Simulation and Experimentation of Complementary-Coded Pulse Radar for Ice Measurement

BY

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Abstract

Sea level is an important indicator of global climate change. An accurate determination of the mass balance of polar ice sheets is required to assess their contribution to sea level rise. Ice thickness, surface and bed topography, interannual layers, and ice velocity are some important parameters for the mass balance estimation of ice sheets.

Information about the ice layers near the ice bed is essential to understand past ice dynamics. The challenges in mapping ice layers near the ice bed are twofold. First, radar signals are much attenuated when they reach the ice bottom and return to the receiver. Second, because the ice bed acts as clutter, the returns from it usually mask the ones from the target ice layers near it. Complementary-coded pulse radar may be a solution in the sense that it transmits complementary-coded pulses and compresses the received pulses so that highly effective transmitted power, fine-range resolution, and low sidelobes level can be achieved simultaneously.

In this thesis, computer simulations and experimentation in the laboratory have been performed to study the sidelobe cancellation of complementary-coded pulse radar for the purpose of building a prototype. A simulation model was first built that includes waveform generator, modulators, targets, demodulators, waveform compressors, and envelope detector. With the simulation model, sidelobe cancellation was verified for ideal
case. In addition, the effects of phase shift and amplitude imbalances on sidelobe level were evaluated. It was found that the magnitude of the maximum sidelobe peak increases linearly as phase shift increases. Doppler phase shift is not a concern for complementary-coded radar to be carried on an airplane for ice measurement. The peak sidelobe level (PSL) was found to increase nonlinearly as amplitude imbalances increase. Experiments were then performed in the laboratory to understand the effects of hardware imperfections on sidelobe cancellation. The waveform amplitude imbalances caused by the waveform generator, modulator, and demodulator were found to be the main obstacle to sidelobe cancellation. Simulations that included the averaged amplitude imbalance of the measurements were used to verify the experimental results. Finally a prototype of complementary-coded pulse radar was proposed for future work.
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Chapter 1: Introduction

1.1 Background of Ice Measurement

Atmosphere, oceans, polar ice sheets, and glaciers are important variables having close interactions with each other in the ecosystem of Earth. The facts show that the planet is warming, sea level is rising, and polar ice sheets and glaciers are shrinking and melting. During the last century the average global temperature has climbed about 0.6 degree Celsius, and the average global sea level has risen about 15 centimeters. Arctic sea ice area has shrunk by about 9% per decade and thinned as well since 1978. Now, fewer than 30, greatly shrunken glaciers remain in Montana’s Glacier National Park, which had some 150 glaciers in 1910 when the park was designated [1].

A major concern is that as the process continues, it may cross the threshold beyond which the ecosystem will never recover. A direct, devastating consequence of sea level rise is the flooding of the coastal regions, resulting in migration of people from these regions and a huge loss of land and property. Worldwide, about 100 million people live within 1 meter of sea level. Typically a 50-centimeter sea level rise would cause a coastline retreat of 50 meters in flat coastal areas. Greenland’s ice sheet alone holds enough water to raise sea level by about 7 meters.

Therefore, the significance of sea level rise observation and prediction is apparent. Scientists believe that about half of the current sea level rise can be attributed to thermal
expansion of the ocean and melting of mountain glaciers [2]. Scientists have speculated that the major ice sheets in Antarctic and Greenland have contributed the remainder of the sea level rise as they account for 80% of Earth’s fresh water. An accurate determination of the mass balance—net gain or loss of ice—of these ice sheets is required to assess their contribution to sea level rise. Ice thickness, surface and bed topography, interannual layers, and ice velocity are important parameters for the mass balance estimation of ice sheets.

1.2 KU’s Contribution to Ice Measurement

Since the 1980s, the University of Kansas (KU) has developed radar depth sounder, airborne radar depth sounder, airborne accumulation radar, snow radar, and synthetic aperture radar for the purpose of ice measurement [3] [4] [5] [6] [7].

In the 1980s the Radar Systems and Remote Sensing Laboratory (RSL) at KU built its first polar radar—the Coherent Antarctic Radar Depth Sounder (CARDS)—for measuring the thickness of the ice sheets in Antarctica. An airborne pulse radar, CARDS used complementary analog expansion and compression techniques. Its data processing system consisted of two 8-bit A/D converters and a signal processing system capable of doing coherent integrations. It was then the only radar with minimal transmitted power (20 W), high loop sensitivity (186 dB), and high resolution (5 m), according to reference [3]. In 1993 RSL joined the NASA program and used CARDS aboard a NASA P3 aircraft over the Greenland ice sheet to collect ice thickness data. The performance of CARDS was evaluated in the field experiment, and the design problems were understood.
The CARDS loop sensitivity was found to be 20 dB less than its expected value. CARDS was therefore redesigned to eliminate the problems and to optimize performance. The new system is called the Improved Coherent Antarctic and Arctic Radar Depth Sounder (ICARDS). Its transmitted power was increased to 200 W, and its loop sensitivity was improved to 196 dB. ICARDS was used in the 1995 Greenland ice sheet mapping missions and was able to generate useful, good-quality data about the ice sheet thickness and some internal layers. In 1996, ICARDS was further improved by using radio frequency integrated circuits (RFICs) and microwave monolithic integrated circuits (MMICs) to replace the bulky connectorized components. Including a data acquisition system with 12-bit A/D converters [8] resulted the Next-Generation Coherent Radar Depth Sounder (NG-CORDS). NG-CORDS has been used in field experiments since 1996. All the data collected in field experiments with the KU radars were processed, distributed to scientists worldwide, and stored on the server at the University of Kansas (http://tornado.rsl.ku.edu/Greenlanddata.htm).

In fall 2001, KU received a grant of $8.7 million from the National Science Foundation (NSF) and the National Aeronautics and Space Administration (NASA) to start a new project called Polar Radar for Ice Sheet Measurement (PRISM) to develop intelligent, ground-level radar sensors capable of measuring more key ice sheet parameters such as bedrock conditions and internal layers in both shallow and deep ice in addition to ice thickness. For the project, two radar systems are being developed. The first is a monostatic/bistatic synthetic aperture radar (SAR) to provide a 2D picture of basal conditions of the polar ice sheet. The second is a wideband, dual-mode radar to
measure ice thickness and map near-surface and deep internal layers. Also in
development are a semiautonomous ground rover to carry the radars and a wireless
communication system that allows the rover and radars to communicate with one another
as well as to transmit near-real-time data back to other researchers and educators in
distant locations. More general information about PRISM can be found on its website at
http://ku-prism.org. Recent progress and field experimental results can be found in
references [9] and [10].

In April 2005, NSF granted $19 million to establish the Center for Remote
Sensing of Ice Sheets (CReSIS) at KU in order to better understand the mass balance of
the polar ice sheets and its contribution to sea level rise and global climate change.
CReSIS is working to create new technologies for studying polar ice and new means of
interpreting the data. The effort focuses on remote sensing technology and integrates
expertise in electrical engineering, information technology, aerospace engineering,
glaciology, and geophysics. More information about CReSIS can be found at
http://cresis.ku.edu.

1.3 Thesis Objectives and Organization

The objective of this thesis is to study the technology required to map the ice
layers near the ice bed. Information about ice layers near the ice bed is essential to
understanding past ice dynamics. The challenges in mapping ice layers near the ice bed
are twofold. First, radar signals are very weak because of huge attenuation when they
reach the ice bottom and return to the receiver. Second, ice bed echoes are much stronger
compared to those from layers near the interface, so their sidelobes usually mask the returns from the target ice layers near the ice bed. One possible solution is to transmit complementary-coded pulses and to compress the received pulses so that high effective transmitted power, fine-range resolution, and low sidelobes level can be achieved simultaneously.

The thesis consists of six chapters. Chapter 2 defines complementary codes and their synthesis methods, and introduces the principle of complementary-coded pulse radar. Chapter 3 describes a Visual System Simulator (VSS) system simulation model of complementary-coded pulse radar and presents simulation results. Chapter 4 presents experimental lab results. Chapter 5 proposes an architecture for a prototypic complementary-coded pulse radar. Finally, Chapter 6 summarizes the results and proposes recommendations for future work.
2.1 Complementary Codes

Complementary codes were first studied comprehensively by Marcel J. E. Golay in 1960; they are therefore also called Golay codes. In his paper [11], Golay created the definition of complementary codes, and studied their properties and synthesis methods.

Golay defined a pair of complementary codes as two equally long finite sequences of two kinds of elements, which have the property that the number of pairs of like elements with any given separation in one sequence is equal to the number of pairs of unlike elements with the same separation in the other sequence. Figure 1 shows a pair of complementary codes, code I (00010010) and code II (00011101), to help clarify the definition. Each code has a length of 8 bits, and the two kinds of elements are 1 and 0. In the figure, L is used to denote a pair of like elements and U a pair of unlike elements in code I or code II. For the given separation of 2, it is shown that code I has one pair of unlike elements and four pairs of like elements, and code II has four pairs of unlike elements and one pair of like elements. Any other separations (0, 1, 3, and so forth) can be used to check out the property of the pair of complementary codes.
The properties of complementary codes can be used to generate new complementary codes once a pair of complementary codes is known. A pair of complementary codes remains complementary after the following operations:

- **Operation 1:** Either code or both codes are reversed.
- **Operation 2:** The two kinds of elements in either code or both codes are interchanged.
- **Operation 3:** The kind of elements at even order in both codes is altered to the other kind.

Appending and interleaving operations can be used to synthesize longer complementary codes. Assuming codes $A = a_1a_2...a_N$ and $B = b_1b_2...b_N$ are a pair of complementary codes of length $N$ bits, and $B' = b'_1b'_2...b'_N$ is the altered version of $B$ where the prime of a bit denotes the change of the element kind from one to the other, then

- **Operation 4:** $C = AB = a_1a_2...a_Nb_1b_2...b_N$ and $D = AB' = a_1a_2...a_Nb'_1b'_2...b'_N$ results in a pair of complementary codes of length $2N$ bits. And
• Operation 5: \( E = a_1b_1a_2b_2...a_Nb_N \) and \( F = a'_1b'_1a'_2b'_2...a'_Nb'_N \) also results in a pair of complementary codes of length \( 2N \) bits.

### 2.2 Principle of Complementary-Coded Pulse Radar

A pulse radar detects objects by measuring the time delay between transmitted and reflected pulses. As illustrated in Figure 2 [12], the maximum unambiguous range \( R_u \) and the range resolution \( \Delta R \) are determined by

\[
R_u = \frac{cT}{2} = \frac{c}{2PRF} \tag{1}
\]

\[
\Delta R = \frac{c\tau}{2} \tag{2}
\]

![Figure 2: Unambiguous Range and Range Resolution of Pulse Radar](image-url)
where \( c \) is the travel speed of electromagnetic waves in medium, \( T \) is the period of transmitted pulses, and its reciprocal is known as pulse repeat frequency (PRF). \( \tau \) is the pulse width.

According to radar equation, the maximum radar range \( R_{\text{max}} \) is

\[
R_{\text{max}} = \left[ \frac{P_t G \sigma A_e}{(4\pi)^3 S_{\text{min}}} \right]^{1/4}
\]

(3)

where \( P_t \) is the power of the transmitted pulse, \( G \) is the antenna gain, \( \sigma \) is the radar cross section (RCS), \( A_e \) is the antenna’s effective aperture, \( S_{\text{min}} \) is the minimum detectable signal power and

\[
P_t = P_{\text{max}} \tau \ \text{PRF}
\]

(4)

From equation (2), it is seen that fine-range resolution (small \( \Delta R \)) requires short pulses (small \( \tau \)), and from equations (3) and (4) for the detection of distant targets, the high pulse peak power \( P_{\text{max}} \), long pulse (bit \( \tau \)), or high PRF are required once \( G \), \( \sigma \), \( A_e \), and \( S_{\text{min}} \) are set. But high PRF reduces the maximum unambiguous range according to equation (1). Thus short pulses with high peak power are needed for both long-range detection and fine-range resolution. However, it is difficult to generate high-peak power pulses, and high-peak power pulses may result in dielectric breakdown of transmission lines. For this reason, the technique of pulse compression is used, in which frequency- or phase-coded long pulses are transmitted, and received pulses are decoded to obtain short pulses with high-pulse peak power.

The pulse compression is implemented by matched filter. As illustrated in Figure 3, a pulse with magnitude of 1 is coded with an 8-bit binary code. At the output of the matched filter, the waveform is compressed into a main lobe and symmetric sidelobes on
each side. The magnitude of the main lobe is increased from 1 to 8, and its width is 2/8 of the original pulse length.

\[
\begin{align*}
\text{Coded Pulse} & \quad \rightarrow \quad \text{Matched Filter} \quad \rightarrow \quad \text{Compressed Pulse}
\end{align*}
\]

**Figure 3: Pulse Compression**

The following parameters describe the property of pulse compression [13]:

\[
PSL = 10 \log \left[ \frac{\text{Max}(x_i^2)}{x_0^2} \right]
\]  

(5)

\[
ISL = 10 \log \left[ \sum_{i=0}^{N} \frac{x_i^2}{x_0^2} \right]
\]  

(6)

where \( x_i \) represents the magnitude of all sidelobes and \( x_0 \) is the magnitude of the main lobe. \( PSL \), the peak sidelobe level, is a measure of the largest sidelobe as compared with the main lobe. \( ISL \), the integrated sidelobe level, is a measure of the total power in the sidelobes as compared with the main lobe power.

Sidelobes are an undesirable outcome of pulse compression either in range or time. Since the sidelobes of a strong target may mask weak returns from a nearby target, it is preferable to delete or reduce the sidelobes. Complementary codes discussed in the previous section are therefore employed for this purpose. For a pair of complementary
codes with elements either 1 or –1, the sidelobes of the compressed waveform from one
code is the inverse of the ones from the other code. When the compressed waveforms
from two codes are added together, the sidelobes will be totally canceled in the ideal case,
and the magnitude of the main lobe will double. This property of complementary codes
can be explained in terms of their autocorrelation functions. The autocorrelation function
of codes $A$ and $B$ is defined by equations (7) and (8), respectively:

$$R_A(j) = \sum_{i=1}^{N-j} a_ia_{i+j}$$ (7)

$$R_B(j) = \sum_{i=1}^{N-j} b_ib_{i+j}$$ (8)

where $N$ is code length in bits and $-(N-1) \leq j \leq (N-1)$. If codes $A$ and $B$ are a pair of
complementary codes and the elements in the codes are either 1 or –1, then

$$R_A(j) + R_B(j) = \begin{cases} 2N & j = 0 \\ 0 & j \neq 0 \end{cases}$$ (9)

Figure 4: Autocorrelation Functions $R_A$ (red), $R_B$ (blue), and $R_A + R_B$ (black)
For example, after the 0s in codes I and II are replaced with –1s, their autocorrelation functions are calculated and illustrated in Figure 4, in which the red plot is the autocorrelation function of code I, the blue plot is the autocorrelation function of code II, and the black plot is the sum of the two autocorrelation functions. Figure 4 shows that the waveform of each code has a main lobe located at the center with a magnitude of 8, which is the code length, and the sidelobes of two codes are opposed to each other so that when summed together the sidelobes are canceled and the peak of the main lobe is increased to 16, which is twice the code length.

2.3 Application Limitations

Reference [14] reported a real application of complementary-coded pulse radar to the measurements of the structure and the dynamics of mesosphere.

There are two application limitations considered in this thesis. The first one is what is called the Doppler phase shift. The second one is amplitude imbalances caused by imperfections in radar hardware. Since the radar is to be carried on an airplane, after it transmits the first pulse coded by the first code in the complementary pair, it will move to a different place when transmitting the second pulse coded by the other code. The moving radar results in what is called the Doppler phase shift between the two pulses. Also because hardware cannot be built perfectly, any imperfection may cause signal amplitude or phase imbalances. For example, different charging and discharging characteristics of capacitors make it impossible to get exactly the same amplitude when generating one positive volt and one negative volt.
What is the effect of Doppler phase shift on sidelobes? How sensitive are complementary codes to hardware imperfections? These questions have to be answered before building the complementary-coded pulse radar. Chapters 3 and 4 estimate the effects of phase shift and amplitude imbalances on sidelobe levels either by system simulations with software or by hardware experiments in the laboratory.
Chapter 3: ADS and VSS System Simulation

3.1 Introduction

The objective of this chapter is to perform system simulation of complementary-coded pulse radar. The system parameters are summarized and the simulation model is described in detail in section 3.2. The sidelobe cancellation, Doppler phase shift, and hardware amplitude imbalance effects are studied in section 3.2, and the results are presented and discussed in section 3.3.

The software tools employed for the simulation are the Advance Design System (ADS) from Agilent Technologies and the Visual System Simulator (VSS) from Applied Wave Research, Inc. The basic use of ADS is discussed in reference [15]. The basic use of VSS is discussed in reference [16]. The architecture of the simulation model in VSS is presented here because of personal preference, although the models built with the two software tools are equivalent.

3.2 Simulation Model

The important parameters of the complementary-coded pulse radar are as follows:

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
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<tr>
<td>Carrier Frequency</td>
<td>150 MHz</td>
</tr>
<tr>
<td>Transmission Pulse Width</td>
<td>2 µs</td>
</tr>
<tr>
<td>Pulse Repetition Frequency</td>
<td>10 kHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>32 MHz</td>
</tr>
<tr>
<td>Simulation Time Step</td>
<td>2/320 µs</td>
</tr>
<tr>
<td>Code Length</td>
<td>32 bits</td>
</tr>
</tbody>
</table>
A pair of known complementary codes of 16-bit length 1001000010100011 and 100100001011100 is given in reference [11], so Operation 4 in section 2.1 generates the following complementary codes of 32-bit length required by the simulation:

code A 10010000101000111001000001011100

code B 10010000101000110110111110100011

The top layer of the simulation model is shown in Figure 5. The model consists of blocks for transmitter, targets, IQ demodulator, I and Q channel digital compressors, and envelope detection and display, which are shown in Figures 5-1, 5-2, 5-3 and 5-4, respectively, for clarity. The model performs the following operations in order:

1) Generates pulse waveforms coded by complementary codes A and B in a time span of $2\mu s$;

2) Modulates the carrier with the complementary codes by quadrature multiplexing;

3) Transmits the signal modulated by code A first, and transmits the signal modulated by code B after $0.1ms$;

4) Demodulates returned signal from targets;

5) Compresses the demodulated signal;

6) Delays the compressed signal from code A by $0.1ms$, and then adds it to the compressed signal from code B;

7) Calculates the envelope of the summed compressed signal;

8) Repeats the process from steps 1 to 7 for the next pulse period.
Figure 5: Top Layer of Simulation Model
Figure 5-1: Transmitter, TargetsModel, and IQ Modulator
I Channel Digital Compressor

Figure 5-2: I Channel Digital Compressor
Figure 5-3: Q Channel Digital Compressor
Envelope Detection and Display

From I Channel Digital Compressor

From Q Channel Digital Compressor

Figure 5-4: Envelope Detection and Display
The subsystem blocks and their sublayers in the hierarchy as well as their functions are explained next.

As shown in Figure 6, the Transmitter block consists of waveform generator and IQ modulator. The waveform generator repeatedly generates pulse waveforms coded by code A (top) and code B (bottom) using Digital Source and Square Wave Generator components. The waveform of each pulse takes up $2 \mu s$ and is multiplied with the timing signal from the CodeClock block that implements the transmitting alternation between waveforms coded by codes A and B. The CodeClock also acts as the blanking signals in a radar system. Following the waveform generator, the IQ modulator performs quadrature multiplexing of both codes using a carrier tone of 150 MHz and its $90^\circ$-phase-shifted version and transmits the modulated signals to targets.

Figure 7 shows the CodeClock block in which a pulse train of unit magnitude is generated at a rate of 5 kHz with a duty cycle of 1% to serve as the timing signal for waveforms coded by code A. The same pulse train is delayed by 0.1 ms to serve as the timing signal for waveforms coded by code B. Figure 8 shows the waveforms coded by codes A and B. Figure 9 illustrates the output timing signals from CodeClock. Figure 10 shows the transmitted signals.
Two targets are modeled in the Targets block shown in Figure 11 in which Delay blocks are used to account for the two-way propagation time and Scale blocks are used to account for the signal attenuation.

Figures 12 and 13 are the IQ demodulators for codes A and B respectively, in which a carrier tone of 150 MHz and its 90°-phase-shifted version are used to demodulate the in-phase and the quadrature components of the received signal. The only difference between these two IQ demodulators is that the one for code B has a phase shifter right after the carrier tone, which makes it possible to estimate the phase shift effect. The final application of the complementary-coded pulse radar will be onboard a moving vehicle such as an uncrewed aerial vehicle (UAV). The vehicle is at some specific location in space when the first coded pulse is transmitted and will have moved to a new location in space when the second coded pulse is transmitted 0.1 ms later. Thus the traveled distances of the two complementary coded pulses to the same target are different. This difference in traveled distances can be seen as a phase shift of the second coded pulse with respect to the first.

To compress the pulses, the outputs of the demodulators for both codes are multiplied by the timing signals from the CodeClock in I channel and Q channel so that the pulse coded by code A is sent to the digital compressor for code A and the pulse coded by code B is sent to the digital compressor for code B.
Figure 14 shows the architecture of the correlators for pulse compression. The first row consists of 31 Delay blocks. The delay time of each Delay block is one-bit time, which is \(2/32\mu s\) in the simulation model. The second row consists of 32 Scale and Offset blocks with the offsets set to zero and the scale set to 1 or \(-1\). The Summer at the bottom adds all signals from the 32 taps. The output of the Summer is the compressed waveform. When the incoming signal enters the digital filter from the left port, for the first tap it is scaled by 1 or \(-1\) and is sent to the Summer without any delay. For the second tap the signal is delayed by one-bit time and is scaled again by 1 or \(-1\) and is sent to the Summer. The process is repeated until it passes through all 32 taps. The scale used for each tap is decided by the code. From left to right, the scales of the 32 taps are the reversal of the code itself with 1 mapped to 1 and 0 mapped to \(-1\). So for different codes, the architecture of the correlators is similar; the only difference is the scale of each tap.

Since the pulse coded with code A is transmitted first, in each channel the compressed waveform for code A is delayed by a pulse period and then added to that for code B to cancel the sidelobes.

Finally the summed waveforms from channels I and Q are squared, added together, and the square root is taken to calculate the envelope. The final signal is used to evaluate system performance in terms of peak sidelobe level (PSL) defined by equation (5).
Transmitter

Waveform Generator

IQ Modulator

Figure 6: Transmitter Block
Figure 7: CodeClock Block

Figure 8: Pulse Waveforms Coded by Code A (left) and Code B (right)

Figure 9: Timing Signals from CodeClock
Figure 10: Transmitted Signals

Figure 11: Targets Block
Figure 12: IQ Demodulator Block for Code A

Figure 13: IQ Demodulator Block for Code B
Digital Compressor

Figure 14: Architecture of Correlator
3.3 Simulation Results

Simulation results are reported in this section. Simulations were performed for the following cases and purposes:

1) Ideal case
2) Doppler phase shift effects
3) Hardware bias effects

3.3.1 Ideal Case

In the ideal case simulation, no Doppler phase shift, noise, and signal distortion by hardware imperfection were included.

The TargetsModel was first bypassed, so the transmitted signal was sent directly into the IQ demodulators. Figure 15 shows the first two compressed pulses right after the digital processors in the in-phase channel for signals coded by codes A and B, respectively. The first compressed pulse was delayed by $2\mu s$ and added with the second compressed pulse. The summed signal went through envelope detection. The result is displayed in Figure 16, which shows that the sidelobes had been completely canceled.

![Figure 15: Compressed Pulses](image)
The TargetsModel was then included. The two targets were assumed to be $R_1 = 2001m$ and $R_2 = 2022m$ away from the transmitter, which were converted in the target model to equivalent two-way propagation delays of

$$DLY_1 = \frac{R_1}{c} = \frac{2001m}{(3 \times 10^8 m/s)} = 6.67 \mu s$$

$$DLY_2 = \frac{R_2}{c} = \frac{2022m}{(3 \times 10^8 m/s)} = 6.74 \mu s. \quad (10)$$

The radar cross sections of the two targets were set to $\sigma_1 = 1$ and $\sigma_2 = 0.5$, so the returned signals from the two targets were scaled in the target model by

$$SCL_1 = \sqrt{\frac{\sigma_1}{(4\pi R_1^2)^2}} \quad (11)$$

$$SCL_2 = \sqrt{\frac{\sigma_2}{(4\pi R_2^2)^2}} \quad (12)$$

Figure 17 shows the first two compressed pulses right after the digital processors in the in-phase channel for signals coded by codes A and B, respectively. In the compressed
pulses two main lobes were observed, with the higher one corresponding to the first target and the lower one to the second target. The first compressed pulse was again delayed by $2 \mu s$ and added with the second compressed pulse. The summed signal went through envelope detection. The result is displayed in Figure 18, which shows the sidelobes cancellation.

![Figure 17: Compressed Pulses—Two Targets](image1)

![Figure 18: Sidelobe Cancellation—Two Targets](image2)

The simulation results for the ideal case verified the principle of complementary-coded pulse radar and assured that the model was working properly.
3.3.2 Doppler Phase Shift Effect

As discussed in section 3.1, since the complementary-coded pulse radar is to be onboard a moving vehicle and the pulses coded by complementary codes are to be transmitted alternatively, the effects of Doppler phase shift on pulse compression and sidelobes cancellation have to be studied. Simulations were run by introducing phase shift of $0.5^\circ$, $1^\circ$, $5^\circ$, $10^\circ$, $15^\circ$, $20^\circ$, and $25^\circ$, respectively, in the IQ demodulation for code B. It was expected that the main lobe peak would decrease and the sidelobes would appear when phase shift was included. Figure 19 illustrates the sidelobes produced by $0.5^\circ$ phase shift. Figure 20 shows the main lobe magnitude of the compressed pulse drop a little as phase shift increases and the variation is nonlinear. Figure 21 shows the maximum lobe magnitude of the compressed pulse increase linearly as phase shift increases. The PSL of compressed pulse was calculated according to equation (5) for different phase shifts, and the results are listed in Table 2 and plotted in Figure 22. The maximum sidelobe is about 60 dB down the main lobe for a phase shift of $0.5^\circ$ and the sidelobe level climbs to about 26 dB for a phase shift of $25^\circ$.

![Figure 19: Sidelobes with 0.5° Phase Shift](image-url)
Figure 20: Main Lobe Magnitude Versus Phase Shift

Figure 21: Maximum Sidelobe Magnitude Versus Phase Shift
Table 2: PSL Values for Different Phase Shift

<table>
<thead>
<tr>
<th>Phase Shift (degree)</th>
<th>0.5</th>
<th>1</th>
<th>5</th>
<th>10</th>
<th>15</th>
<th>20</th>
<th>25</th>
</tr>
</thead>
<tbody>
<tr>
<td>PSL (dB)</td>
<td>-60.41</td>
<td>-54.38</td>
<td>-40.40</td>
<td>-34.36</td>
<td>-30.81</td>
<td>-28.27</td>
<td>-26.29</td>
</tr>
</tbody>
</table>

Figure 22: PSL Versus Phase Shift

The question now is what the phase shift is in the real application. As illustrated in Figure 23, suppose the vehicle carrying the radar is flying 500m over the target, its speed is \( V = 150m/s \), and it is at position 1 when it transmits the first coded pulse. The distance from the target is \( R_1 \) at position 1. It will move to position 2 when transmitting the second coded pulse phase 0.1ms later. The distance from the target is \( R_2 \) at position 2. The phase shift is calculated as follows:

\[
d = VT = 150m/s \times 0.1ms = 0.015m
\]

\[
R_2 = \sqrt{R_1^2 + d^2} = R_1 \sqrt{1 + \left( \frac{d}{R_1} \right)^2} \approx R_1 + \frac{d^2}{2R_1}
\]

\[
\delta R = R_2 - R_1 = \frac{d^2}{2R_1}
\]
\begin{align*}
\Delta \phi &= \frac{2\pi f_c (2\delta R)}{c} = \frac{2\pi f_c d^2}{cR_1} = \frac{2\pi (150 \times 10^6)(0.015)^2}{3 \times 10^8 \times 500} = 1.41 \times 10^{-6} \text{ rad} = 0.000081^\circ \\
\end{align*}

According to Figure 22, this amount of Doppler phase shift is very small and the sidelobes caused by it would be much less than –60 dB. Therefore in the real application Doppler phase shift is not a concern for the complementary-coded pulse radar.

3.3.3 Amplitude Imbalance Effect

Any hardware imperfection in the radar system would result in signal distortion. It has been found in laboratory experiments that the amplitude imbalances between positive and negative voltages were the main contributor to sidelobes. For instance, Figure 24 is the waveform of code A generated using the Polynomial Waveform Synthesizer of Model 2045 by Analogic Data Precision. The device is supposed to generate 10 volts for bits of 1, and –10 volts for bits of 0 in the code. But the actual measurements turned out to be
9.606 volts on average for bits of 1, and –9.570 on average for bits of 0. If 9.57 is chosen as the reference voltage, the amplitude imbalance is calculated as $\frac{9.606 - 9.570}{9.570} = 0.38\%$.

Figure 24: Amplitude Imbalances in Hardware

Figure 25: Amplitude Imbalance Model
The amplitude imbalance effect of the waveform generator can be modeled by thinking of the imbalances between positive and negative voltages as a constant bias in the device, as shown in Figure 25. For instance, to simulate the effect of 1% bias, 1.01 volts is generated instead of 1 volt in the Square Wave Generator for bits of 1, –1 volt is generated for bit of 0 as the ideal situation. Simulations were performed with different bias levels (0.1%, 0.5%, 1%, 2%, 3%, 4%, 5%, 10%, 20%), and their effects on PSL are listed in Table 3 and are shown in Figure 26. The PSL increases non-linearly as the bias increases. The PSL is –80.09 dB for 0.1% bias and increases to –38.79 dB for 20% bias.

<table>
<thead>
<tr>
<th>Amplitude Imbalance (%)</th>
<th>0.1</th>
<th>0.5</th>
<th>1</th>
<th>2</th>
<th>3</th>
</tr>
</thead>
<tbody>
<tr>
<td>PSL (dB)</td>
<td>–84.09</td>
<td>–70.12</td>
<td>–64.12</td>
<td>–58.14</td>
<td>–54.65</td>
</tr>
<tr>
<td>Amplitude Imbalance (%)</td>
<td>4</td>
<td>5</td>
<td>10</td>
<td>20</td>
<td></td>
</tr>
<tr>
<td>PSL (dB)</td>
<td>–52.19</td>
<td>–50.29</td>
<td>–44.44</td>
<td>–38.79</td>
<td></td>
</tr>
</tbody>
</table>

Figure 26: Amplitude Imbalance Effects on PSL
Chapter 4: Laboratory Experimentation

4.1 Experiment Setup and Procedures

In addition to the system simulations performed in Chapter 3, physical experiments were performed in the laboratory to study the performance of sidelobe cancellation with complementary coded pulses.

There are three devices available in RSL that can be used to generate complementary-coded pulses. These devices are as follows:

1) The Model 2045 Arbitrary Waveform Synthesizer from Analogic Data Precision. This device has a maximum clock rate of 800 MHz, an 8-bit data acquisition card (DAC), and 512 K memory. Its maximum output level is 10 V p-p with bandwidth of 200 MHz. Figure 27 shows the front panel of the waveform synthesizer, which includes all of the keys required to program the device. Reference [17] describes its specifications in detail and presents examples to illustrate the methods to generate desired waveforms.

2) The DBS 2050 VXI Waveform Generator from Analogic Corporation. This device has a maximum sampling rate of 2 GS/s with 8-bit resolution, its bandwidth is greater than 850 MHz, output amplitude is 0.5 V p-p, and memory is 8 M. It has two channels and can be operated as two 1-GS/s generators. Figure 28 shows the front panel

![Figure 27: Model 2045 Arbitrary Waveform Synthesizer](image188x210.png)

![Figure 28: DBS 2050 VXI Waveform Generator](image184x455.png)

3) AWG Card for Coherent Radar Depth Sounder developed by RSL. This device is designed to produce signals with a maximum bandwidth of 60 MHz and maximum
frequency of 400 MHz. Its highest sampling frequency is 150 MHz, and the ADC resolution is 14 bits. Figure 29 shows the front panel of the waveform generator. Reference [19] discusses its design in detail.

![Waveform Generator for Coherent Radar Depth Sounder](image)

Figure 29: Waveform Generator for Coherent Radar Depth Sounder

All three devices were used to generate complementary-coded pulses, and their performances were evaluated and compared. Other devices and components used in experimentation include the following:

1) A digital oscilloscope of 8-bit resolution, used to record and display experimental results
2) A signal generator, used to produce LO signal of 150 MHz for modulation and demodulation

3) Two mixers ZFM-3 by Minicircuits, used for modulation and demodulation

4) A delay line of 50 ns

![Experiment Setup Diagram]

Figure 30: Experiment Setup

The complete experiment setup is shown in Figure 30. The procedures in the experimentation are as follows:

**Step 1:** Pulses coded by codes A and B were directly programmed into the Model 2045 Arbitrary Waveform Synthesizer if this device was used, or were prepared in the proper format with MATLAB codes and C++ codes and loaded in the DBS 2050 VXI Waveform Generator or the AWG Card for Coherent Radar Depth Sounder if these devices were used. The duration of each coded pulse was $2 \mu s$.

**Step 2:** The output of the AWG device was first sent directly to the oscilloscope to see what the sidelobe level would be. Since the AWG device is to be the signal source in complementary-coded pulse radar, its performance will determine the lowest sidelobe level one can expect from the radar.
**Step 3:** The output of the AWG device was modulated on a carrier of 150 MHz from the signal generator, then passed the signal through the delay line of 50 ns, and demodulated with the same carrier. The demodulated signal was sent to the digital oscilloscope. The purpose of this step was to see the effects of other stages of the radar system on sidelobe levels.

**Step 4:** Repeat Step 2 using the two channels of the DBS 2050 VXI Waveform Generator. One channel was used to generate the pulse coded by code A, and the other channel was used to generate the pulse coded by code B. The purpose of this step was to study the effect of channel imbalances.

**Step 5:** All data were recorded by the oscilloscope and were processed in ADS to compress the pulses and calculate the PSL.

### 4.2 Experiment Results

#### 4.2.1 AWG Output

The parameter settings and results obtained in Step 2 for the three waveform generation devices are summarized in the following table:

<table>
<thead>
<tr>
<th></th>
<th>Model 2045</th>
<th>DBS 2050 VXI</th>
<th>AWG Card</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Sampling Frequency</strong></td>
<td>500 MHz</td>
<td>2 GHz</td>
<td>120 MHz</td>
</tr>
<tr>
<td><strong>Output Amplitude</strong></td>
<td>–10 V to 10 V</td>
<td>–0.5 V to 0.5 V</td>
<td>–2 V to 2 V</td>
</tr>
<tr>
<td><strong>PSL</strong></td>
<td>–72.56 dB</td>
<td>–64.89 dB</td>
<td>–59.71 dB</td>
</tr>
</tbody>
</table>
The PSL value turns out to be $-72.56\, \text{dB}$ from the outputs of the Model 2045 Arbitrary Waveform Synthesizer. The sampling frequency and the output amplitude were set to 500 MHz and $-10\, \text{V}$ to $10\, \text{V}$ in the experiment. Figure 31 shows the output waveforms recorded from the Model 2045 Arbitrary Waveform Synthesizer. Figure 32 shows the corresponding compressed waveforms. Figure 33 shows the dB levels of the main lobe and sidelobes when the two compressed pulses were summed together. The main lobe level is $111.6\,\text{dB}$ and the maximum sidelobe level is $39.04\,\text{dB}$.

![Figure 31: Output Waveforms from Model 2045](image)

The PSL value is $-64.89\, \text{dB}$ from the outputs of the DBS 2050 VXI Waveform Generator. The sampling frequency and the output amplitude were set to 2 GHz and $-0.5\, \text{V}$ and $0.5\, \text{V}$ in the experiment.
The PSL value is –59.71 dB from the outputs of the AWG card. The sampling frequency and the output amplitude were 120 MHz and –2 V and 2 V in the experiment.

Figure 32: Compressed Waveforms from Model 2045
4.2.2 Modulation and Demodulation Effects

Figure 34, Figure 35, and Figure 36 are the experimental results from Step 3 with the DBS 2050 VXI Waveform Generator. Compared to Figure 31, Figure 34 shows that the high-frequency components introduced in the modulation and demodulation were superposed onto the waveforms. Figure 35 shows the corresponding compressed waveforms. Figure 36 shows the dB levels of the main lobe and sidelobes when the two compressed pulses were summed together. The main lobe level is 39.49 dB, and the maximum sidelobe level is 0.04 dB. Therefore the PSL is −39.45 dB, which is about 25 dB higher compared to the value obtained directly from the waveform generator’s output.
Rather than the high-frequency components, it is concluded after analysis that the PSL rise is caused by the amplitude imbalances introduced in the modulation and demodulation. Figure 37 illustrates the amplitude imbalances by overlapping the two waveforms in Figure 34. The positive amplitude of the waveform for code B is obviously bigger than that for code A. This conclusion can be further verified by including the amplitude imbalances into the simulation model described in Chapter 3. For the waveform for code A, the averaged positive amplitude is calculated to be 0.130 V, and the averaged negative amplitude is –0.157 V. The averaged positive and negative amplitudes for code B are calculated to be 0.148 V and –0.153 V, respectively. When these amplitudes are entered in the simulation model, the PSL turns out to be
37.18 dB. The difference between the simulation and the measurements is only 2.27 dB. It is also observed that the amplitude imbalances are present not only between the two waveforms for codes A and B but also between the positive and negative amplitudes for the same waveform.

Figure 35: Compressed Waveforms after Modulation and Demodulation
Figure 36: Main and Sidelobe Levels after Modulation and Demodulation

Figure 37: Amplitude Imbalances after Modulation and Demodulation
4.2.3 Channel Imbalance Effect

Figure 38, Figure 39, and Figure 40 are the experimental results from Step 4 with the DBS 2050 VXI Waveform Generator. Figure 38 shows the output waveforms recorded from the two channels of the DBS 2050 VXI Waveform Generator. Figure 39 shows the corresponding compressed waveforms. Figure 40 shows the dB levels of the main lobe and sidelobes when the two compressed pulses were summed together. The main lobe level is 60.3 dB, and the maximum sidelobe level is 20.51 dB. Therefore, the PSL is –38.79 dB, which is about 26 dB higher than the result of a single channel.

![Figure 38: Waveforms from Two Channels](image)

Again, the amplitude imbalances between two channels are the reason for the PSL rise. Figure 41 illustrates the amplitude imbalances by overlapping the two waveforms in Figure 38. In this case, the negative amplitude of the waveform for code B is obviously
bigger than that for code A. For the waveform for code A, the averaged positive amplitude is calculated to be 0.489 V, and the averaged negative amplitude is –0.478 V. And the averaged positive and negative amplitudes for code B are calculated to be 0.500 V and –0.528 V, respectively. When these amplitudes are entered in the simulation model, the PSL turns out to be –37.82 dB. The difference between the simulation and the measurements is only 0.92 dB.

Figure 39: Compressed Waveforms from Two Channels
Figure 40: Main and Sidelobe Levels for Two Channels

Figure 41: Amplitude Imbalances between Two Channels
Chapter 5: Prototype System Design

A prototype of complementary-coded radar is proposed in this chapter. The subsystems and system operation are described first in section 5.1. The transmitter and receiver architectures are discussed in detail in section 5.2 and section 5.3, respectively.

5.1 System Overview

As shown in Figure 42, the architecture of the proposed radar prototype is similar to that of the Advanced Coherent Radar Depth Sounder (ACORDS) [20], consisting of a digital system, transmitter, receiver, antennas, and clock synchronization circuit.

![Figure 42: System Block Diagram of Prototype](image)

The digital system includes a rack-mount host computer, an AWG card, a data acquisition card (DAC), and the trigger circuit. The host computer operates the radar using the preinstalled software and stores the data from the DAC. The AWG card, to be
constructed with field programmable gate arrays, generates complementary binary phase-coded waveforms for the transmitter. The DAC digitizes and integrates the returned signals collected by the receiver. The trigger circuit generates three control signals, including transmitter blank, receiver blank, and attenuator control.

The transmitter is to filter the complementary binary phase-coded waveforms, to amplify them to the proper level, and to feed the signals into the Tx antennas for transmitting. The transmitter design is discussed in detail in section 5.2. The receiver is to amplify and filter the returned signals from targets and send them to the DAC to be digitized and integrated. The receiver design is discussed in detail in section 5.3. The antennas are to employ a half-wave-length dipole array, which has been widely used for previous measurements over the polar ice sheets.

The synchronization circuit generates two clock signals: (1) a base clock of 10 MHz as the reference of the whole system; (2) a sampling clock of 120 MHz for both the AWG card and the DAC. In the synchronization circuit, a 10-MHz stable oscillator (STALO) is to be used to generate the 10-MHz base clock signal, and a 120 MHz frequency synthesizer is to be phase-locked to the STALO to generate the 120 MHz sampling frequency.

The system parameters of the prototype are shown in Table 5. The operation of the prototype is illustrated in the time diagram of Figure 43.
Table 5: System Parameters of Prototype

<table>
<thead>
<tr>
<th>Description</th>
<th>Characteristic</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radar Type</td>
<td>Pulse Compression</td>
<td>—</td>
</tr>
<tr>
<td>Waveform</td>
<td>Complementary binary phase coding</td>
<td>—</td>
</tr>
<tr>
<td>Code Length</td>
<td>32</td>
<td>bit</td>
</tr>
<tr>
<td>RF Carrier Frequency</td>
<td>150</td>
<td>MHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>32</td>
<td>MHz</td>
</tr>
<tr>
<td>Transmitted Pulse Width</td>
<td>2</td>
<td>usec</td>
</tr>
<tr>
<td>Compressed Pulse Width</td>
<td>62.5</td>
<td>nsec</td>
</tr>
<tr>
<td>Pulse Repetition Frequency</td>
<td>2-10</td>
<td>kHz</td>
</tr>
<tr>
<td>Range Sidelobes</td>
<td>&lt;60</td>
<td>dB</td>
</tr>
<tr>
<td>Peak Transmitting Power</td>
<td>200</td>
<td>W</td>
</tr>
<tr>
<td>A/D Dynamic Range</td>
<td>12-bit, 72</td>
<td>dB</td>
</tr>
<tr>
<td>Receiver Dynamic Range</td>
<td>&gt;110</td>
<td>dB</td>
</tr>
<tr>
<td>Range Resolution</td>
<td>~3.82</td>
<td>m</td>
</tr>
<tr>
<td>Antennas</td>
<td>$\lambda$ / 2 dipole array</td>
<td>—</td>
</tr>
</tbody>
</table>

Figure 43: Timing Diagram of Prototype Operation
The trigger circuits generate the pulse repetition frequency (PRF), the transmitter and the receiver blanking signals with reference to the base clock. The AWG synthesizes a binary phase-coded complementary pulse of 2 µs each time as triggered by the PRF. When the transmitter is on, the receiver is off to avoid damage that may be caused by the leakage signal from the transmitter. When the transmitter is off, the receiver is set to On to receive the echoes. The process repeats after each pulse repetition period as long as the radar is working.

5.2 Transmitter

The function of the transmitter is to amplify waveforms to the desired peak power level, which is 200 W (53 dBm), as shown in Table 5. Figure 44 shows the block diagram of the transmitter.

![Figure 44: Block Diagram of Prototype Transmitter](image)
Figure 45: Complementary Binary Phase-Coded Signals

The complementary codes A and B are binary phase-coded with a sinusoid signal of 30 Hz, as shown in Figure 45. The power spectrum of the coded signal with code A is shown in Figure 46. The main lobe, centered at 30 MHz, extends from 14 MHz to 46 MHz, corresponding to a bandwidth of 32 MHz. The base-band-coded signals are digitized and loaded in the AWG. A digital-to-analog (D/A) converter inside the AWG reads the loaded signals at a frequency of 120 MHz to generate the desired waveforms. The attenuator after the AWG serves to reduce reflections and adjust the power level to prevent saturation. A bandpass filter (BPF) with good stopband attenuation is then inserted to up-convert the base band signal to 150 MHz, which is the operation frequency of the prototype. The passband of the filter is from 134 MHz to 166 MHz. The frequency up conversion is illustrated in Figure 47, in which the blue plot is the power spectrum of
Figure 46: Power Spectrum of Binary Phase-Coded Signal

Figure 47: Frequency Up Conversion
the output of the AWG, and the pink, thicker plot is the power spectrum of the signal after the BPF (1). The up-converted signals are preamplified by the first amplifier PreAmp. The Tx Blank switch turns the transmitter on or off according to the control signal from the digital system. The second bandpass filter, which has the same specifications as the first, filters out any out-of-band harmonics, if present. The second attenuator after BPF (2) has the same function as the first attenuator. The filtered signals are finally amplified to 200 W by the power amplifier and fed into antennas to be transmitted.

5.3 Receiver

The function of the receiver is to amplify the weak returns from the ice layers near the ice bed to the level of 4 dBm, which is the maximum signal level that can be digitized by the DAC. Figure 48 shows the block diagram of the receiver.

![Figure 48: Block Diagram of Prototype Receiver](image-url)
The antenna array receives the echo signals from the ice layers. Connected with the antenna array is the receiver front end, which consists of a BPF, limiter, switch, and low noise amplifier (LNA). Because the noise figure of the front end would dominate the noise figure of the receiver, components with low noise figure are to be used in the receiver front end to reduce the receiver noise. The BPF (1) filters the out-of-band signals in echoes, and the limiter following the BPF protects the receiver from strong echoes. The Rx Blank switch is used to turn off the receiver as the radar is transmitting. The LNA (1) has a high gain and a high 1-dB compression point. A similar LNA is used at the second amplifier stage. The BPF (2) is used to filter harmonics after the receiver front end. The attenuator placed before the second LNA is a digital attenuator of 6-bit, 63 dB total attenuation in steps of 1 dB. A third BPF, which is the same as BPF (2), is used to filter harmonics after LNA (2) before sending the signals to DAC to be processed and stored.
Chapter 6: Summary and Recommendations

6.1 Summary

The research achievements in the thesis are summarized as follows:

1) A simulation model was built for complementary-coded pulse radar. The model includes waveform generator, IQ modulators, multitargets model, IQ demodulators, digital waveform compressors, and envelope detector.

2) Sidelobe cancellation was verified using the simulation model for ideal case. Doppler phase shift and amplitude imbalances were introduced in the model to evaluate their effects on sidelobe level. The magnitude of the maximum sidelobe peak was found to increase linearly as Doppler phase shift increases. The PSL is below –60 dB for a 0.5° phase shift. For complementary-coded radar to be carried on an airplane for ice measurement, Doppler phase shift is not a concern. The PSL was found to increase nonlinearly as amplitude imbalances increase. The PSL is below –60 dB for a 1% amplitude imbalance.

3) Experiments were performed in the laboratory to understand the effects of hardware imperfections on sidelobe cancellation. The waveform amplitude imbalances caused by waveform generator, modulator, and demodulator were found to be the main obstacle to sidelobe cancellation. The same channel should be used for complementary codes to avoid the effect of amplitude imbalances between channels. The experimental results were verified by simulations with the
averaged amplitude imbalance of the measurements included in the simulation model.

4) A prototype of complementary-coded pulse radar was proposed. The architectures of the prototype transmitter and receiver were discussed.

6.2 Recommendations

All the simulation and experimental results presented in this thesis are for base band signals. Since binary phase-coded signals of 30 MHz are to be transmitted as proposed for the prototype, the simulation model should be modified accordingly, and down conversion blocks should be inserted between demodulators and the digital compressors. Using the AWG card for Coherent Radar Depth Sounder, the best PSL is about –40 dB based on experiments on binary phase-coded signals of 30 MHz, which is about 20 dB higher compared to the value for base band signals. The reason for this big difference needs to be further analyzed to determine whether it is from the card distortion or the down conversion distortion in data processing or from both. Based on simulation and experimental experience, the challenge and key point for a successful complementary-coded pulse radar is to reduce signal distortions caused by hardware or find a way to compensate for distortions in signal processing. As to the method of signal processing, there are two options. The first is to use a calibration target to find the corrections and then apply the corrections to other targets to see if it works. The second is to determine distortions stage by stage in the radar system. No matter what method is to be used, a prototype needs to be built for either experimentation in the laboratory or in the field.
References


University of Kansas, 2003.


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