Evaluation and Application of LTE, DVB, and DAB Signals of Opportunity for Passive Bistatic SAR Imaging

Aaron S. Evers
Wright State University

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Evaluation and Application of LTE, DVB, and DAB Signals of Opportunity for Passive Bistatic SAR Imaging

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in Engineering

by

Aaron S. Evers
B.S.E.E., Wright State University, 2013

2014
Wright State University
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Julie Jackson, Ph.D.
Thesis Co-Director

Brian Rigling, Ph.D.
Thesis Co-Director

Kefu Xue, Ph.D.
Department Chair of Electrical Engineering

Julie Jackson, Ph.D.

Brian Rigling, Ph.D.

Zhiqiang Wu, Ph.D.

Robert E.W. Fyffe, Ph.D.
Vice President for Research and Dean of the Graduate School
ABSTRACT


Due to the many advantages of passive radar and ubiquity of commercial broadcast transmitters, interest in passive bistatic radar (PBR) applications has continued to grow. More specifically, sources studying commercial orthogonal frequency division multiplexing (OFDM) waveforms for passive bistatic synthetic aperture radar (SAR) imaging have become more common. This work evaluates and applies long term evolution (LTE), digital video broadcast (DVB), and digital audio broadcast (DAB) signals of opportunity for passive bistatic SAR imaging.

First, implications of the structure and properties of each of the signal of opportunity’s transmitted waveform are characterized by examining the waveform’s self- and cross-ambiguity functions (AFs). In addition to deriving waveform properties, link budget analysis is completed using pessimistic values intrinsic to LTE, DVB, and DAB transmissions for predicting performance of potential passive bistatic SAR imaging scenarios. Small-scale, passive bistatic SAR imaging experiments are carried out using signals structured similarly to LTE, DVB, and DAB signals, demonstrating the merits of the considered processing schemes for passive bistatic SAR image generation.
# Contents

1 Introduction  
   1.1 Problem Description .............................................. 1  
   1.2 Research Goals and Methodology ................................. 3  
   1.3 Thesis Organization .............................................. 3  

2 Background  
   2.1 Previous Work ................................................... 5  
      2.1.1 Ambiguity Analysis ........................................... 6  
      2.1.2 Link Budget Analysis ........................................ 7  
      2.1.3 Passive Bistatic Synthetic Aperture Radar Imaging .......... 8  
   2.2 Passive Bistatic Radar ........................................... 8  
   2.3 Orthogonal Frequency Division Multiplexing ................... 14  
   2.4 Long Term Evolution ............................................ 17  
      2.4.1 Primary and Secondary Synchronization Signals .......... 19  
      2.4.2 Cell-Specific Reference Elements ......................... 21  
   2.5 Digital Video Broadcast ........................................ 21  
      2.5.1 Scattered Pilots ............................................. 23  
      2.5.2 Continual Pilots ........................................... 25  
   2.6 Digital Audio Broadcast ....................................... 26  
      2.6.1 Synchronization Channel .................................... 28
2.7 Synthetic Aperture Radar Imaging ................................................. 28
  2.7.1 Phase History Model for Orthogonal Frequency Division Multiplexing Signals ................................................. 30
2.8 Chapter Conclusion ................................................................. 33

3 Ambiguity Analysis ................................................................. 36
  3.1 Ambiguity Function ............................................................... 36
  3.2 Frequency Division Duplexing Long Term Evolution Downlink ....... 38
    3.2.1 Self-Ambiguity Function ................................................. 38
    3.2.2 Cell-Specific Reference Element Cross-Ambiguity Function .... 40
    3.2.3 Synchronization Signal Cross-Ambiguity Function ............... 42
  3.3 8k Digital Video Broadcast ................................................... 44
    3.3.1 Self-Ambiguity Function ................................................. 44
    3.3.2 Scattered Pilot Cross-Ambiguity Function ......................... 47
    3.3.3 Continual Pilot Cross-Ambiguity Function ......................... 49
  3.4 Transmission Mode I Digital Audio Broadcast ......................... 51
    3.4.1 Self-Ambiguity Function ................................................. 52
    3.4.2 Synchronization Signal Cross-Ambiguity Function ............... 54
  3.5 Geometry and Signal Parameter Considerations ....................... 55
  3.6 Application to Passive Bistatic Synthetic Aperture Radar Imaging .... 57
  3.7 Additional Results .................................................................. 60
    3.7.1 Frequency Division Duplexing Long Term Evolution Downlink ... 60
    3.7.2 Frequency Division Duplexing Long Term Evolution Downlink Synchronization Signal ................................. 61
    3.7.3 8k Digital Video Broadcast ................................................. 62
    3.7.4 8k Digital Video Broadcast Scattered and Continual Pilots ...... 62
    3.7.5 Transmission Mode I Digital Audio Broadcast ..................... 63
3.7.6 Transmission Mode I Digital Audio Broadcast Synchronization Symbol ..................................................... 64

3.8 Chapter Conclusion ................................................................. 66

4 Link Budget Analysis ................................................................. 67

4.1 Performance Prediction for Passive Bistatic Synthetic Aperture Radar Imaging ................................................. 67

4.1.1 Transmitter ......................................................................... 68

4.1.2 Receiver ............................................................................. 69

4.1.3 Environment ...................................................................... 71

4.1.4 Bistatic Radar Cross-Section .................................................. 71

4.2 Proposed Real-World Scenario .................................................. 73

4.3 Application to Passive Bistatic Synthetic Aperture Radar Imaging ................................................................. 76

4.4 Additional Results ................................................................. 78

4.4.1 Frequency Division Duplexing Long Term Evolution Downlink ............................................................. 78

4.4.2 Transmission Mode I Digital Audio Broadcast ................................................................. 78

4.5 Chapter Conclusion ................................................................. 80

5 Passive Synthetic Aperture Radar Imaging Results .................................................................................. 82

5.1 Experimental Setup ................................................................. 82

5.1.1 Hardware Setup .................................................................. 83

5.1.2 Software Setup .................................................................. 84

5.1.3 Scene Setup ........................................................................ 87

5.2 Experimental Results .............................................................. 88

5.2.1 Demonstration of “Actual” Resolution Properties ........................................................................ 90

5.3 Chapter Conclusion ................................................................. 92

6 Conclusion .............................................................................................. 93
## List of Figures

2.1 Potential PBR scenarios. ........................................ 9  
2.2 North-referenced coordinate system. ............................. 10  
2.3 Transformation of an isorange contour. .......................... 11  
2.4 Oval of Cassini versus an isorange contour. ...................... 13  
2.5 Assembly of an OFDM waveform. ................................. 15  
2.6 CP extension of an OFDM symbol. ................................. 16  
2.7 FDD LTE\textsubscript{e} DL signal structure. ...................... 20  
2.8 Time-frequency grid of a CSRE signal. ........................... 22  
2.9 8k DVB signal structure. ......................................... 24  
2.10 Time-frequency grid of a scattered pilot signal. ............... 25  
2.11 TxI DAB signal structure. ....................................... 27  
2.12 Basic spotlight mode SAR imaging scenario. .................... 30  
2.13 Timing and implementation of the AMF. .......................... 34  

3.1 SAF of an FDD LTE\textsubscript{e} DL signal. ......................... 39  
3.2 CAF of an FDD LTE\textsubscript{e} DL CSRE signal. .................. 41  
3.3 Top down view of the FDD LTE\textsubscript{e} DL-CSRE CAF result. .... 42  
3.4 CAF of an FDD LTE\textsubscript{e} DL signal and its PSS and SSS. .... 43  
3.5 SAF of an 8k DVB signal. ....................................... 46  
3.6 Top down view of the 8k DVB signal’s SAF. ...................... 46
3.7 CAF of an 8k DVB scattered pilot signal. .......................... 48
3.8 Top down view of the 8k DVB-scattered pilot CAF result. ........ 49
3.9 CAF of an 8k DVB continual pilot signal. .......................... 50
3.10 Top down view of the 8k DVB-continual pilot CAF result. ........ 51
3.11 SAF of a TxI DAB signal. ............................................ 53
3.12 Top down view of the TxI DAB signal’s SAF. ....................... 53
3.13 CAF of a TxI DAB SS signal. ......................................... 55
3.14 Locations of bistatic peaks inferred from the FDD LTE DL-CSRE CAF monostatic result. ................................................. 56
3.15 The first additional peak arising from the structure of an FDD LTE DL-CSRE signal. ................................................. 58
3.16 Range profile generated at the center pulse of the receive aperture with 
\[ s_{\text{ref}}[n] = s[n]. \] .................................................. 61
3.17 Range profile generated by partially matched filtering with an FDD LTE DL synchronization signal \( (s_{\text{ref}}[n] = s_{\text{PSS}}[n] + s_{\text{SSS}}[n]). \) ................................. 62
3.18 Range profile generated at the center pulse of the receive aperture with 
\[ s_{\text{ref}}[n] = s[n]. \] .................................................. 63
3.19 Range profile generated at the center pulse of the receive aperture with 
\[ s_{\text{ref}}[n] = s_{\text{SP}}[n] + s_{\text{CP}}[n]. \] .................................. 64
3.20 Range profile generated at the center pulse of the receive aperture with 
\[ s_{\text{ref}}[n] = s[n]. \] .................................................. 65
3.21 Range profile generated at the center pulse of the receive aperture with 
\[ s_{\text{ref}}[n] = s_{\text{SS}}[n]. \] ................................................ 65
4.1 Trade-off between resolution and bistatic RCS. ....................... 72
4.2 Proposed scenario for passive bistatic SAR imaging. ................ 74
4.3 Performance regions for an over-the-shoulder scenario. ............ 75
4.4 Demonstration of SAR image quality produced by matched filtering with an 8k DVB signal ($s_{\text{ref}}[n] = s[n]$). .................................................... 77

4.5 Demonstration of SAR image quality produced by matched filtering with an FDD LTEe DL signal ($s_{\text{ref}}[n] = s[n]$). .................................................... 79

4.6 Demonstration of SAR image quality produced by matched filtering with a TxI DAB signal ($s_{\text{ref}}[n] = s[n]$). .................................................... 81

5.1 Test setup block diagram. ............................................................... 84

5.2 OFDM phase history processing model. ............................................. 86

5.3 Test setup used for experimentation. ................................................... 87

5.4 Experimental SAR images produced using a full AMF with a segment duration of $T_{\text{sym}}$. ................................................................. 88

5.5 Experimental SAR images produced using a partial AMF with a segment duration of $T_{\text{sym}}$. ................................................................. 89

5.6 Experimental SAR image of a target scaled to illustrate the “actual” down-range and crossrange resolution inherent to the respective signal. .......... 91
# List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Bistatic radar range equation parameters and definitions.</td>
<td>12</td>
</tr>
<tr>
<td>2.2</td>
<td>Parameters of signals considered for this work.</td>
<td>17</td>
</tr>
<tr>
<td>3.1</td>
<td>Quantitative metrics for one, discrete instantiation of the SAF and CAFs of</td>
<td></td>
</tr>
<tr>
<td></td>
<td>an FDD LTE&lt;sub&gt;e&lt;/sub&gt; DL signal.</td>
<td>38</td>
</tr>
<tr>
<td>3.2</td>
<td>Quantitative metrics for one, discrete instantiation of the SAF and CAFs of</td>
<td></td>
</tr>
<tr>
<td></td>
<td>an 8k DVB signal.</td>
<td>44</td>
</tr>
<tr>
<td>3.3</td>
<td>Quantitative metrics for the SAF and CAFs of a TxI DAB signal.</td>
<td>52</td>
</tr>
<tr>
<td>5.1</td>
<td>Parameters of signals, structured similar to FDD LTE&lt;sub&gt;e&lt;/sub&gt; DL, 8k DVB,</td>
<td></td>
</tr>
<tr>
<td></td>
<td>and TxI DAB signals, utilized for experimental testing.</td>
<td>85</td>
</tr>
<tr>
<td>5.2</td>
<td>Resolution values for scaled target signatures.</td>
<td>91</td>
</tr>
<tr>
<td>5.3</td>
<td>Resolution required for mapping applications.</td>
<td>91</td>
</tr>
<tr>
<td>A.1</td>
<td>Induced Doppler calculations for each respective signal type.</td>
<td>102</td>
</tr>
</tbody>
</table>
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Dedicated to

my wife, family, and friends
Chapter 1

Introduction

1.1 Problem Description

Historically, interest in bistatic radar has come in surges due to advances and developments in technology and rediscovery of specific bistatic applications [1]. In the most recent surge, a bistatic application referred to as passive bistatic radar (PBR) has seen an upsurge in interest. This interest is in large part a result of the improved practicality of PBR applications due to a number of factors, including, but not limited to: 1) increased processing power, 2) more precise equipment, and 3) the availability of signals of opportunity (i.e., commercial terrestrial broadcasters) [2]. Exploitation of commercial broadcast transmitters was first considered in the 1980s by exploiting a television transmitter [2]. A wide variety of radar applications have been considered for exploitation of commercial broadcast transmitters including target detection and tracking [3–7] and synthetic aperture radar (SAR) imaging [8–11]. PBR is referred to by numerous names, including: passive coherent location (PCL), parasitic radar, piggyback radar, or passive covert radar (PCR) [2, 12–14].

The increased deployment of commercial broadcast transmitters has increased the complexity of the global spectrum [15]. This complexity has increased the appeal of PBR
as passive radar systems make use of existing spectrum. Utilization of existing spectrum enables the development of radar systems operating in frequency bands typically not available, such as very high frequency (VHF) and ultra high frequency (UHF) [2]. In addition, non-cooperative illuminators of opportunity (hostile or neutral resources) are oblivious to when their transmitted signal is being exploited. The lack of a transmit waveform allows the radar to remain stealthy within the spectrum [2]. To more efficiently make use of the global spectrum for digital communications, global partnerships (e.g., 3rd generation partnership project) have been formed to assemble well-defined digital communication standards such as long term evolution (LTE) [16]. Consequently, these well-defined digital communication standards also aid in the development of processing schemes seeking to exploit these signals of opportunity.

Although there are factors encouraging the continued development of PBR, it is not without immense challenges. PBR systems demand a large dynamic range, require appropriate processing schemes, and require measures to mitigate direct-path signal interference [12]. Additionally, with PBR, the radar engineer may have no control of the employed transmitted waveform [2]. Signals of opportunity are generally communication waveforms and may employ features or have properties which limit the signal’s utility as a radar waveform (i.e., relatively small bandwidth compared to that typically used for radar applications). These properties include the transmitter’s location, operating frequency, operational schedule, waveform, and radiation pattern. Any one of the transmitter’s properties may hinder the usefulness of an illuminator of opportunity for PBR applications. To characterize the potential of PBR, waveform properties and operational limitations inherent to an illuminator of opportunity must be understood and evaluated.
1.2 Research Goals and Methodology

With the intent of application to passive bistatic SAR imaging, LTE, digital video broadcast (DVB), and digital audio broadcast (DAB) signals’ potential is assessed in this research. The signals’ potential is characterized by examining waveform properties and predicting detection and coverage regions for potential passive bistatic SAR imaging scenarios.

Within this work, the structure and signal parameters provided in the standards [16–18] for LTE, DVB, and DAB transmissions are presented. The implications of signal structure and properties are examined using the ambiguity function (AF) defined in [19]. Then, similar to analysis performed in [2, 13, 20–24], the self- and cross-AFs (SAF and CAF) of each signal are evaluated to demonstrate attainable capabilities and shortcomings ensuing from potential processing schemes utilizing LTE, DVB, and DAB transmissions. Next, adhering to work in [12, 25], link budget analysis for passive bistatic SAR imaging systems exploiting LTE, DVB, and DAB signals is performed using the bistatic radar range equation. Using the performance prediction results, a real-world passive bistatic SAR imaging scenario is proposed. Then, a small-scale setup mimicking the proposed imaging scenario is used with signals structured similarly to LTE, DVB, and DAB signals to experimentally exhibit the merits of the considered processing schemes for passive bistatic SAR imaging.

1.3 Thesis Organization

Chapter 2 begins with a recap of previous work related to this research. The summary of related research is followed by an introduction to the PBR configuration assumed for this work, along with a description of the north-referenced coordinate system typically used to describe bistatic geometry. Derived from the bistatic geometry, isorange contours and ovals of Cassini are introduced for describing bistatic timing and link budgets, respectively.
Next, the structure and signal parameters for the considered LTE, DVB, and DAB signals are presented. Chapter 2 concludes with a basic introduction to SAR imaging, a derivation of the orthogonal frequency division multiplexing (OFDM) phase history model from [11], and a description of techniques employed for generating SAR images shown within this work. In Chapter 3, the implications of LTE, DVB, and DAB signal structure and properties are examined on each signal’s SAF and CAF. A simulated passive bistatic SAR image is then produced and used to demonstrate the effects of a transmitted waveform’s structure and properties on the image. Chapter 4 contains detailed descriptions of components of the bistatic radar range equation as well as proposed parameters for analyzing the link budget of a passive bistatic SAR imaging scenario. Based upon derived performance, a real-world imaging scenario is proposed and analyzed using typical transmitter properties for LTE, DVB, and DAB transmit sites. In Chapter 5, a small-scale experimental setup is employed with signals structured similarly to LTE, DVB, and DAB signals of opportunity to demonstrate considered processing schemes for passive bistatic SAR imaging. Chapter 6 concludes with a summary of the objective and findings of this work and a discussion on possibilities for future work.
Chapter 2

Background

The intent of this chapter is to summarize the findings of previous related research and provide necessary background information for the developments presented in Chapters 3–5. This chapter begins with a brief discussion on previous related works. The summary of related research is followed by a description of the PBR configuration assumed for this work, along with the north-referenced coordinate system typically used for describing bistatic geometry. Then, isorange contours and ovals of Cassini are presented as a means for describing bistatic timing and constant signal to noise ratio (SNR) contours. The review of bistatic geometry is followed by a summary of LTE, DVB, and DAB signal structure and parameters. The chapter concludes with a brief summary of SAR imaging concepts, the OFDM phase history model rederived, and techniques used for SAR image generation within this work.

2.1 Previous Work

In Chapter 1, previous research related to this work were briefly mentioned. Here, the findings of these previous works are elaborated upon and discussed in greater detail.
2.1.1 Ambiguity Analysis

Works such as [13, 20–24] have examined the AF of many signals of opportunity

\[ |\chi(\tau, \nu)|^2 = E \left\{ \frac{1}{E_s} \left| \int_{-\infty}^{+\infty} s(t) s^*_\text{ref}(t + \tau) e^{j2\pi\nu t} dt \right| \right\}^2, \]

(2.1)

where \( E\{\cdot\} \) is the expected value operator, \( s(t) \) is the transmitted signal with energy \( E_s \), \( s^*_\text{ref}(t) \) is the reference signal (matched filter), \( \tau \) is the relative time delay, and \( \nu \) is the relative Doppler shift [19]. For this research, we estimate the SAF and CAF from one discrete time instantiation

\[ |\chi(m, k)|^2 = \left| \frac{1}{E_s} \sum_{n=1}^{N} s[n] s^*_\text{ref}[n + m] e^{j2\pi kn/N} \right|^2, \]

(2.2)

where \( m \) is the relative time delay, \( k \) is the relative Doppler shift, and \( N \) is the number of samples.

The primary method for examining signals of opportunity has been SAF analysis, where the reference signal is the ideal transmitted signal \( (s^*_\text{ref}(t) = s(t)) \) in (2.1). Some CAF analysis has been conducted, where the reference signal is comprised solely of the deterministic feature of interest \( (s^*_\text{ref}(t) = s(t)) \) in (2.1). The objective of the CAF analysis was design of methods to mitigate the effects of deterministic features on the SAF. We also consider the SAF and CAF for waveform characterization. The primary focus of our CAF analysis is for understanding the effects signals’ deterministic features have on the SAF and to provide insight for potential mismatched filter design.

In [13], frequency modulated (FM), global system for mobile communications (GSM), analog UHF television, DVB, and DAB signals are experimentally collected and used to produce their respective SAF. In addition, discussion is provided on the variance of the waveform properties due to the time variance of the signals’ content. In [20], the effects of
the deterministic features of a 2k\textsuperscript{1} DVB signal on its SAF are presented. Using these results, modifications mitigating the deterministic features’ affects are made to the matched filter waveform to produce a more favorable SAF result. The work in [20] is extended further in [21] to an 8k\textsuperscript{1} DVB signal. Analogous research to [13, 20, 21] is presented for worldwide interoperability for microwave access (WiMAX) signals in [22–24]. These considered previous works, [13, 20–24], have successfully derived waveform properties of the aforementioned signals and shown that with an understanding of these properties, these signals may successfully be employed for applications such as target detection and tracking. In Chapter 3, we seek to perform similar analysis, but also extend upon previous results by considering analysis of signals’ deterministic features, all with application toward passive bistatic SAR imaging.

\subsection{2.1.2 Link Budget Analysis}

Relatively few publications have performed link budget analysis for PBR systems. Moreover, little has been published to date examining link budgets for passive bistatic SAR imaging. In [12], detection range and coverage are predicted for a variety of FM, GSM, and DAB transmission sites. The work in [12] includes a detailed discussion on the need for a large dynamic range, the suppression required of the direct-path signal for PBR systems, and justification for typical PBR parameters used to evaluate the bistatic radar range equation. The work in [25] extends the results in [12] further, to predict detection range of a naval ship attainable using a WiMAX signal of opportunity. Both of the considered works, [12, 25], illustrate that exploitation of FM, GSM, DAB, and WiMAX signals is feasible from a link budget perspective for given scenarios. We seek to expand upon these results in Chapter 4 by examining the link budget of LTE, DVB, and DAB signals, specifically for proposed passive bistatic SAR imaging scenarios.

\textsuperscript{1}The operating mode specifies the number of subcarriers comprising a symbol (see Section 2.5 for a description of the available operating modes) [18].
2.1.3 Passive Bistatic Synthetic Aperture Radar Imaging

While publications on passive bistatic SAR imaging are not as prevalent as those on passive target detection and tracking, passive bistatic SAR imaging continues to draw interest and become more common. Most recently, works such as [8–11, 26] have evaluated passive bistatic SAR imaging capabilities through simulation and experimentation. In [8], ultra-wideband (UWB) OFDM signals are utilized for strip-map SAR imaging through simulation and experimentation. Similarly, in [9–11] generic OFDM and WiMAX signals are successfully employed to produce simulated and experimental passive spotlight mode SAR images. The work in [9–11] also includes the derivation of the OFDM phase history model utilized in this work. Moreover, the authors of [9–11] developed an experimental OFDM radar system which is used to collect experimental data for this research. The work in [26] deviates from [8–11] by deriving metrics for emitter selection (multistatic case) for passive SAR imaging.

2.2 Passive Bistatic Radar

As alluded to previously, PBR systems operate using an existing transmitter. The transmitter may be categorized as one of two types, cooperative or non-cooperative [2]. For this research a non-cooperative scenario is assumed, meaning the transmitter is considered to be a hostile or neutral resource and the transmitter’s properties are out of the control of the radar engineer’s hands. Furthermore, this work assumes a scenario consisting of an airborne receiver and a stationary broadcast transmitter as shown in Figure 2.1. PBR naturally lends itself to two processing schemes based on whether a priori knowledge is available concerning the transmit signal at the receive end (see Figure 2.1). Figure 2.1(a) depicts the scenario where no a priori knowledge is available to the receiver about the transmit signal, requiring a direct-path channel to construct a reference waveform for processing. On the other hand, Figure 2.1(b) portrays the case where a priori transmit signal knowl-
(a) PBR scenario assuming *a priori* knowledge, concerning the transmit signal, is not available at the receive end. 

(b) PBR scenario assuming *a priori* knowledge, concerning the transmit signal, is available at the receive end.

Figure 2.1: Different PBR scenarios based on *a priori* knowledge of the transmit signal at the receive end.

edge is available at the receive end, alleviating the need for a direct-path signal to perform processing. As depicted by the scenarios shown in Figure 2.1, bistatic setups consist of a transmitter and receiver which are not collocated. In general, the timing and geometry associated with bistatic setups is more complex than that encountered with monostatic setups [1]. To describe the potentially complex geometry of bistatic setups, a north-referenced coordinate system is used. The coordinate system is defined in the bistatic plane, which is the plane containing the transmitter, receiver, and target [27]. The north-referenced coordinate system is shown in Figure 2.2 and described by the following parameters: $R_T$ is the distance from the transmitter to the target, $\theta_T$ is the transmitter look angle measured positive clockwise from the North axis, $R_R$ is the distance from the receiver to the target, $\theta_R$ is the receiver look angle measured positive clockwise from the North axis, $L$ is the distance between the transmitter and receiver and referred to as the bistatic baseline, $\beta$ is the bistatic angle between the transmitter and receiver, $V$ is the velocity vector of the target, and $\phi$ is the angle between the bistatic bisector and the velocity vector of the target. From Figure 2.2, it is clear the geometry of a bistatic setup reduces to a monostatic setup by letting $L = 0$. The “collapse” of the bistatic description to the monostatic case is suitting as bistatic
geometry is more general and encompasses the monostatic case as its simplest case. For a monostatic setup, the distance traveled to and from a target is simply $2R_R$. However, for bistatic scenarios the transmitter and receiver are not collocated, requiring accountability of the distance of two independent legs of a bistatic triangle. As a result, the total path time delay (transmitter-to-target-to-receiver time delay) for a bistatic setup is

$$\tau = \frac{R_T + R_R}{c},$$

(2.3)

where $c$ is the speed of light. For a monostatic setup ($R_T = R_R$), a target with a total path time delay of $\tau$, lies on a circle with radius $R_R$, centered about the receiver in the bistatic plane. The circle formed centered about the receiver is referred to as an isorange contour. In general, for bistatic geometries, isorange contours no longer form circles. Rather, the circles elongate into ellipses as depicted in Figure 2.3 [1]. The major axis of an isorange contour formed for a bistatic setup is [1]

$$a = \frac{R_T + R_R}{2}.$$  (2.4)
The minor axis of the ellipse is $[1]$

$$b = \sqrt{\left( \frac{R_T + R_R}{2} \right)^2 - \left( \frac{L}{2} \right)^2}.$$

(2.5)

Once again, from Equations (2.4) and (2.5), it is evident the bistatic description for isorange contours reduces to the monostatic result by letting $L = 0$ (the major axis is equivalent to the minor axis). Separation of transmitter and receiver not only alters the timing of a radar system, but separation also alters the operating regions of a transmitter-receiver pair [1]. This alteration is derived from the bistatic radar range equation described in [12] as (assuming the signal and receiver bandwidths are equivalent)

$$\left( R_R R_T \right)^2 = \frac{P_t G_t \sigma_b G_r \lambda^2 L_{sys} T_i}{(4\pi)^3 \text{SNR}_{min} k T_0 F},$$

(2.6)

for the parameters tabulated and defined in Table 2.1.
Table 2.1: Bistatic radar range equation parameters and definitions.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Defines</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_t$</td>
<td>Transmit power</td>
</tr>
<tr>
<td>$G_t$</td>
<td>Transmit antenna gain</td>
</tr>
<tr>
<td>$\sigma_b$</td>
<td>Bistatic radar cross-section (RCS)</td>
</tr>
<tr>
<td>$G_r$</td>
<td>Receive antenna gain</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>Signal wavelength</td>
</tr>
<tr>
<td>$L_{sys}$</td>
<td>System losses (≤ 1)</td>
</tr>
<tr>
<td>$T_i$</td>
<td>Coherent integration time</td>
</tr>
<tr>
<td>$\text{SNR}_{\text{min}}$</td>
<td>Minimum detectable signal to noise ratio</td>
</tr>
<tr>
<td>$k$</td>
<td>Boltzmann’s constant, $1.38 \times 10^{-23}$ J/K</td>
</tr>
<tr>
<td>$T_0$</td>
<td>Noise reference temperature, 290 K</td>
</tr>
<tr>
<td>$F$</td>
<td>Receiver effective noise figure.</td>
</tr>
</tbody>
</table>

As opposed to an isorange contour, which is dependent upon a constant range sum, a constant SNR contour is dependent upon a constant range product (see Equations (2.3) and (2.6)) [1]. Constant SNR contours are conveniently represented using ovals of Cassini with foci corresponding to the transmitter and receiver (see Figure 2.4(a)) [2]. While isorange and constant SNR contours both lie in the bistatic plane (coplanar), the two are not collinear as illustrated in Figure 2.4(b) [1]. To successfully perform passive bistatic SAR imaging, an understanding of bistatic geometry, timing, and link budget must be had. The north-referenced coordinate system, isorange contours, and ovals of Cassini provide the means necessary for describing the geometry, timing, and operational limitations associated with a passive bistatic SAR imaging system. In Chapters 3–5, these definitions are
(a) Oval of Cassini with a bistatic baseline of $L = 15$ km.

(b) Illustration of the lack of collinearity between isorange and constant SNR contours.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_t$</td>
<td>1</td>
<td>kW</td>
</tr>
<tr>
<td>$G_t$</td>
<td>0</td>
<td>dBi</td>
</tr>
<tr>
<td>$\sigma_b$</td>
<td>1</td>
<td>m$^2$</td>
</tr>
<tr>
<td>$G_r$</td>
<td>0</td>
<td>dBi</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>0.41</td>
<td>m</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{sys}$</td>
<td>-5</td>
<td>dB</td>
</tr>
<tr>
<td>$T_i$</td>
<td>3</td>
<td>s</td>
</tr>
<tr>
<td>$\text{SNR}_{\min}$</td>
<td>10</td>
<td>dB</td>
</tr>
<tr>
<td>$T_0$</td>
<td>290</td>
<td>K</td>
</tr>
<tr>
<td>$F$</td>
<td>25</td>
<td>dB</td>
</tr>
</tbody>
</table>

Figure 2.4: Oval of Cassini versus an isorange contour.
used to examine the timing and link budget of bistatic scenarios and better interpret generated SAR images.

2.3 Orthogonal Frequency Division Multiplexing

As mentioned in Chapter 1, to fully exploit a signal of opportunity, the radar engineer must be aware of the transmitted signal’s structure. LTE, DVB, and DAB standards each employ OFDM waveforms. Therefore, in the following, the fundamentals of OFDM waveforms are explained to provide insight into the underlying structure shared by the signals considered in this research.

OFDM is a type of multi-carrier modulation where data are assembled in parallel by modulating a set of \( N \) orthogonal subcarrier frequencies \([28]\). Each subcarrier within the band of \( N \) subcarriers is a multiple of the lowest subcarrier, making them mutually orthogonal. Using an inverse discrete Fourier transform (IDFT) the time domain signal is assembled, creating a superposition of \( N \) modulated subcarriers (see Figure 2.5) \([10]\). The baseband sampling frequency or spectrum bandwidth of an OFDM waveform is determined by the number of subcarriers, \( N \), and the subcarrier separation, \( \Delta f \), as \([28]\)

\[
B = N \Delta f. \tag{2.7}
\]

As will be further evidenced in the proceeding sections, communication signals do not typically occupy the entire allotted spectrum, utilizing only an effective bandwidth\(^2\) of

\[
B_{\text{eff}} = (N_a + 1) \Delta f, \tag{2.8}
\]

\(^2\)The span of \((N_a + 1) \Delta f\) accounts for the direct current (DC) subcarrier and the data carrying subcarriers.
Figure 2.5: The superposition of $N$ mutually orthogonal, modulated subcarriers defining a time domain OFDM waveform [10].

where $N_a$ is the number of data carrying subcarriers. The data symbol duration,

$$T_u = \frac{1}{\Delta f},$$

(2.9)

is inversely related to the subcarrier separation. Although, the data symbol duration for wireless communication standards is typically extended by a fraction of the data symbol duration to mitigate the effects of multipath [28]. This extension is referred to as a cyclic extension [28]. Each of the considered standards, LTE, DVB, and DAB, employ signals with a cyclic prefix (CP) implemented as shown in Figure 2.6. The extended duration is referred to as the symbol duration and is

$$T_{sym} = T_g + T_u = \Delta_g T_u + T_u,$$

(2.10)

where $T_g$ is the cyclic prefix duration and $\Delta_g$ is the cyclic prefix ratio. Using the previously
discussed parameters, an OFDM symbol generated with no cyclic extension is

\[
s(t) = \sum_{n=0}^{N-1} d_n e^{j2\pi \Delta f n t}, \quad T_g \leq t \leq T_{\text{sym}} \tag{2.11}
\]

where \(d_n\) is the subcarrier dependent complex modulation value [10]. The complex modulation value, \(d_n\), is typically assigned a value from the constellation of a modulation scheme employed for the transmitted signal. Modulation schemes defined for the signal standards considered in this work include: binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), differential quadrature phase shift keying (DQPSK), and 16 and 64 quadrature amplitude modulation (QAM). Insertion of a CP produces an OFDM symbol

\[
s'(t) = \begin{cases} 
  s(t + T_u) & \text{if } 0 \leq t < T_g \\
  s(t) & \text{if } T_g \leq t \leq T_{\text{sym}}
\end{cases} \tag{2.12}
\]

where \(s(t)\) is defined in (2.11). Although the combination of Equations (2.11) and (2.12) only define one OFDM symbol, the result may be extended to define multiple symbol transmissions by constructing a time series of OFDM symbols generated with independent complex modulation values [10].
Table 2.2: Parameters of signals considered for this work.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>LTE [16,29]</th>
<th>DVB [18,30]</th>
<th>DAB [17,31]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating mode</td>
<td>-</td>
<td>FDD</td>
<td>8k</td>
<td>TxI</td>
</tr>
<tr>
<td>Tx power(^1) (kW)</td>
<td>(P_t G_t)</td>
<td>1</td>
<td>5.55</td>
<td>2.32</td>
</tr>
<tr>
<td>Carrier frequency (MHz)</td>
<td>(f_c)</td>
<td>728</td>
<td>474</td>
<td>47</td>
</tr>
<tr>
<td>Data symbol duration ((\mu)s)</td>
<td>(T_u)</td>
<td>66.66</td>
<td>896</td>
<td>1000</td>
</tr>
<tr>
<td>CP Ratio</td>
<td>(\Delta g)</td>
<td>1/4</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Symbol duration ((\mu)s)</td>
<td>(T_{\text{sym}})</td>
<td>83.33</td>
<td>1120</td>
<td>1246</td>
</tr>
<tr>
<td>Subcarrier separation (kHz)</td>
<td>(\Delta f)</td>
<td>15</td>
<td>1.116</td>
<td>1</td>
</tr>
<tr>
<td>Data subcarriers per symbol</td>
<td>(N_a)</td>
<td>1320</td>
<td>6816</td>
<td>1536</td>
</tr>
<tr>
<td>Subcarriers per symbol</td>
<td>(N)</td>
<td>2048</td>
<td>8192</td>
<td>2048</td>
</tr>
<tr>
<td>Effective bandwidth (MHz)</td>
<td>(B_{\text{eff}})</td>
<td>19.815</td>
<td>7.608</td>
<td>1.537</td>
</tr>
<tr>
<td>Bandwidth (MHz)</td>
<td>(B)</td>
<td>30.720</td>
<td>9.143</td>
<td>2.048</td>
</tr>
<tr>
<td>Symbols per frame</td>
<td>-</td>
<td>120</td>
<td>68</td>
<td>77(^2)</td>
</tr>
<tr>
<td>Modulation scheme</td>
<td>-</td>
<td>64 QAM</td>
<td>DQPSK</td>
<td></td>
</tr>
</tbody>
</table>

\(^1\) Note, this value represents the equivalent radiated power (ERP).  
\(^2\) Includes a null symbol with duration 1296.88 \(\mu\)s.

In this section, the underlying structure of OFDM signals is provided. However, it is important to note that the description represents the simplest of forms for an OFDM signal. As will be shown in the proceeding sections, LTE, DVB, and DAB standards employ a more complex scheme, where subcarriers may be assigned complex modulation values from multiple modulation schemes or a complex modulation value from a coding sequence (e.g., Zadoff-Chu sequence [16]).

### 2.4 Long Term Evolution

The following is a brief summary of the LTE physical (PHY) layer. For a more detailed description, the reader is referred to [15,16]. Also, for the reader’s reference the parameters of the LTE signal considered for this work are tabulated in Table 2.2.

The 3rd generation partnership project (3GPP) is a partnership project composed of different standards-developing bodies, which first began developing LTE in December 2004 [15]. As of February 2014, 274 operators provide LTE services in 101 countries [32].
LTE was developed to provide reliable mobile-communications across multiple contiguous bands by offering scalable channel bandwidths of 1.4, 3, 5, 10, 15, or 20 MHz [15]. The LTE standard was developed to accommodate an operating frequency range from 450 MHz up to 2.6 GHz [15]. Coverage areas for an LTE basestation range from 5 to 100 km with 5 km cell sizes being the most common [15]. The maximum transmit power for LTE transmissions varies from an equivalent radiated power (ERP) of $\approx 1.26 \text{ kW}$ for channel bandwidths up to 5 MHz to $\approx 2.5 \text{ kW}$ for channel bandwidths of 10 MHz or more [29]. This work assumes a typical ERP of 1 kW for link budget analysis of LTE transmissions in Chapter 4 as typically less than the maximum transmit power is used [2].

LTE makes use of two transmission methods: time division duplexing (TDD) and frequency division duplexing (FDD) [16]. TDD operates with the uplink (UL) and downlink (DL) on the same frequency band with non-overlapping time slots, while FDD has separate frequency bands for the UL and DL transmissions [16]. This work only examines a DL transmission using FDD. Regardless of the channel size, the channel bandwidth is divided into subcarriers with a separation of $\Delta f = 15 \text{ kHz}$ or $\Delta f = 7.5 \text{ kHz}$. Both are supported for LTE [16]. This work only considers an LTE signal with a subcarrier separation of 15 kHz. LTE symbols may contain 2048 or 4096 subcarriers and may employ a normal CP or an extended CP [16]. The normal CP is employed using a variable CP ratio dependent upon the symbol number, while the extended CP uses a constant CP ratio of $\Delta_g = 1/4$ for all symbols [16]. This research only examines symbols composed of $N = 2048$ subcarriers and implemented using an extended CP. For the remainder of this work, the considered LTE signal is referred to as FDD LTE$_e$ DL$^3$ to denote the specific type of LTE signal being examined.

Figure 2.7 illustrates the time and frequency domain structures of the FDD LTE$_e$ DL signal. The FDD LTE$_e$ DL signal has a hierarchical structure in the time domain, with the most basic component being the data symbol with duration $T_u$. The data symbol is

---

$^3$Subscript $_e$ denotes an extended CP.
comprised of a superposition of $N = 2048$ mutually orthogonal subcarriers, 1320 of which carry information. The remaining 728 subcarriers are allocated as guard bands and a DC subcarrier (see Figure 2.7). FDD LTE$_e$ DL data symbols are concatenated in the time domain to form a 0.5 ms slot, 1 ms subframe, and 10 ms radio frame. Slots, subframes, and radio frames each have a constant duration. However, the total number of symbols varies depending on the type of cyclic prefix used. A slot is composed of $N_{\text{sym}} = 7$ symbols for a normal CP or $N_{\text{sym}} = 6$ symbols for an extended CP.

Thus far, the basic structure of an FDD LTE$_e$ DL signal has been described. Synchronization signals and cell-specific reference elements (CSREs) are explained next.

### 2.4.1 Primary and Secondary Synchronization Signals

LTE uses 504 PHY layer cell identities (CIDs), which are assigned to transmitters in different cells so a mobile user may separate information received from multiple transmitters across different cells [16]. Each PHY layer CID is uniquely defined by a primary and secondary synchronization signal (PSS and SSS) pair [15].

The PSS is defined as a 62-length Zadoff-Chu sequence, dependent upon the CID number or root index, $u \in \{25, 29, 34\}$ [16]. The expression for the Zadoff-Chu sequence is provided in [16, pp. 92]. The values from the Zadoff-Chu sequence are mapped to subcarriers (31 occupied subcarriers directly to the left and right of the DC subcarrier) on the last symbol of slots zero and ten [16]. For the symbols containing a PSS, the remaining subcarriers are modulated in a normal fashion. Meaning, guard bands are assigned a complex modulation value of zero, DC subcarrier is assigned zero, and the data is modulated using the modulation scheme.

The SSS is generated by interleaving two 31-length binary sequences, then scrambling using two scrambling sequences [16]. The expressions for the two 31-length binary sequences, scrambling sequences, and implementation are provided in [16, pp. 93-94].
(a) FDD LTEc DL time domain structure.

(b) FDD LTEc DL baseband frequency domain structure for \( N = 2048 \) subcarriers.

Figure 2.7: FDD LTEc DL signal structure.
Similar to the PSS, the SSS has a length of 62 and is mapped to 62 subcarriers centered about the DC subcarrier on the second last symbol of slots zero and ten [16]. The remaining subcarriers of symbols containing SSSs are modulated in the normal fashion.

2.4.2 Cell-Specific Reference Elements

In order for the mobile user to discern which antenna the received signal was transmitted from, CSREs are employed. A pseudorandom 31-length Gold sequence is used within an expression provided in [16, pp. 99-100] to produce complex modulation values for each CSRE. The CSRE values take on those used for a QPSK modulation scheme. Although the complex modulation values for the CSREs are pseudorandom, they may be determined or known \textit{a priori} by knowing the PHY layer CID. Within this work, it is assumed the PHY layer CID is known. However, if this was not the case, cell search procedures using the synchronization signals, PSS and SSS, could be performed as described in [15] to determine the CID.

The CSREs occupy symbols zero and three when using an extended CP [16]. The position of the first occupied subcarrier for the CSREs is determined from the PHY layer CID [16]. CSREs are placed on every six subcarriers, with consecutive symbols containing CSREs having a frequency offset of three subcarriers. The positions of the occupied subcarriers are given by expressions provided in [16, pp. 75-76]. Figure 2.8 is an illustration of the location of CSREs for a signal using an extended CP and having a CID of zero.

2.5 Digital Video Broadcast

The following is a brief summary regarding the structure and properties of DVB signals. A more detailed description may be found in [18, 33, 34]. The reader is referred to Table 2.2 for a tabulation of the properties of the DVB signal considered for this work.
Figure 2.8: Time-frequency grid of a CSRE signal with an extended CP and CID = 0. Colored squares, repeated every $3T_{\text{syn}}$ seconds and every $6\Delta f$ Hz within a symbol, indicate symbols and subcarriers containing the CSREs.

DVB is a technical standard specified for broadcasting digital television. In the early 1990s, DVB was started as a European project [33]. Since then, the technology has spread quickly, and as of February 2014, is now adopted or deployed in 146 countries [34]. The operating frequency range for DVB includes VHF and UHF bands [18]. The transmit power for DVB transmissions may vary significantly based upon the area of operation and desired coverage range (20 mW - 200 kW) [30]. Thus, this work assumes the average ERP, 5.55 kW, of the data set of available DVB transmit powers provided in [30] for link budget analysis in Chapter 4.

The DVB standard accommodates three operating modes: 2k, 4k, and 8k [18]. The type of mode determines the number of subcarriers used for an OFDM symbol, 2048, 4096, or 8192, respectively [18]. The number of subcarriers determines the subcarrier separation as 4.464, 2.232, or 1.116 kHz, respectively [18]. This work only examines an 8k mode DVB signal. It is straightforward to extend the results for other operating modes. To provide reliable service in a variety of environments, DVB offers channel bandwidths of 6, 7, or 8 MHz and CP ratios of 1/4, 1/8, 1/16, or 1/32 [18]. For this research, only a
signal using a CP ratio of $\Delta_g = 1/4$ is considered. Results are easily extended for other CP durations. Throughout the rest of this work, the considered DVB signal is referred to as 8k DVB to denote the type of DVB signal being examined.

An 8k DVB signal is comprised of OFDM symbols containing $N = 8192$ subcarriers with subcarrier separation $\Delta f = 1.116$ kHz (see Figure 2.9(b)). Of the 8192 subcarriers, 6816 are assigned to carry information, modulated using a non-zero complex modulation value. The remaining subcarriers are allotted as guard bands and a DC subcarrier. For 8k DVB signals, the information carrying subcarriers may be modulated using QPSK, 16 QAM, or 64 QAM [18]. In addition, 8k DVB offers variations of 16 and 64 QAM where non-uniform constellations are used as specified in [18]. 8k DVB symbols are arranged in time series to produce frames and superframes [18]. A frame is comprised of 68 symbols but may vary in duration based upon the selected 8k DVB signal parameters (e.g., number of subcarriers comprising a symbol or CP ratio) [18]. Similarly, a superframe may vary in duration and consists of four frames [18]. Within this section, the basic structure of an 8k DVB signal was described. In the following, 8k DVB reference features are described.

### 2.5.1 Scattered Pilots

8k DVB employs scattered pilots as a reference feature for synchronization and channel estimation [33]. The complex modulation value for each subcarrier corresponding to a scattered pilot is known *a priori* and determined from a pseudorandom binary sequence (PRBS). The initialization sequence and polynomial for the PRBS are provided in [18, pp. 27-28]. The complex modulation values assigned to scattered pilots are consistent with those of a BPSK constellation. The modulation values are transmitted with a 2.5 dB power “boost” relative to a typical BPSK modulation scheme (i.e., constellation with unit average energy) [18]. To maintain a sufficient data rate, scattered pilots are sparsely placed within a symbol as shown in Figure 2.10. Scattered pilots are uniformly spaced in the frequency
(a) 8k DVB time domain structure.

(b) 8k DVB baseband frequency domain structure.

Figure 2.9: 8k DVB signal structure.
Figure 2.10: Time-frequency grid of a scattered pilot signal. Colored squares, repeated every $4T_{sym}$ seconds and every $12\Delta f$ Hz within a symbol, indicate symbols and subcarriers containing scattered pilots.

domain occupying every 12th subcarrier. Every symbol contains scattered pilots. Although, adjacent symbols’ scattered pilots have a frequency offset of three subcarriers [18]. The sequence of the position of scattered pilots repeats every fourth symbol as illustrated in Figure 2.10.

### 2.5.2 Continual Pilots

Continual pilots are another reference feature utilized in the 8k DVB standard. Similar to scattered pilots, continual pilots are assigned complex modulation values that are consistent with a BPSK constellation and determined from a PRBS. Moreover, continual pilots are transmitted 2.5 dB higher than typical BPSK constellation values and their positions may coincide with scattered pilots. Continual pilots, also, serve the same purpose as scattered pilots though their structures differ.

Unlike scattered pilots, continual pilots occupy the same subcarriers within every symbol (i.e., no frequency offset between continual pilots of adjacent symbols) [18]. Additionally, continual pilots do not occupy uniformly separated subcarriers. Instead, continual pilots are placed on a total of 177 subcarriers [18]. The indices of the occupied subcarriers are provided in [18, pp. 29] and demonstrate no apparent pattern except each is an integer
multiple of three (not consecutive integers).

2.6 Digital Audio Broadcast

In the following, a summary of the DAB standard is provided. For a more detailed discussion, the reader is referred to [17, 35]. Also, the reader is referred to Table 2.2 for a tabulation of the DAB signal’s properties considered for this research.

In 1987, Eureka Project 147 was founded to develop a system which provided broadcast services for audio program and data services to fixed, portable, and mobile receivers [17]. The work completed for Eureka Project 147 resulted in a standard referred to as DAB. The DAB standard has been accepted worldwide and provides a service which enables spectrum and power efficient broadcasts to be transmitted in the VHF and UHF bands [17, 35]. Depending upon the size of the desired area of coverage and operating environment, the ERP for DAB transmissions may fluctuate largely (3 W - 10 kW) [31]. For this work, the average ERP, 2.32 kW, of a data set of DAB transmit powers provided in [31] is assumed for link budget analysis in Chapter 4.

Similar to FDD LTE, DL and 8k DVB, DAB utilizes multiple operating modes. Multiple operating modes enable DAB systems to be deployed in a number of network configurations and types of environments [17]. The DAB standard defines four transmission modes. Only transmission mode I (TxI) is considered for this research. Unlike FDD LTE, DL and 8k DVB, DAB systems utilizing TxI only employ one channel bandwidth and CP ratio, 1.537 MHz and 63/256, respectively [17]. Throughout the remainder of this work, the considered DAB signal is referred to as TxI DAB to denote the type of DAB signal being examined.

TxI DAB symbols are comprised of \( N = 2048 \) subcarriers and concatenated in the time domain to form a DAB frame. A TxI DAB frame contains 77 symbols, including a null symbol which has duration 1296.88 \( \mu s \) (see Figure 2.11(a)). Although each OFDM symbol
is comprised of 2048 subcarriers, only 1536 are assigned a non-zero complex modulation value from a DQPSK modulation scheme. The remaining 512 subcarriers are designated as guard bands and a DC subcarrier as shown in Figure 2.11(b) [17]. To this point, the basic structure of a TxI DAB signal has been discussed. In the proceeding section, the reference feature employed for TxI DAB signals is discussed.
2.6.1 Synchronization Channel

As previously noted, TxI DAB employs a DQPSK modulation scheme. For demodulation, DQPSK requires initial phase information [36]. Consequently, TxI DAB employs a synchronization channel as a phase reference to the DQPSK demodulator [17]. The synchronization channel, as depicted in Figure 2.11(a), is composed of a null symbol and a synchronization symbol (SS) [17]. The null symbol contains no information, has a duration of 1296.88 $\mu$s, and produces a period which is analogous to the transmitter being “off.” The SS occupies all 1536 information carrying subcarriers, repeats every 77 symbols, and provides the phase reference for DQPSK demodulation [17]. The complex modulation values for the SS are determined by an expression provided in [17, pp. 147-148]. Unlike the deterministic features employed for FDD LTE, DL and 8k DVB transmissions, TxI DAB’s deterministic feature occupies all the information carrying subcarriers contrary to being sparsely inserted.

2.7 Synthetic Aperture Radar Imaging

The purpose of this section is to introduce basic spotlight mode SAR concepts, introduce the OFDM phase history model presented in [10, 11], and describe the processing used for spotlight mode SAR image generation. The focus of this research is not the development of bistatic SAR imaging. This work is concerned with evaluating FDD LTE, DL, 8k DVB, and TxI DAB signals of opportunity for application to passive bistatic SAR imaging. For a more detailed discussion on the development of bistatic SAR imaging the reader is referred to [2, Ch. 10], [37].

The objective of SAR is to synthesize a large aperture, yielding fine crossrange resolution, by coherently combining echoes from a scene of interest collected across the aperture (see Figure 2.12) [2, Ch. 10]. Processing of the pulses for coherent combination requires precise estimates of timing for the collection period, knowledge of the transmitted signal,
and precise knowledge of motion of the transmitter and receiver [38]. The culmination of appropriately processed pulses is referred to as phase history data. Phase history data may be used with common imaging algorithms such as polar reformatting or convolution backprojection (CBP) to produce SAR images [39], [40]. For this research, all SAR images are produced using CBP. The resolution properties of a SAR image are determined by the bandwidth of phase history data in range and crossrange dimensions (assuming a 2D ground plane SAR image). The bandwidth of generated phase history data is dependent upon the geometry of a system for the collection period and properties of the exploited signal. Specifically, the range resolution of a bistatic SAR image is [2, Ch. 10], [37]

\[
\rho_r = \frac{c}{2B \cos(\beta/2)},
\]

and the crossrange resolution is [41]

\[
\rho_x = \frac{\lambda}{4 \sin(\frac{\Delta \phi}{2}) \cos(\tilde{\theta}_b) \cos(\beta/2)},
\]

where \(\Delta \phi\) is the azimuthal extent of the phase history data and \(\tilde{\theta}_b\) is the elevation angle of the bistatic SAR line-of-sight (see [2, pp. 327] for a definition of the bistatic SAR line-of-sight and how it differs from the standard bistatic radar line-of-sight) for the center pulse of the synthetic aperture [2, Ch. 10]. As noted previously, generation of SAR images requires phase history data. The processing necessary to generate phase history data is dependent upon the transmitted signal [40]. For instance, the processing required for a short rectangular pulse differs from that required for a linear frequency modulation (LFM) pulse [40]. In order to exploit OFDM signals for passive bistatic SAR imaging, a processing model for generating phase history is necessary. This model is rederived in the following section from material presented in [10, 11].
Figure 2.12: Basic spotlight mode SAR imaging scenario.

2.7.1 Phase History Model for Orthogonal Frequency Division Multiplexing Signals

Assume a time series of OFDM symbols as the transmitted signal

$$s(t) = \Re\{ e^{j\omega_0 t} \sum_l \sum_n d_{l,n} e^{j\Delta \omega n (t-t_0)} \}_{s_I(t)+js_Q(t)}, \quad (2.15)$$

where $\Re\{\cdot\}$ is the real operator, $\omega_0$ is the angular carrier frequency, $l$ is the symbol number, $n$ is the subcarrier number, $d_{l,n}$ is the complex modulation value for the $n$th subcarrier of the $l$th symbol, $\Delta \omega$ is the angular subcarrier separation, $t_0$ depends on $l$, and $s_I(t)$ and $s_Q(t)$ are the in-phase and quadrature-phase portions of the signal, respectively [10]. The transmitted signal scatters from a scene of interest, assumed to be composed of a continuum
of isotropic point scatterers, and is observed by the receiver as

\[ s_r(t) = \int_{u_1}^{u_2} g(u) \left[ s_f(t - \tau_0 - \tau_u) \cos(\omega_0(t - \tau_0 - \tau_u)) - \right. \]
\[ \left. s_Q(t - \tau_0 - \tau_u) \sin(\omega_0(t - \tau_0 - \tau_u)) \right] du, \]  

(2.16)

where \( u_1 \) and \( u_2 \) are the lower and upper limits of the scene of interest, \( g(u) \) is the aspect dependent, complex reflectivity function of the scene, \( \tau_0 \) is the bistatic time delay from the transmitter-to-scene center-to-receiver, and \( \tau_u \) is the bistatic differential time delay. The received signal, \( s_r(t) \), is basebanded by mixing with \( 2e^{-j\omega_0(t-\tau_0)} \), then passing the received signal through a low-pass filter (LPF) producing

\[ s_r(t) = \int_{u_1}^{u_2} g(u) \left[ s_f(t - \tau_0 - \tau_u) + j s_Q(t - \tau_0 - \tau_u) \right] e^{-j\omega_0 \tau_u} du. \]  

(2.17)

Substituting in the complex, baseband form of (2.15) in for \( s_{BB}(t - \tau_0 - \tau_u) \) of (2.17) yields

\[ s_r(t) = \int_{u_1}^{u_2} g(u) e^{-j\omega_0 \tau_u} \sum_l \sum_n d_{l,n} e^{j\Delta\omega n(t-t_0-\tau_0-\tau_u)} du. \]  

(2.18)

Equation (2.18) is rearranged to yield

\[ s_r(t) = \sum_l \sum_n d_{l,n} e^{j\Delta\omega n(t-t_0-\tau_0)} \int_{u_1}^{u_2} g(u) e^{-j(\omega_0 + \Delta\omega n) \tau_u} du, \]  

(2.19)
where $G[U_n]$ is the discrete Fourier transform (DFT) of the aspect dependent, complex reflectivity function $g(u)$ [10]. By rearranging summations and letting

$$A_n = \sum_l d_{l,n} e^{-j\Delta\omega n(t_0 + \tau_0)} G[U_n],$$

(2.20)

$s_r(t)$ may be represented as a Fourier series

$$s_r(t) = \sum_n A_n e^{j\Delta\omega nt}.$$  

(2.21)

From Fourier series relations, the $n$th coefficient is

$$A_n = \frac{1}{T} \int_0^T s_r(t) e^{-j\Delta\omega nt} dt,$$

(2.22)

where $T$ is the integration period. Equating Equations (2.20) and (2.22) yields [10]

$$\sum_l d_{l,n} e^{-j\Delta\omega n(t_0 + \tau_0)} G[U_n] = S_r[n].$$

(2.23)

Recognizing the left hand side of (2.23) is the product of the $n$th coefficient of the DFT of the ideal transmitted signal, delayed to scene center and $G[U_n]$, $G[U_n]$ may be solved for as

$$G[U_n] = \frac{S_r[n]}{S_{ref}[n]},$$

(2.24)

where $S_{ref}[n]$ is the $n$th coefficient of the spectrum of the ideal transmitted signal, delayed to scene center [10]. When $S_{ref}[n] = 0$, a divide by zero is encountered. Consequently, a
pseudo inverse may be taken where [10]

\[
G[U_n] = \begin{cases} 
S_r[n] & \text{if } S_{\text{ref}}[n] \neq 0 \\
S_{\text{ref}}[n] & \text{if } S_{\text{ref}}[n] = 0.
\end{cases}
\]

(2.25)

From (2.25), it may be seen that a matched filter may be constructed by knowing the complex modulation value for each subcarrier of the transmitted signal. In [10, 11], a variation of the matched filter in (2.25) is proposed to mitigate the effects of clutter and noise. The proposed scheme is referred to as an averaged matched filter (AMF). The AMF is a coherent average over segments of each received pulse of the synthetic aperture using an arbitrary segment length (see Figure 2.13). When using the full, ideal transmitted signal (i.e., signal comprised of “user” data and known signaling features) as the matched filter, we refer to the AMF as a full AMF. Conversely, when the matched filter is comprised only of a signal’s deterministic features, the AMF is referred to as a partial AMF. The full and partial AMF are employed in Chapter 5 for generating small-scale, experimental passive SAR imaging results.

2.8 Chapter Conclusion

In conclusion, this chapter provides a summary of previous related research and describes the background material necessary for the results presented in Chapters 3–5. The background material includes: 1) bistatic geometry, timing, and link budget; 2) summary of the signal structure and properties of FDD LTE, DL, 8k DVB, and TxI DAB signals; 3) basic SAR imaging concepts, OFDM phase history model rederived, and techniques employed for SAR image generation. This background information is helpful in understanding the effects of signal structure on a signal’s AFs in Chapter 3, evaluating link budget for LTE,
One pulse of the synthetic aperture

(a) AMF timing for one pulse of a synthetic aperture [11].

(b) AMF [11].

Figure 2.13: Timing and implementation of the AMF.
DVB, and DAB transmission sites in Chapter 4, and for better interpretation of generated SAR images in Chapter 5.
Chapter 3

Ambiguity Analysis

In this chapter, attainable capabilities and shortcomings derived from FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB signals’ transmitted waveform are evaluated. These properties are characterized by the self- and cross- AFs. From the AFs, the effects of signal structure on matched filtered outputs are determined. The chapter concludes by demonstrating the effects of signal structure on SAR image output, produced by exploiting a simulated FDD LTE<sub>e</sub> DL transmission.

3.1 Ambiguity Function

An extension of the monostatic AF to the bistatic case is developed in [42]. The bistatic AF should be applied in the passive radar case. However, the bistatic AF depends on the geometry of the transmitter, receiver, and target. In this work, the primary concern is with the intrinsic properties of the considered signals than with the bistatic geometry dependence. Therefore, we evaluate the simplest geometry and report a discrete instantiation (see (2.2)) of the monostatic AF described in (2.1). Results for the bistatic AF may be inferred from the monostatic AF. This point is discussed further in Section 3.5. Peaks due to structure
within the transmitted signal will appear in both the monostatic and bistatic AFs. (Note, in this work the term peaks refers to distinguishable energy levels located outside the main lobe of the AF. These distinguishable energy levels are not referred to as ambiguities as the term ambiguity is dependent upon the threshold value for a given radar application.)

For this research, a simulated signal $s[n]$ is generated for each signal standard using MATLAB® with the parameters shown in Table 2.2. Each signal contains random “user” data (i.e., the complex modulation value for data carrying subcarriers is randomly selected from the signal’s modulation constellation) and deterministic features as described in Sections 2.4.1, 2.4.2, 2.5.1, 2.5.2 and 2.6.1. The reference signal, $s_{\text{ref}}[n]$, is used for matched filtering the received signal. For AF analysis, a single target at zero-range ($\tau = 0$) and zero-Doppler ($\nu = 0$) is assumed, so the received signal is equal to the transmitted signal $s[n]$ [19]. It is assumed the reference signal is perfectly constructed either by knowing $s[n]$ a priori or collecting a direct-path signal to build the ideal waveform. To reiterate, when $s_{\text{ref}}[n] = s[n]$, we refer to (2.2) as the SAF. As alluded to in Chapter 2, one may wish to examine the effects of known signaling features, such as the PSS, SSS, CSREs, scattered or continual pilots, or SS on the SAF result. In such cases, $s_{\text{ref}}[n] \neq s[n]$, and (2.2) is referred to as the CAF. In the following, we compute one discrete instantiation of the SAF and CAFs for the considered communication signal standards and their respective deterministic features using (2.2). For each AF result pertaining to a particular signal type (i.e., FDD LTEe DL), the same signal $s[n]$ is used. We compare each SAF and CAF result qualitatively as well as quantitatively using integrated sidelobe level (ISL) and peak sidelobe level (PSL).
Table 3.1: Quantitative metrics for one, discrete instantiation of the SAF and CAFs of an FDD LTE<sub>e</sub> DL signal.

<table>
<thead>
<tr>
<th>Result</th>
<th>$\chi(0, 0)^2$ (dB)</th>
<th>ISL (dB)</th>
<th>PSL (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SAF</td>
<td>0</td>
<td>19.05</td>
<td>-14.14</td>
</tr>
<tr>
<td>CAF (CSRE)</td>
<td>-12.55</td>
<td>30.80</td>
<td>-1.68</td>
</tr>
<tr>
<td>CAF (PSS and SSS)</td>
<td>-28.06</td>
<td>46.46</td>
<td>-0.02</td>
</tr>
</tbody>
</table>

### 3.2 Frequency Division Duplexing Long Term Evolution Downlink

In the proceeding FDD LTE<sub>e</sub> DL analysis, much of the material presented is referenced from [43]. For the analysis, a radio frame comprised of 120 FDD LTE<sub>e</sub> DL symbols, each containing random “user” data chosen from a 64 QAM constellation, is considered one radar pulse. Consequently, SAF analysis is completed by matched filtering with a full radio frame. Conversely, CAF analysis is achieved by partially matched filtering with a radio frame containing only CSREs or synchronization signals (PSS and SSS). For the parameters of the considered FDD LTE<sub>e</sub> DL signal, the reader is referred to Table 2.2.

#### 3.2.1 Self-Ambiguity Function

Figure 3.1 plots the result of computing (2.2) for a discrete instantiation of the SAF, $s_{ref}[n] = s[n]$. Since in Table I $s[n]$ is defined to be one full radio frame or DL transmission, the result corresponds to matched filtering a single radar pulse. To reiterate, the result shown in Figure 3.1 is one discrete instantiation of the AF described in (2.1). On average, the two most prominent peaks seen in Figure 3.1 will remain at the energy levels shown. As more instantiations are considered, the interference floor from random “user” data will continue to degrade and peaks resulting from CSRE, PSS, and SSS structure will become more apparent (i.e., peaks from signal structure will rise above the interference floor).
The SAF has significant peaks around two locations: 1) the zero-delay, zero-Doppler reference and 2) at zero-Doppler, $T_u = 66.67 \, \mu s$ delay. As expected, the secondary peak is due to the alignment of the CP portion of each symbol with its corresponding copy within the original symbol duration. From FDD LTE\textsubscript{e} DL signal properties described in Section 2.4, the unambiguous radar range (URR) for the monostatic case is $\frac{c T_u}{2} \approx 10$ km. The 19.815 MHz effective signal bandwidth (see Table 2.2) results in a range resolution of $\approx 7.56$ m. Dividing the URR by range resolution yields 1321 range bins. Table 3.1 provides a listing of the amplitude of the zero-delay, zero-Doppler peak as well as the ISL and PSL for the SAF. From the discrete instantiation shown in Figure 3.1, it is not apparent where the peaks from CSRE, PSS, or SSS structure occur. Therefore, in Sections 3.2.2 and 3.2.3, we give consideration to the radar response for simpler processing that constructs a partially matched filter from the CSREs, PSS, and SSS described in Sections 2.4.1 and 2.4.2.
3.2.2 Cell-Specific Reference Element

Cross-Ambiguity Function

Figure 3.2 shows a discrete instantiation of the CAF result when $s_{\text{ref}}[n]$ is built using $s_{\text{CSRE}}[n]$ containing only the CSREs (and zeros in the non-CSRE subcarriers and symbols) described in Section 2.4.2. As mentioned in Section 3.2.1, on average, peaks from CSRE structure will become more apparent in the FDD LTE DL SAF. From the result shown in Figure 3.1, this structure is not evident. Thus, to better understand how CSRE structure effects the SAF on average, the CAF is considered.

Figure 3.3 shows the top view of Figure 3.2 along with peak locations predicted by evaluating the CSRE intervals. As discussed in Section 2.4.2, CSREs are uniformly spaced every six subcarriers within a symbol and occupy every third symbol with successive symbols containing CSREs having a frequency offset of three subcarriers. The timing of the CSRE signal makes it analogous to a pulsed radar system having a pulsewidth of $T_{\text{sym}}$ and pulse repetition interval (PRI) of $3T_{\text{sym}}$ (see Figure 2.8). According to [19], the $M$-period periodic AF for a stepped-frequency train of coherent pulses is

$$|\chi(\tau, \nu)| = |\chi_{M=1}(\tau, \nu + k_s\tau)| \left| \frac{\sin(M\pi(\nu + k_s\tau)T_p)}{M\sin(\pi(\nu + k_s\tau)T_p)} \right|,$$

(3.1)

where $\chi_{M=1}(\tau, \nu + k_s\tau)$ represents the AF produced using one period of the signal, $T_p$ is the PRI of the signal, and $k_s$ is the slope of the stepped-frequencies. From Figure 2.8, we note the slope of the stepped-frequency for CSREs is

$$k_s = \pm \frac{3\Delta f}{3T_{\text{sym}}},$$

(3.2)
resulting in peaks of the example FDD LTE<sub>e</sub> DL-CSRE CAF at delay-Doppler locations

\[
(\tau', \nu') = \left( \frac{n_\tau T_u}{6} + 6zT_{\text{sym}}, \frac{n_\nu}{3T_{\text{sym}}} - \frac{n_\tau}{6T_{\text{sym}}} - 6\Delta f z \right)
\]

\[
= \left( (11.11n_\tau + 500z) \mu s, (4n_\nu - 2n_\tau - 90z) \text{kHz} \right),
\]

where \( n_\tau, z, \) and \( n_\nu \) are integers. The locations predicted using (3.3) have been overlaid on the delay-Doppler plane of the CAF in Figure 3.3 for ease of comparison. It is clear from Figure 3.3 the peaks seen within the CAF are a result of the structure of the CSRE signal. As mentioned in Section 3.2.1, the peaks arising in the CAF from the structure of the CSRE signal are not apparent in the FDD LTE<sub>e</sub> DL SAF. The lack of these peaks is a result of them being masked by the interference floor from the random “user” data, since only one instantiation is considered for the SAF shown in Figure 3.1. In addition, the total energy contained in \( s_{\text{CSRE}}[n] \) is significantly less than that of \( s[n] \), as evidenced by a 12.55 dB reduction (see Table 3.1) of main lobe energy of the FDD LTE<sub>e</sub> DL-CSRE CAF.
Figure 3.3: Locations of peaks $(\tau',\nu')$ in the delay-Doppler plane arising from the structure of the CSRE signal as predicted by (3.3) with $n_\tau \in [0,6]$, $z = 0$, and $n_\nu \in [-1,4]$. Additional peaks are predicted by (3.3) but occur outside the region shown. Only peaks within $\approx 15$ dB of the zero delay, zero Doppler peak of the FDD LTE$_e$ DL-CSRE CAF ($> -30$ dB) result are highlighted.

compared to the FDD LTE$_e$ DL SAF.

Figures 3.2 and 3.3, along with (3.3), provide insight into the effects of FDD LTE$_e$ DL’s CSREs on the FDD LTE$_e$ DL SAF. Moreover, the previously shown results provide a better understanding for the development of potential mismatched filtering schemes.

### 3.2.3 Synchronization Signal Cross-Ambiguity Function

The reference waveform used to compute a discrete instantiation of the CAF in Figure 3.4 contains both the PSS and SSS ($s_{\text{ref}}[n] = s_{\text{PSS}}[n] + s_{\text{SSS}}[n]$). Due to their low energy levels, both synchronization signals are considered together as one *a priori* signal feature. The unstructured shape and low peaks of the non-zero delay region may be attributed to the favorable autocorrelation sequences of the Zadoff-Chu and binary sequences defining the synchronization signals [19]. However, the relatively large PRI of the synchronization
signals results in peaks every \( \frac{1}{607\text{sym}} \approx 182 \text{ Hz} \) along the zero delay cut, as a result of the Dirichlet sinc term in (3.1).

As was the case with FDD LTEe DL CSREs, peaks from the synchronization signals are not apparent in the FDD LTEe DL SAF shown in Figure 3.1, since only one instantiation of the SAF is considered. Moreover, the synchronization signals only comprise a small portion of the energy contained within FDD LTEe DL signals, as evidenced by the estimate of the FDD LTEe DL-synchronization signal CAF main lobe energy, \(-28.06 \text{ dB}\), shown in Table 3.1.

From Figure 3.4 and the aforementioned synchronization signal intervals, the effects of FDD LTEe DL’s synchronization signals on the FDD LTEe DL SAF may be inferred. These results aid in the development of potential mismatched filtering schemes for FDD LTEe DL signals by providing insight into how the synchronization signals affect the FDD LTEe DL SAF result.
Table 3.2: Quantitative metrics for one, discrete instantiation of the SAF and CAFs of an 8k DVB signal.

| Result                          | $|\chi(0,0)|^2$ (dB) | ISL (dB) | PSL (dB) |
|---------------------------------|----------------|----------|----------|
| SAF                             | 0              | 19.65    | -13.99   |
| CAF (scattered pilots)          | -8.63          | 28.31    | -2.37    |
| CAF (continual pilots)          | -13.7          | 33.53    | -0.5     |

3.3 8k Digital Video Broadcast

For SAF analysis of an 8k DVB signal, a frame comprised of 68 symbols containing all features and random “user” data is considered one radar pulse. CAF analysis is performed by partially matched filtering with a frame comprised of 68 symbols containing only scattered pilots or continual pilots. The parameters of the considered 8k DVB signal are provided in Table 2.2 for the reader’s reference.

3.3.1 Self-Ambiguity Function

The SAF for an 8k DVB signal, computed by setting $s_{ref}[n] = s[n]$ in (2.2), is shown in Figure 3.5. Recall, the result shown in Figure 3.5 corresponds to one discrete instantiation of the AF described in (2.1). Deduced from the 8k DVB signal structure discussion in Section 2.5, SAF peaks are expected in numerous locations resulting from the following: 1) target at zero-delay, zero-Doppler, 2) scattered pilots, 3) continual pilots, and 4) CP. In particular, the peaks witnessed in Figure 3.5 prior to $T_u = 896 \mu s$ delay may be attributed to the scattered and continual pilots. Although peaks from the pilots are already evident in one instantiation of the 8k DVB SAF, as more iterations are averaged, the pilot peaks will become even more discernible (i.e., the interference floor from the random “user” data will continue to degrade). The location and occurrence of these peaks are discussed further in Sections 3.3.2 and 3.3.3.

As may be inferred from 8k DVB’s implementation, the most prominent secondary peaks in the SAF arise along $T_u = 896 \mu s$ delay cut from the CP portions of each symbol.
aligning with its corresponding copy. The location of the peaks are made evident by evaluating (3.1) using an 8k DVB signal’s intervals ($\tau = T_u$, $\tau_p = T_{sym}$, and $k_s = 0$), producing peaks at the following delay-Doppler locations

$$(\tau', \nu') = \left( T_u, \frac{n\nu}{T_{sym}} \right)$$

$$= (896 \, \mu s, \ 892.86n\nu \, \text{Hz}), \quad (3.4)$$

where $n\nu$ is an integer. In Figure 3.6, peaks greater than -25 dB\textsuperscript{1} are highlighted illustrating the periodicity of the CP peaks for an 8k DVB signal as described by (3.4) ($n\nu \in [-5, 5]$). Peaks along $\tau = T_u$ delay cut are evident in Figures 3.5 and 3.6 but not along $\tau = 0$ delay cut as one may expect. This lack of peaks along $\tau = 0$ delay cut is a result of (3.1) being equivalent to zero for non-zero integer values of $n\nu$ and $\tau = 0$. Therefore, peaks only occur at integer multiples of $\frac{1}{T_{sym}}$ along the $\tau = T_u$ delay cut due to the Dirichlet sinc term in (3.1). These same repetitive peaks for the chosen FDD LTE\textsubscript{e} DL signal along the $\tau = T_u$ delay cut were not made apparent in Figure 3.1 as the repetitive peaks appear at multiples of 12 kHz, requiring excessive computational power. From 8k DVB signal parameters, the URR for the monostatic case is $\frac{cT_u}{2} \approx 134$ km. The 7.608 MHz signal bandwidth (see Table 2.2) results in a range resolution of $\approx 19.70$ m. The number of range bins, the quotient of the URR and range resolution, is 6817 bins.

As previously mentioned, peaks prior to $T_u = 896 \, \mu s$ delay cut are a result of 8k DVB’s pilots. Therefore, in Sections 3.3.2 and 3.3.3, we analyze the effects of these features in greater detail by examining the CAF results.

\textsuperscript{1}For SAF results, peaks $> -25$ dB are highlighted, while for CAF results, peaks within $\approx -15$ dB of the zero delay, zero Doppler CAF peak are highlighted.
Figure 3.5: SAF of an 8k DVB signal. Peaks are apparent from the CP as well as the scattered and continual pilots.

Figure 3.6: Top down view of the 8k DVB signal’s SAF. The most apparent peaks are a result of the CP and the signal’s intervals.
3.3.2 Scattered Pilot Cross-Ambiguity Function

As remarked in Section 2.5.1, 8k DVB scattered pilots are uniformly spaced every 12 subcarriers within a symbol, and occupy every symbol with successive symbols containing scattered pilots having a frequency offset of three subcarriers. Referring to Sections 2.4.2 and 2.5.1, it is apparent that 8k DVB scattered pilots are analogous to FDD LTE DL CSREs with different intervals. Therefore, a result similar to that seen in Figures 3.2 and 3.3 is expected but with scattered pilot intervals.

As pointed out in Section 3.3.1, peaks prior to $T_u$ delay in the 8k DVB SAF are a consequence of 8k DVB pilot structure. To better understand why and where these peaks occur, the 8k DVB-scattered pilot CAF is considered. Figure 3.7 depicts a discrete instantiation of the CAF produced using a scattered pilot reference signal, $s_{\text{ref}}[n] = s_{\text{SP}}[n]$, in (2.2). From Figure 2.10, we note the scattered pilot signal has a pulsewidth and PRI of $T_{\text{sym}}$ (i.e., continuous wave radar system) and a stepped-frequency slope of

$$k_s = \pm \frac{3\Delta f}{T_{\text{sym}}}. \tag{3.5}$$

Substituting the scattered pilot intervals into (3.1) results in peaks located in the delay-Doppler plane at

$$(\tau', \nu') = \left( \frac{n_\tau T_u}{12} + 4z T_{\text{sym}}, \frac{n_\nu}{T_{\text{sym}}} - \frac{n_\tau}{4T_{\text{sym}}} - 12\Delta f z \right)$$

$$= \left( (74.67n_\tau + 4480z) \mu s, (0.89n_\nu - 0.22n_\tau - 13.39z) \text{ kHz} \right), \tag{3.6}$$

where $n_\tau$, $z$, and $n_\nu$ are integers. Figure 3.8 shows the top view of Figure 3.7 with peak positions highlighted and locations predicted from (3.6) circled. From Figure 3.8, it is clear that peaks do not occur at every circled location, but all peaks that are present coincide with locations predicted by (3.6). The lack of peaks at each circled location may be attributed to
unpredictable inter-symbol interference (ISI), noting that scattered pilots are placed within every symbol. As noted previously, peaks arise in the SAF due to scattered pilots. These peaks are a result of the scattered pilots occupying every symbol and being transmitted with a 2.5 dB power “boost” relative to a typical BPSK modulation scheme. These properties produce a CAF zero delay, zero Doppler peak $\approx -8.63$ dB below the zero delay, zero Doppler peak of the 8k DVB SAF. The 8k DVB-scattered pilot CAF peak is $\approx 4$ dB higher than the FDD LTE$_e$ DL-CSRE CAF peak (see Tables 3.1 and 3.2), further demonstrating why the CSREs do not produce peaks in the FDD LTE$_e$ DL SAF, but the scattered pilots do produce peaks in the 8k DVB SAF.

As alluded to with previous CAF results, 8k DVB-scattered pilot CAF analysis enables improved development of potential mismatched filters and provides insight on the effects of 8k DVB scattered pilots on the 8k DVB SAF.
Figure 3.8: Locations of peaks \((\tau', \nu')\) in the delay-Doppler plane arising from the structure of the scattered pilot signal as predicted by (3.6) with \(n_\tau \in [1, 12], z = 0, \) and \(n_\nu \in [-5, 8]\). Only peaks within \(\approx 15\) dB of the zero delay, zero Doppler peak of the 8k DVB-scattered pilot CAF \((> -25\) dB) result are highlighted.

### 3.3.3 Continual Pilot Cross-Ambiguity Function

As is the case with 8k DVB scattered pilots, continual pilots have well-defined modulation values determined from a PRBS and occupy subcarriers specified in [18]. The structure of the continual pilots result in a less favorable 8k DVB SAF which exhibits additional peaks outside the zero delay, zero Doppler peak. A discrete instantiation of the CAF produced using \(s_{\text{ref}}[n] = s_{\text{CP}}[n]\) is shown in Figure 3.9.

In Section 2.5.2, it was mentioned that continual pilots occupy 177 subcarriers within every symbol but are not inserted on uniformly separated subcarriers. Moreover, continual pilot subcarrier indices demonstrate no apparent pattern except that each index is an integer multiple of three (not consecutive integers). Combining the fundamental period shared by occupied subcarriers with a pulsewidth and PRI of \(T_{\text{sym}}\) \(\) (i.e., continuous radar wave system)
Figure 3.9: CAF of an 8k DVB continual pilot signal. The peaks are a result of the structure of the continual pilot signal and ISI.

results in predicted peaks at the following delay-Doppler plane locations

\[
(\tau', \nu') = \left( \frac{n_\tau T_u}{3} + z T_{\text{sym}}, \frac{n_\nu}{T_{\text{sym}}} \right)
= \left( \left(298.67n_\tau + 1120z\right) \mu s, \ 892.86n_\nu \text{ Hz} \right), \quad (3.7)
\]

where \( n_\tau, z, \) and \( n_\nu \) are integers. Figure 3.10 depicts the locations predicted by (3.7) overlaid on the 8k DVB-continual pilot CAF shown in Figure 3.9. While some peaks occur at locations predicted by (3.7), it is evident that all peaks do not coincide with these locations. This lack of predictability is a result of elevated ISI due to adjacent symbols sharing the same continual pilot subcarriers. While ISI was present in the 8k DVB-scattered pilot CAF, the ISI level was lower in that case due to only every fourth symbol sharing the same scattered pilot subcarriers as adjacent symbols have a scattered pilot frequency offset. The elevated levels of ISI for the continual pilots is further made evident by the numerous peaks, which occur along the zero Doppler cut shown in Figures 3.9 and 3.10. Similar to
scattered pilots, continual pilots produce noticeable peaks within the 8k DVB SAF. As was the case with scattered pilots, these peaks are a result of continual pilots occupying every symbol and being transmitted with a 2.5 dB power “boost” relative to a typical BPSK modulation scheme. Although the continual pilots contribute to the peaks witnessed in the 8k DVB SAF, the continual pilots contribution is less significant than the scattered pilots as made evident by the continual pilots reduced zero delay, zero Doppler energy level shown in Table 3.2.

### 3.4 Transmission Mode I Digital Audio Broadcast

For SAF and CAF analysis of a TxI DAB signal, a 77-symbol signal comprised of random “user” data chosen from a DQPSK constellation is considered one radar pulse. SAF analysis is achieved by matched filtering with a full frame, while CAF analysis is completed by partially matched filtering with a full frame containing only a SS. The properties of the
Table 3.3: Quantitative metrics for the SAF and CAFs of a TxI DAB signal.

| Result | $|\chi(0,0)|^2$ (dB) | ISL (dB) | PSL (dB) |
|--------|--------------------|----------|----------|
| SAF    | 0                  | 19.51    | -14.12   |
| CAF (SS)| -18.82             | 35.35    | -9.18    |

considered TxI DAB signal are tabulated in Table 2.2 for the reader’s reference.

3.4.1 Self-Ambiguity Function

A discrete instantiation of the SAF for a TxI DAB signal is shown in Figure 3.11 and generated using $s_{ref}[n] = s[n]$ in (2.2). Similar to the 8k DVB SAF, the TxI DAB SAF exhibits multiple peaks at the $\tau = T_u$ delay cut. As was the case in Section 3.3.1, these peaks are expected and are a result of the signal’s CP. The result derived for the 8k DVB SAF in (3.4) is written in terms of the signal’s parameters. Thus, the same result applies using a TxI DAB signal’s parameters (see Table 2.2), resulting in peaks located at the following locations in the delay-Doppler plane

$$(\tau', \nu') = (1 \text{ ms}, 802.51n_\nu \text{ Hz}), \quad (3.8)$$

where $n_\nu$ is an integer. Figure 3.12 highlights peaks within -25 dB$^2$ of the zero delay, zero Doppler peak shown in Figure 3.11 ($n_\nu \in [-6, 6]$). As described by (3.8), the peaks occur at intervals consistent with TxI DAB signal intervals. As explained in Section 3.3.1, peaks are not present along the $\tau = 0$ delay cut due to the Dirichlet sinc function being equivalent to zero at non-zero integer multiples of $\nu = \frac{1}{T_{\text{sym}}}$ and $\tau = 0$. From TxI DAB signal parameters, the URR is $\frac{cT_u}{2} \approx 150 \text{ km}$. A TxI DAB signal occupies a channel bandwidth of 1.537 MHz providing a range resolution of $\approx 97.53 \text{ m}$. The combination of

---

$^2$For SAF results, peaks $>-25$ dB are highlighted, while for CAF results, peaks within $\approx -15$ dB of the zero delay, zero Doppler CAF peak are highlighted.
Figure 3.11: SAF of a TxI DAB signal. Peaks are apparent from the CP.

Figure 3.12: Top down view of the TxI DAB signal’s SAF. The most apparent peaks are a result of the CP and the signal’s intervals.
the URR and range resolution provide 1537 range bins.

As performed with FDD LTE\textsubscript{e} DL and 8k DVB deterministic features, the TxI DAB CAF is considered in the following to gain a better understanding of the impact of TxI DAB’s deterministic signaling feature on the SAF.

### 3.4.2 Synchronization Signal Cross-Ambiguity Function

As stated for previous CAF results, CAF analysis is performed to gain a better understanding of how the signal’s structure impacts the SAF and to aid in the development of potential mismatched filtering schemes. A discrete instantiation of the TxI DAB-SS CAF is shown in Figure 3.13 and is generated by partially matched filtering with a TxI DAB SS ($s_{\text{ref}}[n]=s_{SS}[n]$). Unlike the deterministic features of the signals considered in the preceding sections, SSs are not periodic within one frame of a TxI DAB signal (only one SS is present in every 77 symbols). Therefore, the effects of the Dirichlet sinc function witnessed in previous results is not evident (the result shown corresponds to $|\chi(\tau,\nu)_{M=1}|^2$ in (3.1)). For example, SSs have a periodicity of $77T_{\text{sym}}$ resulting in expected non-zero Doppler peaks at $\approx 33.24$ m/s. However, the result in Figure 3.13 does not illustrate peaks at these intervals as the considered signal, a TxI DAB frame (77 symbols), only contains one SS and does not exhibit a periodic envelope. On the other hand, if a TxI DAB signal containing two SSs is considered, the non-zero Doppler peaks will be present as the signal has a periodic envelope. Also, as made evident by Figure 3.13, the lack of periodicity produces peaks with apparent “degraded” Doppler resolution. The apparent “degraded” Doppler resolution is a result of no spectral lines resulting from the Dirichlet sinc term in (3.1). The “degraded” Doppler resolution is most apparent at the zero delay, zero Doppler peak as well as along the $\tau = T_u$ delay cut. In the TxI DAB SAF shown in Figure 3.12, peaks are evident along the $\tau = T_u$ delay cut at intervals of $\frac{1}{T_{\text{sym}}}$. These same peaks may be seen in the TxI DAB-SS CAF shown in Figure 3.13. Although, due to the peaks exhibiting “degraded” Doppler resolution, the peaks overlap. In Figure 3.13, this overlap is made
apparent by the mound which occurs along the $\tau = T_u$ delay cut, extending to $\pm 1.51$ kHz centered about $\nu = 0$.

### 3.5 Geometry and Signal Parameter

#### Considerations

The monostatic AFs presented thus far represent a best-case scenario. The bistatic AF exhibits degraded range and Doppler resolution, inversely proportional to the cosine of half the bistatic angle [27]. Using the expressions derived for the bistatic AF in [42], peaks may be inferred from the monostatic result with respect to the receiver by

\[
\left( \frac{R'_R}{c}, \nu'_b \right) = \left( \frac{(\tau + \tau')^2 - L^2}{2((\tau + \tau') + L \sin(\theta_R))}, \nu \cos(\beta/2) \right).
\]  

(3.9)
Figure 3.14: Locations of bistatic peaks inferred from the FDD LTE\textsubscript{e} DL-CSRE CAF monostatic result. Peaks are normalized to zero time delay with respect to the receiver by $(R'_R - R_R, \nu'_b)$ from (3.9) $(n_\tau \in [0, 6], z = 0, \text{ and } n_\nu \in [-1, 4])$.

Figure 3.14 depicts the bistatic FDD LTE\textsubscript{e} DL-CSRE CAF with respect to the receiver assuming the shown tabulated geometry. To demonstrate the inference of bistatic peaks resulting from signal structure, peaks greater than $-30$ dB have been highlighted and locations predicted by (3.9) circled. From the results, it is clear bistatic peaks may be inferred from the monostatic result as each location is accurately predicted. In addition to antenna geometry, variations in signal parameters affect the AF results. For each of the considered signals, the bandwidth is not increased for an increased value of subcarriers. Instead, the same allocated bandwidth is divided among the increased number of subcarriers resulting in reduced subcarrier separation. The reduced subcarrier separation results in a longer pulse duration, which for many of the results shown in the preceding sections yields an increase in URR. Of the three standards considered, only FDD LTE\textsubscript{e} DL and 8k DVB offer

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variable channel bandwidths. For the previously shown results, only the largest available channel bandwidth was considered. Therefore, for smaller channel bandwidths a reduction in range resolution is expected. Of the many parameters settings possible, decreased sub-carrier separation is desirable. Resolution and ambiguity trade-offs must be balanced if the radar engineer has the ability to choose signal parameters.

3.6 Application to Passive Bistatic Synthetic Aperture Radar Imaging

As previously noted, it is important the radar engineer understand limitations arising from the structure of a signal of opportunity’s transmitted waveform. In the preceding sections, peaks arising from the different features of FDD LTEe DL, 8k DVB, and TxI DAB signals were presented. In this section, a simulated applied example is shown demonstrating the effects of FDD LTEe DL CSRE peaks on a passive bistatic SAR image. To demonstrate an application of the results shown in the preceding sections to passive bistatic SAR imaging, a simple bistatic scenario is assumed (see Figure 3.15(a)). The scenario consists of a stationary broadcast transmitter located at (0 km, -5 km) in the ground plane. The assumed signal of opportunity is an FDD LTEe DL transmission comprised of random “user” data chosen from a 64 QAM constellation (see Table 2.2). A receiver is assumed in an over-the-shoulder configuration with a standoff distance of 20 km to the transmitter. For simplicity, the transmitter and receiver are assumed to be located in the ground plane with the target to produce a result in the bistatic plane (for sake of demonstration). This result may be extended further to the three dimensional problem if need be by analyzing the isorange contours of the system. The scene is assumed to consist of one isotropic point scatterer located at (0 km, 5 km) in the ground plane. Additionally, the scenario is assumed to be noiseless to isolate the effects of the transmitted waveform on the passive bistatic SAR
Figure 3.15: The first additional peak arising from the structure of an FDD LTE_e DL CSRE signal in a simulated passive bistatic SAR image. Due to the nature of the peaks witnessed in the results shown in the preceding sections, peaks arising from the structure of the signal appear as moving targets.
Figure 3.15(c) depicts the simulated passive bistatic SAR image formed by partially matched filtering with an FDD LTE DL CSRE signal. In the range profile of the center pulse shown in Figure 3.15(b), it is evident the actual target appears at a range of 5 km with secondary peaks occurring at integer multiples of 3.33 km relative to the target due to the CSREs structure. Because the secondary peaks arise at a constant differential range relative to the target in each range profile, the secondary peaks act similar to moving targets in Figure 3.15(c). In Figure 3.15(c), it is apparent the first additional peak from CSRE structure produces a streak in the image analogous to a moving target at a range of $\approx 3.33$ km relative to the target.

The simple demonstration shown in Figure 3.15 illustrates how the peaks observed in the preceding sections may affect passive bistatic SAR imaging results. The predictability of secondary peaks for a simple geometry is straightforward as demonstrated by the previous example. However, as the transmitter and receiver are moved out of the ground plane, predictability becomes more complex. Even so, by examining the isorange contours, secondary peaks may be located for more complex system configurations.

For the result shown in Figure 3.15, a single isotropic point scatterer in a noiseless environment was considered. Thus, the result does not illustrate many of the challenges encountered with real-world SAR imaging applications. Because the scenario considered is noiseless, there is no speckle seen within Figure 3.15 (the effects of thermal noise on SAR image output is considered in greater detail in Section 4.3). Although the effects of thermal noise on SAR image output must be considered, thermal noise is typically not the most pressing concern for producing quality SAR images [44]. In general, clutter background from surrounding terrain is a more daunting obstacle than thermal noise is for producing quality SAR images. Moreover, as a set of point targets are considered, point targets’ sidelobes may constructively combine to mask actual target returns. Both clutter returns and constructively combined sidelobes may reduce the dynamic range of the system to where
the effects of thermal noise are negligible (i.e., the level of the thermal noise falls outside the examined dynamic range) [44]. To mitigate the effects of clutter returns and constructively combining sidelobes, many processing schemes may be employed. In particular, the sidelobe levels of the impulse response function may be reduced to a desirable level by applying a window function to the phase history data (e.g., Taylor window [40]), while the effects of clutter may be reduced by employing appropriate filtering schemes discussed in [44].

3.7 Additional Results

In the following, the effects of signal structure on SAR image output are illustrated for matched filtering schemes not considered in Section 3.6. For the following results, only the range profile from the center pulse of the synthetic aperture is examined. Range profiles are used since all of the results require considerable computational power (i.e., utilizing CBP, which is $O(N^3)$ [2], with a scene extent $> 10$ km). The configuration of the system assumed for the following analysis is equivalent to that shown in Section 3.6 (see Figure 3.15(a)).

3.7.1 Frequency Division Duplexing Long Term Evolution Downlink

In this section, a range profile is produced by matched filtering with an FDD LTE$_e$ DL signal comprised of random “user” data chosen from a 64 QAM constellation ($s_{ref}[n] = s[n]$). From the FDD LTE$_e$ DL SAF shown in Figure 3.1, the only prominent peak outside the actual target occurs at a range of $\frac{cT_u}{2} \approx 10$ km from the target due to the CP. Consistent with the result shown in Figure 3.1, these peaks are clearly evident in the range profile shown in Figure 3.16.
Figure 3.16: Range profile generated at the center pulse of the receive aperture with $s_{\text{ref}}[n] = s[n]$. The peaks from the FDD LTE e DL CP appear at a range of $\frac{cT_u}{2} \approx 10$ km relative to the actual target.

3.7.2 Frequency Division Duplexing Long Term Evolution Downlink

Synchronization Signal

In this section, additional peaks resulting from FDD LTE e DL synchronization signal structure (PSS and SSS) are illustrated. As explained in Section 3.2.3, FDD LTE e DL synchronization signals occur every 66 symbols. Thus, as shown in Figure 3.17(a), the most prominent peaks outside the actual target occur from the repetitiveness of the synchronization signals ($\frac{c66T_{\text{sym}}}{2} \approx 824$ km). An additional peak also occurs at a range corresponding to a differential time delay of $T_u$ ($\frac{cT_u}{2} \approx 10$ km), arising from the CP as shown in Figure 3.17(b). As expected, the most prominent peaks outside the actual target illustrated in Figures 3.17(a) and 3.17(b) may be attributed to the structure of FDD LTE e DL synchronization signals.
(a) Range profile generated at the center pulse of the receive aperture with $s_{\text{ref}}[n] = s_{\text{PSS}}[n] + s_{\text{SSS}}[n]$. The most apparent peaks aside from the actual target are a result of the synchronization signals repeating every $66T_{\text{sym}}$.

(b) Zoomed in range profile generated at the center pulse of the receive aperture with $s_{\text{ref}}[n] = s_{\text{PSS}}[n] + s_{\text{SSS}}[n]$. The additional peaks caused by the CP occur at a range of $\frac{L}{2} \approx 10$ km relative to the target.

Figure 3.17: Range profile generated by partially matched filtering with an FDD LTE DL synchronization signal ($s_{\text{ref}}[n] = s_{\text{PSS}}[n] + s_{\text{SSS}}[n]$).

### 3.7.3 8k Digital Video Broadcast

Similar to the previous section’s results, a range profile is used to illustrate additional peaks arising from the signal structure of an 8k DVB signal, comprised of random “user” data chosen from a 64 QAM constellation. Consistent with the results derived in Section 3.3.1, the most pronounced peaks aside from the target are a result of the CP and occur at a distance of $\approx 134$ km with respect to the target (see Figure 3.18). In addition, as witnessed in the 8k DVB SAF, less significant peaks appear because of 8k DVB pilot’s structure and properties (see Figure 3.5). These same peaks are illustrated in Figure 3.18.

### 3.7.4 8k Digital Video Broadcast Scattered and Continual Pilots

Here, additional peaks resulting from the scattered and continual pilots of an 8k DVB signal are illustrated in a range profile ($s_{\text{ref}}[n] = s_{\text{SP}}[n] + s_{\text{CP}}[n]$). Both pilots are considered together because of the poor URR attained exploiting only the continual pilots (see Figure 3.10). From Figure 3.19, it is clear additional peaks outside the actual target arise from
Figure 3.18: Range profile generated at the center pulse of the receive aperture with \( s_{\text{ref}}[n] = s[n] \). The most prominent additional peaks from 8k DVB signal structure occur at a distance of \( \frac{CT_u}{2} \approx 134 \text{ km} \) relative to the target due to the CP. Other, less distinguishable peaks are also present and are a result of 8k DVB pilots as explained in Sections 3.3.2 and 3.3.3.

The scattered and continual pilots. The most apparent pilot peaks correspond to a differential time delay of \( T_u \) (see Equations (3.6) and (3.7)). The least distinguishable additional pilot peaks in Figure 3.19 are solely a result of continual pilot ISI, as demonstrated by the continual pilot CAF result shown in Figure 3.10 (i.e., peaks arise just prior to time delay multiples of \( \frac{T_u}{3} \)).

### 3.7.5 Transmission Mode I Digital Audio Broadcast

As performed for previous additional results, a range profile is used to illustrate the effects of TxI DAB signal structure on matched filter output. The range profile produced exploiting a TxI DAB signal comprised of random “user” data chosen from a DQPSK constellation, for the geometry described in Figure 4.2, is shown in Figure 3.20. As witnessed in the results in Section 3.4.1, the only prominent, additional peak outside the actual target arises from the CP at a differential delay of \( T_u \). It may also be expected that another prominent
Figure 3.19: Range profile generated at the center pulse of the receive aperture with $s_{\text{rcf}}[n] = s_{\text{SP}}[n] + s_{\text{CP}}[n]$. The most prominent additional pilot peaks occur at range multiples of $\frac{cT_u}{6} \approx 45$ km relative to the target. Other, less distinguishable peaks are also present and are a result of continual pilot ISI as shown in Figure 3.10.

A peak would occur from the periodic nature of TxI DAB’s SS. However, this peak occurs at a range of $\frac{cTT_{\text{sym}}}{2} \approx 14382$ km relative to the target, which is outside our area of interest and not shown in Figure 3.20.

### 3.7.6 Transmission Mode I Digital Audio Broadcast Synchronization

#### Symbol

The range profile produced by partially matched filtering with a TxI DAB SS is shown in Figure 3.21. The result is similar, in the sense additional peaks occur at the same intervals, to the matched filter result shown in Figure 3.20. Additional peaks, aside from the target, are apparent at a distance of $\frac{cT_u}{2} \approx 150$ km relative to the target from the CP (see Figure 3.21). As mentioned in Section 3.7.5, another additional peak is expected at a distance of $\frac{cTT_{\text{sym}}}{2} \approx 14382$ km relative to the target but is not shown in Figure 3.21, as this extent
Figure 3.20: Range profile generated at the center pulse of the receive aperture with $s_{\text{ref}}[n] = s[n]$. The only apparent additional peak outside the actual target shown is a result of the TxI DAB CP (i.e., distance of $\frac{cT_u}{2} \approx 150$ km relative to the target).

Figure 3.21: Range profile generated at the center pulse of the receive aperture with $s_{\text{ref}}[n] = s_{SS}[n]$. The only apparent, additional peak outside the actual target arises from the TxI DAB CP (i.e., distance of $\frac{cT_u}{2} \approx 150$ km relative to the target).
exceeds our area of interest.

3.8 Chapter Conclusion

In this chapter, attainable performance derived from FDD LTE\textsubscript{c} DL, 8k DVB, and TxI DAB waveforms was evaluated by examining each respective signal’s SAF and CAFs. In addition, the derived results were used to provide insight into how the structure of the matched filter waveform affects passive bistatic SAR image output exploiting FDD LTE\textsubscript{c} DL, 8k DVB, or TxI DAB signals of opportunity.
Chapter 4

Link Budget Analysis

In this chapter, link budget analysis is performed for passive bistatic SAR imaging systems exploiting FDD LTE_e DL, 8k DVB, and TxI DAB signals of opportunity. Discussion begins with justification for only evaluating the performance of the surveillance channel of a PBR system. Then, relatively conservative values are established for the parameters of the bistatic radar range equation assuming a passive bistatic SAR imaging scenario. Using the established conservative values and proposed imaging scenario, detection and coverage ranges exploiting FDD LTE_e DL, 8k DVB, and TxI DAB signals of opportunity are predicted. The chapter concludes with simulated passive bistatic SAR imaging results produced exploiting an 8k DVB signal in the presence of additive white Gaussian noise (AWGN), confirming the predicted performance for the proposed imaging scenario.

4.1 Performance Prediction for Passive Bistatic Synthetic Aperture Radar Imaging

In Chapter 3, the waveform properties of FDD LTE_e DL, 8k DVB, and TxI DAB transmissions were presented. While the waveform properties of the transmitted signal are impor-
tant, equally important is performance prediction of coverage and detection ranges [12]. To gain a better understanding of operational constraints and attainable performance, analysis of the bistatic radar range equation, (2.6), is required. Equation (2.6) has several degrees of freedom producing an impractical amount of scenarios to be analyzed. Therefore, this work only seeks to elaborate upon the facets of (2.6) a radar engineer must consider to accurately assess performance for passive bistatic SAR imaging scenarios. Only performance of the surveillance channel is analyzed (see Figure 2.1), as this channel is assumed to be the limiting case between the surveillance and direct-path channels. The basis for this assumption follows from the fact that for the surveillance channel, the received signal propagates over a longer distance (assumed to not be line-of-sight) compared to the signal received in the direct-path channel (assumed to be line-of-sight). In addition, it is assumed that each receive channel has an appropriate directivity pattern to mitigate any interference from the other channel. For these reasons, only performance of the surveillance channel is evaluated.

In the following, parameters of the bistatic radar range equation are discussed and typical values are established for predicting performance of a proposed passive bistatic SAR imaging scenario.

4.1.1 Transmitter

Generally, broadcast/terrestrial systems have substantial transmit power to service inefficient receiving antennas that are not line of sight [2]. These substantial power levels are the transmit power levels assumed for FDD LTEe DL, 8k DVB, and TxI DAB signals from [29–31] (see Table 2.2). Naturally, part of the system losses shown in (2.6) are attributed to the transmit antenna. However, the typical transmit power levels described in Table 2.2 already include antenna losses. Therefore, the level of expected system losses used in this work are tempered accordingly. This point is reiterated in Section 4.1.2 for establishing a reasonable value for receive antenna and signal processing losses.
Another property dictated by the transmitter is the carrier frequency of the signal. Here, the carrier frequencies assumed for the considered FDD LTE, DL, 8k DVB, and TxI DAB signals are those shown in Table 2.2. These frequencies fall within the VHF and UHF bands and are convenient for this work as the radio frequency (RF) environment has been characterized for these bands [2].

For passive bistatic SAR imaging exploiting continuous wave (CW) signals of opportunity, the radar engineer must be aware of the pulse repetition frequency (PRF) induced by partitioning the echo returns for the collection interval into pseudo pulses, where pseudo pulses are meant to refer to the particular partition of the return considered for processing of one pulse of the synthetic aperture. For clarity throughout the remainder of this work, pseudo pulses is used in turn of pulses to make clear that we are considering exploitation of CW waveforms, and that the collected returns for a collection interval are partitioned to construct pseudo pulses. As dictated by Nyquist criterion, the PRF of the pseudo pulses of the synthetic aperture must exceed the Doppler bandwidth [2]. Thus, appropriate receiver flight path planning and signal processing steps (i.e., removing the along-track Doppler bandwidth [2]) must be performed to ensure sufficient sampling in crossrange is attainable.

In general, the radar engineer has no control of the transmitter’s properties for non-cooperative passive radar applications. However, the radar engineer may have available a set of transmitters. Consequently, the radar engineer may have some degrees of freedom for the transmitter’s properties based on emitter selection [26].

4.1.2 Receiver

According to [2], covert PBR receive antennas are typically limited to smaller antennas such as Yagi or compact circular arrays, enabling PBR systems to remain mobile and cost effective. These smaller antennas typically offer modest gains ranging from 6-10 dBi [2]. To conservatively predict performance, a receive antenna gain of 6 dBi is assumed for this research.
As mentioned in Section 4.1.1, the transmit power level assumed for FDD LTE, 8k DVB, and TxI DAB transmissions already include transmit antenna losses. Therefore, the overall system losses shown in (2.6) are reduced accordingly. This research assumes system losses of 5 dB, equivalent to that utilized in [12], ascribed solely to receive antenna and signal processing losses.

From (2.6), it may be seen that SNR is proportional to coherent integration time. A common approximation for the limit of coherent integration time is

\[ T_{i,\text{max}} = \sqrt{\frac{\lambda}{a_r}}, \]  

(4.1)

where \( a_r \) is the maximum radial acceleration of a target [12]. Typically for SAR imaging, targets are assumed to be stationary [2, 40]. However, due to motion of the transmitter or receiver Doppler may be induced on target returns. Recalling the system configuration assumed for this work (see Figure 2.1), it may be seen that, at each instance along the aperture, the Doppler shift induced by the receiver changes, thereby resulting in relative acceleration. Here, it is assumed an appropriate flight path is flown so that a modest coherent integration time of 20 ms\(^1\) is attained. This coherent integration time is assumed so that at every pseudo pulse along the synthetic aperture at least one complete symbol of each type of signal may be received (see Table 2.2). Moreover, a coherent integration time of 20 ms enables range profiles with a sufficient scene extent to be produced at each pseudo pulse along the synthetic aperture.

In [2], it is stated that a minimum SNR of 13 dB provides a probability of detection of \( \approx 0.5 \) and probability of false alarm of \( \approx 10^{-6} \) for all Swerling target fluctuation models. Because values for the bistatic RCS have such a large degree of variability and bistatic RCS is not the primary topic of this research, the same minimum SNR of 13 dB (total SNR for the produced SAR image) is assumed here for performance prediction. Although

\(^1\)See Appendix A for the calculation ensuring the PRF exceeds the spotlighted Doppler bandwidth.
a constant RCS is assumed, a brief discussion is provided in Section 4.1.4 on the variability of the bistatic RCS and its possible effects on performance.

As stated in Section 2.7, a synthetic aperture is formed by coherently combining pseudo pulses. Thus, the minimum SNR is subject to pseudo pulse integration across the synthetic aperture, where the minimum SNR per pseudo pulse is equivalent to the total minimum SNR divided by the number of coherent pseudo pulses. (Pseudo pulse integration is not to be confused with coherent integration time, as coherent integration time limits the duration that may be referred to as one pseudo pulse of the synthetic aperture.) As will be seen in Section 4.2, a configuration is assumed in which 501 pseudo pulses are received, thereby reducing the minimum SNR per pseudo pulse to $\approx -14$ dB.

### 4.1.3 Environment

PBR systems experience thermal and environmental noise similar to active radar systems. Although, the main form of interference for the surveillance channel of PBR systems is line-of-sight signal interference from the transmitter [12]. To increase the effectiveness of a PBR system, line-of-sight signal interference must be mitigated. Mitigation of interference may be achieved using any of the following techniques: physical shielding, Doppler processing, high gain antennas, adaptive beamforming, or adaptive filtering [12]. Even after applying techniques to suppress the effects of the line-of-sight signal interference, a receiver noise figure of $F = 25$ dB is not unrealistic for an operating frequency in VHF or UHF bands [12]. Thus, for performance prediction a noise figure of $F = 25$ dB is assumed.

### 4.1.4 Bistatic Radar Cross-Section

One of the key advantages of bistatic radar is the ability to gain aspect diversity, as many targets’ RCS value fluctuates with aspect angle. In particular, for SAR mapping applications (e.g., terrain mapping) it may be advantageous to induce poorer resolution (i.e., poorer
signal to clutter ratio) by increasing the bistatic angle in exchange for an improved bistatic RCS, where the target is considered to be a clutter cell. This trade-off is made apparent in Figure 4.1, where it is assumed the bistatic RCS is equivalent to the area of a resolution cell scaled by the bistatic coefficient per unit area of soybean foliage provided in [1] (i.e., target is considered to be one isolated resolution cell comprised of soybean foliage).

For both results shown in Figure 4.1, the transmitter and receiver have a standoff distance of $R_T = R_R = 5.75$ km. In Figure 4.1, it is shown the operational range expands (i.e., the constant SNR contour expands) as a result of an increased bistatic RCS. The increase in operational range is a consequence of the receiver moving closer to the specular ridge point (the scattering coefficient at a bistatic angle of $160^\circ$ for the considered experiment.

**Figure 4.1: Illustration of the possible trade-off between resolution and bistatic RCS.**

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</table>
described in [1]) of the scattering cell [1]. The parameters selected for (2.6) to produce the results shown in Figure 4.1 are not meant to mimic a real-world scenario. Instead, the parameters were chosen to clearly demonstrate that a potential trade-off between bistatic angle and bistatic RCS may exist. Determination of the bistatic RCS for a given target is a rich topic in itself but is not discussed in further detail here. Instead, for evaluation of (2.6), a constant bistatic RCS of 1 m² is assumed.

4.2 Proposed Real-World Scenario

For the following analysis, it is important the reader understand that there is a large degree of flexibility in the values selected to evaluate (2.6) and that (2.6) must be re-evaluated for any given scenario or system. Nonetheless, using the bistatic radar range equation and values established in Section 4.1, a potential real-world passive bistatic SAR imaging scenario exploiting FDD LTE_e DL, 8k DVB, and TxI DAB signals is proposed and evaluated in the following.

With SAR imaging, resolution capabilities are of extreme importance. To maintain adequate resolution capabilities, flight path planning must be performed to attain an appropriate bistatic angle. This work proposes an over-the-shoulder approach (see Figure 4.2) which provides pseudomonostatic performance ($\beta \approx 6.3^\circ$) [2]. The transmitter is assumed to have a height of 45 m and a ground plane stand-off distance to scene center of 5 km. The receiver is assumed to be an airborne platform moving at a velocity of 150 m/s parallel to the x-axis. Moving at this velocity and assuming a coherent collection interval of approximately 10 s, enables 501 pseudo pulses to be collected (each pseudo pulse has a duration of 20 ms for each communication signal standard). The receive platform has a ground plane standoff distance of 25 km relative to scene center and 20 km relative to the transmitter. The altitude of the flight path is assumed constant at 3 km. The predicted
coverage areas for the respective signal standards are shown in Figure 4.3 with a potential scene overlaid. From Figure 4.3, it may be seen that even for the pessimistic values established in Section 4.1, an operable region exists for each transmission type. (The transmitter and receiver form one oval of Cassini with a minimum SNR of $-14$ dB large enough to contain a scene of interest as illustrated in Figures 4.3(d)–4.3(f).) Although, for the case exploiting an FDD LTE$_c$ DL signal, the scene of interest is limited to a relatively small area located around the transmit site. This relatively limited operable region may be attributed to the considered FDD LTE$_c$ DL signal having the shortest wavelength and lowest transmit power level. While the considered FDD LTE$_c$ DL signal offers the smallest operable region, the signal offers the finest resolution capabilities. Thus, depending upon the specific SAR imaging application, the radar engineer may have to weigh the trade-offs between performance region and resolution properties.

From the results in Figure 4.3, it is not readily apparent the quality (in the sense of background noise) produced SAR images will have. Therefore, in the following section, simulated passive bistatic SAR images are produced using the proposed configuration.
Figure 4.3: Performance regions for a proposed over-the-shoulder configuration exploiting FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB signals of opportunity.
4.3 Application to Passive Bistatic Synthetic Aperture Radar Imaging

In the following, passive bistatic SAR image quality is demonstrated assuming a scene of interest located on two of the constant, per pseudo pulse SNR contours, $-14$ and $-17$ dB (total SNR of 13 and 10 dB, respectively), shown in Figure 4.3. The geometry assumed for the simulated scenario is shown in Figure 4.2 with a target located at scene center. For the simulated results, an 8k DVB signal is assumed to be the signal of opportunity. In addition, it is assumed one 8k DVB symbol is received for each pseudo pulse of the synthetic aperture. Furthermore, to produce phase history data, it is assumed direct-path channel data is processed and used to perfectly construct the ideal transmitted signal for matched filtering with the full signal.

Figure 4.4 depicts SAR images with a total SNR of 10 and 13 dB, produced by matched filtering with the full 8k DVB signal. Qualitatively, it appears a total SNR of 13 dB is adequate to produce SAR images with a distinguishable target and tolerable background noise (see Figures 4.4(b) and 4.4(d)). However, the same cannot be said for the SAR image produced with a total SNR level of 10 dB, as the target is not readily discernible (see Figures 4.4(a) and 4.4(c)). In either case, whether 10 or 13 dB of total SNR is considered, Figure 4.4 only provides a qualitative assessment of the quality of SAR images produced with the respective total SNR. In general, more extensive analysis is needed to determine a suitable minimum SNR for a particular application based upon desired performance criteria (e.g., minimum SNR to provide a desired probability of detection and probability of false alarm). In addition, as mentioned in Section 3.6, the effects of clutter and constructively combining sidelobes may dominate the effects of thermal noise and present a more pressing issue.

The results in Figure 4.4 not only confirm the application of pseudo pulse integra-
(a) SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 10 dB.

(b) SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 13 dB.

(c) Zoomed in SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 10 dB.

(d) Zoomed in SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 13 dB.

Figure 4.4: Demonstration of SAR image quality produced by matched filtering with an 8k DVB signal ($s_{\text{ref}}[n] = s[n]$). Each SAR image is normalized to the maximum value in the scene.
tion for evaluating (2.6) for passive bistatic SAR imaging applications, but also suggest that real-world exploitation of FDD LTE, DL, 8k DVB, and TxI DAB signals for passive bistatic SAR imaging is feasible by adhering to predicted performance regions.

4.4 Additional Results

In the proceeding, SAR image quality is demonstrated for matched filtered schemes exploiting FDD LTE, DL and TxI DAB signals of opportunity. The geometry assumed for the analysis is that shown in Figure 4.2.

4.4.1 Frequency Division Duplexing Long Term Evolution Downlink

For SAR images generated by matched filtering with a full FDD LTE, DL frame (120 symbols), total SNR levels of 10 and 13 dB are considered. The SAR images produced by matched filtering are shown in Figure 4.5. As witnessed with 8k DVB SAR images, 13 dB appears to be an adequate total SNR to produce distinguishable targets with tolerable background noise (see Figures 4.5(b) and 4.5(d)). Although, as noted in the previous section, a more detailed analysis is required for determining a minimum SNR that is suitable for desired performance criteria.

4.4.2 Transmission Mode I Digital Audio Broadcast

Similar to previous analysis, a total SNR of 13 dB is assumed for matched filtering with a TxI DAB signal. The resultant SAR images exploiting a TxI DAB signal are shown in Figure 4.6. Consistent with previously shown results, 13 dB appears to yield distinguishable targets with acceptable background noise (see Figures 4.6(b) and 4.6(d)). However, to reiterate again, further examination is needed to determine an appropriate total SNR based
(a) SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 10 dB.

(b) SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 13 dB.

(c) Zoomed in SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 10 dB.

(d) Zoomed in SAR image produced by matched filtering ($s_{\text{ref}}[n] = s[n]$), yielding a total SNR of 13 dB.

Figure 4.5: Demonstration of SAR image quality produced by matched filtering with an FDD LTEu DL signal ($s_{\text{ref}}[n] = s[n]$). Each SAR image is normalized to the maximum value in the scene.
4.5 Chapter Conclusion

In this chapter, components of the bistatic radar range equation were discussed in detail
and justification was provided for establishing pessimistic values for link budget analysis.
Using these values, detection and coverage regions were predicted for a proposed passive
bistatic SAR imaging scenario. Then, simulated SAR images were produced employing
matched filtering for phase history generation. The simulated SAR image quality for the
processing scheme advocate exploitation of FDD LTE, 8k DVB, and TxD DAB signals
of opportunity for passive bistatic SAR imaging is feasible.
Figure 4.6: Demonstration of SAR image quality produced by matched filtering with a TxI DAB signal ($s_{\text{ref}}[n] = s[n]$). Each SAR image is normalized to the maximum value in the scene.
Chapter 5

Passive Synthetic Aperture Radar

Imaging Results

In the following, an experimental setup mimicking the proposed passive bistatic SAR imaging scenario presented in Chapter 4 is considered for small-scale experimental testing. The chapter begins with a description of the experimental equipment, the setup used for testing, and the software employed for processing. Then, small-scale results produced by exploiting signals structured similar to FDD LTEc DL, 8k DVB, and TxI DAB signals of opportunity are shown and discussed. Experimental results are produced using a full AMF and partial AMF, indicating the utility of the matched filtered processing scheme considered in this research for passive bistatic SAR imaging.

5.1 Experimental Setup

For this research, there is not means for testing the scenario proposed in Chapter 4. Instead, a small-scale setup mimicking the proposed scenario is used to exhibit the utility of the matched filtering scheme considered in this research for passive bistatic SAR imaging.
The experimental system employed for testing was created in the Radar Instrumentation Laboratory (RAIL) at the Air Force Institute of Technology and has been used in previous passive bistatic SAR imaging experiments [11]. In the proceeding section, the hardware used for the experimental setup is described along with the scene setup and processing schemes employed.

### 5.1.1 Hardware Setup

The hardware setup utilized for experimental testing is depicted in Figure 5.1. The hardware setup consists of the following devices: Tektronix arbitrary waveform generator (AWG) 7102, Mini-Circuits ZKL-2R7+ coaxial amplifier, AirMax 2G-16-90 sector antennas, RAIL’s linear track, Mini-Circuits ZRL-3500 low noise amplifier (LNA), and Tektronix TDS6124 digital storage oscilloscope (DSO).

First, signals with an increased bandwidth and carrier frequency, structured similar to FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB transmissions, are generated using MATLAB®. Then, the signals are loaded and generated on the AWG ready for transmission. Upon initiating a transmit pulse, the signal is passed through the coaxial amplifier to the sector antenna illuminating a scene of interest. The return energy from the scene is observed using a sector antenna. The echo return is passed through the LNA and onto the DSO for the data to be saved. This procedure generates data constituting one pulse of the synthetic aperture and is repeated multiple times, as the receive antenna is moved along the length of RAIL’s linear track, creating a synthetic aperture. Once all pulses are collected, the data is transferred from the DSO to a portable storage device for post-processing to be performed using MATLAB®.
5.1.2 Software Setup

To imitate a passive scenario, signals with an increased bandwidth and carrier frequency, structured on FDD LTE\textsubscript{e} DL, 8k DVB, and TxI DAB standards are generated containing random “user” data. The properties of the modified signals are shown in Table 5.1. From the descriptions in Sections 2.4–2.6, it is clear that in general the considered communication signal standards do not utilize the same bandwidth or operate at the same carrier frequency. However, for testing, signals generated with a bandwidth of $B = 300$ MHz and a carrier frequency $f_c = 2.5$ GHz are employed.

Ideally, testing would be performed utilizing simulated FDD LTE\textsubscript{e} DL, 8k DVB, and TxI DAB signals at an appropriate carrier frequency and with an appropriate bandwidth as specified by the standards. However, for the configuration of the scene used in this work, this would require a testing area with a downrange scene extent on the order of 300 m. This extent exceeds not only our available testing area, but also requires significant power. Therefore, due to power constraints and limited operating region, signals with a bandwidth of 300 MHz (i.e., 0.5 m range resolution) are necessary to produce a sufficient image of
Table 5.1: Parameters of signals, structured similar to FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB signals, utilized for experimental testing.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>LTE</th>
<th>DVB</th>
<th>DAB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating mode</td>
<td>-</td>
<td>FDD</td>
<td>8k</td>
<td>TxI</td>
</tr>
<tr>
<td>Carrier frequency (GHz)</td>
<td>f&lt;sub&gt;c&lt;/sub&gt;</td>
<td>2.5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Data symbol duration (µs)</td>
<td>T&lt;sub&gt;u&lt;/sub&gt;</td>
<td>6.8</td>
<td>27.3</td>
<td>6.8</td>
</tr>
<tr>
<td>CP Ratio</td>
<td>∆&lt;sub&gt;g&lt;/sub&gt;</td>
<td>1/4</td>
<td>63/256</td>
<td>63/256</td>
</tr>
<tr>
<td>Symbol duration (µs)</td>
<td>T&lt;sub&gt;sym&lt;/sub&gt;</td>
<td>8.5</td>
<td>34.1</td>
<td>8.5</td>
</tr>
<tr>
<td>Subcarrier separation (kHz)</td>
<td>∆f</td>
<td>146.5</td>
<td>36.6</td>
<td>146.5</td>
</tr>
<tr>
<td>Data subcarriers</td>
<td>N&lt;sub&gt;a&lt;/sub&gt;</td>
<td>1320</td>
<td>6816</td>
<td>1536</td>
</tr>
<tr>
<td>Subcarriers per symbol</td>
<td>N</td>
<td>2048</td>
<td>8192</td>
<td>2048</td>
</tr>
<tr>
<td>Effective bandwidth (MHz)</td>
<td>B&lt;sub&gt;eff&lt;/sub&gt;</td>
<td>193.51</td>
<td>249.65</td>
<td>225.15</td>
</tr>
<tr>
<td>Bandwidth (MHz)</td>
<td>B</td>
<td>300</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Symbols&lt;sup&gt;1&lt;/sup&gt;</td>
<td>-</td>
<td>23</td>
<td>5</td>
<td>23&lt;sup&gt;2&lt;/sup&gt;</td>
</tr>
<tr>
<td>Signal duration (µs)</td>
<td>-</td>
<td>196</td>
<td>170.7</td>
<td>196.3</td>
</tr>
<tr>
<td>Modulation scheme</td>
<td>-</td>
<td>64 QAM</td>
<td>DQPSK</td>
<td></td>
</tr>
</tbody>
</table>

1 Meeting Nyquist criterion, the maximum collection interval is limited to 200 µs due to memory constraints.

2 Includes a null symbol with duration 8.85 µs.

A small scene (< 20 m downrange). To structure experimental signals after FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB standards, signals constructed according to the descriptions in Sections 2.4–2.6 are used but with an increased subcarrier separation of ∆f = B<sub>eff</sub> / N. By increasing the subcarrier separation, the same number of subcarriers, N, may be used for each of the experimental signals as specified by the respective communication signal standard. Moreover, the placement of deterministic signaling features may be maintained. Also, in general, FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB transmissions do not typically operate on the same frequency bands. However, for this work, each of the experimental signals is assigned the same carrier frequency due to hardware constraints (e.g., antenna operating frequency range of 2.3-2.7 GHz). As previously alluded to, for this research, there is no means available for performing passive bistatic SAR imaging experiments employing actual FDD LTE<sub>e</sub> DL, 8k DVB, and TxI DAB signals. Instead, within the constraints of the utilized hardware and proposed experiment, signals structured as closely as possible to
resemble FDD LTE, DL, 8k DVB, and TxI DAB signals are employed. Figure 5.2 depicts the processing block diagram performed during post-processing to form OFDM phase history data from collected data. The procedures performed for processing are based upon the OFDM phase history model rederived in Section 2.7.1. First, the received data are shifted to baseband by mixing the data with an exponential, delayed to scene center, oscillating at the carrier frequency. A LPF is applied to the data, and then, sampling rate conversion is performed to convert the sampling rate of the received data to the signal’s baseband sampling rate. Once sampling rate conversion is completed, a fast Fourier transform (FFT) is taken. Then, matched or partially matched filtering is performed for each segment duration and averaged, as denoted in Figure 5.2, to generate phase history data. Recall, the segment duration is an arbitrary segment length or window for which matched filtering is performed (see Figure 2.13). The phase history data is then used with the CBP algorithm to generate an image of the scene of interest.
5.1.3 Scene Setup

The setup used for experimental testing is shown in Figure 5.3. The transmitter is stationary as is typical for exploitation of broadcast/terrestrial signals of opportunity. For data collection, a move-stop-move process is needed to successfully collect data for post-processing (i.e., dictated by hardware limitations, a pulsed system is employed to successfully gather and store data for post-processing). The receiver is moved across approximately a 2 m aperture in 6 cm increments, producing 33 pulses. At each position along the receive aperture, the full signal duration (see Table 2.2) of each signal is transmitted and received. Based upon the setup of the transmitter and receiver, the bistatic angle is $\beta \approx 3^\circ$, the azimuthal extent of the aperture is $\Delta \phi \approx 4.22^\circ$, and the bistatic elevation look angle is $\theta_b \approx 12.29^\circ$ according to the definitions in [2]. Based on the geometry of the experimental setup and signal parameters used for the transmitted signals, the range and crossrange resolutions are 0.5 and 0.83 m, respectively, as calculated using Equations (2.13) and (2.14). Within the scene of interest, four 0.4 m$^2$ plate targets are setup as shown in Figure 5.3(a).
5.2 Experimental Results

Much of the material presented in the following discussion is pulled from [45]. Experimental SAR images produced using the setup described in the previous section are shown in Figures 5.4 and 5.5. The images are produced using the matched filtering scheme discussed in Section 2.7 and the CBP algorithm. Each of the images is normalized to the peak reflection within the scene and shown on a dBsm scale.

Figure 5.4 consists of images produced using a full AMF (i.e., $S_{\text{ref}}[n]$ in (2.25) is the baseband, delayed to scene center, full, ideal transmitted signal) with a segment duration of $T_{\text{sym}}$. Although Figures 5.4(a)–5.4(c) are generated using different experimental signals, the produced images are very similar in the sense the targets are clearly distinguishable. This is expected as each experimental signal has the same bandwidth and is transmitted at the same carrier frequency (i.e., each signal exhibits the same range and crossrange resolution). However, some small differences appear in Figures 5.4(a)–5.4(c). These differences may be attributed to the random nature of the signals (i.e., the “user” data consists of randomly generated complex modulation values) and the differences in their structure and implementation (see Sections 2.4–2.6 and Table 2.2). Some of the small differences include varying sidelobe levels around crossrange, range coordinate pairs (1 m, 0 m), (1 m, -4 m), (1 m, 2 m), and (0 m, -1 m). Figure 5.5 depicts SAR images produced using a
Figure 5.5: Experimental SAR images produced using a partial AMF with a segment duration of $T_{sym}$.

From Figure 5.5, Figure 5.5(b) most closely resembles the SAR images produced using a full AMF. The closer resemblance may be attested to the structure and properties of 8k DVB’s pilot tones. 8k DVB’s pilot tones are contained within every symbol and are transmitted 2.5 dB higher than a typical BPSK modulation scheme. Thus, for each iteration of the AMF (i.e., each 1D range profile) the partial reference signal has a strong enough correlation with the received signal to produce distinguishable targets. The same strong correlation was also evidenced in Chapter 3, as the CAF zero delay, zero Doppler peaks for 8k DVB features were higher than those witnessed for FDD LTEe DL and TxI DAB CAF results (see Tables 3.1–3.3). As explained in Chapter 3, the strong correlation is a result of the 2.5 dB power increase and the density at which the pilots are inserted in the signal. Each segment is averaged, mitigating the effects of noise and clutter. The mitigation of noise and clutter is evident in Figure 5.5 as Figure 5.5(b) has the least background scattering compared to Figures 5.5(a) and 5.5(c). On the other hand, Figures 5.5(a) and 5.5(c) exhibit more significant background scattering. Based upon the zero delay, zero Doppler peaks
for FDD LTE<sub>e</sub> DL and TxD DAB CAF results, the higher background scattering compared to the 8k DVB partially matched filtered SAR image is expected (see Tables 3.1–3.3). The relatively high background may be attributed to the structure and properties of FDD LTE<sub>e</sub> DL’s and TxD DAB’s known features. FDD LTE<sub>e</sub> DL’s features are contained within multiple symbols but are not transmitted at a higher power such as 8k DVB’s pilot tones. In addition, FDD LTE<sub>e</sub> DL’s features do not occupy a full symbol such as TxD DAB’s SS. Thus, FDD LTE<sub>e</sub> DL’s known features do not correlate as well with the received signal to produce clearly distinguishable targets within each 1D range profile. The poorer correlation is evident from Figure 5.5(a) as the background reflectivity along the downrange dimension is significantly higher than that shown in Figures 5.5(b) and 5.5(c).

While Figure 5.5(c) exhibits higher background than Figure 5.5(b), it is less extreme than that shown in Figure 5.5(a). The improved quality is a result of TxD DAB’s SS occupying a full symbol. Thus, aside from the noise and distortion of the channel, the SS perfectly correlates with its echo return symbol. On the other hand, FDD LTE<sub>e</sub> DL’s <em>a priori</em> features never exhibit perfect correlation because the features do not occupy all active subcarriers.

As discussed in Section 5.1.2, each of the communication standards’ signals considered in this research do not generally operate at the same carrier frequency, utilize the same bandwidth, or are employed in a pulsed system. Thus, the results shown in Figures 5.4 and 5.5 are not indicative of the resolution capabilities offered by the examined signals and do not illustrate the levels of ISI expected when exploiting a CW signal of opportunity.

### 5.2.1 Demonstration of “Actual” Resolution Properties

As indicated in Section 5.1.2, Figures 5.4 and 5.5 do not illustrate real-world resolution capabilities attainable exploiting FDD LTE<sub>e</sub> DL, 8k DVB, or TxD DAB signals. However, target signatures within Figures 5.4 and 5.5 may be scaled to demonstrate the “actual” resolution offered by the signals. Figure 5.6 depicts SAR images of a target’s return produced using the full AMF with a segment duration of $T_{sym}$ and respective signal of opportunity.
Table 5.2: Resolution values for scaled target signatures.

<table>
<thead>
<tr>
<th>Signal</th>
<th>Downrange Resolution (m)</th>
<th>Crossrange Resolution (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FDD LTE_e DL</td>
<td>7.56</td>
<td>2.84</td>
</tr>
<tr>
<td>8k DVB</td>
<td>19.70</td>
<td>4.36</td>
</tr>
<tr>
<td>TxI DAB</td>
<td>97.53</td>
<td>43.96</td>
</tr>
</tbody>
</table>

(a) Scaled target signature using an FDD LTE_e DL signal with a downrange and crossrange resolution of 7.56 m and 2.84 m, respectively.

(b) Scaled target signature using an 8k DVB signal with a downrange and crossrange resolution of 19.70 m and 4.36 m, respectively.

(c) Scaled target signature using a TxI DAB signal with a downrange and crossrange resolution of 97.53 m and 43.96 m, respectively.

Figure 5.6: Experimental SAR image of a target scaled to illustrate the “actual” downrange and crossrange resolution inherent to the respective signal.

For each of the images in Figure 5.6, the theoretical resolution is indicated with a rectangle. The scaled signal parameters of the signals are those shown in Table 2.2 yielding the resolution capabilities tabulated in Table 5.2. Based upon the signals’ resolution capabilities and the rough baseline values of resolution required for mapping applications shown in Table 5.3 [46], each of the considered signals appear to be viable candidates for mapping applications.

Table 5.3: Resolution required for mapping applications [46].

<table>
<thead>
<tr>
<th>Features to be Resolved</th>
<th>Cell Size(m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coast lines, large cities, and the outlines of mountains</td>
<td>152.4</td>
</tr>
<tr>
<td>Major highways, variations in fields</td>
<td>18.3-30.5</td>
</tr>
<tr>
<td>“Road map” details: city streets, large buildings, small airfields</td>
<td>9.1-15.2</td>
</tr>
</tbody>
</table>

1 The downrange and crossrange resolution are assumed to be equivalent, so the values tabulated for cell size only specify one dimension of a resolution cell.
5.3 Chapter Conclusion

In summary, this chapter included descriptions of the hardware, software, and scene setup used to perform small-scale passive bistatic SAR imaging experiments. Experimental results produced using signals structured similar to FDD LTE_{c} DL, 8k DVB, and TxI DAB signals indicate the utility of the considered matched filtered processing scheme for passive bistatic SAR imaging.
Chapter 6

Conclusion

The objective of this research was to evaluate and apply FDD LTE\textsubscript{e} DL, 8k DVB, and TxI DAB signals of opportunity for passive bistatic SAR imaging. Analysis consisted of understanding the signals’ structure and their inherent waveform properties, performing link budget analysis for a proposed passive bistatic SAR imaging scenario, and conducting small-scale experiments employing signals structured similar to FDD LTE\textsubscript{e} DL, 8k DVB, and TxI DAB signals of opportunity.

The signal structure for each of the standards considered in this work were presented. Based upon the signals’ standard, the effects of signaling features on the SAF result were inferred. The inference of these effects were confirmed by examining each signal’s CAF result. In addition, peaks arising due to signal structure outside an actual target were shown to appear in SAR images at an expected location based upon the waveform used for matched filtering, suggesting that awareness of the AF results enables better interpretation of produced SAR images.

Link budget analysis was performed to determine the feasibility of passive bistatic SAR imaging exploiting FDD LTE\textsubscript{e} DL, 8k DVB, and TxI DAB signals of opportunity. Evaluation of the bistatic radar range equation suggests the application is feasible. Although when exploiting FDD LTE\textsubscript{e} DL, the scene of interest may be limited to a relatively
small area located around the transmit site.

Small-scale experiments were performed to verify the considered matched filtering technique. The technique was successfully employed with signals structured similar to FDD LTEe DL, 8k DVB, and TxI DAB signals to generate passive bistatic SAR images.

6.1 Future Work

A natural extension to the work presented here is testing a full-scale passive bistatic SAR imaging scenario using a controlled environment. Prior to performing such testing, the capability to become synchronous with non-cooperative transmitters transmitting FDD LTEe DL, 8k DVB, or TxI DAB signals must be demonstrated. Once the capability to synchronize is obtained, a receive system must be configured and accurately modeled. In conjunction with configuration of a receive system, an appropriate flight path for the receive platform must be determined. Then, performance prediction may be carried out to determine an operating region which provides the desired performance. Upon predicting performance, full-scale testing may be carried out as proof of concept.

The bistatic case is the simplest form of a multistatic case. This work only considered the bistatic case and may be extended for future research to investigate a more general multistatic scenario. This multistatic scenario may encompass multiple transmitters with different signal types such as FDD LTEe DL and 8k DVB, or multiple transmitters using the same signal type on the same or different frequency bands.
Bibliography


Appendix A

Spotlighted Doppler Bandwidth

In this chapter, calculations are carried out to verify the coherent integration time of 20 ms assumed in Section 4.2 satisfies azimuthal sampling requirements. The calculation is computed using results derived in [2]. For the computation, it is assumed the along-track Doppler induced on echo returns by the receiver are removed (i.e. Doppler induced on a target located at scene center). As shown in Figure 4.2, a stationary transmitter is assumed. Thus, to satisfy azimuthal sampling requirements, the PRF for the synthetic aperture must exceed the Doppler bandwidth induced by motion of the receiver. The spotlighted Doppler bandwidth from [2] is (assuming a transmitter and receiver with constant elevation)

\[
B_d = \frac{2}{\lambda} \left[ v_y \cos(\phi) \sin \left( \frac{\Delta\phi}{2} \right) \cos(\theta) + v_y \cos(\phi) \sin \left( \frac{\Delta\phi}{2} \right) \cos(\theta) - v_x \sin(\phi) \sin \left( \frac{\Delta\phi}{2} \right) \cos(\theta) - v_x \sin(\phi) \sin \left( \frac{\Delta\phi}{2} \right) \cos(\theta) \right] \tag{A.1}
\]

where \(v_y\) and \(v_x\) are the velocities along the y- and x-axis, respectively, \(\phi\) is the line-of-sight angle to the origin, \(\Delta\phi\) is the beamwidth, and \(\theta\) is the depression angle from the platform to scene center (subscript \(t\) and \(r\) denote the transmitter and receiver, respectively). Based upon the assumed geometry shown in Figure 4.2, the largest Doppler induced by the receiver occurs at the first pseudo pulse of the synthetic aperture. To determine the induced Doppler bandwidth using (A.1), the following parameters for the collection geometry are
needed (no parameters are provided for the transmitter since it is stationary).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Units</th>
<th>Parameter</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>$v_y r$</td>
<td>0</td>
<td>m/s</td>
<td>$\Delta \phi'_r$</td>
<td>20</td>
<td>$^\circ$</td>
</tr>
<tr>
<td>$v_x r$</td>
<td>150</td>
<td>m/s</td>
<td>$\theta_r$</td>
<td>83.16</td>
<td>$^\circ$</td>
</tr>
<tr>
<td>$\phi_r$</td>
<td>-88.28</td>
<td>$^\circ$</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Using the previously described parameters, the calculated induced Doppler for each signal type and scenario are tabulated in Table A.1.

Table A.1: Induced Doppler calculations for each respective signal type.

<table>
<thead>
<tr>
<th>Signal Type</th>
<th>Wavelength (m)</th>
<th>PRF$^1$ (Hz)</th>
<th>Induced Doppler (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>FDD LTE$_{e}$ DL</td>
<td>0.41</td>
<td>50</td>
<td>15.06</td>
</tr>
<tr>
<td>8k DVB</td>
<td>0.63</td>
<td>50</td>
<td>9.81</td>
</tr>
<tr>
<td>TxI DAB</td>
<td>6.38</td>
<td>50</td>
<td>0.97</td>
</tr>
</tbody>
</table>

$^1$ Inverse of the coherent integration time or duration of a pseudo pulse.