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ULTRA–WIDEBAND ORTHOGONAL FREQUENCY CODED SAW CORRELATORS

by

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B.S. University of Central Florida, 2003

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

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Major Professor: Donald C. Malocha
ABSTRACT

Ultra–wideband (UWB) communication new technology with ability to share the FCC allocated frequency spectrum, large channel capacity and data rate, simple transceiver architecture and high performance in noisy environments. Such communication advantages have paved the way for emerging wireless technologies such as wireless high definition video streaming, wireless sensor networks and more. This thesis examines orthogonal frequency coded surface acoustic wave (SAW) correlators for use in advanced UWB communication systems.

Orthogonal frequency coding (OFC) and pseudo-noise (PN) coding provides a means for UWB spreading of data. The use of OFC spectrally spreads a PN sequence beyond that of CDMA because of the increased bandwidth; allowing for improved correlation gain. The transceiver approach is still very similar to that of the CDMA approach but provides greater code diversity. Use of SAW correlators eliminates many of the costly components that are needed in the IF block in the transmitter and receiver, and reduces much of the signal processing requirements.

The OFC SAW correlator device consists of a dispersive OFC transducer and a wideband output transducer. The dispersive filter was designed using seven contiguous chip frequencies within the transducer. Each chip is weighted in the transducer to account for the varying
conductance of the chips and to compensate for the output transducer apodization. Experimental correlator results of an OFC SAW correlation filter are presented. The dispersive filter is designed using seven contiguous chip frequencies within the transducer. SAW correlators with fractional bandwidth of approximately 29% were fabricated on lithium niobate (LiNbO$_3$) having a center frequency of 250 MHz and the filter has a processing gain of 49. A coupling of modes (COM) model is used to predict the experimental SAW filter response. Discussion of the filter design, analysis and measurements are presented. Results are shown for operation in a matched filter correlator for use in an UWB communication system and compared to predictions.
To Elizabeth “Bette” Gallagher and Teresa Donohoe
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I would like to thank my family for their love and continued support throughout my endeavors. The sacrifices my parents have made throughout my life affording me any, and every, opportunity and to enable me to excel in my education has been selfless and wholehearted. I am eternally gracious to them for everything.

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<th>Description</th>
</tr>
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<tbody>
<tr>
<td>AM</td>
<td>Amplitude Modulation</td>
</tr>
<tr>
<td>bps</td>
<td>bits per second</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CMOS</td>
<td>Complimentary Metal Oxide Semiconductor</td>
</tr>
<tr>
<td>dB</td>
<td>Decibel</td>
</tr>
<tr>
<td>dBm</td>
<td>Decibel in milliwatts</td>
</tr>
<tr>
<td>DC</td>
<td>Direct current</td>
</tr>
<tr>
<td>DDS</td>
<td>Direct Digital Synthesis</td>
</tr>
<tr>
<td>DS/SS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>DS-UWB</td>
<td>Direct Sequence Ultra-Wideband</td>
</tr>
<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FH</td>
<td>Frequency Hopping</td>
</tr>
<tr>
<td>FSK</td>
<td>Frequency shift keying</td>
</tr>
<tr>
<td>LiNbO₃</td>
<td>Lithium Niobate</td>
</tr>
<tr>
<td>LiTaO₃</td>
<td>Lithium Tantalate</td>
</tr>
<tr>
<td>GHz</td>
<td>Giga Hertz (billion cycles / sec)</td>
</tr>
<tr>
<td>IC</td>
<td>Integrated circuit</td>
</tr>
<tr>
<td>IDT</td>
<td>Interdigital Transducer</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter-Symbol Interference</td>
</tr>
<tr>
<td>ISM</td>
<td>Industrial, Scientific, and Medical frequency band</td>
</tr>
<tr>
<td>LNA</td>
<td>Low noise amplifier</td>
</tr>
<tr>
<td>MATLAB</td>
<td>Simulation software available from Math Works</td>
</tr>
<tr>
<td>Mbps</td>
<td>Mega bits per second</td>
</tr>
<tr>
<td>M-FSK</td>
<td>M-ary Frequency Shift Keying</td>
</tr>
<tr>
<td>MHz</td>
<td>Mega Hertz (million cycles / sec)</td>
</tr>
<tr>
<td>OFC</td>
<td>Orthogonal frequency coding</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>OOK</td>
<td>On-Off Keying</td>
</tr>
<tr>
<td>PA</td>
<td>Power Amplifier</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>--------------</td>
<td>-----------------------------------</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed circuit board</td>
</tr>
<tr>
<td>PG</td>
<td>Processing Gain</td>
</tr>
<tr>
<td>PN</td>
<td>Pseudorandom Noise</td>
</tr>
<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RFID</td>
<td>Radio Frequency Identification</td>
</tr>
<tr>
<td>SAW</td>
<td>Surface Acoustic Wave</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>UWB</td>
<td>Ultra-Wideband</td>
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CHAPTER 1
INTRODUCTION

Ultra–wideband (UWB) radio communications is expected to play a revolutionary role in the future of wireless communication systems. The interest in UWB communications has recently been sparked by FCC rulings in 2002 allowing unlicensed commercial radio communications meeting UWB specifications. UWB communication is an emerging technology with ability to share the FCC allocated frequency spectrum, large channel capacity and data rate, simple transceiver architecture and high performance in noisy environments. Such communication advantages have paved the way for emerging wireless technologies such as wireless high definition video streaming, wireless sensor networks and more.

A conventional UWB system utilizes extremely narrow RF pulses to achieve very wide bandwidths and low spectral power density. Among the many numerous attractive features of UWB communications arise numerous deployment challenges. The use of various spread spectrum techniques can be implemented in UWB communications to help overcome many of these technical challenges without the need for overly complicated communication algorithms. However, the more complicated modulations method increases the processing requirements such that very fast DSP computations are required as well as significantly complication the simplistic receiver with low power requirements. Surface acoustic wave are capable of
increasing the processing power without complicating the receiver architecture or high speed silicon devices.

Orthogonal frequency coding (OFC) is a spread spectrum coding technique that has been successfully implemented in surface acoustic wave (SAW) tags and sensors using reflective structures [1]. The use of OFC in a SAW transducer allows multiple access in UWB systems as well as overcoming many of the conventional UWB complications such as pulse echoes from multi-path transmission. The OFC technique offers both frequency and phase coding producing a significant number of unique code sets. This thesis presents the background, development and evolution of SAW correlation filters using OFC for use in UWB communication systems. The UWB OFC filter design is also shown to function in a prototype UWB communication system and demonstrates its feasibility for multiple access UWB communications.

An overview of various aspects of communication theory relevant to this thesis are outlined in Chapter 2. The important communication benefits of spread spectrum communications are discussed; including access of a common communication channel by more than one user and a resistance to interference and jamming. The methods of data acquisition and synchronization for spread spectrum communications are discussed. The chapter concludes with a background discussion of Ultra-wideband communications.

Chapter 3 highlights the need for advanced spread spectrum techniques in UWB communication systems and discusses the solutions that SAW technology provides. The numerous advantages of SAW devices for UWB communication transceivers were recently demonstrated
using a pseudo-noise (PN) SAW transducer to implement a CDMA coded signal on a single frequency RF carrier[2]. The use of OFC in the SAW device offers many advantages beyond this approach. The analytical development of the orthogonal frequency definition is discussed.

The UWB OFC device design and implementation is discussed in Chapter 4. The development of the dispersive transducer and the equations used to normalize the chip conductance is outlined in detail. The development of the wideband input transducer are presented for both a polarity weighted wideband transducer and an apodized transducer with an inverse cosine envelope. The investigation of piezo-electric material selection is discussed.

Chapter 5 presents device design results achieved using a wafer-level RF probe station and data acquisition system. The evolution of the final device design, with a discussion of problems and solutions, are presented. The final UWB OFC device design is compared to coupling-of-modes (COM) model predictions and experimental data is correlated with its matched filter and compared to the ideal compressed pulse. Finally, results for a double dispersive OFC device are presented.

Chapter 6 presents system results using the final device design. The device is bonded and packaged for use in a prototype system configuration using numerous RF components and the resulting device performance is presented. The matched filter correlation is obtained using a vector network analyzer and compared to the RF probed and ideal compressed pulse obtained earlier. Prototype system results are then presented using a digital scope output and the auto-correlation and cross-correlation are shown on similar scalings for comparison.

In Chapter 7, a summary of important conclusions of the research are discussed.
2.1 Spread Spectrum Communications

Spread spectrum (SS) communication systems have been used since its conception in the 1920’s [3]. Modern spread spectrum communication systems have evolved a long way since using top-secret noise wheels, however the overall concept and inherent communication benefits remain the same.

The term “spread spectrum” can be defined as:

“Spread spectrum is a means of transmission in which the signal occupies a bandwidth in excess of the minimum necessary to send the information; the band spread is accomplished by means of a code which is independent of the data, and a synchronized reception with the code at the receiver is used for despreading and subsequent data recovery.”

This definition sums up to one important point; the transmitted signal is a greater bandwidth than needed for the data being transmitted. Spread spectrum techniques were initially used for military applications such as RADAR and secure communications. This is
due to the number of inherent benefits spread spectrum communication techniques have over conventional systems. Some of the distinct benefits include [4]:

- Anti-jamming
- Anti-interference
- Low probability of intercept
- Multiple user random access communications with selective addressing capability
- High resolution ranging
- Accurate universal timing

Spread spectrum communications provides security, immunity to jamming and is less sensitive to unwanted signal interference. These benefits are been key ensuring the security of military communications since the second world war. The same concept can be applied to commercial communication systems. In this case, the jamming signal is other communication devices using the same channel. For instance, modern cellular phones use a variety of spread spectrum communication techniques to ensure that there is no cross talk with other nearby phones.

The data signal is spread using a number of methods including “direct sequence”, “frequency hopping” and “time hopping.” It is also common to use a combination of these methods to spread the signal further and gain a higher level of coding.

The spread spectrum techniques discussed here lay the foundation and background for the techniques used throughout this thesis.
2.2 Direct Sequence Spread Spectrum Communications (DS/SS)

In direct sequence modulation, a fast pseudorandomly generated, or pseudorandom noise (PN), sequence produces phase transitions in the carrier containing data due to the binary chipping sequence of +1 or −1 being mapped into the phase of the modulated signal. This common coding technique for DS/SS is called binary phase shift keying (BPSK). Figure 2.1 shows an example of a typical DS/SS transceiver. The original signal, \( d(t) \), is capable of being received only when the chipping sequence, \( c(t) \), is known. The spreading sequence is chosen to have properties that facilitate proper demodulation by the intended receiver, make demodulation by an unintended receiver as difficult as possible and make the transmitted signal distinguishable from a jamming signal [5].

![Figure 2.1: DS/SS Communication System Block Diagram](image)

This concept of spreading the signal is illustrated in Figure 2.2 for an unspread data signal and DS/SS modulated signal. By comparing both time and frequency domain, we see that the data time length is identical, yet the bandwidth needed to transmit the code signal \( c(t) \) is much larger than the data signal. Each chip in the code signal is \( \tau_c \) long and the data
Figure 2.2: Spreading of Spectrum Plot. $d(t)$ is the data $c(t)$ is the spreading PN code. The bandwidth of the spreading signal to the right is $N$ times greater than the data signal, where $N$ is the number of chips.

Bit is $\tau_B$ long where:

$$\tau_B = N \cdot \tau_c \quad (2.1)$$

Therefore, the bandwidth of the spread signal is $N$ times longer after being mixed with the code sequence, $c(t)$. The processing gain (PG) of the spread spectrum system is defined as the bandwidth consumed by the transmitted SS signal by the bandwidth required to transmit the unspread signal, as shown in Equation 2.2. For the example shown in Figure 2.2, the PG is equal to the number of chips, or $N$.

$$PG = \frac{\text{RF Bandwidth}}{\text{Information Bandwidth}} = N \quad (2.2)$$
The processing gain defines the amount of performance improvement that is achieved through the use of spread spectrum. The PG can also be gauged visually by comparing the correlation output of a spread signal to the unspread data signal. The processing gain also provides an improvement to the peak-to-sidelobe (PSL); an measure of detectability

2.2.1 Code Division Multiple Access (CDMA)

An alternative coding method to BPSK is CDMA. CDMA is a type of direct sequence spread spectrum in which each user, or transceiver/receiver pair has a unique PN code for transmission over a common channel bandwidth [6]. Typically, PN sequences are chosen specifically for their auto-correlation and cross-correlation properties. For CDMA, it would be ideal for each PN sequence to be optimized so that the interference observed by a single transceiver is minimal limiting the number of users by the available processing gain and the orthogonality of the PN codes used. CDMA is commonly used by cellular phone providers today; enabling each phone to transmit independently of another while utilizing the same carrier frequency.

2.3 M-ary Frequency Shift Keying

In a frequency-hopped spread spectrum (FH/SS) communications system the channel bandwidth is subdivided into a number of contiguous frequency slots where the transmitted signal
occupies one or more of the available frequency slots [6]. In other words, the communication spectrum is widened by changing the frequency of the carrier at a certain periodicity. Each frequency selection is made pseudorandomly based on the output from a PN generator. The modulation method typically used is either binary or M-ary frequency shift keying (FSK). In binary FSK, one of two frequencies is rapidly selected for signal modulation. A typical FH/SS communication system block diagram can be seen in Figure 2.3.

![FSS (FSK) Communication System Block Diagram](image)

**Figure 2.3:** FH/SS (FSK) Communication System Block Diagram.

In Figure 2.3, the data sequence is passed through a FSK modulator and mixed with a randomly generated frequency before transmission. The received signal is synchronously mixed with the same random frequency and demodulated into the original data signal. One method for frequency generation is to use a pulsed SAW device for each desired slot frequency; two devices in total for binary FSK. Non-coherent FSK detection is usually employed due to the difficulty in phase synchronization between one hopped frequency to another. A direct digital synthesizer (DDS), however, can be used to generate coherent, high rate, frequencies for M-ary FSK modulation.
Reception of the spread signal is inherently secure and reception is only possible by knowing the spreading code, \( c(t) \). At the receiver, the correlator uses a replica of the PN code to de-spread the coding signal. Although the code sequence is known, synchronization of a receiver to a specified code can be a daunting task. The use of active correlator structures or matched filters make it possible to synchronize to and demodulate the incoming waveform. The active correlator is useful for signals that are easily synchronized and is realized using multiple correlators or a sliding correlator structure. The sliding correlator utilizes an integrator to produce a ramp wave form which follows the polarity of the PN code of \( \pm 1 \). If the integrator output falls below the threshold and deemed too small, the correlator is determined to be out of synchronization and is repeated after time adjustments. The time adjustments are generally a two step process involving course synchronization and time-base tracking. The sliding correlator approach becomes even more complicated with M-ary FSK systems. An overview of this approach are outlined in detail in [7].

The reception of received signal where synchronization is more difficult is accomplished using a matched filter technique. A matched filter can take many forms which basic principle is the same. The matched filter impulse response is a time reversal, or complex conjugate, of the original coded signal. The matched filter is correlated with the incoming signal, \( y(t) \) as defined by Equation 2.3. The resulting output of the matched filter, \( \psi \), is a series of
compressed with a phase of $\pm \pi$, or polarity of $\pm 1$, representative of the original data signal, $d(t)$. A more detailed description of matched filters and their properties can be found in [8].

$$\psi = y(t) * c(t) = \int_{-\infty}^{\infty} y(t) c(\tau - t) d\tau \quad (2.3)$$

The matched filter correlator, through the use of surface wave devices, maximizes the signal-to-noise ratio (SNR) at the bit decision instant; also known as the epoch instant [9]. The instant correlation of the signal provides a number of advantages for complex signal detection and high data transmission rates; such as those detailed in this thesis. Also, a complicated synchronization procedure is not required.

### 2.5 Ultra-Wideband Communications

With the advent of new wireless technologies, the demand for bandwidth in the RF spectrum has dramatically increased. Ultra–wideband (UWB) technology offers an effective solution during this time of growing demand for personal wireless consumer technology. UWB communications is an revolutionary wireless technology offering numerous communications advantages. The ability to share the FCC allocated frequency spectrum, large channel capacity and data rate, simple transceiver architecture and high performance in noisy environments has paved the way for emerging wireless technologies such as wireless high definition video streaming, penetrating RADAR and imaging, wireless sensor networks and more. The ability
of UWB to coexist with current radio systems with minimal or no interference also provides
the distinct advantage of avoiding expensive spectrum licensing fees which other RF services
must pay to the Federal Communication Commission (FCC).

The early form of UWB communication was first employed by Guglielmo Marconi in 1901
with the transmission of Morse code sequences across the Atlantic Ocean using spark gap
radio transmitters [10]. The use of ultra-short pulses has become the widely accepted method
for achieving the very wide bandwidths and low power spectral density needed for UWB
communications. Modern pulse-based transmission, in the form of impulse radars and other
applications which were exclusive to Department of Defense (DoD), resurfaced in the 1960’s.
UWB technology was referred to as baseband, carrier-free or impulse technology, as the term
ultra wideband was not used until 1989 by the U.S. DoD. It was not until 2002, however,
that the FCC approved the commercial use of UWB technology.

The FCC defines UWB as any communication technology that occupies greater than
500 MHz of bandwidth, or fractional bandwidth greater than 25% of the operating center
frequency [11]. Typical narrow band systems occupy only about 10% fractional bandwidth
and at greater power levels.

Fractional bandwidth is determined using the equation:

\[ BW\% = \frac{BW}{f_c} = \frac{f_h - f_l}{f_c}, \quad (2.4) \]
where \( f_h \) and \( f_l \) are the cut-off frequencies measured at the \(-10\)dB points of the UWB signal spectrum.

The typical UWB receiver, shown in Figure 2.4, contains a pulse generator for synchronization and mixing with the received signal rather than the typical carrier. The transmission of short impulses make high speed data transmission possible. A carrier free technology significantly lowers the component count and the power needed for the operation of the UWB receiver. This is quite advantageous for many personal wireless communication devices for the consumer market.

![Typical UWB Receiver](image)

**Figure 2.4:** Typical UWB Receiver

UWB can be modulated using routines such as pulse position modulation (PPM) \([12]\); using impulses at high data rates which are not evenly spaced in time. The intervals are spaced randomly to generate a noiselike signal. While there are many advantages to the pulse based UWB communication technology, it can be significantly enhanced by using a combination of UWB techniques and spread spectrum coding techniques of DS/SS or FH/SS. In order to implement these spread spectrum coding techniques in UWB communications, the simplistic UWB receiver is likely to become much more complicated. In some cases,
processing capabilities cannot keep up with the demand of the high speed communications. The use of SAW devices allow advanced spread spectrum coding at high data rates without significantly complicating the transceiver architecture. This thesis examines the design and use of a unique approach using spread spectrum coded UWB SAW devices.
CHAPTER 3
SAW DEVICES FOR UWB / SS COMMUNICATIONS

3.1 Surface Acoustic Wave Devices for UWB Communications

The “Impulse Radio” method of UWB communications is effective and simple; however the deployment of such systems is not without complications. In many environments, the UWB signal becomes significantly distorted during transmission due to reflections of pulses and interference with other systems such as GPS [13]. A sub-nanosecond pulse may reverberate in an indoor environment, making proper reception of the UWB signal difficult. Such problems are typically overcome using complicated post processing algorithms and becomes even more complicated with the deployment of a large number of communication systems. The use of spread spectrum communication techniques in addition to UWB permits large scale deployment. The significant benefit this provides to UWB communications comes at the price of increased processing power requirements. The implementation of more complex signals, such as continuous wave (CW) or CDMA UWB, is not feasible with current silicon technologies.

Surface acoustic wave (SAW) devices, however, allow for simple generation and detection of complex UWB communication. The numerous advantages of SAW devices for UWB
communication transceivers were recently demonstrated using a pseudo-noise (PN) coded SAW transducer to implement a CDMA coded signal on a single frequency RF carrier \[2, 14\]. In these devices, the information is encoded by pulse-phase-modulation; which is used to excite the SAW which can be amplified and then transmitted. Reception is achieved by correlation of the matched filter received response and base-band envelope detection to extract the pulse phase.

![Conceptual block diagram of UWB OFC transmitter and receiver. The OFC SAW filter is used as a code generator in the transmitter and correlation filter in the receiver.](image)

**Figure 3.1:** Conceptual block diagram of UWB OFC transmitter and receiver. The OFC SAW filter is used as a code generator in the transmitter and correlation filter in the receiver.

The UWB OFC SAW device is capable of being used as both code generator and correlator in an UWB transceiver as shown in Figure 3.1. The use of a SAW correlator eliminates the need for costly high speed silicon CMOS devices as well as many of the costly components needed in the IF section. The optional power amplifier, shown in the transmitter section, is used to increase the range of the UWB communication. The receiver architecture for the proposed device remains simplistic, has a low component count and the power requirements are minimal; as compared to the conventional UWB receiver seen previously in Figure 2.4.
However, the use of a SAW correlator in the receiver allows for use of spread spectrum coding without complicating the receiver architecture. In fact, through advanced coding techniques, the processing gain of the UWB system can be increased well beyond that achieved with a CDMA coded SAW correlator. The SAW correlator can also be excited with a pulse to operate as an efficient code generator in the transmitter side of the UWB communication system.

### 3.2 Orthogonal Frequency Coding

The introduction of OFC in UWB SAW correlators provides several advantages over CDMA including an increased range due to enhanced processing gain, increased data rate resulting from reduced compressed pulse ambiguity and greater multiple access operation due to greater code diversity [15]. OFC offers all advantages inherent to conventional spread spectrum communications including enhanced processing gain and lower power spectral density (PSD). OFC is a spread spectrum coding technique that has been successfully implemented in SAW tags and sensors using reflective structures [16, 1, 17, 18]. The technique uses multiple orthogonal chips, each $\tau_{\text{chip}}$ long. In the frequency domain, the local chip frequencies are separated by $1/\tau_{\text{chip}}$. The final criteria requires that $f_{\text{chip}} \cdot \tau_{\text{chip}}$ must equal an integer number of half carrier cycles. A well known example signal is a linear stepped chirp, which contains a series of local chips with contiguous orthogonal frequencies and linear group delay. For OFC, a level of coding is achieved by shuffling the chips in time such that the adjacent chip
carrier frequencies are no longer contiguous in time; as illustrated in Figure 3.2. Additionally, PN coding is available for an even higher level of code diversity. The OFC waveform is implemented in a SAW device embodiment using interdigital transducers or periodic reflector gratings with local center frequencies and electrode counts necessary to meet the orthogonality conditions outlined above. The OFC technique permits multiple signals to occupy the same bandwidth with the data contained in the signal phase.

Figure 3.2: Coding is achieved in OFC by randomly mixing the orthogonal frequencies. The colors of the individual frequencies are the same for each subfigure.

The use of orthogonal frequencies and the orthogonal frequency coding theory has been previously published [15], and is reviewed here.
3.3 Orthogonal Frequency Definitions and Review

Consider a time limited, nonzero time function defined as:

\[ h(t) = \sum_{n=0}^{N} a_n \cdot \varphi_n(t) \cdot \text{rect} \left( \frac{t}{\tau} \right), \]

where:
\[ \varphi_n(t) = \cos \left( \frac{n \pi t}{\tau} \right) \]
\[ \text{rect}(x) = \begin{cases} 1, & |x| \leq \frac{1}{2} \\ 0, & \text{otherwise} \end{cases} \]

The function \( \varphi_n(t) \), represents a complete orthogonal basis set with real coefficients \( a_n \).

The members of the basis set are orthogonal over the given time interval if:

\[ \int_{-\frac{\tau}{2}}^{\frac{\tau}{2}} \varphi_n(t) \cdot \varphi_m(t) dt = \begin{cases} K_n, & n = m \\ 0, & n \neq m \end{cases} \]

Given the basis set and constraints, two functional descriptions are obtained that have the forms:

\[ h_1(t) = \sum_{n=0}^{N} a_n \cdot \cos \left( \frac{2n \pi t}{\tau} \right) \cdot \text{rect} \left( \frac{t}{\tau} \right), \] \hspace{1cm} (3.3a)

\[ h_2(t) = \sum_{m=0}^{M} b_m \cdot \cos \left( \frac{(2m+1) \pi t}{\tau} \right) \cdot \text{rect} \left( \frac{t}{\tau} \right). \] \hspace{1cm} (3.3b)
Each cosine term in the summations in (3.3) represents a time-gated sinusoid whose local center frequencies are given by:

\[ f_n = \frac{n}{\tau} \quad \text{and} \quad f_m = \frac{(2m + 1)}{2\tau}. \quad (3.4) \]

In the frequency domain, the basis terms are the well-known sampling functions with center frequencies given in (3.4). From (3.4), \( f_n \cdot \tau \) must be an integer, which requires an integer number of wavelengths at frequency \( f_n \), and similarly there must be an integer number of half wavelengths at \( f_m \). These are required conditions for the orthogonality of the basis functions. Given that each basis term is a frequency domain sampling function, then the null bandwidth is known to be \( 2 \cdot \tau^{-1} \). The overall frequency function is defined given the choice of the even or odd time function in (3.3a) and (3.3b), respectively, the basis frequencies of interest, the weight of the basis function, and either the bandwidth or the time length.

The desired signal will have both frequency and time diversity. The time function, \( g_{bit}(t) \), will have a time length \( \tau_B \) as the defined bit length. The bit will be divided into an integer number of chips such that:

\[ \tau_B = J \cdot \tau_{chip}, \quad (3.5) \]

where \( J = \text{Num of chips} \).
The chip interval, \( \tau_{\text{chip}} \), is set as the time interval in (3.2) for the basis set. Given the definition of each chip as \( h_{cj} \), a bit is then defined as a sum of \( J \) chips as:

\[
g_{\text{bit}}(t) = \sum_{j=1}^{J} w_j \cdot h_{cj}(t - j \cdot \tau_{\text{chip}})
\]  

(3.6)

Each orthogonal chip is contiguous without time overlap and \( w_j \) is the bit weight and the functional form for the chip definition, \( h_{cj}(t - j \cdot \tau_{\text{chip}}) \), is given in (3.3). In general, multiple carrier frequencies are possible in each chip, depending on their weighting coefficient. Assuming a chip uses a basis set in (3.3a), similar results are obtained using (3.3b). Then, in general:

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

(3.7)

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

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

(3.8)
The form in (3.8) shows that each chip has a single local carrier frequency, \( f_{cj} = \frac{2C_j + 1}{2\tau_{chip}} \) and \( b_j \) is the chip weight. In order to form the desired time function \( b_j = \pm 1 \) for all \( j \), the bit null bandwidth is \( BW_{bit} = J \cdot 2^{-1} \cdot 2\tau_{chip} \) and \( C_j \) is a sequence of unique integers which means that \( f_{cj} \) form a contiguous, non-repetitive set. The level of coding is achieved by shuffling the contiguous chip frequencies in time.

The development of orthogonal frequency coding, reviewed and outlined here, is developed in detail for SAW tagging and sensors in [15, 16]. More detail of the benefits of orthogonal frequency coding are also shown in these references.
CHAPTER 4
UWB OFC DEVICE DESIGN AND IMPLEMENTATION

4.1 Device Design Parameters and Measurement

All the devices presented have the same basic design parameters. Devices were fabricated on YZ LiNbO$_3$ with aluminum electrodes and acoustic beam width ($W_a$) of 20 mil. The filters were designed with a center frequency of 250 MHz, a fractional bandwidth of 29% and an insertion loss of 30 dB. This exceeds the FCC defined bandwidth for a UWB signal with fractional bandwidth greater than 25% measured at the 10 dB points. The center frequency of 250 MHz is chosen for proof of concept and ease of implementation using conventional contact photolithography techniques. The devices can easily be scaled to higher frequencies using advanced fabrication techniques such as e-beam lithography.

The SAW devices for this work are all two-port devices including a wideband input transducer and a dispersive transducer. The dispersive, OFC, transducer had seven chips with seven different chip frequencies which yields a time bandwidth product of 49. Each chip is $\tau_{chip} = 96\,\text{ns}$ long. A wideband input transducer was used in conjunction with the OFC or stepped chirped transducer as shown in Figure 4.1 and 4.2, respectively. An input transducer was also placed on either side and at equal distance from the OFC transducer to allow the
filter and its matched filter responses to be obtained using the same device as seen on the final device layout. All final device layouts are shown to scale in Appendix A.

4.2 UWB Dispersive Transducer Design

To meet orthogonality conditions, every chip is a constant length ($\tau_{\text{chip}}$). In order to keep the chip length constant, the number of electrodes in the chip must increase proportionately as the chip frequency increases. The conductance is proportional to the chip frequency ($f_{\text{chip}}$), the beam width ($W_a$) and the chip electrode count. Therefore, it is necessary to apodize the chips in order to obtain a uniform conductance for each chip across the inline dispersive transducer. The chip apodization is shown in Figure 4.3 for a chirp configuration and Figure 4.4 for an OFC configuration. The tap weights for each chip can be determined.
by considering the conductance relationship between adjacent chip frequencies.

To the first order, the center frequency acoustic conductance for a SAW transducer is given as:

$$G_0 = 8k^2 f_0 N_p^2 C_s W_a,$$  \hspace{1cm} (4.1)

where $k^2$ and $C_s$ are material properties of the chosen piezoelectric substrate, $f_0$ is the center frequency of the transducer and $N_p$ is the effective number of electrode pairs.

For each OFC chip,

$$N_p = \frac{1}{BW\%} = \left( \frac{BW_{chip}}{f_{chip}} \right)^{-1} = \tau_{chip} \cdot f_{chip},$$  \hspace{1cm} (4.2)
Substituting Equation 4.2 into Equation 4.1, the acoustic conductance for each chip is:

\[
G_{0,\text{chip}} = 8k^2\tau_{\text{chip}}^2 f_{\text{chip}}^3 C_s W_{a,\text{chip}}.
\] (4.3)

From Equation 4.3, the chip beam width, \(W_{a,\text{chip}}\), is the only available parameter for tuning the chip’s acoustic conductance. The remaining parameters are constants defined by the orthogonality conditions and material properties. Choosing the lowest frequency chip as the reference, the beam width of all subsequent chips will be calculated as a ratio to normalize the overall conductance.

The ratio between adjacent chip frequencies is:

\[
\frac{G_{0_{n-1}}}{G_{0_n}} = \frac{f_{n-1}^3 W_{a_{n-1}}}{f_n^3 W_{a_n}},
\] (4.4)
where the subscript \( n \) is the chip number.

From Equation 4.4, the ratio of conductances for adjacent chips is proportional to the cubed ratio of those chip frequencies given as:

\[
\frac{G_{0_{n-1}}}{G_{0_n}} \propto \frac{f_{n-1}^3}{f_n^3}.
\]  

(4.5)

Therefore, to achieve constant conductance over the entire frequency band, the beam width for each chip can be calculated as:

\[
\frac{W_{a_{n-1}}}{W_{a_n}} = \frac{f_{n}^3}{f_{n-1}^3}.
\]  

(4.6)

The lowest frequency chip, \( f_1 \), will have a tap weight of unity; or \( W_{a_1} = 1 \). Using Equation 4.6, and the known value for \( W_{a_1} \), all remaining apodization tap weights are able to be determined.

The effectiveness of this procedure can be witnessed in Figure 4.5. Prior to adjusting the chip apodization, the conductance of each chip increases with its local frequency and will produce a skewed frequency response shown in 4.5(a). Using the relative beam width \((W_{a_n})\) value, as defined by Equation 4.6, the conductance remains constant for each chip frequency in the dispersive transducer. The conductance for each chip frequency after apodization is shown in 4.5(b).
Figure 4.5: Dispersive transducer conductance for each chip before and after chip apodization.
4.3 Wideband Input Transducer Design

A typical uniform, unweighted interdigitated transducer (IDT) will produce a frequency response resembling a typical sampling function as result of the rectangular time response and will cause the overall frequency response to roll off due to this shape; thereby distorting the desired UWB signal. Additionally, in order to achieve a wide bandwidth the unweighted IDT will have a small number of electrodes; causing additional conductance and impedance matching complications. It is therefore necessary to use a wideband input transducer that will maximize the conductance and minimize the distortion of the overall desired signal. For this project, two different wideband input transducer designs were investigated; the polarity weighted and inverse cosine apodized transducer. Each transducer type provides distinct advantages.

4.3.1 Polarity Weighted

The polarity weighted transducer is capable of flattening the conductance curve while maximizing the usable transducer bandwidth [19]. This is accomplished by using a main response IDT, of length \( \tau_1 \), with two adjacent IDTs with opposite polarity of length \( \tau_2 \) connected electrically in parallel as shown in Figure 4.6. The transducers are separated symmetrically on each side by an offset of \( \tau_{12} \). It is also possible to adjust the offset of each transducer independent of the other, but for this design \( \tau_{12} \) is held constant.
Figure 4.6: Polarity weighted input transducer concept. The side transducers of length $\tau_2$ are of opposite polarity to the main transducer with length $\tau_1$ and are separated symmetrically on each side by $\tau_{12}$.

The envelope polarity weighted transducer is shown in Equation 4.7, and is applied to the desired center frequency of input transducer as shown in Equation 4.8.

$$e(t) = \text{rect} \left( \frac{t}{\tau_1} \right) - \text{rect} \left( \frac{t - \tau_{12}}{\tau_2} \right) - \text{rect} \left( \frac{t + \tau_{12}}{\tau_2} \right)$$  \hspace{1cm} (4.7)

$$h(t) = e(t) \cdot \cos(2\pi f_0 t)$$  \hspace{1cm} (4.8)

By manipulating the parameters $\tau_{12}$, $\tau_1$ and $\tau_2$, the designer is capable of increasing the bandwidth of the input transducer and flattening or altering the shape of the transducer response in order to minimize distortion of the desired overall signal. The effect of the
Table 4.1: Polarity weighted input transducer design values.

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>$\tau_1$</td>
<td>24 ns</td>
</tr>
<tr>
<td>$\tau_2$</td>
<td>4 ns</td>
</tr>
<tr>
<td>$\tau_{12}$</td>
<td>14 ns</td>
</tr>
<tr>
<td>$N_{p1}$</td>
<td>6 pair</td>
</tr>
<tr>
<td>$N_{p2}$</td>
<td>1 pair</td>
</tr>
<tr>
<td>$N_{eff}$</td>
<td>4 pair</td>
</tr>
</tbody>
</table>

manipulating individual parameters can be determined using Equation 4.9; which is found by taking the analytical Fourier transform of Equation 4.8

$$H(f) = \tau_1 \frac{\sin(2\pi(f - f_o)\tau_1)}{2\pi(f - f_o)\tau_1} - \tau_2 \frac{\sin(2\pi(f - f_o)\tau_2)}{2\pi(f - f_o)\tau_2} e^{-j2\pi f \tau_{12}} - \tau_2 \frac{\sin(2\pi(f - f_o)\tau_2)}{2\pi(f - f_o)\tau_2} e^{+j2\pi f \tau_{12}}$$

(4.9)

In the transducer's transfer function, the terms $\tau_1$ and $\tau_2$ adjust the scaling of the frequency response due to their constant term outside of the sinc function. The $\tau_{12}$ term is only in the exponential terms only therefore changes in this parameter will change the phase of the polarity weighting.

The design values for the polarity weighted wideband input transducer used for this project are summarized in Table 4.1. The actual IDT layout used is shown to scale in the appendix in Figure A.10.
4.3.2 Apodized

The second transducer used in this project was an apodized transducer using inverse cosine windowing. The inverse cosine envelope, as defined in Equation 4.10, generates a cosine weighted transverse beam profile used to couple to the first symmetrically transverse mode. This ensures the elimination of any possible losses due to transverse waveguide moding.

\[ w(t) = \frac{2}{\pi} \arccos \left( \frac{2|t|}{\tau} \right) \text{rect} \left( \frac{t}{\tau} \right) \]  

\[ (4.10) \]

\[ \text{Figure 4.7: Inverse cosine weighted IDT apodization. The sampling phase is chosen so that the positive and negative tap values are equal resulting in SAW transduction over the entire transducer aperture.} \]

For the apodized transducer design, the sampling phase is chosen to maximize the positive and negative tap values and for a symmetric transducer layout. It is important to determine the proper phase of the samples in order to ensure the maximum electrode overlap as well as transducer symmetry. Figure 4.7 shows the desired impulse response for the apodized input
transducer and the inverse cosine window used. The electrode samples are also marked for the $3f_o$ sampled transducer used in this project. The actual IDT layout used is shown to scale in the appendix in Figure A.11.

### 4.4 Material Considerations

A preliminary investigation of various piezoelectric substrate properties was performed in order to determine the proper material choice. Previous development of OFC sensor technologies required a substrate with a number of properties due to the broad range of sensor applications investigated. The material properties considered are summarized in Table 4.2.

<table>
<thead>
<tr>
<th>Material</th>
<th>Velocity (m/sec)</th>
<th>$k^2$ (%)</th>
<th>TCD (ppm/°C)</th>
<th>BW% for 0dB Loss</th>
<th>Commercially Avail. Size</th>
<th>Operating Temp.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Quartz (ST)</td>
<td>3159</td>
<td>0.16</td>
<td>0</td>
<td>4.514 %</td>
<td>3” – 6”</td>
<td>&lt; 450°C</td>
</tr>
<tr>
<td>LiNbO$_3$ (Y – Z)</td>
<td>3488</td>
<td>4.5</td>
<td>+94</td>
<td>23.937 %</td>
<td>3” – 6”</td>
<td>&lt; 450°C</td>
</tr>
<tr>
<td>LiNbO$_3$ 128°</td>
<td>3978</td>
<td>5.3</td>
<td>+75</td>
<td>25.977 %</td>
<td>3” – 6”</td>
<td>&lt; 450°C</td>
</tr>
<tr>
<td>LiTaO$_3$ (Y – Z)</td>
<td>3230</td>
<td>0.74</td>
<td>+35</td>
<td>9.707 %</td>
<td>3” – 6”</td>
<td>&lt; 450°C</td>
</tr>
<tr>
<td>LGT (X cut – Y prop)</td>
<td>2292</td>
<td>0.589</td>
<td>-40.6</td>
<td>8.66 %</td>
<td>Not Avail</td>
<td>&lt; 1200°C</td>
</tr>
<tr>
<td>LGT (Y cut – X prop)</td>
<td>2210.6</td>
<td>0.423</td>
<td>-65</td>
<td>7.33 %</td>
<td>Not Avail</td>
<td>&lt; 1200°C</td>
</tr>
<tr>
<td>LGS (0°, 138.5°, 26.6°)</td>
<td>2730</td>
<td>0.350</td>
<td>0</td>
<td>6.676 %</td>
<td>2” – 3”</td>
<td>&lt; 1200°C</td>
</tr>
<tr>
<td>LGN (X cut – Y prop)</td>
<td>2309</td>
<td>0.474</td>
<td>-80</td>
<td>7.769 %</td>
<td>Not Avail</td>
<td>&lt; 1200°C</td>
</tr>
<tr>
<td>LGN (Y cut – X prop)</td>
<td>2242</td>
<td>0.415</td>
<td>-63</td>
<td>7.269 %</td>
<td>Not Avail</td>
<td>&lt; 1200°C</td>
</tr>
</tbody>
</table>

For sensing applications, the temperature coefficient of delay (TCD) is an important parameter. For temperature sensing, a high TCD provides an increased sensitivity to temperature changes. However, when the sensor is to be used to detect pressure variations,
changes in temperature should not affect the device performance. For this application, a substrate with a zero TCD should be used, such as the temperature compensated ST Quartz. High temperature operation is possible up to 1200 °C using materials such as langatate (LGT), langasite (LGS) and langanite (LGN). Unfortunately, these materials are rare, expensive and often have to be custom grown; making it infeasible for use during early technology development. For multi-channel device operation, it is desirable to utilize a large fractional bandwidth. Higher operating fractional bandwidths (BW%) are achievable by using high coupling substrates. In Figure 4.8, the minimum achievable insertion loss (IL) versus fractional bandwidth was plotted for common substrates. The plot assumed bidirectional transducers which were tuned and conjugately matched. Lithium niobate (LiNbO$_3$) offers the highest coupling coefficient ($k^2$) of the materials surveyed and therefore is capable of higher fractional bandwidth before suffering higher insertion loss. The LiNbO$_3$ 128°-cut offers the highest coupling coefficient but can be more difficult to work with in the fabrication process than Y–Z LiNbO$_3$ due to pyroelectric effects. Finally, the maximum center frequency associated with lithography resolution is an important consideration for material selection. The maximum device center frequencies for a given lithography resolution are summarized in Table 4.3 for quarter-wavelength electrode widths. Device features less than 1.5 $\mu$m are difficult to obtain using conventional contact lithography and must be achieved with advanced fabrication methods such as steppers or e-beam lithography.

The piezoelectric substrate selected for this project was Y–Z LiNbO$_3$. This material offers favorable properties for sensing temperature and was commercially available at an affordable
Above plot assumes bidirectional transducers tuned and conjugately matched.

**Figure 4.8:** Minimum insertion loss for common piezoelectric substrates versus fractional bandwidth for bidirectional operation.

unit price. The high coupling coefficient allows for relatively large fractional bandwidths without having a high insertion loss. Although LGT, LGS and LGN allowed for operation at a higher temperature range, Y–Z LiNbO₃ offered several advantages: higher velocity (producing higher frequency for a given line resolution); greater coupling (resulting in lower device loss); lower cost; and ability to operate over a significant temperature range (cryogenic to < 350°C).

Early UWB OFC devices were fabricated on a ST Quartz substrate with quarter-wavelength electrode transducers. These devices demonstrated very high insertion loss
Table 4.3: Maximum center frequency for given lithographic line resolution as calculated for quarter-wavelength electrodes.

<table>
<thead>
<tr>
<th>Material</th>
<th>2 µm (MHz)</th>
<th>1 µm (MHz)</th>
<th>0.5 µm (GHz)</th>
<th>0.25 µm (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Quartz (ST)</td>
<td>394.9</td>
<td>789.75</td>
<td>1.579</td>
<td>3.159</td>
</tr>
<tr>
<td>LiNbO₃ (Y – Z)</td>
<td>436</td>
<td>872</td>
<td>1.744</td>
<td>3.488</td>
</tr>
<tr>
<td>LiNbO₃ 128°</td>
<td>497.25</td>
<td>994.5</td>
<td>1.989</td>
<td>3.978</td>
</tr>
<tr>
<td>LiTaO₃ (Y – Z)</td>
<td>403.75</td>
<td>807.5</td>
<td>1.615</td>
<td>3.230</td>
</tr>
<tr>
<td>LGT (X cut – Y prop)</td>
<td>286.5</td>
<td>573</td>
<td>1.146</td>
<td>2.292</td>
</tr>
<tr>
<td>LGT (Y cut – X prop)</td>
<td>276.33</td>
<td>552.65</td>
<td>1.105</td>
<td>2.211</td>
</tr>
<tr>
<td>LGS (0°, 138.5°, 26.6°)</td>
<td>341.25</td>
<td>682.5</td>
<td>1.365</td>
<td>2.730</td>
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<tr>
<td>LGN (X cut – Y prop)</td>
<td>288.63</td>
<td>577.25</td>
<td>1.155</td>
<td>2.309</td>
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<tr>
<td>LGN (Y cut – X prop)</td>
<td>280.25</td>
<td>560.5</td>
<td>1.121</td>
<td>2.242</td>
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</tbody>
</table>

(IL) which would be highly infeasible for operational in a low power communication system. Later devices also used sixth-wavelength electrodes in the transducers, as seen in Section 5.1.2, pushing the limits of conventional contact lithography methods for a 250 MHz design center frequency. The in the UWB OFC device, the chip of highest frequency is also the one with the smallest features. Therefore, the corresponding electrode width must be considered in lithography resolution limitations. The electrode width at the highest frequency used in this project, \( f_7 \), is displayed in Table 4.4 for different electrode periodicity.

Table 4.4: Smallest electrode widths for IDT at \( f_7 \).

<table>
<thead>
<tr>
<th>Sampling Electrodes</th>
<th>electrode width</th>
</tr>
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<tbody>
<tr>
<td>( 2f_o )</td>
<td>( \lambda/4 )</td>
</tr>
<tr>
<td>( 3f_o )</td>
<td>( \lambda/6 )</td>
</tr>
<tr>
<td>( 4f_o )</td>
<td>( \lambda/8 )</td>
</tr>
</tbody>
</table>

\( S_e = 2 \) 3.1 µm  
\( S_e = 3 \) 2.1 µm  
\( S_e = 4 \) 1.5 µm
CHAPTER 5
UWB CORRELATION FILTER DEVICE RESULTS

The UWB OFC device design used the various principles discussed in earlier chapters, however various modifications were needed to perfect the device design. The development process utilized a linear stepped chirp as a benchmark to identify design problems and to determine solutions due to its relative simplicity compared to an OFC device implementation. Using up-chirp and down-chirp signals facilitates the identification of issues such as bulk mode problems and other necessary transducer design modifications.

![Figure 5.1: Wafer-level RF probing of early UWB OFC device design fabricated on ST Quartz.](image)

The frequency response data in this chapter was obtained using an RF probe, as depicted in Figure 5.1, to measure the two port S-parameters. The time domain response was extracted via a fast Fourier transform (FFT).
5.1 UWB OFC Device Design Evolution

5.1.1 Initial Chirp Device Layout

The initial chirp transducer design, shown in Figure 5.2, was implemented using quarter-wavelength electrodes and was paired with a wideband, polarity weighted, output transducer. Polarity weighting was used in the output transducer to achieve higher conductance, broader bandwidth and lower insertion loss through the device. The experimental results for the up-chirp and down-chirp directions are presented in Figure 5.4. The frequency response for the up-chirp indicates that the inter-electrode reflections were not problematic for inline chips with orthogonal local center frequencies; a suspected issue with quarter-wavelength electrodes in the transducer design.

Figure 5.2: Microscopic image of initial linear stepped chirp dispersive transducer design with quarter-wavelength electrodes. Chip weightings are visible across the device as electrode periodicity increases with frequency.
For this device design, however, an important consequence in the design of wideband chirp filters was observed. The device suffered from significant high frequency attenuation when operated in an up-chirp configuration. For up-chirps with fractional bandwidths above about 25%, bulk mode conversion effects within the dispersive transducer begin to cause the frequency response to roll off [20]. Bulk mode conversion occurs for high bandwidths since the bulk acoustic wave (BAW), which travels at a higher velocity than the SAW, can phase match the electrode pattern at the low frequency device side when a BAW mode frequency is higher than the synchronous SAW mode of the local electrode pattern. The radiating angle of the bulk wave varies with frequency; producing coherent bulk wave radiation that cause a BAW loss for an up-chirp device at the higher frequencies, which does not occur in down-chirp devices. The principle of coherent bulk wave generation in asynchronous arrays is compared to a synchronous array in Figure 5.3.

Figure 5.3: Bulk mode conversion in electrode arrays.
The effect of coherent bulk mode conversion is evident in the experimental up-chirp frequency response shown in Figure 5.4. The effect is even more defined when comparing the time domain magnitude shown in Figure 5.5. The bulk mode conversion problem does not occur in down-chirp devices. For the proper correlation of the received signal and for optimum peak to side-lobe (PSL) ratio, the up and down chirp should be reciprocal.

![Figure 5.4: Experimentally measured frequency response ($S_{21}$) of the initial chirp device design utilizing quarter-wavelength electrodes and polarity weighted output transducer. Frequency response obtained using RF probe station and network analyzer.]

5.1.2 Revised Chirp Device Layout

The attenuation resulting from bulk mode conversion can be eliminated by using more than two electrodes per period or by using a slanted geometry. Continuing with the in-line transducer geometry, all subsequent device designs use sixth-wavelength electrodes. Since
Figure 5.5: Experimental time response of initial chirp device design. The time response is obtained using the FFT of the frequency response shown in Figure 5.4. The time axis is normalized to chip length, $\tau_{chip}$.

Each chip in the OFC transducer has a different local carrier frequency, inter-electrode reflections are greatly reduced compared to a single carrier frequency. In general, this allows quarter-wavelength electrodes to be used in narrow bandwidth correlator designs; however, for bandwidths above 25%, as in the ultra-wideband case, sixth-wavelength electrodes are used to eliminate bulk mode conversion effects within the transducer.

The revised chirp design experimental frequency response is shown in Figure 5.6 and the experimental time response is shown in Figure 5.7. The up-chirp and down-chirp frequency response are nearly reciprocal in frequency and time; showing that sixth-wavelength electrodes effectively eliminated the bulk mode conversion. Using previous experimental data, the velocity under the $3f_0$ grating was experimentally determined to be $3418\text{m/s}$. A known velocity
is an important design parameter needed in centering the output transducer to prevent unintentional in-band frequency attenuation.

![Frequency Response Graph](image)

**Figure 5.6:** Experimentally measured frequency response ($S_{21}$) of revised chirp device design with sixth-wavelength electrodes and apodized ($\cos^{-1}$) output transducer.

Various test devices were fabricated using a $3f_0$ sampled polarity weighted and an inverse cosine apodized output transducer to ensure wave-guiding parasitics were not present in the device. Improvement in the device response from using a polarity weighted or apodized transducer was minimal, however, the use of the apodized output transducer was selected for subsequent designs due to its well defined time domain symmetry. Although the revised design produces a reciprocal time response, the frequency response is distorted by the apodized output transducer at the passband edges.
Figure 5.7: Experimental time response of revised chirp device design. Time response obtained via FFT. The time axis is normalized to chip length, $\tau_{\text{chip}}$.

5.1.3 Final Chirp Device Layout

Rather than attempt to flatten the response by redesigning the output transducer window, the pre-existing apodization on the dispersive transducer chips were used to compensate for the output transducer window and flatten the overall response. To accomplish the compensation, the chip weights calculated using (4.6) are adjusted. The resulting apodization profile is evident in the device shown in Figure 5.8.

The final, compensated, linear stepped chirp frequency response is shown in Figure 5.9. The frequency response magnitude for the up-chirp correlates very well with the down-chirp showing that bulk mode conversion effects were effectively eliminated. The chirp filter pass-band was effectively flattened; compensating for the non-uniform conductance in the output transducer. The improved pass-band shape is better observed in the time response,
Figure 5.8: Microscopic image of final, compensated, linear stepped chirp dispersive transducer design with sixth-wavelength electrodes and apodized input transducer. The compensated chip weightings are visible across the device.

shown in Figure 5.10. The linear time response, shown in 5.10(b), displays the continuous transition in chip cycles over the span of the stepped chirp which occur at the zero crossings. The continuous chip transition is characteristic of orthogonal frequencies and the stepped chirp configuration yields a constant phase throughout the time response.

5.2 OFC Device Results

Using the insight provided through troubleshooting the linear stepped chirp design, the OFC device was implemented by shuffling the chip frequencies as described in Section 3.3. The resulting device, with OFC sequence of \( \{f_6, f_3, f_7, f_1, f_4, f_5, f_2\} \), is shown in Figure 5.11. The distinct colors, produced by the optical diffraction grating effect, revealing the differing OFC chip frequencies. Multiple code sequences were implemented without considering any
Figure 5.9: Experimentally measured frequency response ($S_{21}$) of final stepped chirp device design using compensated apodization weighting on each chip.

Figure 5.10: Experimental time response of final chirp device design. The time axis is normalized to chip length, $\tau_{chip}$. 
code optimization. The OFC device also has output transducers on either side of the OFC transducer; allowing for the up and down OFC direction to be measured on a single device. An additional level of pseudo-noise (PN) phase coding could be implemented in addition to OFC by changing the polarity of the individual chips; adding to the code diversity.

Figure 5.11: Microscopic image of OFC dispersive transducer design. The optical diffraction grating effect reveals the chips. The colors, produced by the wavelength of the transducer electrodes, depict the differing chip frequencies in the code.
5.2.1 OFC Device Experimental Results

The experimental OFC frequency responses for the up and down directions are shown in Figure 5.12. Comparing the up and down directions of the OFC sequence, the frequency response magnitudes compare extremely well at all frequencies and is nearly reciprocal as desired for proper matched filtering. The time response of the OFC device is shown in Figure 5.13. The normalized magnitude response, shown in 5.13(a) in dB, shows a relatively flat response as desired. The linear time response, shown in 5.13(b), enables the determination the code sequence by considering the relative carrier cycles of the experimentally measured signal. The transition of the chip frequencies occurs at the zero crossing for the OFC device, as with the chirp device, due to the properties of the orthogonal frequencies.

![Figure 5.12](image-url)

**Figure 5.12**: Experimental frequency response ($S_{21}$) of a seven chip UWB OFC device with center frequency of 250 MHz and fractional bandwidth of 29%. The up and down directions of the OFC code sequence, as shown in Figure 4.4 are compared in frequency.
Figure 5.13: Experimental time response of UWB OFC device design. The time axis is normalized to chip length, $\tau_{\text{chip}}$. The OFC sequence of $\{f_6, f_3, f_7, f_1, f_4, f_5, f_2\}$ is labeled under each chip in the linear time plot and can be seen by observing the relative number of cycles in each chip.

5.2.2 Coupling of Modes (COM) Simulation Results

The device was simulated using a COM model and compared to experimental results for each stage of design troubleshooting. An observational comparison of model prediction and experimental results were used to easily detect any device design issues. The COM model simulated frequency response is compared to the final experimental OFC device response in
Figure 5.14. The COM model prediction compares very well with the experimental frequency response; nearly matching the experimental OFC response across the entire pass-band.

![Figure 5.14: Coupling of modes simulated device response compared to the experimentally measured OFC device frequency response. Both responses are the up-direction OFC code sequence.](image)

The experimental up-direction OFC time magnitude response is compared with the COM model in Figure 5.15 with the time axis on this figure has been normalized to the chip length. The spike in magnitude occurring at the transition between the third and fourth chips in the experimental up-direction OFC data shows evidence of inter-symbol interference (ISI) in the time domain from the transition between the highest and lowest frequency. This is a result of the OFC sequence chosen for the device which is accurately predicted by the model simulation.
Figure 5.15: Coupling of modes simulated device response compared to the experimentally measured OFC device time response. The time response, shown in normalized magnitude, is obtained using an FFT of the frequency response shown in Figure 5.14. The time axis is normalized to chip length, $\tau_{chip}$. 
5.2.3 Experimental OFC Correlation Results

The experimental OFC correlation between the up and down directions is shown in Figure 5.16 and is compared with the ideal OFC correlation. The experimental correlation results were obtained using S-parameter data from RF probe station measurements of both up-direction and down-direction OFC frequency responses. The ideal correlation response was generated using the mathematically ideal OFC time response. Experimental correlation results are in agreement with ideal predictions with respect to the time ambiguity, pulse width and sidelobe level.

The correlation produces a compressed pulse approximately,

\[
\tau_{\text{pulse}} = 2 \cdot N_{\text{chip}}^{-1} \cdot \tau_{\text{chip}},
\]

or approximately \(0.28 \cdot \tau_{\text{chip}}\) long, which corresponds to a processing gain of 49 resulting from the seven chip, seven frequency OFC signal. This is seven times greater than the processing gain of a PN sequence of the same time length using a single frequency carrier. The traces shown in 5.16(a) and 5.16(b) are narrow and wide time window scalings of the same correlation result. The figures show that the experimental correlation result coincides with the ideal case very well across both time ranges.
Figure 5.16: Correlation results of experimental up and down direction OFC device data compared to ideal correlation response. Experimental device data is obtained from the same dispersive OFC transducer with an apodized output transducer located at equal distance on each side. The two plots show alternate time axis scalings of the same data.
5.3 Double Dispersive Device Results

To further examine the operation of dispersive transducer, a device was fabricated using two dispersive OFC transducers and omitting the wideband input transducer completely. Each dispersive transducer used a differing OFC code. The resulting time domain signal is spread to twice the length of the device with a single dispersive transducer. The frequency response resembles that of a signal using orthogonal frequency division multiplexing (OFDM). The properties of the double dispersive device increase the level of code diversity beyond what is offered by OFC alone.

![Double Dispersive OFC Device](image)

**Figure 5.17:** Double Dispersive OFC Device

The fabricated double dispersive OFC device is shown in Figure 5.17. The transducer design is identical to those used in the final OFC device layouts, including the conductance compensation. One of the transducers is rotated to aid RF probe placement. The resulting frequency response is shown in Figure 5.18 and is compared to the COM model simulation.
The response shape is augmented when compared to previous frequency responses. This is a result of adjusting the apodization of dispersive transducer to compensate for the apodized input transducer.

**Figure 5.18:** Experimental frequency response ($S_{21}$) of a double dispersive OFC device compared to COM model simulated response. Each dispersive transducer contains differing seven chip OFC codes device with center frequency of 250 MHz.
CHAPTER 6
UWB OFC CORRELATOR AND SYSTEM

The UWB OFC device design was optimized using an RF probe station to detect device faults. Each transducer has $3f_0$ electrode sampling in order to eliminate bulk mode conversion effects. The device was then bonded and packaged in SMT packaging and soldered to a custom printed circuit board. The performance of the packaged device was then measured for comparison to the results found using wafer-level RF probing. The performance of the resulting correlator circuit is first measured using the vector network analyzer, then in a prototype system, similar to a time-domain test set, built using various RF components.

The packaged device measurements presented here are obtained using a network analyzer for frequency domain data and using the FFT to achieve the time domain response. The final time domain response is obtained with a prototype RF system configuration using an Agilent DSO6032A digital sampling oscilloscope as an analog to digital converter and output. The system output produces a compressed pulse through matched filtering the coded signal and the system results demonstrate the performance of the OFC correlator in an UWB system.
6.1 Packaged Device Results

Each device was bonded to a surface mount package and soldered onto a custom prototype PCB as shown in Figure 6.1. The PCB was designed to reduce the parasitic RF feed-through by grounding all floating leads and isolating the input and output portions of the device into separate ground planes. The device pictured in Figure 6.1 is configured with connectors to allow the packaged device to be measured in both directions. This eases the testing of matched filter operation with a device in the opposite configuration. The device edges were cut at an angle and coated with an absorber of photoresist to reduce reflections of the device edges.

Figure 6.1: Packaged UWB OFC device in standard SMT packaging. The center connector is attached to the dispersive OFC transducer. The two outer connectors are attached to the wide-band apodized transducers. This permits the packaged device to be operational in either direction; easing the matched filtering operation measurements.
The packaged device was measured directly on a network analyzer and is presented without the gating of RF feed-through or reflections in the device. The frequency domain data is compared to the COM model simulation in Figure 6.2. The packaged device response compares very well across the entire passband of interest and the low frequency side-lobes. The electrical parasitic effects due to packaging and RF feed-through are evident in the upper-band side-lobes but are minimal and not problematic.

**Figure 6.2:** Packaged UWB OFC device experimental frequency response compared to the COM simulated device response. Data acquired using network analyzer.
The time magnitude response of the packaged device obtained using the FFT of the network analyzer data is shown in Figure 6.3. The time response is plotted in dB showing a very flat response across all chips and is predicted very well by the COM model.

![Figure 6.3: Time magnitude response comparison of the experimental packaged UWB OFC device and the COM simulated device. Experimental device results obtained using FFT.](image)
6.2 Experimental OFC Correlation Results

Prior to measuring the UWB OFC device in a system, the correlation for the packaged device was measured using a network analyzer. The compressed pulse is obtained by cascading UWB OFC device in series with its matched filter. This was achieved using a typical SMA coupler to connect an OFC device inline with its matched filter. The correlation peak is obtained by simply taking the Fourier transform of the network analyzer data. In Figure 6.4 the device correlation results are shown for the packaged device. The figure compares the correlation results of the packaged device to the wafer-level RF probe data and the auto-correlation of a mathematically ideal OFC signal obtained previously. The three responses compare favorably and have minimal increases in the side-lobes levels. The responses in 6.4(a) and 6.4(b) offer alternate time scaling of the same correlation results in attempt to show that there is no major increase in side-lobe level further out in time from the compressed pulse peak.
Figure 6.4: Correlation results of packaged UWB OFC device in series with its matched filter as measured on network analyzer (blue); compared to RF probed results and ideal correlation.
6.3 OFC System Measurement

The system response using the OFC UWB correlator module after match filtering was obtained using a prototype system shown in schematic form in Figure 6.5.

![Figure 6.5: UWB OFC test system block diagram. Values used for the system measurement are shown in Table 6.1.](image)

The transmitter end of the system uses a programmable pulse generator with a pulse width which produces a frequency spectrum wide enough to excite the entire bandwidth of the UWB device. This pulse is mixed with the center frequency of the OFC device via double balanced mixers, resulting in a gated RF burst. The power splitter provides an accurate trigger needed later by the digital oscilloscope. A variable attenuator is used to prevent the saturation of the signal amplifiers. The UWB OFC device in the receiver side of the system is used to generate the coded signal with its output connected directly to the receiver side.
of the system (this excludes the propagation path in the prototype system). The receiver portion of the system uses the matched filter to output the resulting correlation peak on the digital oscilloscope.

Table 6.1: System Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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</thead>
<tbody>
<tr>
<td>$f_0$</td>
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</tr>
<tr>
<td>Pulse Width</td>
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</tr>
<tr>
<td>Lead Edge</td>
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</tr>
<tr>
<td>Trail Edge</td>
<td>2 ns</td>
</tr>
<tr>
<td>Period</td>
<td>2 µs</td>
</tr>
</tbody>
</table>

Figure 6.6: UWB OFC test system photo. The packaged SAW devices are visible near the center of the photo.
6.3.1 OFC System Measurement Results

The output of the correlated system response is shown in Figure 6.7. The highest side-lobe level is approximately 16 dB down as expected from the measurements taken from the network analyzer shown in Figure 5.16. The correlation peak is centered precisely at $1.31\mu s$ as measured using the oscilloscope. This value was also observed on the network analyzer FFT response. The center of the resulting compressed pulse in time, $\phi_{\text{pulse}}$, can accurately be predicted using Equation 6.1; where $N_{\text{chip}}$ is the total number of chips and $\tau_{\text{delay}}$ is the device delay caused by the distance between the apodized transducer and the dispersive OFC transducer. The location of the compressed pulse is the time delay between the beginning of the RF feed-through to the pulse peak as shown in Figure 6.8 for the linear stepped chip device. A processing gain of 49 results from the seven chip, seven frequency OFC signal.

$$\phi_{\text{pulse}} = N_{\text{chip}} \cdot \tau_{\text{chip}} + 2 \cdot \tau_{\text{delay}}$$  \hspace{1cm} (6.1)
Figure 6.7: OFC system output waveform and correlation peak as captured using a digital oscilloscope.

Figure 6.8: Linear stepped chirp auto-correlation output. The RF feedthrough of the system is observation on the left side of figure. The time difference from the beginning of the RF feedthrough to the center of the compressed pulse is $1.31\mu s$ as predicted by Equation 6.1.
6.3.2 OFC System Cross-Correlation Measurement

The system cross-correlation is obtained by replacing the matched filter SAW device in the receiver side of the system with a device with a completely different OFC sequence. The cross-correlation is compared with an auto-correlation response in Figure 6.9 using identical axis scaling. The cross-correlation output shows that there is little energy at the location of the correlation peak. The correlator therefore has rejection of dissimilar code sequences as desired in a multiple access spread spectrum communication system. The code sequences of both devices were implemented without considering any code optimization and therefore the code rejection performance may somewhat improve with a properly defined code set.
Figure 6.9: OFC auto-correlation (a) and cross-correlation (b) system output. Both plots shown with identical axis scaling of 50mV/div and 50µs/div are obtained using digital oscilloscope.
6.4 Wireless Operation

A test of the OFC correlator wireless operation was performed using unmatched dipole antennas where the antenna bandwidth was inherently narrower than the UWB device. The test setup used a RF signal amplifier with gain of 12dB was used on the transmit side just before the antenna on port one of the network analyzer. The signal was transmitted wirelessly over a distance of approximately one foot and received via a similar antenna and the OFC matched filter connected to port two of the network analyzer. A defined correlation peak was observed in the time domain of the network analyzer. The resulting peak was attenuated by approximately 10dB due to the highly lossy dipole antennas and slightly shifted due to the additional delay introduced by the propagation path. The wireless test setup is shown in Figure 6.10. The devices and dipole antennas are fixed and held stationary. The compressed pulse obtained is visible on the network analyzer in the photo. The correlation response was obtained on the network analyzer using the built in FTT functionality. This quick test demonstrated the performance of the UWB OFC system in noisy environments.
Figure 6.10: Wireless operation photo. The two dipole antennas used and packaged devices can be seen on either side of the photo. The correlation peak can be seen on the network analyzer.
CHAPTER 7
CONCLUSIONS

This thesis presented the development of SAW correlators using orthogonal frequency coding for use in ultra–wideband spread spectrum communications systems. The OFC SAW correlator offers numerous advantages beyond conventional UWB techniques due to its enhanced processing gain and greater multiple access operation due to greater code diversity. The use of OFC also overcomes many of the challenges in UWB deployment. The conventional sub-nanosecond pulse may reverberate in indoor environments; making proper reception difficult even with complex reception algorithms. The increased code diversity and accurate pulse time predictability ease the synchronization and demodulation of UWB data signals.

The UWB OFC device embodiment consisted of a wideband input transducer and an inline dispersive transducer. Device development utilized a linear stepped chirp filter configuration to aid in the identification of bulk mode problems and necessary transducer design modifications. It is shown that the shape of the transfer function can be enhanced by adjusting the conductance of chips within the dispersive transducer, thereby reducing unintended signal distortion. Additionally, the individual chip weightings may be adjusted to achieve a flat, or even a nearly arbitrary, frequency response.
The experimental results of an UWB OFC SAW device with 250 MHz center frequency, 29% fractional bandwidth was presented. The up and down direction OFC experimental response correlate with each other very well and are nearly reciprocal, as desired. Experimental results were compared to simulated COM model results and agree very well in both frequency and time. The compressed pulse with a processing gain of 49 was verified from experimental data.

The UWB OFC SAW correlator was configured for operation in a prototype UWB system by bonding devices in SMT packages and developing custom printed circuit board designs to minimize RF feedthrough. The packaged device results were presented and compared to simulated COM model showing strong comparison. Experimental correlation was obtained for the packaged devices and had excellent performance; which was similar to the probed wafer device correlation and an ideal correlation response. The resulting compressed pulse from the prototype system was shown for the UWB OFC correlator after matched filtering and was compared to the cross-correlation signal. The cross-correlation was obtained using packaged devices with dissimilar code sequences, while the auto-correlation was obtained using the proper matched filter. The code sequences were selected at random and did not include any optimization. A wireless configuration experiment also yielded a promising compressed pulse output despite the inefficient antenna design.

These results demonstrate the use of UWB OFC correlators is a feasible and viable technique for high performance UWB communication systems.
APPENDIX A
DEVICE LAYOUTS
A.1 UWB OFC Device Layouts

The following section provides the details of transducer layout and design. Each figure is a scale vector figure converted from the actual CAD layouts. When this thesis is viewed in its electronic format, the reader can utilize the pdf reader’s magnification tools (zoom) to observe details of transducers, individual chips and electrodes. The final, optimum device designs from mask DG5 are listed first and are followed by previous, interim design configurations. The large structures surrounding each transducer are the RF probe pads and have an angle of 15° to prevent undesired reflections. The delay between input transducer and dispersive transducer was arbitrarily set to 40 mil for each design.
Figure A.1: Final linear stepped chirp device layout. All transducers have sixth-wavelength electrodes with $3f_0$ sampling. The input transducers are apodized with an inverse cosine ($\cos^{-1}$) window. Mask device label is DG5-A.
Figure A.2: Final layout of UWB OFC Filter. All transducers have sixth-wavelength electrodes with $3f_0$ sampling. The input transducers are apodized with an inverse cosine ($\cos^{-1}$) window. Mask device label is DG5-B.
Figure A.3: Final layout of UWB OFC Filter. All transducers have sixth-wavelength electrodes with $3f_0$ sampling. The input transducers are apodized with an inverse cosine ($\cos^{-1}$) window. Mask device label is DG5-C.
Figure A.4: Early device layout. All transducers have quarter-wavelength electrodes with $2f_0$ sampling. Mask device label is DG2-A.
Figure A.5: Initial linear stepped chirp device layout. Input transducer is polarity weighted. All transducers have quarter-wavelength electrodes with $2f_0$ sampling. Mask device label is DG3-A.
Figure A.6: All transducers have quarter-wavelength electrodes with $2f_0$ sampling. The input transducers are apodized with an inverse cosine ($\cos^{-1}$) window. Mask device label is DG3-B.
Figure A.7: All transducers have sixth-wavelength electrodes with $3f_0$ sampling. The input transducers are polarity weighted. Mask device label is DG3-C.
Figure A.8: Conceptual parallel track device layout. Mask device label is DG3-TRK
Figure A.9: Linear stepped chirp device layout without input transducer conductance compensation. All transducers have sixth-wavelength electrodes with $3f_0$ sampling. The input transducers are apodized with an inverse cosine ($\cos^{-1}$) window. Mask device label is DG4-B.
A.2 Input Transducer Layouts

Figure A.10: Detail of polarity weighted input transducer.
Figure A.11: Detail of Inverse cosine apodized input transducer.
APPENDIX B
DEVICE DESIGN PARAMETERS
### Table B.1: Design parameters of initial device layout.

<table>
<thead>
<tr>
<th>DG3–A</th>
<th>Inline Chirp — Wide-band Polarity Weighted Output Transducer</th>
<th>(\lambda/4) Fingers</th>
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<tr>
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<td>(f_{\text{chip}})</td>
<td>velocity</td>
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<td></td>
<td>MHz</td>
<td>(m/s)</td>
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<td>(f_0)</td>
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<td>(f_2)</td>
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<td>3424.217</td>
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### Table B.2: Design parameters of final device layout.

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<th>(\lambda/6) Fingers</th>
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<td>(f_1)</td>
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</tr>
<tr>
<td>(f_4)</td>
<td>260.417</td>
<td>3418</td>
</tr>
<tr>
<td>(f_5)</td>
<td>270.833</td>
<td>3418</td>
</tr>
<tr>
<td>(f_6)</td>
<td>281.25</td>
<td>3418</td>
</tr>
</tbody>
</table>

Output: 250 | 3418 | 13.672 | 2.2787 | \(\cos^{-1}\) | 37.25 |

Velocity: 3418 m/s | 134.5669 mil/\(\mu\)s

Separation: \(\tau_c \cdot v\) | 12.91843 mil
LIST OF REFERENCES


