M. \quad (5.18)$$

Each of the terms are then summed over all frequency and contains minimum energy when $\tau_{1i} = \tau_{D1}$ shown as

$$H_{\text{sum}_{E_i}} = \sum_f |H_{\text{Error}_i}|^2 \text{ for } i = 1 \ldots M \quad (5.19)$$

Once all estimated delay values are implemented, a decision is made and a tau selected.

$$\text{find } i \text{ for } (\min(H_{\text{sum}_{E_i}}))$$

$$\tau_{1\text{final}} = \tau_{1i}. \quad (5.20)$$

Now with only one estimate for $\tau_{D1}$, the generic output of MF2 could look like
\[ H_R \cdot H_2^* = H_2 \cdot H_2^* \exp(-j2\pi f \cdot \tau_{D_2}) + H_2 \cdot H_1^* \exp(-j2\pi f \cdot \tau_{D_1}) \]
\[ + \sum_{i=3}^{N} H_i \cdot H_2^* \exp(-j2\pi f \cdot \tau_{D_i}) + H_{AWGN} \cdot H_2^* + H_{jam} \cdot H_2^*. \]  

(5.21)

One could then subtract out both the autocorrelation signal of H2 but also the cross-correlation signal of H2 with H1 yielding

\[ H_{Error} = H_R \cdot H_2^* - H_2 \cdot H_2^* \exp(-j2\pi f \cdot \tau_{D_2}) - H_2H_1^* \exp(-j2\pi f \cdot \tau_{1,final}) \]  

(5.22)

The dependency upon the estimated MF delay is critical. Accordingly, it was necessary to determine the significance of an incorrect estimation if the second subtraction were also implemented.

*Estimation Effects*

A study was performed to determine what should be expected if the value of the subtraction differs from the actual delay value, \( \tau_{1e} \neq \tau_{D_1} \). Figure 5-14 contains a plot of a simple scenario; only two devices are present so the sidelobe interactions are not significant. The subtraction is done for three different scenarios: the tau is correct, \( \tau_{1e} = \tau_{D_1} \), a single sample point off, \( \tau_{1e} + dt = \tau_{D_1} \), and two sample points off, \( \tau_{1e} + 2 \cdot dt = \tau_{D_1} \).
Figure 5-14: An example of what happens if the subtraction in Equation (5.17) is done with an inappropriate tau value. The black solid line is the tau that allows for the smallest amount of energy. The blue is if the tau was merely one dt or sample point away from the expected value and the red shows what 2 samples away would look like where a sample point is 0.3 ns. An incorrect subtraction could therefore be very detrimental to obtaining correct device information.

It appears that an incorrect estimation could cause large inconsistencies in the accuracy of the delay.

The next aspect examined focused on the result when the amplitude of the matched filter signal to be subtracted out is incorrect, \( A \cdot H_{1R} \cdot H_1^* = B \cdot H_1 \cdot H_1^* \), where \( A \) is the scaling due to spatial delay and is not equal to \( B \) the scaling input by the matched filter, \( A \neq B \). When the amplitude is incorrectly matched, as seen in Figure 5-15, the value still provides the lowest amount of energy but the signal is clearly not subtracted out.
Figure 5-15: Subtraction in Equation (5.17) is performed but the amplitude of the matched filter estimation does not match that of the signal. The matched filter signal being more in this case dominates the subtraction yielding little difference in amplitude.

As follows, further manipulation would be required if the tau was to be used in subsequent estimations as in Equation (5.22). Based on an initial estimation of tau, amplitude A must also be calculated for implementation into the matched filter. This would increase computation time and possibly diminish the accuracy if implemented incorrectly.

Error Estimation Type Two

An additional method was devised for extraction of the delay, similar to the first. Recall from Equation (5.15) the frequency representation of the signal just before the receiver front end can be represented by

\[ H_R(f) = \sum_{i=1}^{N} H_i(f) \cdot \exp(-j \cdot 2\pi f \cdot \tau_{D_i}) + H_{AWGN}(f) + H_{Jam}(f). \]  \hspace{1cm} (5.23)

Again, if one were to examine only a single matched filter output one would obtain Equation (5.16). The value \( \tau_{D_1} \) is estimated based on simple peak detection. The received matched filter signal is then multiplied by the estimated delay shown as
In an attempt to set the time delay to zero, creating an ideal correlation signal at \( t = 0 \). If \( \tau_{D_{est}} = \tau_{D_1} \) then

\[
H_R \cdot H_1^* \cdot \exp(j2\pi f \cdot \tau_{D_{est}}) = H_1 \exp(-j \cdot 2\pi f \cdot \tau_{D_1}) \cdot H_1^* \exp(j \cdot 2\pi f \cdot \tau_{D_{est}})
\]

\[
+ \sum_{i=2}^{N} H_i \cdot H_i^* \cdot \exp(-j2\pi f \cdot (\tau_{D_i} - \tau_{D_{est}}))
\]

\[
+ (H_{AWGN} + H_{jam}) \cdot H_1^* \cdot \exp(j2\pi f \cdot \tau_{D_{est}})
\]

(5.24)

The first term to the right of the equal sign now represents an ideal autocorrelation with no delay. Therefore, the largest amount of energy can now be found in the real part of the signal. A set of delays are created

\[
\tau_{D_{est},i} = \pm i \Delta t \cdot \tau_{D_{est}} \text{ for } i = 1 \ldots \frac{M}{2}
\]

(5.26)

where \( \Delta t \) is the step parameter. Each of these delays is run against the received signal and the frequency domain signal is then summed over all frequency

\[
H_{sum_{est},i} = \sum_j \left| H_R \cdot H_1^* \cdot \exp(j2\pi f \cdot \tau_{D_{est,i}}) \right|^2 \text{ for } i = 1 \ldots M.
\]

(5.27)

The delay, \( \tau_{D_{est,i}} \), that yields the largest real signal is determined to be the signals delay. This decoding technique would be repeated for each subsequent N targets as shown in Figure 5-16.
Figure 5-16: Estimation technique decoding block diagram of the receiver post processing. After the matched filter and soft decision multiple delayed signals are implemented and the signal is summed in frequency. The desired outcome is the signal with the largest real value.

The results from the two different coding techniques are shown in Figure 5-17. The established criteria for determining what constituted an error had to be adjusted to account for the increased accuracy of these decoding types. The peak larger than a sample point away would now be considered erroneous. Though there is not a significant improvement between the two estimation types, type two is slightly superior in most instances, and both are considerably better than no estimation.
In-Phase and Quadrature Demodulation

While the techniques described above had the ability to increase the accuracy of the estimated time delay to within one sample point of the actual value, additional accuracy is always useful when determining sensed information. This led to a technique to potentially view in between the sample points using the in-phase and quadrature (IQ) information in the signal. The following technique utilizes the phase information at time equal to zero to extract the delay value. A simple block diagram of an IQ demodulator utilizing the phase for extracting information is shown in Figure 5-18.
Figure 5-18: Block diagram of an IQ demodulation technique used to extract the phase information.

After the signal has been matched filtered it goes through a quadrature demodulation. If, $y$, is the signal then the total signal, from [82], is given by

\[ x = \sqrt{(y \cdot \cos(2\pi f_0 t))^2 + (y \cdot \sin(2\pi f_0 t))^2} \quad (5.28) \]

The in-phase portion is

\[ x_i = y \cdot \cos(2\pi f_0 t) \quad (5.29) \]

and the quadrature

\[ x_q = y \cdot \sin(2\pi f_0 t). \quad (5.30) \]

The phase between the I and Q channels is

\[ \text{phase} = \tan^{-1} \left( \frac{x_q}{x_i} \right). \quad (5.31) \]

When the signal contains no time offset between the received signal and the matched filter signal, Figure 5-19 is obtained.
Figure 5-19: A correlation signal with no time offset plotted on the top and the phase of the signal plotted on the bottom. The phase of the signal is zero yielding all of the energy in the in-phase channel at time equal zero.

Figure 5-19 shows that when the delay is estimated correctly the phase at time zero should be zero and the largest amount of energy should be in the I channel. The IQ demodulation is then performed on varying delays within one dt or one sample point of the estimated peak location. The observation of the total signal was to determine the change in the peak shape and amplitude.
Figure 5-20: View of the correlation peak when changing the delay from -1 dt to +1 dt, where a dt is 0.5 ns. Observe the change in amplitude profile. There is a change not only in peak value but also in the shape of the peak.

The change in peak shape due to minor variation in the delay time illustrates that the peak can show up as much as 2 dt away from the expected value due only to I Q interaction. If looking only at the point time zero, one could be as much as 2 sample points off. An alternative was devised.

When implementing within OFC that has been estimated by removing the time delay, it is possible to look across the entire signal as opposed to looking at only a single point as in the previous examples. This technique utilizes the whole signal information when attempting to extract the delay value.

The signal received through the channel is a delayed reflected version of the interrogation signal with the OFC reflector response from the device.

\[
X_r(f) = X(f) \cdot e^{-jft_\Delta}.
\]  

(5.32)
The signal without any delay, when matched filtered, would have no imaginary response. This can be used to the user’s advantage. When the signal is matched filtered and then multiplied by its estimated delay, the result should have zero imaginary part.

\[
\sum_{\tau} \text{imag} \left( F^{-1} \left( X_r(f) \cdot X(f)^* \cdot e^{i2\pi f \tau_{est}} \right) \right) = 0. \tag{5.33}
\]

Zero imaginary would result in a phase of zero as well. The phase can then be extracted utilizing this summed signal.

This extraction technique would then consist of several steps. The signal would go through a correlator to estimate the peak as done previously with the estimation techniques. This estimation would then be multiplied by the signal but oppositely phased in order to remove it. Any remaining delay would result in a slight error in the correctness of the peak, \( \tau_{err} = \tau_{est} - \tau_d \). The \( \tau_{err} \) would then cause some phase which is calculated using

\[
\text{phase} = \tan^{-1} \left( \frac{\sum x_q}{\sum x_i} \right). \tag{5.34}
\]

---

**Figure 5-21**: Block diagram of an IQ demodulation technique used when the signal is summed through each of the separate channels and the phase information then extracted.
When the I and Q channels are separated in this way and the phase is obtained from them, there is a known variation in the phase due to a single sample point offset. The phase value is converted to time to be added or subtracted from the estimated value to create a new extracted value. Extrapolation of this value between the sample points can allow the ascertaining of the real value to within .01 dt.

When a single target is simulated and allowed to vary spatially between each run, a performance analysis can be performed. Figure 5-22 illustrates the potential of the quadrature peak estimation. The analysis in this simulation was done with all ideal signals only the spatial variation introducing any change.

![Figure 5-22: A plot of the percentage of targets observed correctly within a specified time deviation of the expected value. Shown is a plot of one target being varied in space around 1m. The time deviation now measured in femtoseconds showing all targets were within 90 femtoseconds of their desired value. The simulation is performed on a single device using the parameters in Table 5-1 and run ten times each varying the random time parameter.](image-url)
When this technique is compared with the previous estimation techniques, an increase in performance is recognized. The figure of merit is now the maximum deviation from the expected value instead of simply an observed error.

**Design Parameters**

Because of the vast potential scenarios for SAW sensor devices, a bound must be placed on the optimal system design. The system will consist of the following:

- The targets will be stationary during the evaluation period
- Up to 16 targets will be viewed simultaneously by the receiver antenna
- No a priori knowledge will be obtained about the location of a specific target
- The devices will operate utilizing ultra wideband specifications
- The frequency shall not exceed that which is able to be manufactured with current photolithographic technology

The range of the system is clearly directly reliant upon the wavelength or frequency of use. Frequency cannot be the only factor when deciding range however because of the many other parameters that are dependent upon it. When the required signal to noise ratio (SNR) is also taken into account, it becomes clear that a range of 100 meters is unobtainable with frequencies even slightly less than a gigahertz. Therefore, an operating range of 1 meter to 50 meters will be considered, with a maximum range dependent upon the required SNR. Given the general range parameters, it is now possible to assess the dynamic range of the potential system. Given an eight bit ADC, the closest target can be no more than two meters while the furthest target is fifty meters away. This sets the absolute dynamic range. One must consider, however, the greatest possible percentage variation around a specific distance attainable without the cross-correlation drowning out the signal of interest as shown in
As illustrated in Figure 5-5, a variation of more than approximately 20% will start to cause significant errors in the peaks. This can be alleviated with a priori knowledge of the target placement but that is not considered here.

With the most desirable frequency still in question, the target parameters will now be assessed. Antenna design is a key figure of merit when considering a deployable sensor. When determining the necessary bandwidth and requisite to be small, the ESA equations show plainly that higher frequencies are superior.

Once a reliable way to determine the peak is assessed, the remaining details of a reliable system can be evaluated. It is clear that the OFC is superior to that of CDMA due mainly to losses incurred when using CDMA. The coding technique that appeared most promising from the OFC types was OFC-PN-TDD. This can be used in conjunction with SDMA for additional diversity when needed. It is worth noting that an apparent cut-off for the SNR for each of the OFC types as shown in Figure 5-7 is around -6 dB. This will help in the determination of the range. When applying TDD, one must consider the number of frequencies to be used as well as the number of chip offsets that will be allowed. From previously observed testing, there is no great benefit from having more than two or three overlapping chips when using a small number of total chips, i.e. five as in previous examples. The next inspected aspect is the on-device propagation loss and device length. As shown in Figure 5-11, the device length for a two-chip offset design really starts to increase below 400 MHz. Also shown is the propagation loss starting to contribute significantly above 1.6 GHz.

Now that several bounds have been set, it is possible to narrow the possible field of design parameters yielding an optimal result. The minimum SNR for seemingly reliable results when testing is -6 dB. Using this value in Equation (5.5), the result can be compared with the other frequency dependent aspects of the optimal system shown in Figure 5-23. When considering these results coupled with the
antenna parameters, the optimal range appears to be somewhere between 800 MHz and 1 GHz. The 915 MHz ISM band is in this range and is available for UWB usage.

Figure 5-23: Comparison of the major frequency dependant optimal system variables. The propagation loss plotted in blue, the device length in dashed green and the range in dotted red. This plot reproduces the previously used plots from Figure 5-2 and 5-11 and places them on a single constrained plot between 400MHz and 1.6GHz.

The end result is a system that operates within the following parameters as noted in Table 5-5.
Table 5-5: Optimal system design parameters for orthogonal frequency coded SAW tags for use in a multiple sensor environment.

<table>
<thead>
<tr>
<th>Coding Technique</th>
<th>OFC-PN-TDD</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of OFC chips and frequencies</td>
<td>5</td>
</tr>
<tr>
<td>Number of overlapping chips</td>
<td>2 or 3</td>
</tr>
<tr>
<td>Operating Frequency</td>
<td>900 MHz range</td>
</tr>
<tr>
<td>Maximum Range</td>
<td>Less than 40m with a 20% range variation</td>
</tr>
<tr>
<td>Signal to Noise Ratio required for decoding</td>
<td>Greater than -6 dB</td>
</tr>
<tr>
<td>Estimation technique</td>
<td>Delay multiplication</td>
</tr>
<tr>
<td>IQ technique</td>
<td>Summing of signals</td>
</tr>
</tbody>
</table>

A simulation was performed utilizing the parameters and the results are shown in Figure 5-24.

![Figure 5-24: A plot comparison of optimized design parameters with PN-OFC as originally designed. There are 16 devices simulated with a 3dB SNR. The range corresponds to a 20 meter arch with 10% variation allowable. PN-OFC-TDD with 2 overlapping chips is utilized for the optimized results.](image-url)
CHAPTER 6: CONCLUSION

The study, design, and implementation of orthogonal frequency coded (OFC) surface acoustic wave (SAW) sensors for use in a multiple sensor environment was discussed. A detailed look at a sensor-level simulator was presented. The parameters of interest examined were the transmitter, channel characteristics, target, and the receiver output. The simulations employed a SAW device coupling of modes (COM) model in combination with RF system parameters. The simulator was shown to be reliable when utilizing the COM model for SAW devices, implicating its use as a fully encompassing system simulator. The designed simulation tool for the overall SAW system was tested to perform analysis on important parameters, such as signal to noise ratio (SNR), SAW coding type, and range effects. System evaluation of a multi-frequency SAW sensor system with multiple targets was also discussed.

Various coding approaches were considered. The coding of the system with decreased collisions was of primary interest. The techniques presented were used in an attempt to create a working multiple sensor code set. Simple correlation techniques, spread spectrum techniques, and multiplexing schemes were reviewed. The use of multiplexing in frequency and two types of time division multiplexing was explored. Time division multiplexing (TDM) and time division duplexing (TDD) were shown to perform most advantageously of all coding types designed and tested.

The discussion and implementation of an optimized SAW sensor system was then presented. The design consideration affecting the range of the system was addressed, as it played a key role in the usability of the system. The target was evaluated as to how the antenna and device parameters affect the system. The antenna on the target was shown to have significant limitations when attempting to remain small and fulfill the required bandwidth. Different decoding techniques were given for the
extraction of the sensed information. An estimation technique utilizing a delayed version of the signal was shown to improve correlation peak accuracy. The implementation of an in-phase and quadrature demodulation technique further improved accuracy to within a few picoseconds. Ultimately, a set of workable design parameters was established. These parameters provided a usable frequency range with tradeoffs between range, antenna size and gain, and ease of fabrication. Additionally, the type of coding and appropriate signal to noise ratio to implement were also addressed.

Orthogonal frequency coding is a viable and advantageous technique for implementing multiple-access SAW sensors. Future research is warranted and topics which can be included will be discussed. The implementation of actual S-parameter data for antenna and matching networks into the system simulation can address the perspective impact of these device facets. Investigation into the use of this type of coding for other applications, i.e. RFID, could provide benefit to the technology. Allowance of sensor data to be placed upon the device yields observable results for accuracy of extraction techniques. Also the inclusion of a graphical user interface to allow ease of testing various coding techniques would be preferable.
REFERENCES


[77] W. Kester. Taking the mystery out of the infamous formula, "SNR=6.02N+1.76dB," and why you should care.


