

Investigation on Multi-dynamic Radar System: A concept for airborne surveillance application

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By

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DECLARATION

This is to certify that the thesis entitled “**Investigation on Multi-dynamic Radar System: A concept for airborne surveillance application**”, submitted by me to the *Indian Institute of Technology Guwahati*, for the award of the degree of DOCTOR OF PHILOSOPHY, is a bonafide work carried out by me under the supervision of Prof. Ratnajit Bhattacharjee and Prof. S. Christopher. The content of this thesis, in full or in parts, have not been submitted to any other University or Institute for the award of any degree or diploma. I also wish to state that to the best of my knowledge and understanding nothing in this report amounts to plagiarism.

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*To my late father who is always alive in
my heart...*





CERTIFICATE

This is to certify that the thesis entitled “**Investigation on Multi-dynamic Radar System: A concept for airborne surveillance application**”, submitted by Jawed Kumar (126102005), a research scholar in the *Deptt of Electronics and Electrical Engg., Indian Institute of Technology Guwahati*, for the award of the degree of DOCTOR OF PHILOSOPHY, is a record of an original research work carried out by him under our supervision and guidance. The thesis has fulfilled all requirements as per the regulations of the institute and in my opinion has reached the standard needed for submission. The results embodied in this thesis have not been submitted to any other University or Institute for the award of any degree or diploma.

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Sincerely
Jawed Kumar



ABSTRACT

During the last few decades, Stealth technology has proven to be one of the most effective approaches to hiding the target from radar systems. The basic concept of low observable is mainly the reduction of Radar Cross Section (RCS) in direction of the receiver. So, for detecting such targets, the radar systems based on bistatic and multi-static concepts attracted substantial attention. More challenges lie when radar platforms are mobile or airborne, the statistical characteristics of radar signals change dynamically. The TX-Rx pair, in this case, is taken on a moving platform such that Airborne radar with multiple receivers on different UAVs. The system geometry of the Bi-static is considered and simulated with all possible radar performance parameters. This technique is used for detecting and tracking stealth aircraft. The geometrical structures are studied with different spacing of radars which is one of the parameters for Bi-static Radar (Baseline distance between transmitter and receiver) to extend the detection coverage over the mono-static radar. The detection performance will further improve in the case of the Transmitter(Tx) and Receiver(Rx) being at the dynamic location, as in the case of all Airborne radars. The simulation is also made with extended features for multi-dynamic scenarios. Waveform design is an important aspect of multi-dynamic radar. In a radar system, the proper choice of waveforms directly affects the performance parameters of the system like signal to noise ratio (SNR), range resolution, Doppler resolution, and ambiguities in range, Doppler, and side-lobe levels. So the proper selection and analysis of the transmitted waveform is the first priority. To analyze the waveforms ambiguity function is used. It is the time-response of a matched-filter to a given finite energy signal, when the echo is received with time delay τ and a Doppler-shift ν with respect to the nominal values expected by the filter. Usually, the target is detected when the processed target echo via signal processing chain crosses the threshold, and the signal processing chain includes target echo process through matched filter across fast time samples, passes through clutter suppression filter, and then coherently added across slow time samples. Matched filters are also used to measure the time delay of an echo signal based on the position of the peak energy concentration of its output. The choice of radar waveform is considered to be perfect if its matched filter response has a narrow main lobe and low side lobes. Waveform design and ambiguity function are the essential tools for the performance analysis of multi-dynamic radar systems in the presence of additive noise

and ambiguities in range and velocity measurements. The ambiguity function properties of a signal depend on the transmitted signal and the geometrical structure of the radar system. The simulation is made towards the OFDM-LFM based waveform to define ambiguity function for multi-dynamic radar.

Multi-dynamic radar (MDR) operational requirement is brought out which includes the essential requirement of time and phase synchronization across the airborne platform for various transmitters and receivers mounted over air-crafts or UAVs. The phase and time synchronization methodology are brought as a design approach for the phase-time synchronization unit (PTSU) to exploit spatial diversity and other features of MDR by processing the target echo across various airborne receivers. The design approach is brought out such that the dependency of GPS signal may not be required, during strategic scenarios the GPS signal from Global Positioning Signal satellites may not be available to keep the all systems across airborne platforms in sync and as well as due to the low update rate of GPS signal, it is difficult to achieve short interval pulse to pulse time-phase stability requirement for radar waveforms.

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Glossary

ACF	Auto Correlation Function
ADC	Analog to Digital converter
AF	Ambiguity Function
ATC	Air Traffic Control
ATD	Automatic Target Detection
AWGN	Additive White Gaussian Noise
BASK	Binary Amplitude Shift keying
CA-CFAR	Cell Averaging Constant False Alarm Rate
CAF	Complex Ambiguity Function
CDF	Cumulative Distribution Function
CFAR	Constant False Alarm Rate
CNR	Clutter to Noise Ratio
CPI	Coherent Pulse Interval
CRLB	Cramer Rao Lower Bound
CW	Continuous Wave
DAC	Digital to Analog Converter
DoF	Degree of freedom
DPCA	Displaced Phase Center Antenna
DSP	Digital Signal Processing
EVD	Eigen Value Decomposition
FIM	Fisher Information Matrix
FFT	Fast Fourier Transform
FLIR	Front-Looking Infra-Red)
GDOP	Geometric Dilution Of Precision
GPS	Global Positioning System
HPRF	High Pulse Repetition Frequency
IEEE	Institute of Electrical and Electronics Engineers
IFFT	Inverse Fast Fourier Transform
LF	Likelihood Function
LFM	Linear Frequency Modulation
LOS	Line Of Sight
IFF	Identification Of Friend or Foe
MIMO	Multiple-Input Multiple-Output
MMSE	Minimum Mean Square Error
MPRF	Medium Pulse Repetition Frequency
MTI	Moving Target Indicator
MVDR	Minimum Variance Distortionless Response
MWF	Multi-Stage Wiener Filter
NLOS	Non Line Of Sight

OCXO	Oven Controlled Crystal Oscillator
OFDM	Orthogonal Frequency Division Multiplexing
PC	Principal Components
PCSI	Principal-Component Signal Independent
PCSD	Principal-Component Signal Dependent
PDF	Probability Density Function
PLL	Phase Locked Loop
PRF	Pulse Repetition Frequency
PRI	Pulse Repetition Interval
PTSU	Phase and Time Synchronization Unit
RADAR	Radio Detection And Ranging
RCS	Radar Cross Section
RMB	Reed Mallet and Brennan
RMS	Root Mean Square
ROC	Receiver Operator Characteristic
SLAR	Side-Looking Airborne Radar
SMI	Simple Matrix Inverse
SNR	Signal to Noise Ratio
SSR	Secondary Surveillance Radar
STAP	Space Time Adaptive Processing
TDOP	Time Dilution Of Precision
UAV	Unmanned Aerial Vehicle
VCO	Voltage Controlled Oscillator

1

Introduction: Challenges for low observable target

This chapter describes the various challenges of low observable airborne targets and how to tackle this situation using bi-static radar, multi-static radar and extended to multi-dynamic radar. This chapter also discusses the motivation of the research work carried out in this arena to overcome different technical challenges for a better and effective airborne surveillance system. This chapter presents motivation of the thesis, outlines the thesis contribution and also provides the overview of the thesis organization.

1.1 Challenges for radar system for low observable target

The most important characteristic of a target is its Radar Cross Section (RCS). For monostatic radars, RCS is a quantitative characteristic of the target's ability to scatter the energy in the direction opposite to the incident wave direction. During the last few decades, low RCS technology has proven to be one of the most effective approaches for hiding the target from radar systems. The basic concept of the low observable principle is mainly the reduction of RCS, RCS is a measure of how detectable a target is by a radar. It is the ratio of the scattered power by the object (target) in a given direction to the incident power density. RCS of a target depends on the illumination and reception direction. Hence bi-static RCS (σ_b) is not equal to the monostatic RCS (σ). These geometrical characteristics of bi-static radars include the set of a multi-site radar. For Multi-site radars, a more general feature is required, which takes into account the difference between directions from target to transmitting and receiving stations. So the detection of low RCS targets involves multi-static radar configuration to counter the reduction in backscattered energy. However, most targets are aircraft, and radars are airborne, mounted over aircraft or unmanned air vehicles (UAVs) or Drones. Hence, instead of all ground based radars, a group of coordinated airborne radars is considered better for airborne surveillance scenario. It views the target from different aspect angles, and the basic target parameters are obtained through a bi-static system. Earlier, for the detection of the low RCS targets, the monostatic radar was used with the SLAR (Side-Looking Airborne Radar) and FLIR (Front-Looking Infra-Red) equipment in a high-flying aircraft to map out the terrain below and to one side or to the front. If a stealth aircraft operates anywhere beyond the terrain which is being mapped, it would not be observed on the detecting aircraft's output screen. That Spot will appear black or blank, like a silhouette of the stealth aircraft, thereby pinpointing the position of the stealth craft in the field of vision and detecting it would be difficult [1]. The satellite method may also be used on battlefields by looking down from satellites equipped with radar sensors. Though satellites are far away from the battlefield, sometimes they are out of range to detect most of the hostile weapons. Another way for the detection of stealth aircraft would be to use surface-based radar installations which scan the sky at high apertures and with high sensitivity. Since these detectors are surface based and

due to their large and heavy size they are not preferred [1]. H. Kuschelet.[2]al proved in their paper that for the detection of low RCS targets, a multi-static radar system is more efficient than monostatic radar.

1.2 Issue of airborne radar cross section for low observable target

According to the monostatic-bistatic equivalence principle and observed from experimental data, the bi-static RCS (σ_b) of a target usually does not differ drastically from monostatic RCS (σ_m) of the same target at moderate bi-static angles. However, when bi-static angle changes, the situation changes significantly. Various low observable technologies for an aircraft have been utilized to reduce RCS in the direction of radar. This is to avoid detection by radars and has become a vital design goal of a tactical aircraft. The architect of the aircraft is the most important factor that divers electromagnetic energy in various directions and in general, low observable performance for airborne surveillance.

1.3 Bi-static radar a step ahead for low observable target detection

A stealth target scatters energy in all directions. This reduces the radar cross-section (RCS), and the energy back-scattered gets reduced in the radar receiver(Rx) direction. A single monostatic radar or bi-static radar system having one receiver(Rx) and one transmitter(Tx) may not be solution for effective detection of such target in such a scenario. So, multiple monostatic radar or multiple bi-static radar configurations are required to deploy for the detection of such low observable target, such combination is known as multi-static radar, as elucidated by Jinlei et al in 1998 [3]. This radar system has multiple number of in combination transmitters(Txs) and receivers(Rxs) placed at different locations [4]. It has significant advantages, including a large coverage area, immunity to jamming, and spatial diversity [5]. In literature, many techniques for the target detection and tracking in a multi-static scenario along with the transmitter identity have

been proposed. The system geometry of bi-static radar is described in [6] [7] and [8]. A review has been made by H. D. Griffith [9] about the RCS analysis and ambiguity function of bi-static radar. They used Passive Coherent Location (PCL) techniques in which the broadcast signals are the illuminators of opportunity. Bezousek et al [10] described various aspects of signal processing and signal parameters in multi-static arrangement with non-cooperative transmitters. In 2010, Tan et al presented the geometry of passive bi-static radar with the ambiguity function to detect the moving targets [11]. A new method was proposed by Ling et al in 2012, for target detection in bi-static radar networks with dynamic frequency selection [12].

Another key feature that is considered is the coverage area. The detection range or the coverage area of a bi-static radar depends on the location of the transmitter and receiver at constant SNR. It is characterized by the Cassini oval, with the transmitter and receiver being the two foci [6]. A oval of Cassini is represented as a quadratic kind of plane curve defined as the set of points in the quadratic plane such that the product of the distances between the point on the curve to the individual foci is constant [13]. The coverage area of monostatic radar becomes a circle due to the co-location of transmitter and receiver. The multi-static coverage area is obtained as a sum of bi-static or monostatic coverage area depending on the system geometry [14].

A multi-static radar system has two or more transmitters and receivers, which are separated by large distances [1]. Bi-static radar is a basic of multi-static radar system. It is capable of detecting targets in a two-dimension co-ordinate system, with one transmitter and one receiver separated far apart. Multi-static radar has several advantages including invulnerability to external interferences such as jamming. Also, mutual interference between the transmitter and receiver is highly reduced. In such systems, increasing the number of receivers also results in increasing the surveillance area [15]. The main disadvantage is that the synchronization between the transmitter and receiver and also the geometry of the entire system is too complicated.

When the number of transmitters and receivers are two or more, it is very difficult for a particular receiver to identify the signal of any particular transmitter. So there is a need to have a transmitter identity along with the transmitted signal. Hence, in this thesis,

an attempt is made to introduce transmitter identity along with the transmitted signal. Spread spectrum techniques are given in [16] and [17] are used to secure the transmitter identity code, and the resultant binary bits are modulated with LFM signal through BASK technique. After identifying the transmitter of interest from the echo, targets are detected using a simple Cell Averaging Constant False Alarm Rate (CA-CFAR) technique [18]. Bi-static ranges are estimated by multiple receivers pertaining to a given target. For the known bi-static angles, multiple target ranges with respect to the transmitter of interest is geometrically obtained using individual bi-static ranges. Thus there exist multiple ranges pertaining to a single target. Finally, these estimated ranges are fused using MLE technique [19].

The complexity increases further when all radars are considered to be airborne radar and form the Multi-dynamic radar. The waveform design for such a scenario is difficult due to dynamic system geometry. Another major issue is to synchronize time and phase across floating airborne radar or radar transmitter or radar receiver.

1.4 Motivation for the Present Research Work

There are several issues that have motivated me to carry out the present research work. These issues and the research gaps are presented in Appendix-B of the thesis.

1. During the airborne radar performance evaluation, it was observed that even though many aircraft are under airborne radar coverage, the same was also verified through open-source satellite-based info such as flight radar application. Also we do have known aircraft GPS data, which enables these aircraft were in our radar coverage. However, few aircraft are getting detected and few are not getting detected even though nature of the targets are known. As per radar system engineering calculation, the target returns are expected to have sufficient SNR for detection. This detection loss gave me the thought that few aircraft structures may divert the electromagnetic energy away from radar receiver. Hence there would be no RCS of targets with respect to the receiver, and this makes the detection loss. However, if we had placed our receiver at multiple places, then would not be detection loss.

Due to this, the purpose and performance of airborne surveillance systems could not achieve. Hence there could have more number of airborne receiver works in tandem of airborne radar having Tx-Rx then all the diverted electromagnetic energy would have been captured to process the echo. The thought came for Multiple input Multiple output (MIMO) radar, which is characterized by multiple antennas to simultaneously transmit diverse (possibly linearly independent) waveforms and by utilizing multiple antennas to receive the reflected signals. There are two possibilities of MIMO configuration, MIMO radar with co-located antennas [20] and MIMO radar with widely separated antennas [21], which is also referred to as statistical MIMO. The enabling concept for MIMO radar, is the transmission of multiple orthogonal waveforms from different antennas [22], which is usually referred to as the waveform diversity [23]. There is a trade-off in using MIMO versus Phased Array as it does not have coherent gain hence in order to exploit both the features, Phased Array-MIMO concept is utilized where a group of antennas elements is used for TX and RX instead of single antenna elements [24]. A MIMO radar consisting of M transmit and N receive elements extracts MN signals [25]. These signals can be considered, those received by a virtual array of MN elements.

2. The multi-dynamic radar scenario studied in this thesis considers transmitter and receiver to be geographically widespread, and the simulation studies have been performed based on some of the techniques reported in the literature. Simulation includes the detection performance for Transmitter identity as well as Target detection [26][27] [28] [29]. The simulation also has been performed towards the modelling of Multi-dynamic radar model [30] [31] [32] [33] [34] and coverage area for various SNR as well as the design of ambiguity function. This includes the choice of OFDM-LFM based waveform [35] [36] [37] [38]. Simulation have also been performed to study detection performance using IFF waveform [39] [40] [41] [42] [43] [44].
3. In a multi-dynamic-radar scenario, due to the presence of multiple airborne Tx-Rx pairs, it is proposed to use orthogonal frequency-division-multiplexing (OFDM) based linear-frequency-modulated (LFM) waveform [45] [46] [47] [48]. Hence, for

studying MDR performance, the ambiguity function is obtained using OFDM-LFM waveforms [28][29]. The OFDM-LFM waveforms [32] achieve better spectral efficiency by using multi-orthogonal carrier modulation to achieve better range-resolution and Doppler-resolution, as guaranteed by LFM waveform. It is required to have a waveform performance parametric benchmark for choosing the optimal waveform. MDR waveform design performance is determined by deducing Cramer-Rao Lower Bound (CRLB) [33] for delay τ_a and Doppler ν_a of ambiguity function. This includes optimal Tx-Rx pair geometry for the estimation of range/Doppler for a particular target of interest. Fisher Information Matrix (FIM) which gives a parametric benchmark to choose a radar waveform for optimal performance. FIM $\mathbf{J}_M$ for monostatic radar case is available in [34].

1.5 Contribution of the Thesis

This thesis investigates different issues of a multi-dynamic radar scenario which is targeted for airborne surveillance. Studies have been performed on how airborne radars with transmitter and receiver position changing dynamically, can work in a coordinated manner.

The main contributions of this thesis are summarized and listed below:

1. Mathematical derivation and simulation of Bi-static radar configuration which includes the simulation of radar coverage for various SNR level.
2. Study and simulation of multi-static radar, this includes the waveform design with transmitter identity and signal processing for target detection.
3. Study and simulation of multi-dynamic radar as extended concept of multi-static radar. The mathematical derivation and simulation of target SNR computation has been for a typical scenario.
4. Study and simulation of IFF waveform based waveform ID for multi-dynamic radar. Simulation is made how Mode-S waveform of IFF can be used for coded waveform identification.

5. STAP based radar functionality is common for ground-based radar, however this can be extended for multi-dynamic radar. As such if target Angle-Doppler map is effected for one receiver to effect the target detection. This due the presence of airborne clutter and jammer. However, the same target angle-Doppler map for other receivers may be clear to detect the target. The simulation is made with various STAP algorithms for multi-dynamic radar scenario.
6. Ambiguity function derivation for multi-dynamic radar, simulation of ambiguity function for the proposed OFDM-LFM waveform for medium PRF cases. Benchmarking the performance of the waveform by deriving the CRLB function for optimal result.
7. Towards muti-dynamic radar operational requirement for various airborne receivers to process the target echo phase, time synchronization is required. The details of design approach to achieve the phase and time synchronization across airborne Tx and Rx by using Phase and Time Synchronization unit (PTSU) for Doppler processing is studied.

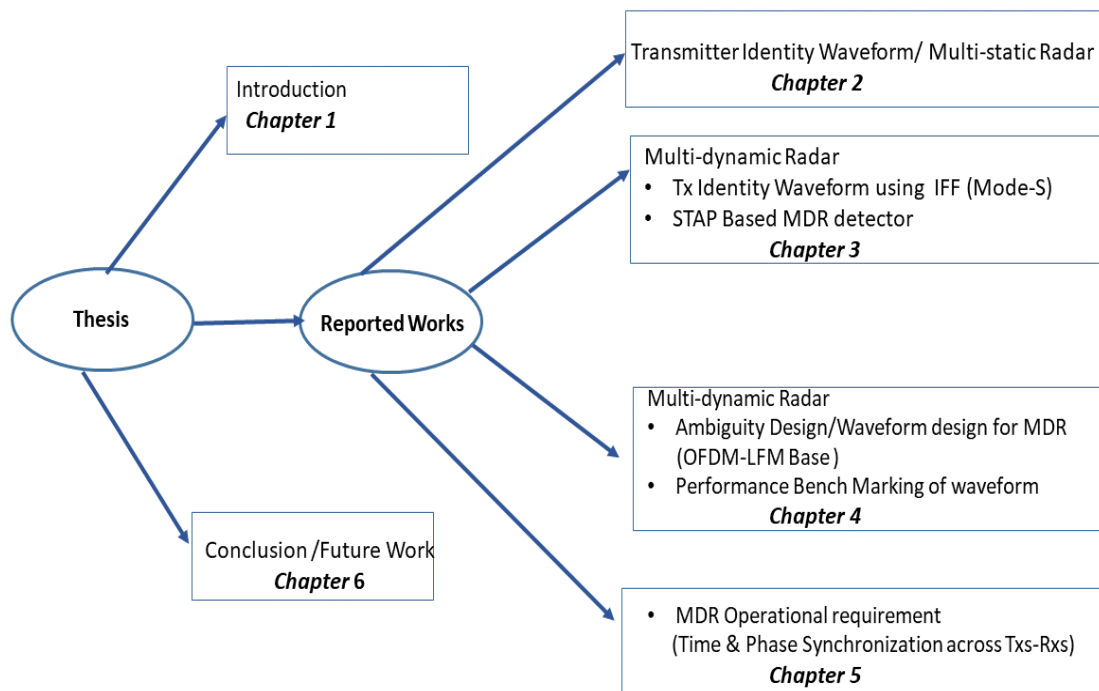


FIGURE 1.1: Thesis organization.

1.6 Thesis Organization

This thesis is organized into six chapters as shown figure (1.1). The summary of each chapter is briefed as follows:

Chapter 1: This chapter presents discussion on challenges for radar systems dealing with low observable targets. Issues with airborne radar cross section for low observable target is discussed next, which is followed by a discussion on bi-static radar configuration and emphasizing some of the issues related to airborne radars operated in multi-dynamic configuration. This chapter also presents the motivation behind the works reported in this thesis and highlights the research contributions. The chapter ends with a brief description on the thesis organization.

Chapter 2: This chapter discusses about multi-static radar configuration which includes the operational geometry with extended concept of bi-static transmitter (Tx) and receiver (Rx) pair. In this chapter the simulation result for multi-static radar configuration is presented which includes transmitter identity waveform design and associated signal and data processing for typical scenarios.

Chapter 3: This chapter presents the details of multi-dynamic radar concepts, requirements for airborne surveillance system. This chapter also includes the IFF Mod-S waveform based waveform identification requirement for primary radar. This is required for multi-dynamic radar to work in a co-ordinated manner. STAP based simulations are also made for multi-dynamic radar in order to utilize the set of receivers in the presence of jammer and clutter.

Chapter 4: In this chapter waveform design aspect of multi-dynamic radar (MDR) is brought out. The detailed derivation of ambiguity function and proposed OFDM-LFM based waveform results are presented. This includes the discussion and performance trade-off over use of number of pulsed OFDM-LFM with respect to peak side-lobe for the matched filter output. The bench-marking of waveform performance as a CRLB derivation for MDR waveform is also included in the result.

Chapter 5: In this chapter, MDR operational requirement is discussed which includes the essential requirement of phase synchronization across the airborne platform for pulse compression. The proposed design method to achieve phase synchronization across platforms is also brought out in this chapter.

Chapter 6: In this chapter, summary of the works carried out has been presented. Scopes of future works for extending the concept of multi-dynamic radar are also discussed.

1.7 Summary

The issue of the detection of a deceptive airborne target can be achieved by a cooperative set of airborne radar forming multi-dynamic radar. The sensitivity for multi-dynamic radar is not only dependent on radar parameters but also on system geometry. Multi-dynamic radar offers more flexible arrangement of system geometry than any traditional monostatic radar, bi-static or multi-static radar providing the possibility to configure radar nodes to form a satisfactory coverage area. Multi-dynamic waveform design tool such as ambiguity function is strongly dependent on the radar system geometry and waveform. It is required to be keep changing time to time as SNR of target changes. More and more airborne radar nodes gives more flexible multi-dynamic radar system geometry and provide the possibility to further improve for deceptive target detection and resolution capability. Overall multi-dynamic radar has several advantages but these do come at the cost of increased design complexity. Major issues in the time and phase synchronization among airborne radar to work in tandem. Additionally, this is still far from matured form of system and there are much more research and development work to be done if multi-dynamic radar is to do routing surveillance in use.

2

Investigation on multi-static radar in bi-static radar Configuration

This chapter contains the details of the mathematical modelling of bi-static radar and its operation. Waveform parameters are very important for radar signal processing. As the geographical location of the transmitter and receiver is far away, and no line of sight exist between them, in this scenario tailored waveform is proposed where waveform parameters are also transmitted as a part of transmitting waveform, so that receiver upon receiving the echo also shall receive the encoded waveform ID. Subsequently, in this chapter it is shown how end-to-end target detection processing can be achieved.

2.1 Introduction

During the last few decades, low Radar Cross Section (RCS) technology has proven to be one of the most effective approaches to hide the target from radar systems. The basic concept of the low observable principle is mainly the reduction of RCS. So, the detection of low RCS air targets involves multi-static radar configuration to counter the reduction in backscattered energy. It views the target from different aspect angles and the basic target parameters are obtained through a bi-static system. Earlier, for the detection of the low RCS targets, the mono-static radar was used with the SLAR (Side-Looking Airborne Radar) and/or FLIR (Front-Looking Infra-Red) equipment in a high-flying aircraft to map out the terrain below and to one side, or to the front. If a stealth aircraft was operating anywhere beyond the terrain which is being mapped, it would not be observed on the detecting aircraft's output screen. That spot will appear black or blank, like a silhouette of the stealth aircraft, thereby pinpointing the position of the stealth aircraft in the field of vision and detecting it would be difficult [49]. A method using satellite may also be employed on battlefields by looking down from satellites equipped with radar sensors. Though satellites are far away from the battlefield, sometimes they are out of range to detect most of the hostile weapons. Another way for the detection of the stealth aircraft would be with the use of surface-based radar installations which scan the sky at high apertures and with high sensitivity. Since these detectors are surface based and due to their large and heavy size they are not preferred [2]. H. Kuschelet.al [2] proved in their paper that for the detection of low RCS targets, multi-static radar system is more efficient than the mono-static radar.

2.2 Bi-static radar modelling

According to the IEEE Standard Radar Definition (IEEE standard 686-1997), a bi-static radar is a radar system that has separate antennas for transmission and reception at sufficiently different locations that the angles or ranges from those locations to the target are significantly different [7]. This has a potential advantage in the detection of stealthy targets which are shaped to scatter energy in the directions away from the transmitter.

figure (2.1) shows the Bi-static configuration with a transmitter (TX) and receiver (RX) being situated at two locations with a baseline separation (L) [4]. The Bi-static angle (β) is one of the important parameters which characterizes Bi-static radar and affects the system performance. The bi-static receiver usually measures the angle of arrival of the target echo and estimates the range. The major subsystems in Bi-static radar are the transmitter (T_X), the receiver (R_X), and the target (T), which make the Bi-static triangle as shown in figure (2.1). By knowing the time delay between the transmitted waveform and echo received, transmitter to the target range (R_T) or Target-to-receiver range (R_R) cannot be measured directly, but can be calculated by solving the Bi-static triangle. A Bi-static radar measures the time delay a signal has travelled from transmitter to target to the receiver. The simulation is made towards the detection performance of Bi-Static Radar for towards the stealth target [50] [51].

2.2.1 Mathematical modelling for Bi-static radar

The receiver in Bi-static radar [15] usually measures the angle of arrival of the target echo and estimates the range. In a bi-static scenario, using time delay (τ) of the echo signal i.e a signal has travelled from transmitter to target to the receiver, the target range represents the sum of R_T and R_R as shown in equation(2.1)

$$R_T + R_R = c\tau \quad (2.1)$$

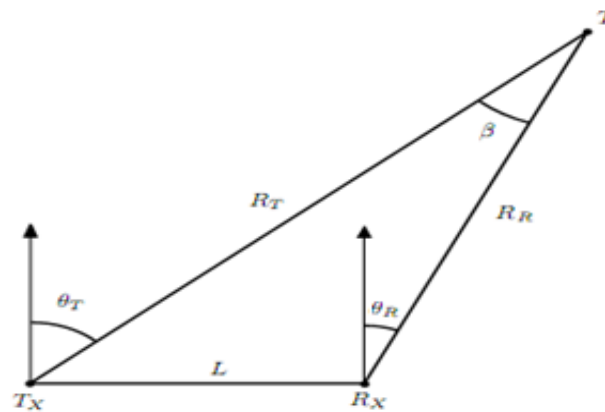


FIGURE 2.1: Bistatic triangle

For bi-static radar, the range coverage is given by [2] as shown in equation (2.2)

$$(R_T R_R)_{\max} = \left[\frac{P_T G_T G_R \lambda^2 \sigma_B^2 F_T^2 F_R^2}{(4\pi)^3 k T_S B_N (S/N)_{\min} L_T L_R} \right]^{1/2} = K \quad (2.2)$$

Where R_T = transmitter-to-target range (m)

R_R = target-to-receiver range (m),

P_T = transmitted power (watts),

G_T = transmitting antenna power gain,

G_R = receiving antenna power gain,

λ = transmitting frequency wavelength (m),

σ_B = bi-static radar cross section (RCS)(m²)

F_T = pattern propagation factor for the transmitter-to-target path,

F_R = pattern propagation factor for the target-to-receiver path,

k = Boltzmanns constant

T_S = receiver noise temperature in Kelvin,

B_N = receiver noise bandwidth (Hz),

$(SNR)_{\min}$ = minimum signal-to-noise power ratio required for detection,

L_T = transmitting system losses (≥ 1),

L_R = receiving system losses (≥ 1),

K = bi-static maximum range product.

Equation (2.2) shows that the maximum range is based on the signal-to-noise ratio $SNR_{\min}$. For the known R_T & R_R , $SNR_{\min}$ can be obtained from equation (2.3) as

$$SNR_{\min} = \left[\frac{P_T G_T G_R \lambda^2 \sigma_B^2 F_T^2 F_R^2}{(4\pi)^3 k T_S B_N R_T^2 R_R^2 L_T L_R} \right] \quad (2.3)$$

In the bi-static, radar the contour at constant SNR forms the Oval of Cassini. The Oval of Cassini is defined as the locus of the vertex of a triangle when the product of the sides adjacent to the vertex is constant. Each oval represents a constant SNR level. R_T and R_R changes as their intersection on a given SNR contour moves. The target position is considered to be this point of intersection. The values of (R_T) and (R_R) change as their

intersection on a given SNR contour moves. The target position is considered to be the points on the ovals of Cassini.

In a bi-static or multi-static scenario, multiple transmitter and receiver combinations will give the bi-static range pertaining to a single target. To get a better estimation pertaining to a single target, the corresponding information estimated by multiple pairs has to be fused. But each pair may be located at different places with different distance apart in the specified multi-static region. Hence, all the estimates have to be geometrically represented with respect to the central unit. Usually transmitter of interest will act as a central unit. Sometimes one of the receivers may also be considered as a central unit. In subsequent section, mathematical derivation is given to get the geometrical range with respect the transmitter (R_T) or with respect to receiver (R_R).

2.2.2 Range estimation in Bi-static scenario

Bi-static coverage area is represented by the Oval of Cassini, as shown in the subsequent section. Without having any information about the direction of the target, i.e. with omni directional antennas, the location of the target will be anywhere on the iso-range contour which, forms an ellipsoid with receiver (R_x) and the transmitter(T_x) as foci. In figure (2.2) β is shown as a bi-static angle between transmitter to target and the target to the receiver. This may vary between 20° to 145° degrees. θ_R and θ_T are the azimuth angles of the receiver and transmitter. Using the law of cosine, (R_T) and (R_R) can be calculated individually, having known L and θ_R or θ_T along with the estimated bi-static range (R_T) + (R_R) from equation (2.1). For the known θ_T , after applying law of cosine on the triangle shown in figure (2.2), we get [52] (R_T) and (R_R) as shown in equations (2.4),(2.5). For the mono-static system, the base line distance $L=0$ and $R_T = R_R$.

$$R_T = \frac{(R_T + R_R)^2 - L^2}{2(R_T + R_R + L \sin \theta_R)} \quad (2.4)$$

$$R_R = \frac{(R_T + R_R)^2 - L^2}{2(R_T + R_R + L \sin \theta_T)} \quad (2.5)$$

The bi-static angle (β) is estimated for the known (R_R), (R_T), θ_T and θ_R as:

$$\beta = \sin^{-1} \left(\frac{L \cos \theta_T}{R_R} \right) \quad (2.6)$$

For Bi-static radar, maximum range coverage is given by equation (2.2) and corresponding signal-to-noise ratio ($\text{SNR}_{\min}$) is given by the equation (2.3) [53]

2.2.3 Bi-static radar coverage - A concept of Cassini (oval)

In bi-static radar, it is interesting to analyse the system performance by plotting the oval of Cassini (shapes like an ellipse) as a function of the parameters considered are ranges R_T and R_R , with the vertices $T_x(-a, 0)$, $R_x(a, 0)$ and $P(x, y)$ of the bi-static-triangle for a particular value of signal-to-noise ratio (SNR). The oval of Cassini for bi-static radar case is defined as a 2D-plane of curve which is depicted by in combination of the locus of the all considered tracing-points of P as presented in figure(2.3). With the locus of all points $P(x, y)$ are taken, such that the product of the euclidean distances of the point $P(x, y)$ from the two considered focus points, i.e transmitter(T_x) and receiver(R_x) is constant. But here our interest is to find the coordinates of the point $P(x, y)$ in order to find the maximum coverage area as in equation (2.7).

Oval of Cassini curve shown in figure (2.3), the two points T_x and R_x are called the foci and the origin of the coordinate is at the midpoint of the two foci. Here a and b are the semi-major and semi-minor axis of the Cassini curve shown in figure (2.3).

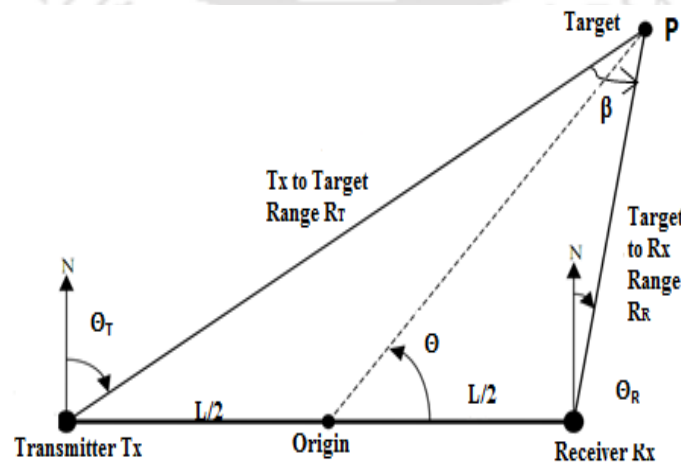


FIGURE 2.2: Bistatic system geometry

$$R_T R_R = b^2 \quad (2.7)$$

The distance (R_T) between transmitter($T_x(-a, 0)$) and the target($P(x, y)$) can be calculated as equation (2.8)

$$\sqrt{(x + a)^2 + y^2} = R_T \quad (2.8)$$

Similarly, the distance (R_R) between the target($P(x, y)$) and the receiver ($R_x(a, 0)$) can be calculated as equation (2.9)

$$\sqrt{(x - a)^2 + y^2} = R_R \quad (2.9)$$

Substituting R_T and R_R in equation (2.7), it becomes

$$\sqrt{(x + a)^2 + y^2} \sqrt{(x - a)^2 + y^2} = b^2 \quad (2.10)$$

Squaring both the sides and on simplification

$$(x^2 + y^2 + a^2)^2 - 4a^2 x^2 = b^4 \quad (2.11)$$

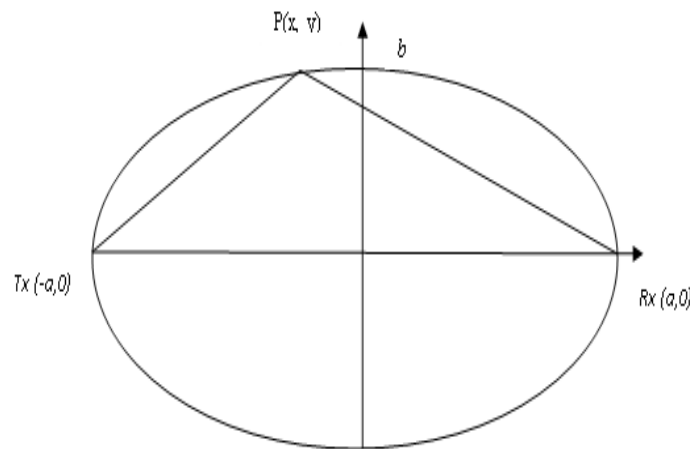


FIGURE 2.3: Oval of cassini curve representing the transmitter and receiver at the two foci

The polar form of the equation (2.11) is represented by substituting,

$$x = r \cos \theta; y = r \sin \theta; x^2 + y^2 = r^2 \quad (2.12)$$

Here, one of the foci is fixed as the origin, with θ as the angular coordinate measured from the major axis. Thus equation (2.12) can be rewritten as:

$$r^4 - 2a^2r^2 \cos(2\theta) + a^4 - b^4 = 0 \quad (2.13)$$

Solving quadratic form of the equation (2.13) results in

$$r^2 = \frac{2a^2 \cos(2\theta) \pm \sqrt{4a^4 \cos^2(2\theta) - 4(a^4 - b^4)}}{2} \quad (2.14)$$

$$r = \pm a \sqrt{\cos(2\theta) \pm \sqrt{\left(\frac{b}{a}\right)^4 - \sin^2(2\theta)}} \quad (2.15)$$

P(x,y) coordinates are obtained by substituting equation (2.15) in $x = r \cos \theta$ and $y = r \sin \theta$

$$x = \left\{ \pm a \sqrt{\cos(2\theta) \pm \sqrt{\left(\frac{b}{a}\right)^4 - \sin^2(2\theta)}} \right\} \cos \theta \quad (2.16)$$

$$y = \left\{ \pm a \sqrt{\cos(2\theta) \pm \sqrt{\left(\frac{b}{a}\right)^4 - \sin^2(2\theta)}} \right\} \sin \theta \quad (2.17)$$

For the known, $R_T R_R = b^2$; b parameter may be obtained. Then Bi-static radar coverage area is obtained by tracing the (x, y) given in the above polar equations (2.16) (2.17) for different θ values as it vary from 0 to 360 degrees [10]. Cassini shapes depend upon the parameters a and b . The shape of the curves depends upon the ratio of $\frac{a}{b}$ is shown in figure (2.4). Cassini ovals with a large value of b^2 approach a circle.

2.2.4 Doppler's shift in the received signal based on target velocity

In general, Doppler phenomenon describes the shift in the center frequency of an incident waveform due to the target motion with respect to the source of radiation. Depending on the direction of the target's motion, this frequency shift may be positive or negative. If the target is a moving target and it is coming towards the radar then the frequency will increase and the Doppler shift is positive. In the same way, when the target is moving away from the radar then the frequency will decrease and it is a negative shift. Here we are considering Doppler shift between consecutive two measures of the echo signal.

The Doppler frequency f_d between the scattered signal and the direct signal in the bi-static radar is proportional to the time rate of change of the total path length of the scattered signal. Doppler frequency shift provides a means for discriminating stationary objects from moving targets but it is not a measure of the radial velocity as it is in the case of the mono-static radar [4]. However, the Doppler frequency f_d is given by the equation (2.18)

$$f_d = \frac{1}{\lambda} \left[\frac{d(R_T + R_R)}{dt} \right] \quad (2.18)$$

The targets velocity vector has magnitude V and aspect angle δ . They are being projected to the transmitter and receiver as V_T , V_R , δ_T and δ_R respectively. A stationary transmitter and receiver will have the projection of the target velocity vector on to the

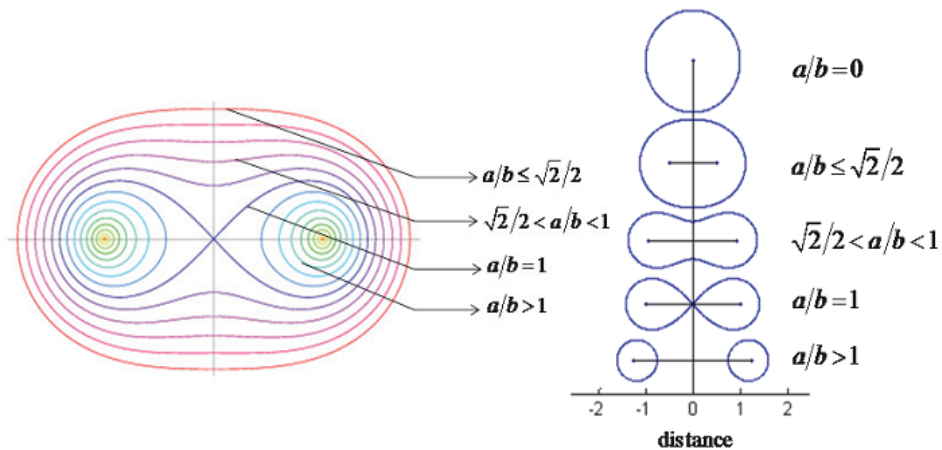


FIGURE 2.4: Cassini curves for different values of a and b

transmitter-to-target line of sight and receiver-to-target line of sight as follows [2]. From the bi-static Doppler geometry figure (2.5), the target velocity with respect to the transmitter is obtained equation (2.19) and (2.20)

$$\frac{dR_T}{dt} = V_T \cos(\delta_T - \beta/2) \quad (2.19)$$

Similarly, target velocity with respect to the receiver:

$$\frac{dR_R}{dt} = V_R \cos(\delta_R + \beta/2) \quad (2.20)$$

Let us assume $\delta_R = \delta_T = \delta$ and $V_T = V_R = V$ By substituting equation (2.19) and (2.20) in equation (2.18), the resulting equation (2.21)

$$f_d = \frac{2V}{\lambda} (\cos \delta) (\cos(\beta/2)) \quad (2.21)$$

Where

f_d = Doppler shift

V = Magnitude of target velocity

V_T = Target velocity with respect to the transmitter

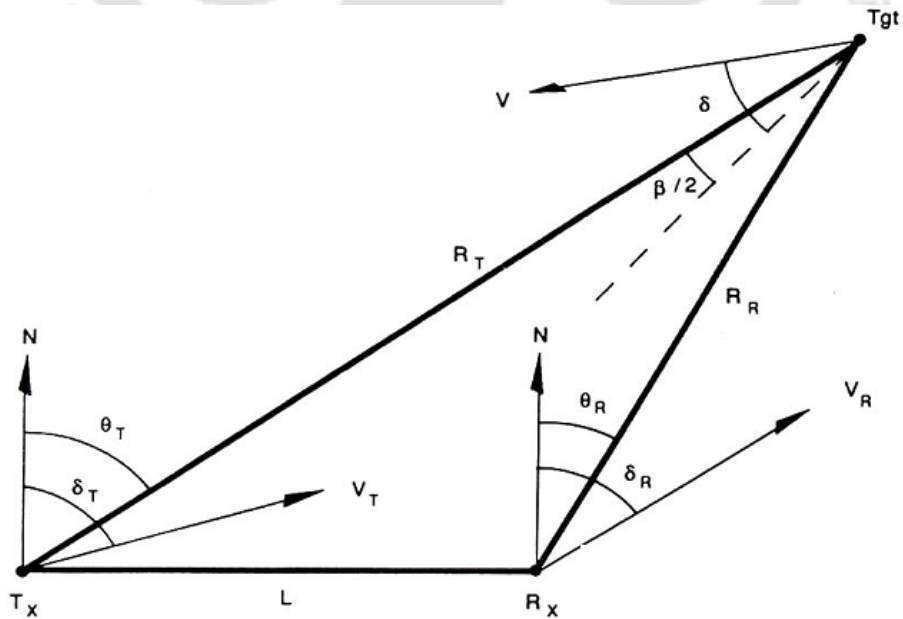


FIGURE 2.5: Bistatic Doppler geometry

V_R = Target velocity with respect to the Receiver δ_R = the azimuthal angle of the receiver

δ_T = the azimuthal angle of the transmitter

λ = wave length of the signal

f_T =transmitting frequency

Equation (2.21) shows that Doppler shift f_d depends not only on the velocity of the target but also on δ angle and β angle.

2.2.5 SNR calculation based on multi-static scenario

This multi-static system is considered to be a co-operative and coherent network of radars, it is assumed that the whole radar-network is in time-synchronized or local-oscillator(LO) synchronized and as well all set of radar works in tandem. it is also assumed that each radar receiver (Rx) on multi-static radar network is capable of receiving target return by the transmitted waveform from any radar transmitter(Tx)on the same radar-network. The target return is assumed to be isotropic, so that constant RCS is assumed in all directions. It is assumed for simulation that there are 'm' number of transmitters (Tx) and 'n' number of receivers (Rx) in the multi-static-radar network, so that the overall $SNR_{overall}$ may be computed by summing up the partial SNR [17], it is given by the equation (2.22)

$$(SNR)_{overall} = \sum_{i=1}^m \sum_{j=1}^n \frac{P_{ti} G_{ti} G_{rj} \lambda_i^2 \sigma_{tj}^2 F_{ti}^2 F_{rj}^2}{(4\pi)^3 k T_S B_{Rj} R_{ti}^2 R_{rj}^2 L_{ij}} \quad (2.22)$$

2.3 Transmitter identity

Two important factors which are required to be considered for radar waveform design are the range resolution and maximum range detection. As defined, the radar range resolution is the ability of the radar waveform to separate closely-spaced targets, and mathematically it is related to the pulse-width of the transmitted waveform. Narrower pulse-width produces better range resolution. However, if the pulse-width of the transmitted waveform is decreased, this will decrease the amount of energy in the pulse, which will lead to the reduction in the maximum range of operation. In order to overcome such

problem, a mathematical technique called pulse-compression is used in radar systems. Here, we consider linear frequency modulation (LFM) technique in which frequency increases linearly across the pulse width, known as chirp signal. A wider pulse width of the transmitted signal using LFM technique will increase the range coverage. In the receiver, the received echo is compressed to get a better resolution in the temporal domain or range. The instantaneous phase of the LFM signal or so called chirp-signal is expressed as [18] in equation (2.23)

$$\phi(t) = 2\pi \left(f_1 t + \frac{kt^2}{2} \right) \quad (2.23)$$

where f_1 is the carrier frequency and k represents the frequency sweeping-rate which is related to the pulse duration T_p and the bandwidth B as equation (2.24) [12]

$$k = \frac{B}{T_p} \quad (2.24)$$

The instantaneous frequency is given by equation (2.25)

$$f(t) = \frac{d}{dt} \left(f_1 t + \frac{kt^2}{2} \right) = f_1 + kt \quad (2.25)$$

width of the compressed pulse (τ_{comp}) depends on the change in the transmitter frequency (ΔF) over the pulse duration (τ) or bandwidth of the transmitted waveform.

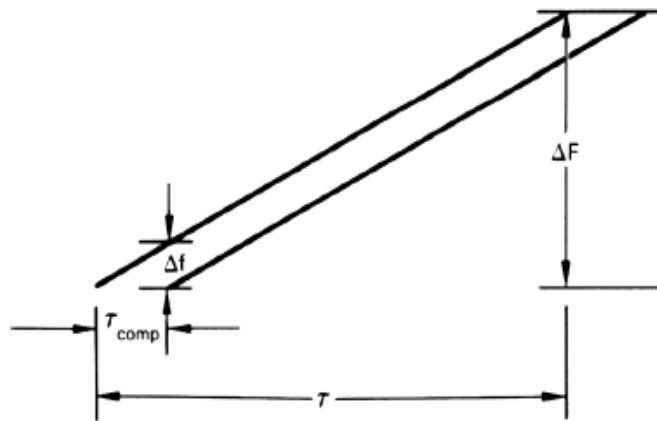


FIGURE 2.6: Ratio of the uncompressed to compressed pulse widths

Thus pulse compression ratio is given by the equation (2.26)

$$= \frac{\tau}{\tau_{\text{comp}}} = \tau \Delta F, \quad (2.26)$$

From equation (2.26) it is clear that the pulse compression ratio is based on the time bandwidth product. These compressed signals are sampled in such a way to get one sample per range bin as shown in figure (2.6).

In a multi-static scenario, if more than one transmitter is present then, estimating the target parameters with respect to the transmitter is a problem [54]. Thus, there is a need to have a transmitter identity in a received echo signal. Here method is proposed by introducing the transmitter identity in the transmitted signal. The identity is a four bit binary data modulated with the digital modulation techniques. Thus, the transmitted pulse signal consists of the LFM signal followed by the transmitter identity.

Transmitter identity along with the transmitted LFM signal in one pulse, is shown in figure (2.7). For example, here the total duration of the pulse is considered to be $25\mu\text{s}$, of which the first $5\mu\text{s}$ is utilized for the transmitter identity, and the next $20\mu\text{s}$ is for the signal. The pulse repetition time (PRT) is considered as $250\mu\text{s}$. The transmitter identity is any 4 bit sequence. For simulation bit sequence is taken (1 0 1 0) which is modulated by any shift keying modulation technique.

For the transmitter identity generation, the digital modulation schemes Binary Phase Shift Keying (BPSK) or binary amplitude shift keying (BASK) are considered. The

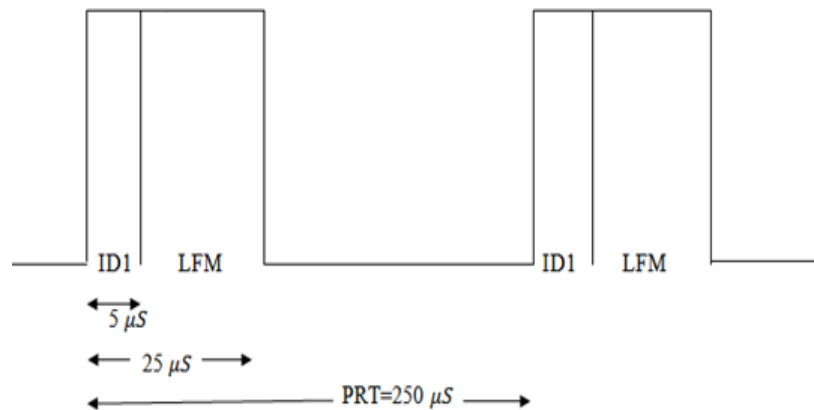


FIGURE 2.7: Transmitted LFM signal with transmitter identity (ID1)

modulated signal is transmitted along with the LFM signal.

The binary sequences are converted in the non-return to zero line coder and then multiplied with the carrier wave to get the BPSK modulated signal as shown in figure (2.8). The BPSK signal representation of the binary data 1010 and 1110 as shown in figure (2.9) and figure (2.10).

In this simulation, for the generation of the LFM signal, the time period that has been taken is 20 millisecond. The starting frequency (f_L) is 100 Hz and final frequency (f_H) is 700 MHz. The frequency is linearly increasing at a rate of 6 Hz. In the multi-static simulation environment, two transmitters with identities (1 0 1 0) and (1 1 1 0) and five

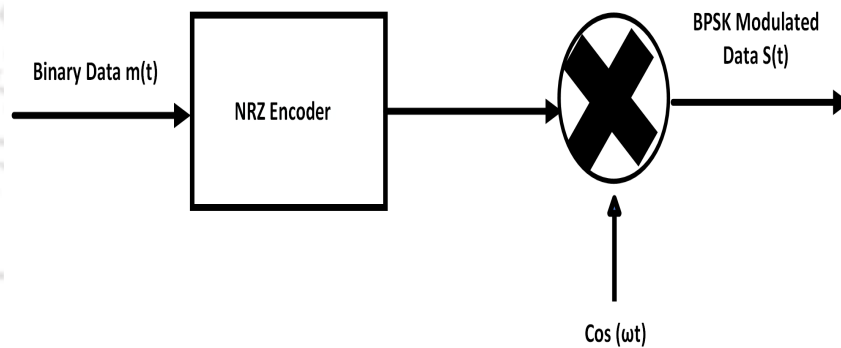


FIGURE 2.8: Generation of BPSK modulated signal

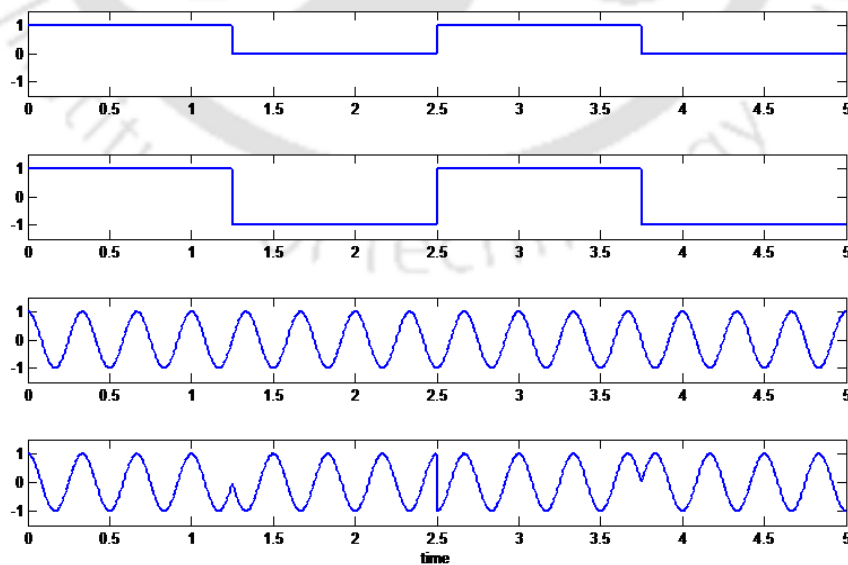


FIGURE 2.9: BPSK signal representation of the binary data 1 0 1 0

receivers are considered. In the receiver, from the echo signal, the transmitter identity is decoded after separating the chirp signal and the BPSK signal. If the decoded identity is our transmitter of interest, then the corresponding chirp signal is used for the estimation the target parameter. The simulated transmitted LFM signal with BPSK modulated transmitter identity (1 0 1 0) is shown in figure (2.11).

The transmitted signal is synthesized for two transmitters with their BPSK modulated

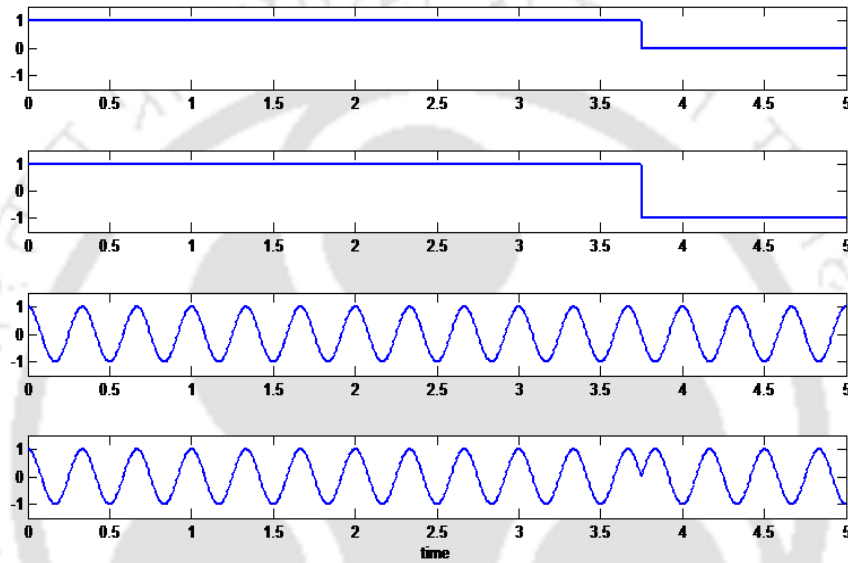


FIGURE 2.10: BPSK signal representation of the binary data 1 1 1 0

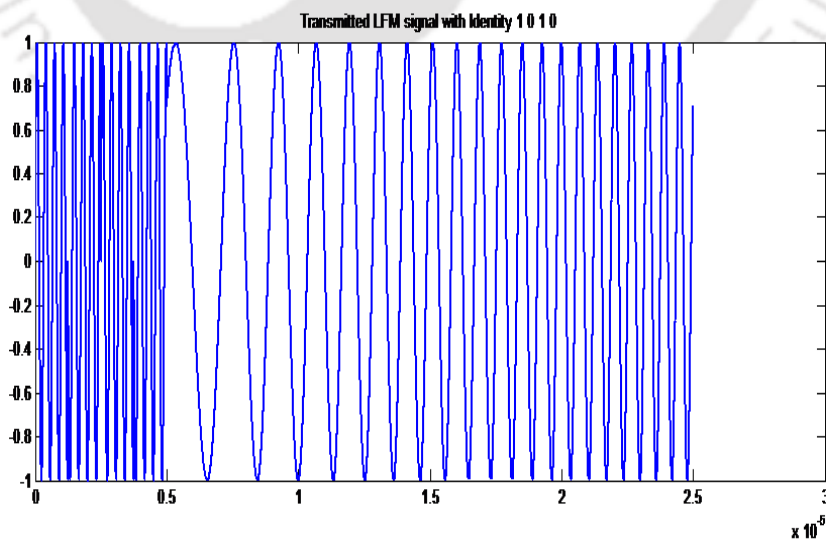


FIGURE 2.11: Transmitted signal (LFM with BPSK modulated identity of transmitter 1)

identity along with the chirp signal is shown in figure (2.12). Synthesised echo signal of figure (2.12) along with delay and noise is shown in figure (2.13).

In the receiver, the noise corrupted transmitter identity signal part (BPSK) is separated from the LFM signal. The separated BPSK modulated identity is shown in figure(2.14) and the LFM is shown in figure (2.15). By using the coherent demodulation technique, target identity is detected from the noisy BPSK modulated signal as shown in figure (2.16)

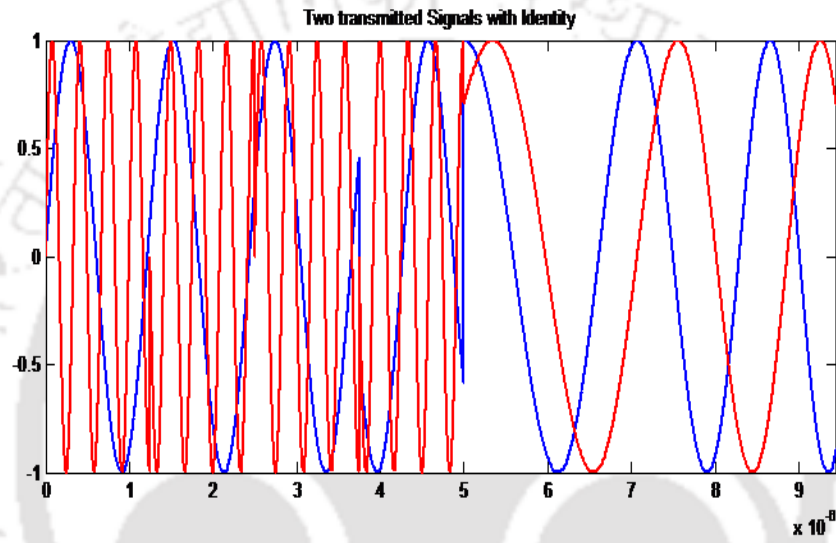


FIGURE 2.12: Transmitted signal of both the transmitters with identity)

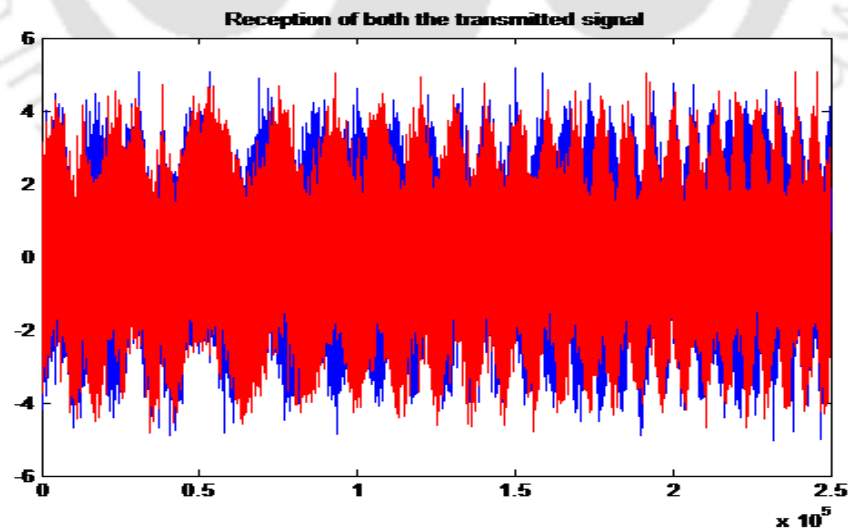


FIGURE 2.13: Reception of both the transmitted signal with identity by a receiver

In BASK technique corresponds to switching the amplitude of the carrier signal in accordance with the modulating (message) bits. Thus, the transmitter identity bits are modulated with the linear frequency carrier signal using BASK technique. In the BASK modulated signal shown in equation (2.27), when the message bit is one, LFM signal is transmitted and when it is zero no signal is transmitted [16]. In this modulation technique, the frequency and the associated phase of the carrier waveform are varied accordingly as per the position of LFM signal in each bit representation.

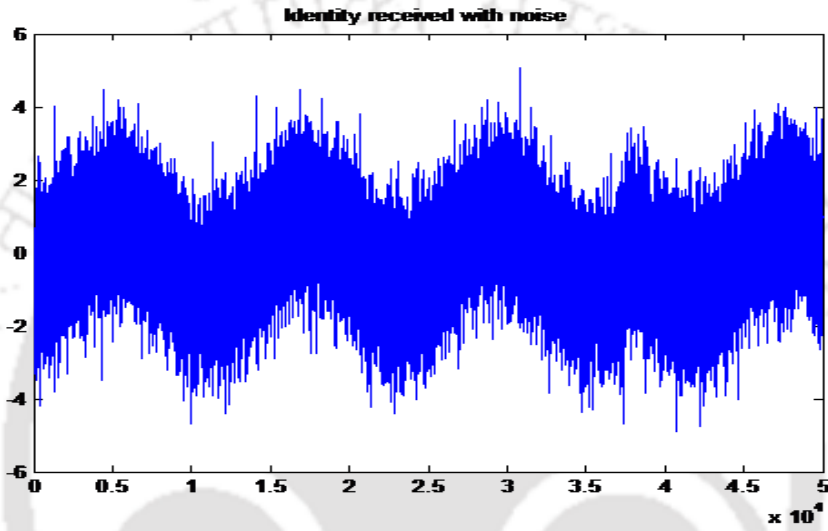


FIGURE 2.14: Separation of the BPSK modulated identity from the received signal

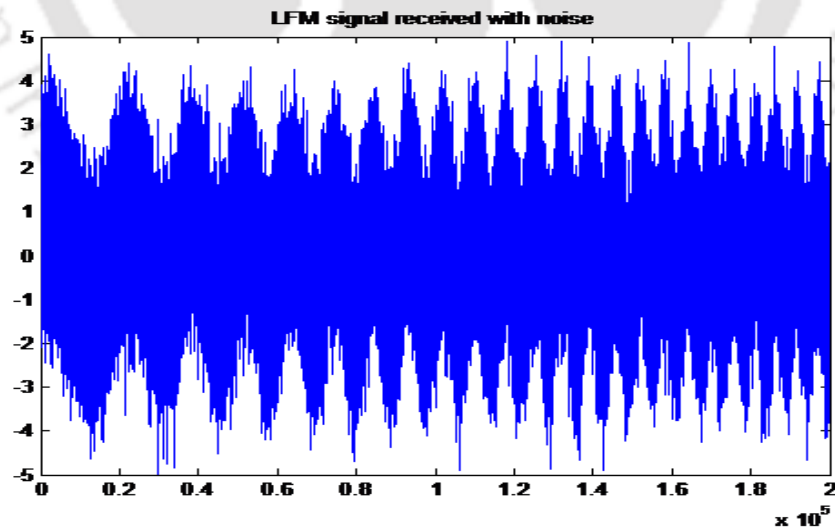


FIGURE 2.15: Separation of the LFM signal from the received signal

Mathematically, BASK modulated LFM can be defined as:

$$S(t) = \begin{cases} \cos(2\pi f_1 t + \pi k t^2) & \text{binary 1} \\ 0 & \text{binary 0} \end{cases} \quad (2.27)$$

Algorithm to transmit and estimate the transmitter identity To Generate the BASK modulated transmitter identity

Step 1. Convert the given four bit transmitter identity in to twelve bits using the technique of repetition codes.

Step 2. Generator polynomial $1 + x + x^2$ is taken so as to satisfy the properties of PN sequence. In our case out of 15 bits being generated, first 3 LSB bits are truncated. Then rest of the 12 bit PN code is XOR with 12 bits obtained in step 1.

Step 3. BASK signal is generated for the resultant 12 bits obtained in step 2 by using LFM as a carrier.

Estimation of transmitter identity from received noise corrupted echo.

Step 1. 10μsec received signal is converted in 12 consecutive sections by multiplying with a window of 0.86μsec and FFT are applied on all these sections.

Step 2. Maximum magnitude in each of the FFT spectrum is obtained.

Step 3. The average value of the data obtained in step 2 is computed.

Step 4. The average value obtained in step 3 is compared with the values from step 2 and the values are assigned as the follows

If the value from step 2 $\geq$ average value assign 1 Otherwise assign 0 Thus we estimate the transmitted bit.

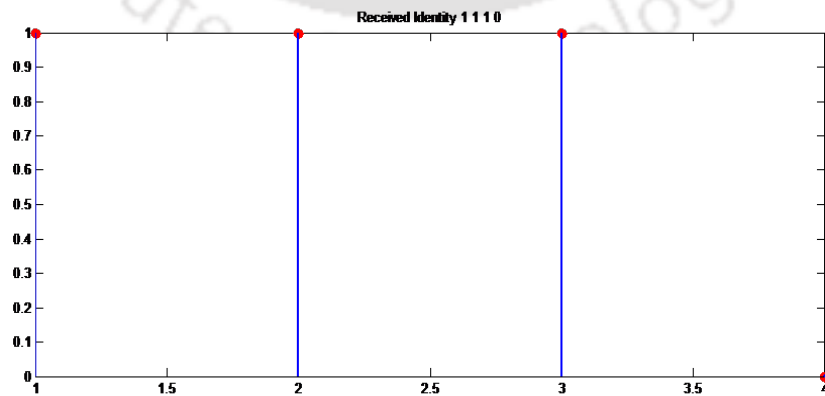


FIGURE 2.16: Detection of transmitter identity of the first transmitted signal from the received signal

Step 5. 12 bits obtained from step 4 is XOR with 12 bit known pseudo code(used in the transmitter)

Step 6. The redundancy embedded with repetition code is used to decode the transmitter identity.

2.4 Target detection

To improve the probability of detection, the threshold should be adapted based on the environment. Constant false alarm rate (CFAR) [55] [56] technique is used to prevent false automatic target detection when radar systems are overloaded with false targets in the presence of an unknown interference environment or clutters. This technique is also used to keep the number of false alarms within the limit, for a particular operating scenario. Cell averaging constant false alarm (CA-CFAR) [18] technique is typically a moving average technique. Left and right training cells are used to find out the adaptive threshold for target detection. There is one guard cell adjacent to cell under test (CUT) both on left and right side as shown in figure (2.17). The main advantage of guard cells are if a target is spread across more than one range cells, the target strength information will not be included in calculating adaptive threshold.

The threshold level is fixed to detect target returns against a background of noise, clutter, and interference. If the received signal is above the threshold, the target is detected.

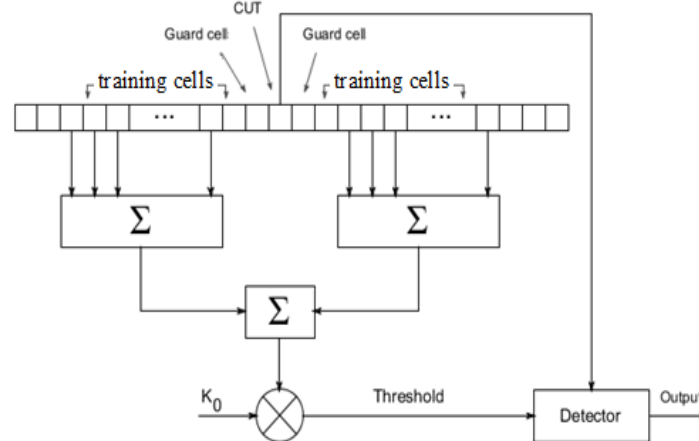


FIGURE 2.17: Block diagram of cell averaging constant false alarm (CA-CFAR) technique

CFAR threshold is computed using the equation (2.28), however improved correlation technique [19] is used to detect the targets and clutter by using probability of false alarm (P_{fa}) and standard deviation (σ) of the samples.

$$T = \sqrt{-2 \times \sigma^2 \times \ln P_{fa}} \quad (2.28)$$

Detected target's range will be estimated whether in a mono-static case or bi-static scenario by estimating time delay(τ), this is based on the position of the range bin containing the target.

2.5 Data fusion

Multi-sensor data fusion refers to the combination of information gathered by various sensors to provide a better understanding of the phenomenon under consideration [57]. Basically, the fusion methods are classified as direct and indirect fusion. In direct method of fusion, raw measurements are combined while, in indirect method, transformations of measurements are fused [58]. The Bayesian statistics approach with Maximum Likelihood Estimator (MLE) is a simple direct fusion method. The MLE states that the desired probability distribution is the one that makes the observed data most likely, which means that one must see the value of the parameter vector that maximizes the likelihood function [59][60].

In another way of definition multi-sensor data fusion [61] refers to associating and combining information gathered by various knowledge sources and sensors to provide a better understanding of the phenomenon under consideration, where all the sensors are independent [58]. When two sources of information from two independent sensors, pertaining to a single target, are given to the central processor, the information is fused using some fusion technique to obtain the optimized result. Here Maximum likelihood estimation (MLE) algorithm is used for the fusion. Likelihood is the joint conditional probability of the measurements for example z_1 and z_2 by two receivers in association with the true value x .

In this section, we describe the maximum likelihood and Bayes law for fusion. Assume that the measurement vector for one receiver in a given observation time t_k be $X = (x_1, x_2, \dots, x_n)$ and the target measurements (e.g. position, velocity) being estimated are θ . Our aim is to find the probability of the existence of target measurements in the available measurements ($p(\theta/X)$) with the help of prior probability ($p(X/\theta)$) that the existence of available measurements at a given target measurements (track). This is obtained using Bayes Law, as shown in equation (2.29) by knowing the probability of the measurement data $p(X)$ and the probability of the track data $p(\theta)$.

$$p(\theta/X) = \frac{p(X/\theta)p(\theta)}{p(X)} \quad (2.29)$$

Assume that in our system, there are k receivers, combining all the receiver measurements observed up to t_k time instant be $X = (X^1, X^2, \dots, X^k)$. When all the X vectors are independent, then the corresponding posteriori likelihood is given by equation (2.30)

$$p(X/\theta) = \prod_{i=1}^k p(X^i/\theta) \quad (2.30)$$

To make the computation simple, logarithmic function is taken to the likelihood function as shown in equation (2.30) the resulting function is given by equation (2.31)

$$L(\theta) = \log[p(X/\theta)] = \sum_{i=1}^k \log[p(X^i/\theta)] \quad (2.31)$$

Assuming all the measurements by all receivers follows independent Gaussian distribution. Thus their joint probability is given by [58] equation (2.32)

$$p(X^i/\theta) = \frac{1}{(2\pi)^{\frac{n}{2}}(|C_i|)^{\frac{1}{2}}} e^{-\frac{1}{2}(X^i-\theta)^T C_i^{-1}(X^i-\theta)} \quad (2.32)$$

Where C_i denotes the covariance matrix. Substitute equation (2.32) in equation (2.31) to get $L(\theta)$. The maximum likelihood ($\hat{\theta}$) of θ may be achieved by equating the derivative of $L(\theta)$ to zero as shown in equation (2.33).

$$\frac{d}{d\theta}L(\theta) = 0 \quad (2.33)$$

(i.e.) the best estimate $\hat{\theta}$ (fused estimate) of θ can be found by differentiating $L(\theta)$ with respect to θ and equate the result to zero as shown in equation (2.34).

$$\hat{\theta} = \frac{\sum_{i=1}^k (\sigma_i^2) X^i}{\sum_{i=1}^k (\sigma_i^2)^{-1}} \quad (2.34)$$

In case the number of receivers are five, then the fused estimate ($\hat{\theta}$) may be found as shown in equations (2.35 or 2.36) by considering $i = 1, 2, 3, 4, 5$ in equation (2.34).

$$\hat{\theta} = \frac{(\sigma_1^2)^{-1}X^1 + (\sigma_2^2)^{-1}X^2 + (\sigma_3^2)^{-1}X^3 + (\sigma_4^2)^{-1}X^4 + (\sigma_5^2)^{-1}X^5}{(\sigma_1^2)^{-1} + (\sigma_2^2)^{-1} + (\sigma_3^2)^{-1} + (\sigma_4^2)^{-1} + (\sigma_5^2)^{-1}} \quad (2.35)$$

$$\hat{\theta} = \frac{\sigma_2^2\sigma_3^2\sigma_4^2\sigma_5^2X^1 + \sigma_1^2\sigma_3^2\sigma_4^2\sigma_5^2X^2 + \sigma_1^2\sigma_2^2\sigma_4^2\sigma_5^2X^3 + \sigma_1^2\sigma_2^2\sigma_3^2\sigma_5^2X^4 + \sigma_1^2\sigma_2^2\sigma_3^2\sigma_4^2X^5}{\sigma_1^2 + \sigma_2^2 + \sigma_3^2 + \sigma_4^2 + \sigma_5^2} \quad (2.36)$$

In our case, let noisy estimated range of the target by receivers 1 and 2 (using equation (2.4)) be R_{tx1} and R_{tx2} respectively. Then the conditional probability of the estimated ranges R_{tx1} and R_{tx2} in association of the true range R_T may be represented by $P(R_{tx1}/R_T)$ and $P(R_{tx2}/R_T)$. Assuming that the noise in the estimated ranges follows the Gaussian distribution with equal mean range R_T and different variances (σ_1^2) and (σ_2^2) then, the conditional probability is given by, equation (2.37) and equation (2.38).

$$P(R_1/R_T) = \frac{1}{\sqrt{2\pi\sigma_1^2}} \exp \left(-\frac{(R_{tx1}-R_T)^2}{2\sigma_1^2} \right) \quad (2.37)$$

$$P(R_2/R_T) = \frac{1}{\sqrt{2\pi\sigma_2^2}} \exp \left(-\frac{(R_{tx2}-R_T)^2}{2\sigma_2^2} \right) \quad (2.38)$$

Likelihood function $L(R_T)$ is the joint conditional probability represented by equation (2.39)

$$L(R_T) = P(R_{tx1}, R_{tx2}/R_T) \quad (2.39)$$

Assuming R_{tx1} and R_{tx2} are independent then Likelihood can be shown as equation (2.40)

$$L(R_T) = P(R_{tx1}/R_T) \times P(R_{tx2}/R_T) \quad (2.40)$$

Substituting equation (2.37) and (2.38) we get equation (2.41)

$$L(R_T) = \frac{1}{2\pi\sigma_1\sigma_2} \exp \left[-\frac{(R_{tx1}-R_T)^2}{2\sigma_1^2} - \frac{(R_{tx2}-R_T)^2}{2\sigma_2^2} \right] \quad (2.41)$$

Maximum likelihood estimator $(L(R_T))_{mle}$ of R_T is the value of R_T which maximizes the likelihood function. This MLE of R_T obtained is equation (2.42)

$$\hat{R}_t = \frac{(R_{tx1} - R_T)^2}{2\sigma_1^2} + \frac{(R_{tx2} - R_T)^2}{2\sigma_2^2} \quad (2.42)$$

2.6 Simulation result and discussion

The simulated transmitted and received LFM signals are shown in figures (2.18) and (2.19). Basic parameters that are assumed for synthesizing the LFM signal is given in Table 2.1 Similarly, parameters required to calculate the SNR, as in equations (2.3) and (2.22), are shown in Table 2.2.

2.6.1 Coverage area in a monostatic and multi-static scenario

In monostatic scenario, time delay (τ) may be obtained by correlating the transmitted and the received signal. Using range delay equation in [1], radar range (R) may be calculated. Mono-static radar [1] coverage area is computed constant SNR circle is plotted with the radar position as the center point and the range (R) as a radius. Coverage area for mono-static scenario, for different SNRs is given in figure (2.20).

TABLE 2.1: System parameters of simulation

Range	R	80 to 200 km
Transmitter Azimuth Angle	θ_T	45 deg
Transmitter Elevation Angle	θ_R	40 deg
Initial Frequency Component	f_L	100 Hz
final Frequency Component	f_H	250 kHz
Pulse Duration	T_P	20 ms
First Baseline distance	L1	40 km
Second Baseline distance	L2	15 km

Using equation (2.4 and 2.5), R_T and R_R are found for the estimated Bi-static range ($R_T + R_R$). SNR of the Bi-static radar is estimated using equation (2.3). Bi-static angle (β) is calculated using equation (2.6). Thus the coverage area of Bi-Static radar with different SNR values is shown in [4]. It has been observed that the coverage area of the Mono-static radar is a circle as shown in figure (2.20) and for the Bi-static radar, it is the oval of Cassini.

Similarly, for a multi-static radar [60], coverage area with one transmitter and two receivers and with three receivers are shown in figures (2.21) and (2.22). Using equation (2.22), SNR is estimated as a sum of the coverage areas of the bi-static radars in the multi static scenario. The simulation results for different system geometry of multi-static radars show that, as the number of transmitters and the receivers increase the coverage area of the multi-static radar also increases. It is also observed that the overall coverage area in a multi-static scenario is the combination of individual receiver's Bi-static area with respect to the transmitter.

2.6.2 Transmitter identity generation for transmitted waveform

Algorithm to estimate the transmitter identity generation of ASK modulated transmitter identity is presented, considering a four bit transmitter identity

Example: Four bit transmitter identity (Tx ID)= [1 0 1 1]

TABLE 2.2: Basic parameters of simulation

Transmitting Power	$P_T(\text{dBW})$	10dBW
Transmitting antenna gain	$G_T(\text{dB})$	8dB
Receiving antenna gain	$G_R(\text{dB})$	8dB
Transmitting frequency	$f(\text{GHz})$	10 GHz
Radar-Cross-section(RCS)	$\sigma(\text{dBsqm})$	10dBsqm
Propagation factor for the Tx-path to target	$F_T(\text{dB})$	-3.5dB
Propagation factor for the Rx-path from target	$F_R(\text{dB})$	-3.5dB
Temperature(Room Temp)	$T_s(\text{K})$	300K
Receiver Noise-bandwidth	$B_n(\text{MHz})$	60MHz
Overall System loss	$L(\text{dB})$	1dB

Step 1 :

a) Converted to 12 bits using repetition of Tx ID = [1 1 1 0 0 0 1 1 1 1 1 1]

b) Generated 12 bit pseudo code (PN Code) = [0 1 1 0 1 1 0 1 0 1 0 1]

Step 2 :

The PN code (Step 1b) XOR with 12 bits Tx ID (Step 1a) = [1 0 0 0 1 1 1 0 1 0 1 0]

Step 3 :

BASK signal, for the resultant 12 bits obtained from step 2, by using LFM as a carrier is as shown in figures (2.23).

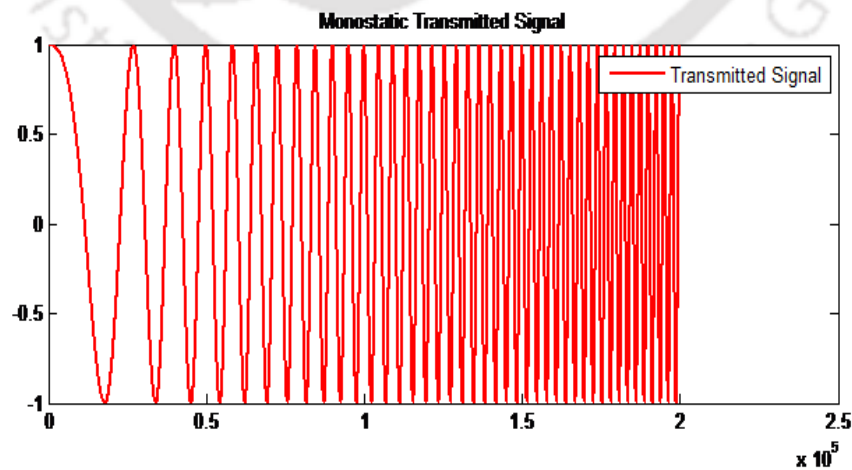


FIGURE 2.18: Transmitted LFM signal

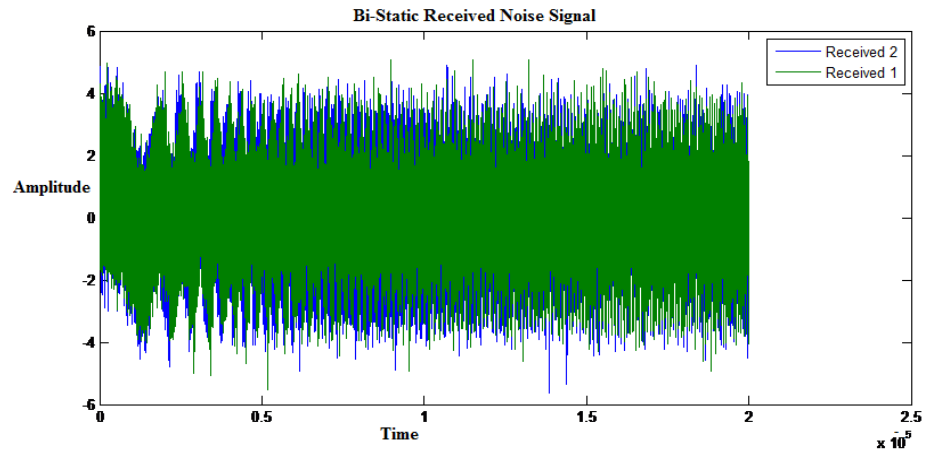


FIGURE 2.19: Received LFM signal

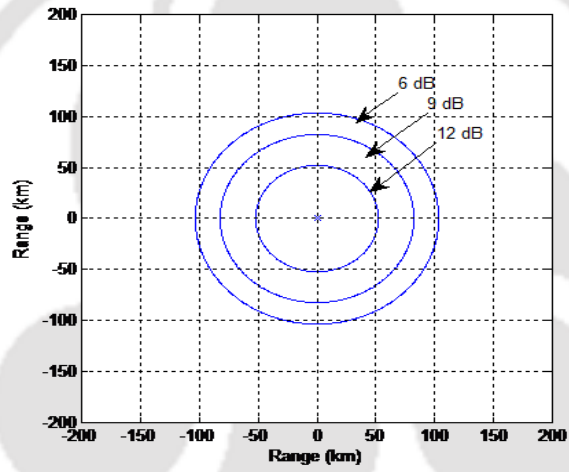


FIGURE 2.20: Monostatic coverage area

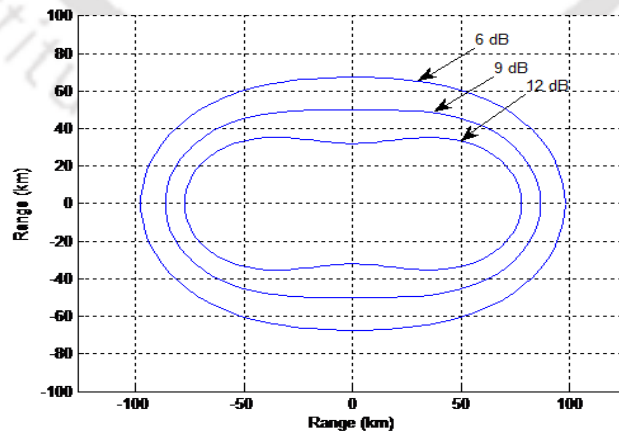


FIGURE 2.21: Multi static coverage area (with two receivers and one transmitter)

2.6.3 Estimation of transmitter identity from received noise corrupted echo

The algorithm consist of the following steps:

Step 1. $10\mu\text{sec}$ received signal shown in figure (2.24) is converted in 12 consecutive

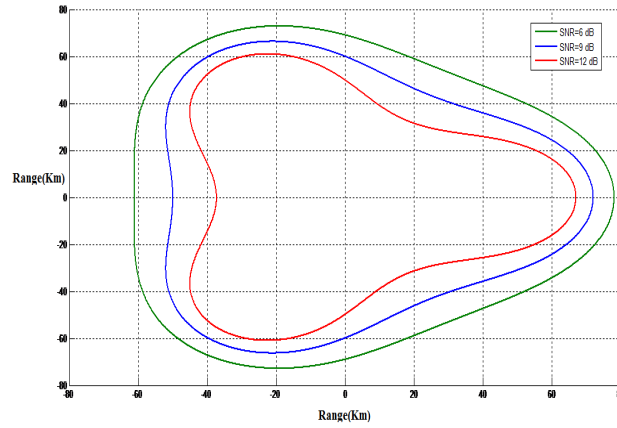


FIGURE 2.22: Multi static coverage area (with three receivers and one transmitter)

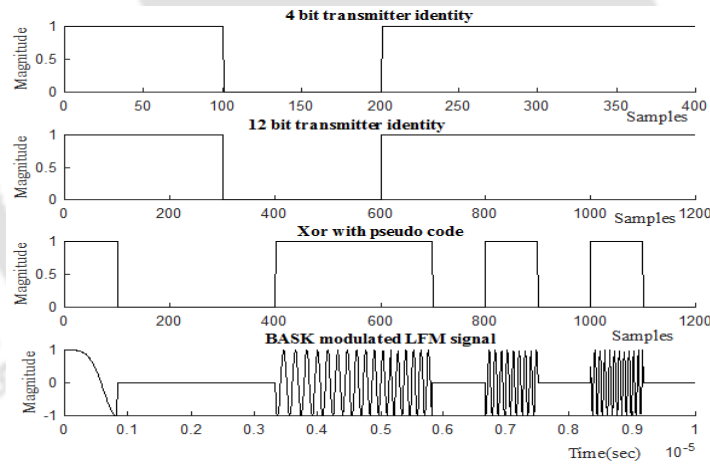


FIGURE 2.23: Modulated transmitted signal with identity

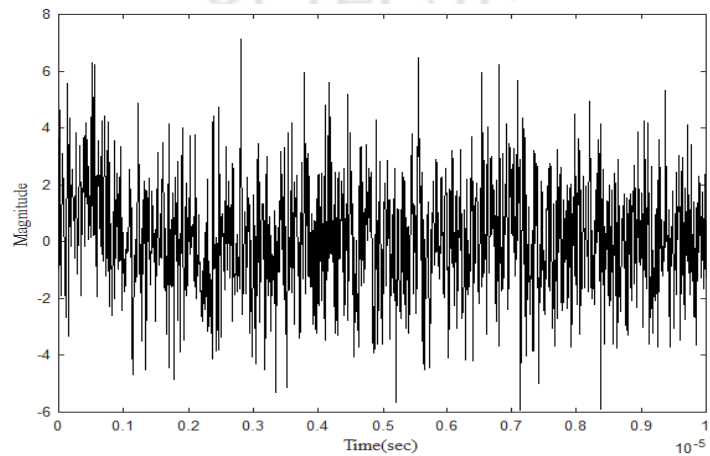


FIGURE 2.24: Received signal with transmitter identity

sections by multiplying with a window of $0.86 \mu\text{sec}$ and FFT are applied on all signals.

Step 2. Maximum magnitude of the each FFT spectrum is given as : 105 47 43 39 58 70 75 38 94 45 83 38.

Step 3. Average value of the data obtained in step 2 is computed = 61

Step 4. Estimated 12 bit binary code by comparing the values from Step 2 with the average value as: 1 0 0 0 0 1 1 0 1 0 1 0

Step 5. 12 bits obtained from step 4 is XOR with 12 bit known pseudo code(used in the transmitter) to get: 1 1 1 0 1 0 1 1 1 1 1 1

Step 6. Estimated the 4 bit transmitter identity using repetition decoder =1011

To validate the above algorithm, Monte Carlo simulation was run for 1000 times with 10 trials to get the average probability of detection (P_d) for each SNR and is plotted in figure (2.25). It is observed that even with 0dB SNR our proposed algorithm shows greater than 0.8 P_d . In general, typical radar system will operate with a P_d of 0.9 and a probability of false alarm (P_{fa}) order of 10^{-6} . The required SNR is 13 dB, as can be read from the graph given in reference [18]. It is observed from the Table 2.3 that even with 11dB SNR, approximately $P_d = 100\%$ is achieved. Since, the algorithm is indented for the detection of transmitter identity in the radar systems, the concept may be very well introduced in radar system to overcome the requirement of multi channel receiver in a multi static scenario. The details of simulation result is tabulated in Table 2.3

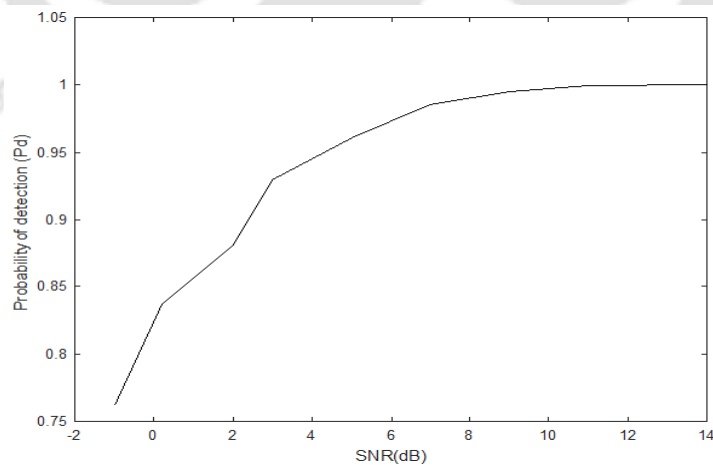


FIGURE 2.25: SNR verses probability of detection of the transmitter identity

TABLE 2.3: Detection probability(P_d) for various SNR-Level

(Monte Carlo simulation for 1000 runs repeated 10 times)									
Transmitter identity detection for 1000 execution at different SNR									
Repetition	14dB	11dB	9dB	7dB	5dB	3dB	2dB	0.2dB	-1dB
1	1000	1000	994	984	961	932	865	850	784
2	1000	998	992	991	954	943	897	821	752
3	1000	1000	998	988	976	929	876	858	755
4	1000	1000	994	988	965	934	879	833	759
5	1000	999	999	975	966	922	901	835	750
6	1000	1000	993	982	960	927	897	847	783
7	1000	1000	997	986	951	918	884	816	755
8	1000	997	995	982	963	934	867	839	761
9	1000	999	993	991	966	925	865	823	767
10	1000	1000	993	986	948	930	874	845	755
Avg	1000	999.3	994.8	985.3	961	929.4	880.5	836.7	762.1
P_d	1	0.9993	0.9948	0.9853	0.961	0.9294	0.8805	0.8367	0.7612

2.6.4 Target detection using CA-CFAR

LFM is used to increase the range resolution and coverage area. To simulate this, it is assumed LFM signal is transmitted with the pulse duration of $10\mu s$ to get a wider coverage area. Then the received echo pulse with $10\mu s$ pulse width is compressed to 200ns and sampled at 5MHz sampling rate to obtain the range resolution of 30 meters. The compressed received signal samples are shown in figure (2.26).

Considering there exist 50 samples and each of the sample may carry information about either the target, clutter or noise. In our simulation as shown in figure (2.27), we have synthesized the samples such that it contains one target information, three clutter information and rest of them are all represent the noise. Let the target amplitude be 0.0035 and the clutter amplitudes are 0.004, 0.0055 and 0.006. The noise added to the signal is random noise with a peak value of 0.003

Threshold (T) is calculated using equation (2.28). By fixing the probability of false alarm (P_{fa}) as 10^{-4} [4] along with the estimated standard deviation ($\sigma=0.0006$) of the received samples, threshold is estimated as 0.0032. Thus, the threshold is more than noise level and less than the target and clutter values. Therefore, the targets are identified as shown in figure (2.27). It is also observed that even though the synthesized received signal has only one target, figure (2.27) shows four targets. The other three are due to clutter. RCS of the detected targets are found using equation (2.2) for the known SNR and other parameters and is compared with the RCS of the clutter database given in the reference [20]. When the RCS matches with the data then it indicates that it is clutter return. Otherwise, it will be considered as a target return. Thus, figure (2.28) shows the final detected target after removal of the clutter.

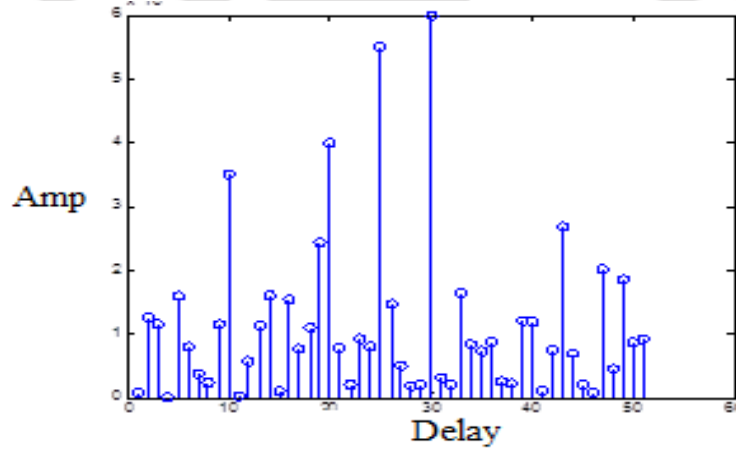


FIGURE 2.26: Compressed signal impulses

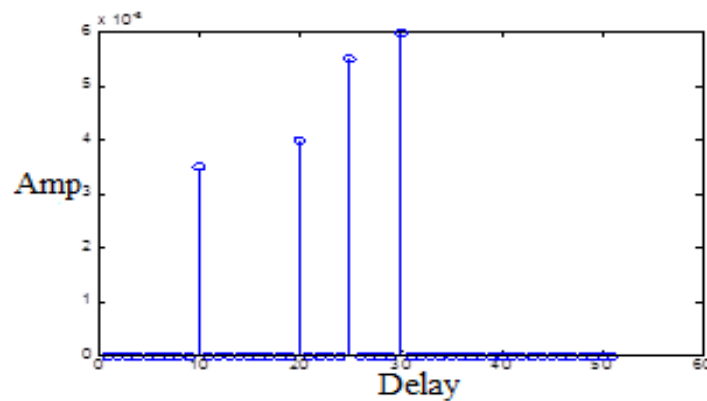


FIGURE 2.27: Target detection based on improved correlation technique

2.6.5 Target tracking using MLE fusion

Target tracking is achieved by detecting the same target for every time instant. In our simulation, it has been assumed that there exist only one target and all the receivers are tracking the same. In this, we used MLE technique (described in Data fusion section) to fuse the tracks by the multiple receivers. figure (2.29) shows a comparison of the actual track with the estimated tracks by two receivers and the fused track. It shows that fused track is closer to the actual track.

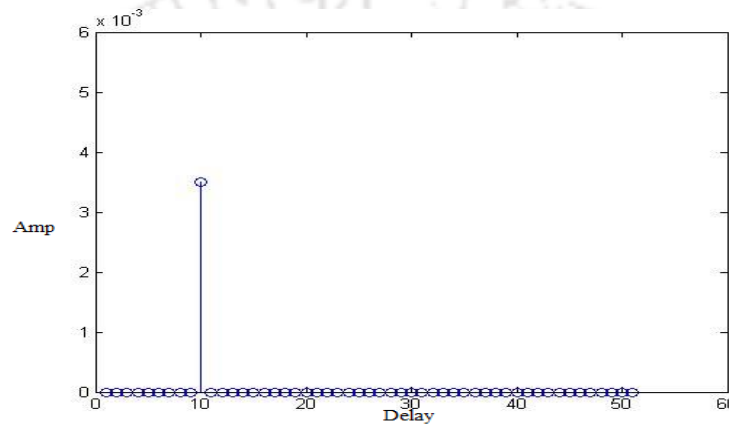


FIGURE 2.28: Target detection after removing the clutter

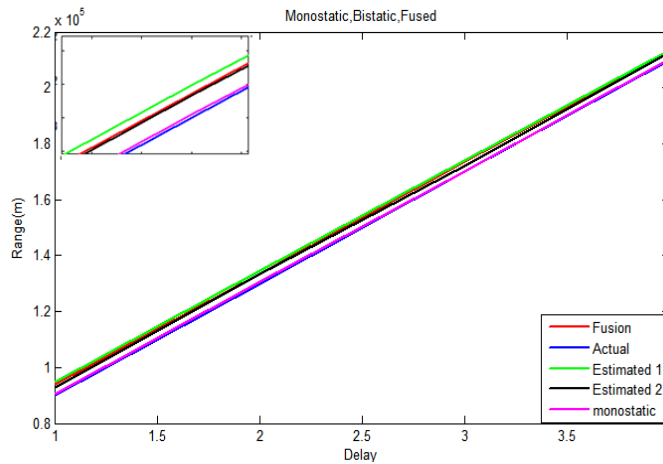


FIGURE 2.29: Comparison of the actual track with the estimated track by individual receiver and fused track in bi-static scenario and estimated track in mono-static scenario (one Tx and two Rx)

2.7 Summary

A passive radar receiver must use a dedicated receiver channel to monitor each transmitter to exploit the multi-static scenario. To overcome this, we have developed an algorithm to introduce a transmitter identity in the transmitted signal. To validate the algorithm, Monte Carlo simulation was performed to find detection probability (P_d) of identity at different SNR. It is observed that up to 11dB SNR, $P_d=0.99$ and at 0dB SNR, a P_d of 0.8.

In general, radar system should maintain 13 dB SNR to operate it with a P_d (target detection) of 0.9 and P_{fa} of 10^{-6} . Thus the proposed algorithm will surely be suitable in a multi-static radar system to track a target through the transmitter identity. To validate the target parameter estimation using the transmitter identity, we have concentrated on radar coverage area, single target detection with multiple receivers and finally obtained the optimized tracking using MLE based fusion.

In general, for the mono-static radar it is observed that the estimated range is approximately equal to the actual range. But in bi-static radar, as the distance between the transmitter and the receiver increases the range error also increases. In future, better techniques need to be developed to reduce this error. Similarly, MLE data fusion gives better optimization only when noise is Gaussian. When the noise is non-Gaussian, better fusion techniques have to be identified. In this chapter only one target is considered with two receivers. In case of multiple targets scenario with multiple receivers, similar target information need to be identified using appropriate association techniques.

3

Investigation on multi-dynamic radar extension of multi-static radar

This chapter describes the complete multi-dynamic radar state of the art concept for airborne surveillance. Investigation on the multi-dynamic radar is presented in the form a case study on the usage of SSR waveform for waveform identification and, subsequently the signal processing algorithm for target detection. Investigation of various multi-dynamic radar configurations in the STAP with various anti clutter processing methodologies is also derived and shown in this chapter.

3.1 Introduction

A non-absorbent stealth target or aircraft scatters electromagnetic energy in all possible directions unequally such that the energy backscattered into the direction of the radar receiver is less or none. Mono-Static or Bi-static radar system may not be viable solution to detect such kind of stealth target however, in such scenario multiple mono-static or multi-static radar systems can gather the diverted electromagnetic energy to detect and track the stealth aircraft. This type of radar system has multiple transmitters and receivers located at different locations. There are numerous advantages to such a surveillance system, such as large coverage area, as well as immunity to jamming and spatial diversity [62] [63]. A review has been made by H. D. Griffiths [9] pertaining to the RCS analysis and ambiguity function of bi-static radar. The detection range of a bi-static radar depends on the location of the transmitter and receiver at constant SNR. It is characterized by the Cassini oval with the transmitter and receiver being the two foci [6]. A Cassini oval is a quadratic plane curve defined as the set of points in the plane such that the product of the distances between the point on the curve to the individual foci is constant [9]. Coverage area of monostatic radar becomes a circle due to the co-location of transmitter and receiver. The multi-dynamic radar coverage area is obtained as a sum of bi-static or mono-static coverage area depending on the system geometry [14].

This chapter focuses on the investigation for the analysis of the system geometry of bi-static and multi-dynamic radar systems, along with the coverage area for constant SNR. It also focuses on the OFDM(Orthogonal Frequency Division Multiplexing)-LFM(Linear Frequency Modulation) based ambiguity function for waveform design as well as detection and tracking of the target.

According to the IEEE standard radar definition (IEEE standard 686-1997), Bi-static radar is a radar system that has separate antennas for transmission and reception at sufficiently different locations to get the potential advantage in the detection of stealthy targets. These stealth targets are designed to scatter energy in the directions away from the radar receiver.

3.2 Multi-dynamic radar system model

As discussed in the earlier chapter the detail about Bi-static radar. The scenario will be more and more complex for detection and classification of the target when more and more targets and sensors come into the picture. Multi-dynamic radar systems are proposed to be used for Airborne surveillance systems. So that it will have better coverage and visibility compared to multi-static radar. Multi-dynamic radar systems have better detection capability as well as tracking for low rcs targets or stealth targets due to their adaptive dynamic geometry which, helps detection ability as well as coverage area as on required. There is no standard fixed signal model for multi-dynamic radar as due to the dynamic of the platform the statistical characteristic of signals changes rapidly and randomly. Multi-dynamic is similar to multi-static or multi-site Radar systems at a given instant of time as shown in figure(3.1). The multi-dynamic radar has advantages to counter the detection of non-cooperative targets, High accuracy of target localization, more Target Resolution capability, more Jammer resistance and more airborne clutter resistance.

The multi-dynamic radar system is considered coherent airborne radar-network, such that the whole radar-network is time and phase synchronized as well as works cooperatively. Each receiver can receive echoes of any waveform transmitted by any transmitter in the network. it is assumed m number of airborne transmitters(Tx) and n number of airborne receivers(Rx) geometrically apart in the network, the overall SNR with summing

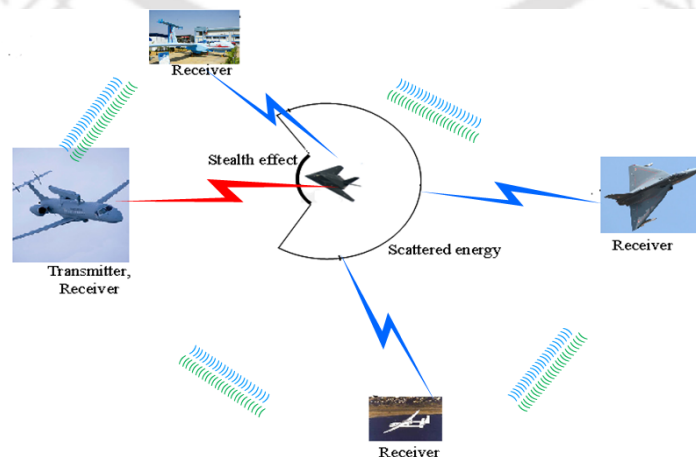


FIGURE 3.1: Multi-dynamic radar system scenario

up partial (individual Tx-Rx combination) SNR [4] as shown in equation (3.1)

$$(\text{SNR})_{\text{overall}} = \sum_{i=1}^m \sum_{j=1}^n \frac{P_{ti} G_{ti} G_{rj} \lambda_i^2 \sigma_{ij}^2 F_{ti}^2 F_{rj}^2}{(4\pi)^3 k T_S B_{Rj} R_{ti}^2 R_{rj}^2 L_{ij}} \quad (3.1)$$

Coverage-area(SNR) for multi-dynamic radar (MDR) system, depends on the considered geometry formed i.e. when the number of airborne transmitters and receivers increases then the combined coverage-area(oval of cassini) of the MDR system also increases. In a special case of consideration when a similar type of airborne transmitters and receivers are used in a multi-dynamic radar system, then the estimation of Signal-to-noise ratio(SNR) is given in equation (3.2) may be simplified as

$$(\text{SNR})_{\text{overall}} = \frac{P_{ti} G_t G_r \lambda^2 \sigma^2 F_t^2 F_r^2}{(4\pi)^3 k T_S B_{Rj}} \sum_{i=1}^m \sum_{j=1}^n \frac{1}{R_{ti}^2 R_{rj}^2 L_{ij}} \quad (3.2)$$

where, P_{ti} = i^{th} Radar Transmitted power and Gain

G_{ti} = i^{th} Radar Antenna Tx Gain

G_{rj} = j^{th} Radar Antenna Rx Gain

σ_{ij} = RCS of the target for i^{th} Tx and j^{th} Rx

λ_i = i^{th} wavelength

R_{ti} = distance of i^{th} Tx to target

R_{rj} = distance from j^{th} Rx from target

L_{ij} = Atmospheric loss towards i^{th} and j^{th} path.

In order to analyse the radar performance by plotting the oval-of-Cassini [6] as a function of R_T and R_R , for bi-static triangle geometry $T_x(-a, 0)$, $R_x(a, 0)$ and $P(x, y)$ for a given SNR. It is a plane of the curve, plotted with the locus of $P(x, y)$ point, in such a way the product of the two distances R_T and R_R from P to two foci points Tx and Rx is constant. The maximum coverage is shown in figure(3.2).

$$R_T R_R = b^2 \quad (3.3)$$

$$(x^2 + y^2 + a^2)^2 - 4a^2 x^2 = b^4 \quad (3.4)$$

$$x = r \cos \theta; y = r \sin \theta; \quad (3.5)$$

$$r^4 - 2a^2r^2 \cos(2\theta) + a^4 - b^4 = 0 \quad (3.6)$$

The coverage area of a multi-dynamic radar, consist in the form of Tx-Rx radar pairs as shown as shown in figure(3.3) to figure(3.8). Signal to Noise Ratio (SNR) for MDR is computed as a sum of the set of coverage areas of the bi-static radars [7]. The coverage area of a multi-dynamic radar consisting 2 Tx-Rx radar pairs.

From the simulation results, it is observed that the coverage-area can be enhanced for given SNR , P_d and P_{fa} by introducing more number of airborne radars Tx-Rx. Here we have used a combination of TX-Rx pairs to obtain the overall coverage-area. Mathematical derivation for the $P(x, y)$ is shown in the equation in earlier chapter equations (2.10) and (2.11), it is assumed that the airborne Tx and Rx are placed on the reference-axis having coordinates $(-a, 0)$ and $(a, 0)$. The simulation result is shown in Table 3.3 for various Tx-Rx pair combinations for different SNRs. Here it is assumed there is one Transmitting(Tx) airborne Radar with multiple radar receivers mounted over small UAVs. The assumed parameters is tabulated below.

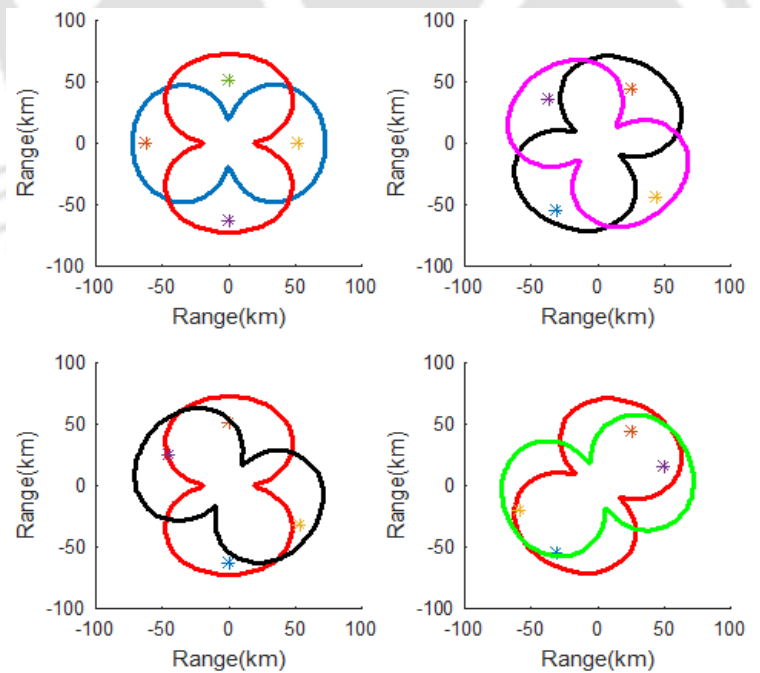


FIGURE 3.2: Multi-dynamic coverage area with 2 Tx-Rx pairs of radars located at different positions

3.3 Target detection with transmitters identity waveform

One of the key problems in the multi-dynamic radar system is the identification of a particular transmitter signal or transmitted waveform parameters from a mixture of signals. Radar and Identification of Friend or Foe (IFF) radar both work in conjunction

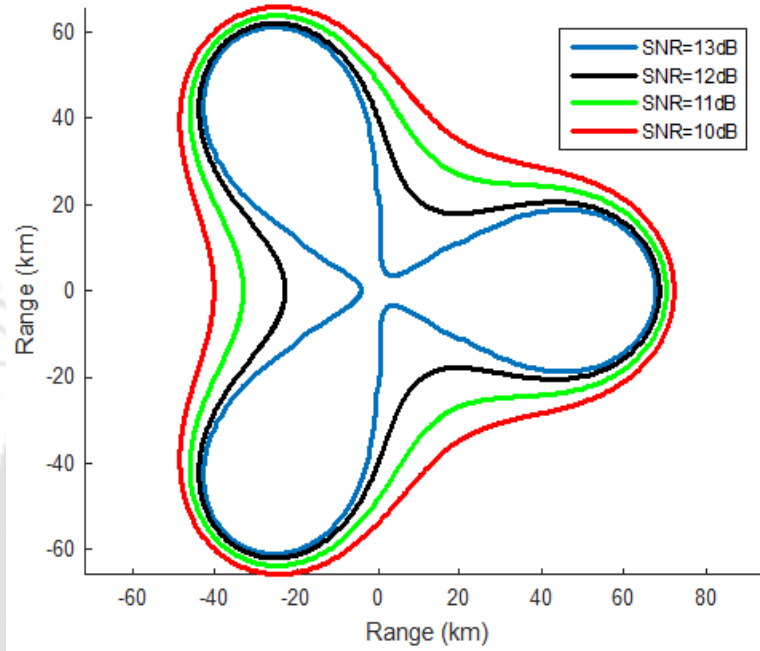


FIGURE 3.3: Three Tx-Rx pairs

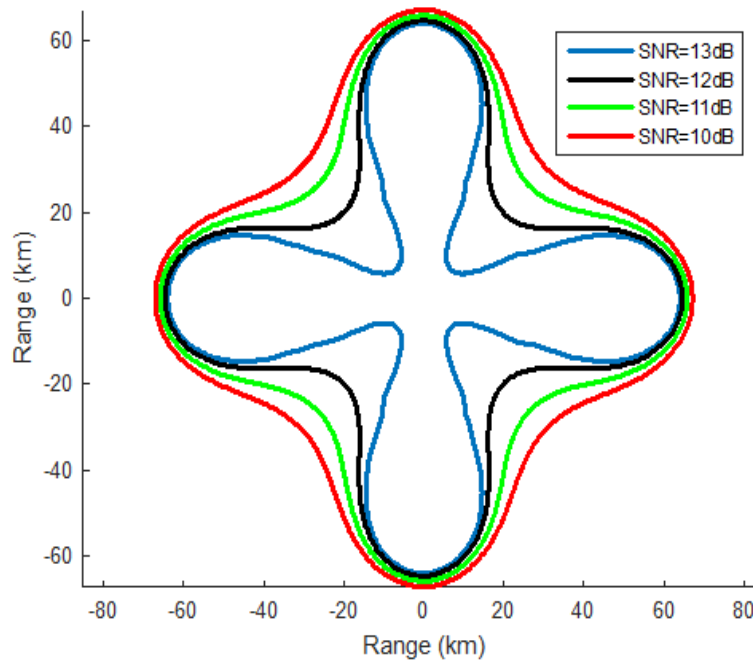


FIGURE 3.4: Four Tx-Rx pairs

for respective non-cooperative and cooperative targets. A method is proposed in which Identification of Friend or Foe (IFF) waveforms is used for Transmitter identity for radar waveform. IFF waveform is detected to decipher the waveform ID of the radar waveform. In general waveform frame format of Mode S mark XII IFF interrogator [64] consists of 56 bits of which 4 bits (10-13) are the interrogator identifier and 16 bits (17-32) are

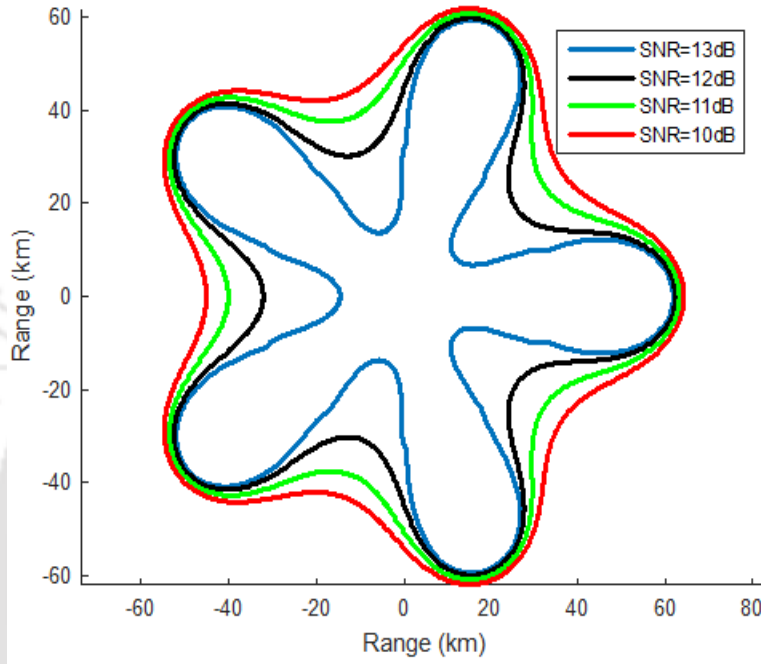


FIGURE 3.5: Five Tx-Rx pairs

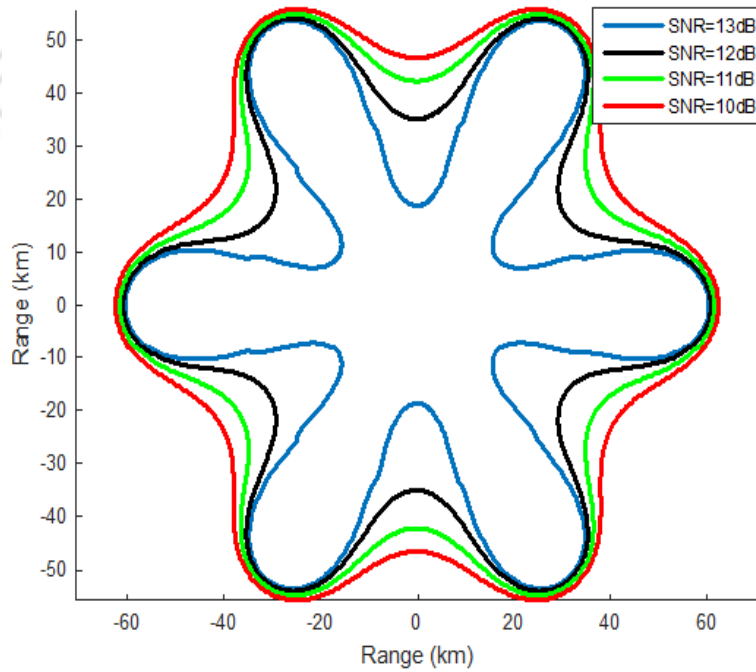


FIGURE 3.6: Six Tx-Rx pairs

the spare bits which is usually left free. These 16 bits are utilized to send the encoded Transmitter waveform identity. The objective is to estimate the Transmitter waveform identity and validate it with the interrogator identifier. Reed Muller algorithm is used to encode the Transmitter waveform identifier. These 56 bits are transmitted using Binary phase shift keying (BPSK) technique. The transponder/receiver will demodulate

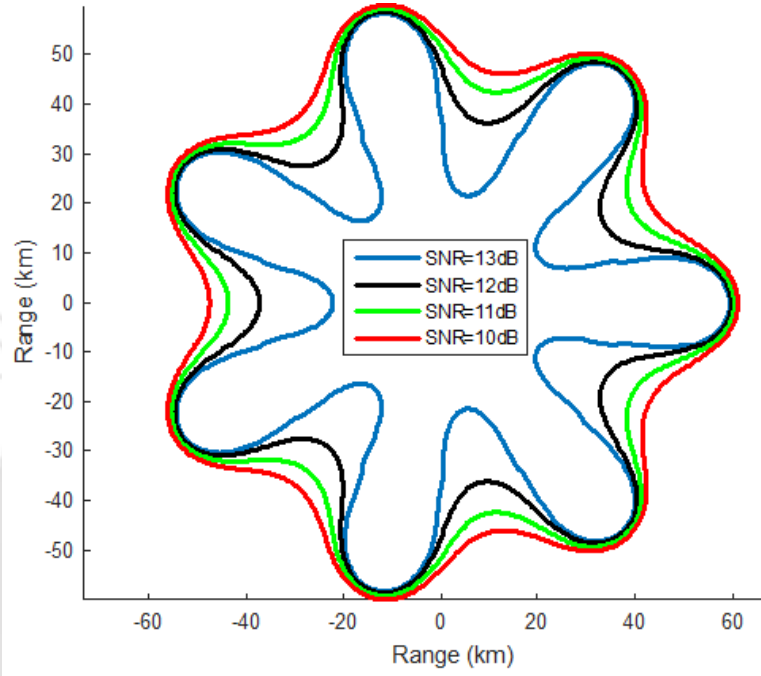


FIGURE 3.7: Seven Tx-Rx pairs

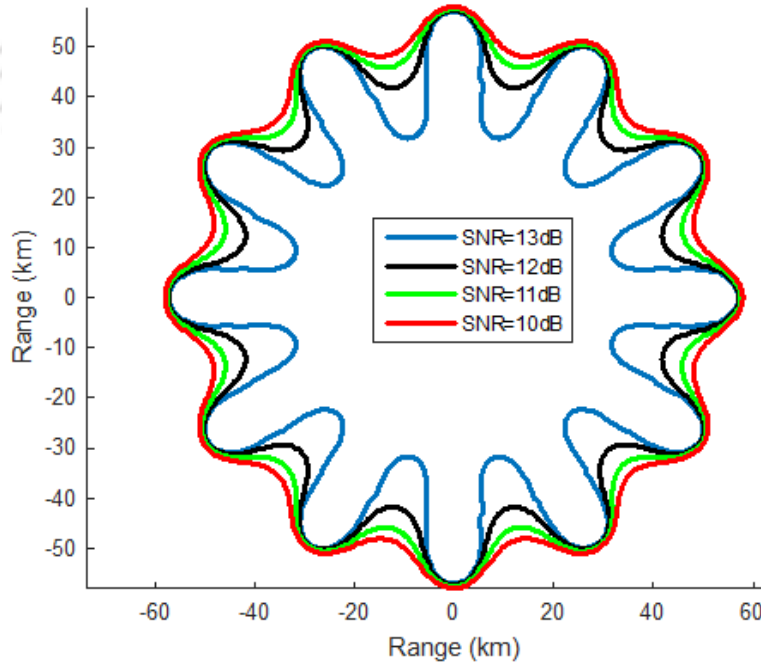


FIGURE 3.8: Twelve Tx-Rx pairs

TABLE 3.1: Simulation parameter

Transmitting Power	$P_T(\text{dBW})$	20 dBW
Transmitting antenna gain	$G_T(\text{dB})$	9 dB
Receiving antenna gain	$G_R(\text{dB})$	9 dB
Transmitting frequency	$f(\text{GHz})$	10 GHz
Radar-Cross-section(RCS)	$\sigma(\text{dBsqm})$	10 dBsqm
Propagation factor for Tx-path	$F_T(\text{dB})$	-3 dB
Propagation factor for Rx-path	$F_R(\text{dB})$	-3 dB
Temperature	$T_s(\text{K})$	300 K
Rx Noise-bandwidth	$B_n(\text{MHz})$	60 MHz
Overl all System loss	$L(\text{dB})$	5 dB

and decode the received signals using Hadamard algorithm to estimate and validate the interrogator identity. Simulated results show that it is possible to estimate the correct Transmitter waveform with up to 3 bits error. Since the probability of detection mainly depends on noise level, the received signal's SNR is also estimated using two standard algorithms, based on the iterative maximum likelihood criterion.

Identification of Friend or Foe (IFF) is the system of interrogator and transponder, which uses the coded signal to identify the friend and enemy aircraft. IFF is used widely in military applications for authenticating the aircraft as friend or foe [65]. The algorithms for the detection of transmitted waveform identity using IFF coded signal are proposed. The IFF signal using a mode S frame format consists of interrogator identifier (II) code [65]. Mode S consists of 56 bits in that 4 bits are used for the interrogator identifier which is used to identify the transmitter. The BPSK IFF coded signals are transmitted. During the long distance transmission, error may occur in the transmitter identity. So the received signal may not be the same as the transmitted signal [66]. To overcome this, the encoding of the IFF signal is done using Reed Muller algorithm. At the receiver side it is decoded using the Hadamard algorithm. The advantage of this decoding method is that even with 3 bit errors in encoded 4 bit messages, it can be retrieved back. In addition to the target signal, the noise present in the received signal may affect the

probability of detection. Thus, the SNR is estimated using two algorithms for better target detection.

3.3.1 Transmitted waveform identity

In Mode S interrogation frame format consists of 56 bits as shown in the below table, is chosen for transmitting the IFF coded signal [67]. The transmitter identity can be any number from (0-15) and the number is converted to four bit binary digits.

TABLE 3.2: Mode-S frame format for IFF

1-5 bits	6-9 bits	10-13 bits	14-16 bits	17-32 bits	33-56 bits
01011	PR	IC	CL	SPARE	AP

All 56 bits in mode S format is divided in to blocks. Each block carries the following binary information

TABLE 3.3: Mode-S 56 bits information

Bit 1-5	conveys whether the interrogator is sending or expecting information
Bit 6-9	(PR) Probability of reply
Bit 10-13	(IC) Interrogator identity (0-15)
Bit 14-16	(CL=000)-indicates the information in the IC field is the interrogator identifier
	(CL=001-100)- indicates the information in the IC field is the Surveillance identifier
Bit 17-32	Spare bits
Bit 33-56	(AP) = all 1's - expect response from all the targets in the region

BPSK modulated signal is used as the transmitted signal in IFF radar. Here it is proposed to generate the BPSK signal using 56 bits Mode S frame format, which consists of transmitted waveform identity. First, including encoded 16 bits transmitter identity in the spare bits block of the Mode S format. Second, synthesizing error prone demodulated Mode S format (56 bits) and decoded to detect the transmitter identity. This BPSK encoded IFF signal is transmitted to identify the waveform identity, then the signal is

received by the transponder, is demodulated and decoded. The simulation has been made for various SNR cases.

3.3.2 Encoding and decoding of interrogator waveform

Four bits transmitted waveform ID is encoded using Reed Muller algorithm. It is one of the oldest multi order error correcting methods in which the first order method is optimized in terms of efficiency and computation [68]. Initially, the four bits identifier encoded to 16 bits using Reed Muller encoding algorithm and are embedded in spare bits block (bits 17-32) of the Mode S frame format.

3.3.3 Encoding using Reed Muller algorithm

General Reed Muller code is represented as $R(r, m)$ where r is the r^{th} order Reed Muller code and m is the number of message bits to be encoded. Encoded binary string length be 2^m . If m number of bits is to be encoded by the Reed Muller matrix then its dimension is given by $k \times 2^m$ [69] where k is given by equation(3.7)

$$k = \sum_{i=0}^r {}^nC_i \quad (3.7)$$

The matrix formation of 1^{st} order polynomial is given by 1^{st} row contains all 1's, 2^{nd} row contains 2^{m-1} 1's followed by 2^{m-1} 0's, k^{th} row consist $2^{m-(k-1)}$ 1's followed by $2^{m-(k-1)}$ 0's repeated until 2^m values have been defined.

The m input message bits are converted $n = m + 1$ bits by padding 0 in MSB. These $[1 \times n]$ bits are multiplied with $[k \times 2^m]$ Reed Muller matrix, to obtained encoded data of length 2^m . For example 1^{st} order 3 bits message Reed Muller matrix $R(1, 3)$ is formed as shown in equation (3.8)

$$R(1, 3) = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ x_1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\ x_2 & 1 & 1 & 0 & 0 & 1 & 1 & 0 \\ x_3 & 1 & 0 & 1 & 0 & 1 & 0 & 1 \end{bmatrix} \quad (3.8)$$

3.3.4 Decoding using Hadamard algorithm

It is a square matrix whose entries are either +1 or -1, with each row are mutually orthogonal [70]. In general, the Hadamard matrix H of order n is an $n \times n$ matrix of 1's and (-1)'s in which $HH^T = nI_n$. In geometric terms, every two different rows in a Hadamard matrix represent two perpendicular vectors. The Hadamard matrix is given by the following matrix values.

$$H_2 = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \quad (3.9)$$

$$H_4 = \begin{bmatrix} H_2 & H_2 \\ H_2 & -H_2 \end{bmatrix} \quad (3.10)$$

$$H_8 = \begin{bmatrix} H_4 & H_4 \\ H_4 & -H_4 \end{bmatrix} \quad (3.11)$$

$$H_{16} = \begin{bmatrix} H_8 & H_8 \\ H_8 & -H_8 \end{bmatrix} \quad (3.12)$$

Equations(3.9),(3.10),(3.11) and (3.12) shows the second, fourth, eight and sixteenth order Hadamard matrix. The Hadamard matrix of order 2^m is used for decoding. The received IFF encoded bits are arranged by representing 1 by -1 and 0 by 1. Then they are multiplied with Hadamard matrix of order equal to the length of incoming bits. The binary equivalent of the position corresponding to the absolute maximum value in the resultant matrix gives the decoded message. Using this algorithm $2^{n-3} - 1$ bit errors can be corrected.

3.3.5 SNR estimation at receiver

In a radar system, target detection is based on three parameters, namely probability of detection (P_d), probability of miss-detection (P_m), and probability of false alarm (P_{fa}). These three parameters are related to SNR [71] as given in equation(3.13) where A and B

is given by equations (3.14),(3.15) for P_{fa} and P_d is taken as mentioned in equation(3.16)

$$SNR = A + 0.12AB + 1.7B \quad (3.13)$$

where

$$A = \ln \left(\frac{0.62}{P_{fa}} \right) \quad (3.14)$$

$$B = \ln \left(\frac{P_d}{1 - P_d} \right) = \ln \left(\frac{P_d}{P_m} \right) \quad (3.15)$$

such that

$$\begin{aligned} 10^{-7} &\leq P_{fa} \leq 10^{-3} \\ 0.1 &\leq P_d \leq 0.9 \end{aligned} \quad (3.16)$$

SNR estimation is a challenging task with only the noise corrupted signal. Earlier methods used are given very good estimation for higher SNR values but fail for low SNR. DiFazio et al [72] shows the improvisation by introducing a new algorithm. They gave a generalized procedure to estimate the negative SNR when the channel is corrupted with AWGN. Similarly, Christopher et al [73] discussed on a new method to estimate the SNR when BPSK signal is heavily corrupted with noise.

These two algorithms are used to detect the SNR in the synthesized AWGN corrupted demodulated encoded 56 bits. Both the algorithms are iterative in nature. In algorithm 1, the mean is refined iteratively based on the estimated error. Detailed explanation for algorithm 1 is given in reference [72]. In algorithm 2, the maximum and minimum of the mean value (A_m) of received signal was refined through an iterative algorithm. a detailed explanation for the same is given in reference [73].

3.3.6 CFAR detector

In a conventional Cell Average or Smaller of/Greater of CFAR detector, the background estimator estimates the noise level is calculated by using the test samples of the reference window with guard window, which depends on target spread. Here it has been considered for swerling case 1 and swerling case 2. [18] [74] [75]

3.3.7 Encoding using Reed-Muller algorithm

Consider the transmitter identity be $(11)_{10}$ and corresponding binary bit stream $(1011)_2$. Thus with $m = 4$, the 1st order Reed-Muller matrix $R(1, 4)$ using Reed-Muller algorithm is obtained as shown in equation (3.17)

$$R(1, 4) = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\ 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 \\ 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 \end{bmatrix} \quad (3.17)$$

Number of rows in the matrix of equation (3.17) is obtained using the equation (3.7). The input message bits are arranged as $m + 1 = n$ bits by padding 0 in MSB.

$$\begin{bmatrix} 0 & 1 & 0 & 1 \end{bmatrix} \quad (3.18)$$

These $[1 \times n]$ bits given in equation (3.18) are multiplied with $[k \times 2^m]$ Reed Muller matrix given in equation (3.17), to get the encoded bits of length 2^m as shown in equation (3.19).

$$\begin{bmatrix} 1 & 0 & 0 & 1 & 1 & 0 & 0 & 1 & 0 & 1 & 1 & 0 & 0 & 1 & 1 & 0 \end{bmatrix} \quad (3.19)$$

These encoded 16 bits are included in spare block of the IFF Mode S format, which is modulated using BPSK technique to transmit for the target detection, as shown in figure (3.9a).

3.3.8 Decoding using Hadamard matrix

Consider the encoded bits given in equation (3.19) are transmitted. Let the demodulated error prone encoded bits in the receiver be as shown in equation (3.20)

$$\begin{bmatrix} 1 & 0 & 1 & 1 & 1 & 0 & 0 & 1 & 0 & 1 & 1 & 0 & 1 & 1 & 1 & 1 \end{bmatrix} \quad (3.20)$$

Compare to equation (3.19), equation (3.20) has the error bits in positions 3, 13 and 16. With this three bits error following steps explains the estimation of the decoded transmitter identity using Hadamard algorithm.

Step 1: The received binary stream in equation (3.20) is rearranged by representing 1 as -1 and 0 as 1 as shown in equation (3.21).

$$\begin{bmatrix} -1 & 1 & -1 & -1 & -1 & 1 & 1 & -1 & 1 & -1 & -1 & 1 & -1 & -1 & -1 & -1 \end{bmatrix} \quad (3.21)$$

Step 2: Select the Hadamard matrix size such that it should be equal to $(2^m \times 2^m)$. Multiply the equation (3.21) with the selected Hadamard matrix.

$$\begin{bmatrix} -6 & -2 & 2 & -2 & 2 & -2 & 2 & 6 & 2 & -2 & 2 & -10 & -6 & -2 & 2 & -2 \end{bmatrix} \quad (3.22)$$

Step 3: Find the maximum absolute weight in equation (3.22). Here we got 10 is the absolute highest weight, whose position is 11.

Step 4: The binary equivalent position obtained in step 3 is considered as the transmitter identity. Hence the decoded data is 1101. This was validated by synthesizing errors in received bits. This works well up to three bit error introduced in the transmitter identity.

3.3.9 SNR estimated output

Figure (3.9) shows the encoded 56 bits input signal and the demodulated IFF signal distorted signal for various values of SNR level. The simulation results are also shown in Table 3.4 for comparing the algorithms.

Table 3.4 shows the simulated result of the mean square error. From Table 3.4 it can be inferred that the efficiency of Algorithm 1 is better than Algorithm 2. For MSE estimation Algorithm 1 gives better results than Algorithm 2. For higher values of SNR both algorithms gives similar results.

TABLE 3.4: Simulation result of mean square error

InitialSNR (dB)	Estimated SNR (Algo-1) (dB)[72]	Estimated SNR (Algo-2) (dB)[69]	MSE (Algo-1)	MSE (Algo-2)
4	4.09537	4.09688	0.6195	0.6425
8	8.0504	8.0539	0.0081	0.0089
12	11.8999	11.9000	0.0026	0.0025
16	16.0217	16.0217	0.0038	0.0038
20	19.9509	19.9372	0.0014	0.0014

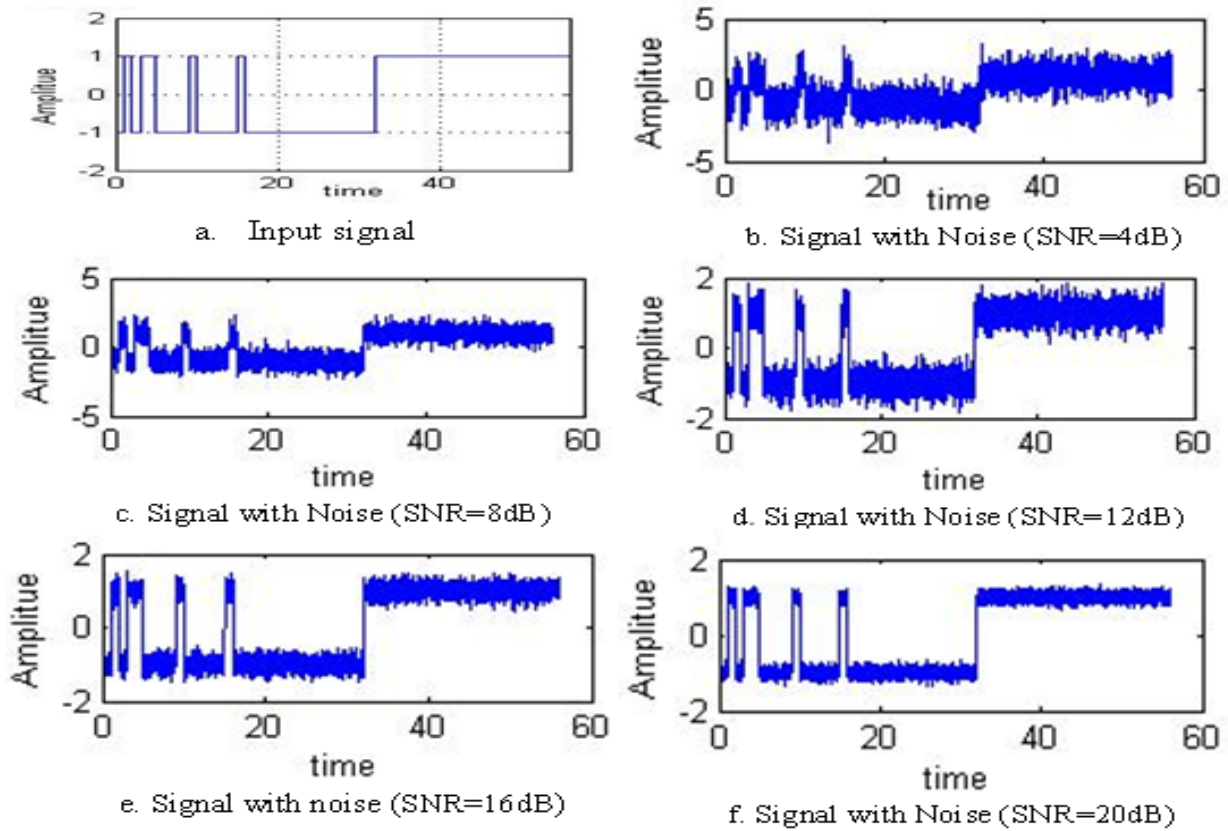


FIGURE 3.9: Distorted signal for different SNR values

3.4 STAP based detector for MDR in homogeneous clutter scenario

Space-Time Adaptive Processing (STAP) [76] refers to multidimensional adaptive filtering algorithms that simultaneously combine and process the signals received from the array elements (spatial domain) and multiple pulses (Temporal domain) of coherent radar waveform to suppress the interference. STAP for multi-dynamic radar scenario improves detection of stealth or low RCS targets obscured by main-lobe clutter and jamming environments as detection performance depends on the set of airborne receivers co-ordinated network. A typical interference scenario involves the detection of a low RCS targets in the presence of a large interfering signal, it is desirable to attenuate the interfering signal while preserving the integrity of low RCS target and an appropriate dynamic range such that receivers should not saturate. This section discusses the reduced rank algorithm to reduce the rank of sample covariance matrix in sample starved homogeneous environment. The two solutions, namely Multi-stage Weiner Filter (MWF) and Principal Component (PC) have been shown in the subsequent section. STAP based new method has been proposed for target detection in multi-dynamic radar scenario. Space-Time Adaptive Processing (STAP) is a signal processing technique mostly used in radar systems to aid in target detection. It uses adaptive processing algorithms to suppress interferences such as jamming, ground clutter and noise to increase system sensitivity to target returns.

Most of the Moving Target Indicator (MTI) active radars with stationary platforms usually encounter slow moving targets. The 2-D nature of the ground clutter makes the detection process complicated because the reflections from ground clutter can be many orders of magnitude greater than the potential targets of interest. The jammer also provides the similar problem in conventional active radar systems. STAP commonly uses a phased-array antenna with multiple elements (i.e. spatial channels). The term "space-time" derives from the coupling of these spatial channels with pulse Doppler waveforms thus creating a 2-D filter to suppress interferences. Adding the statistics of the interference environment (which can be known or unknown), an adaptive STAP weight vector can be estimated to improve target detection.

Multi-dynamic airborne STAP platform used in detecting stealth airborne target in the presence of strong clutter. The multi-static radar is focused as set of bi-statics geometry. Multi-static radar extended to exclusively for Airborne radar for the concept of multi-dynamic radar. Subsequently, section formulates the STAP weight vector by considering the maximum SINR filter for multi-dynamic radar scenario. Then consider the Reduce-Rank Transformation (RRT) methods, which represent the practical approaches for improving the statistical convergence of the adaptive filter and reducing the computational burden. The obtained simulated results are shown in the subsequent section.

3.5 Clutter characteristics for airborne radar

Ground clutter for Airborne radar is a principal performance limiter, the mitigation of the effect of clutter on radar performance fetches far more importance. The clutter returns enter the radar receiver either through main-lobe of the antenna beam and through the side-lobes [77]. The portion of clutter power which enters through main-lobe is called main-lobe clutter and the portion which enters through side-lobe is called the Side lobe clutter. Side-lobe clutter also includes altitude clutter. A target occupies a single point corresponding to a specific angle and at a specific Doppler frequency.

A jammer occupies a particular angle but is temporally considered to be white. There is no standard fixed signal model for multi-dynamic radar, as due to the dynamic of platform the statistical characteristic of signals changes rapidly and randomly.

The clutter occurring due to platform motion occupies the ridge. In 2-D spatial [78], the clutter patch which comes front of the moving platform used to have the highest Doppler frequency, whereas at broader-side used to have zero Doppler. The normalized Doppler shift $\bar{f}_d$ induced on a clutter patch located at an angle θ relative to the array bore-sight is given by equation (3.23).

$$\bar{f}_d = \frac{2vT \sin(\theta)}{\lambda} \quad (3.23)$$

where, v = speed of the radar. T = PRT

The distribution of scattered clutter energy is along a line as shown in figure (3.10). Slope of the clutter ridge can be computed as follows:

$$\bar{f}_d = \frac{2vT}{d} \frac{d}{\lambda} \sin(\theta) \quad (3.24)$$

or,

$$\bar{f}_d = \beta \bar{\theta} \quad (3.25)$$

β represent of the slope of the clutter ridge. Thus, a simple 1-D filtering is not sufficient for separating target from clutter since clutter can come from numerous possible directions (depending on β), hence there is a need for joint space and time (2-D) processing. The detection problem can be defined as in equations (3.26) and (3.27):

$$H_0 : \mathbf{X} = \text{Clutter}(\mathbf{C}) + \text{Jammer}(\mathbf{J}) + \text{Noise}(\mathbf{N}) \quad (3.26)$$

$$H_1 : \mathbf{X} = \text{Ae}(\theta_t, \phi_t) + \text{Clutter}(\mathbf{C}) + \text{Jammer}(\mathbf{J}) + \text{Noise}(\mathbf{N}) \quad (3.27)$$

Where hypothesis H_1 denotes the target present and H_0 shows for target absent hypothesis. $\mathbf{Y} \in \mathbb{C}^{M_K N_K}$ is the received data from K^{th} Radar, The total interference is considered here are $\mathbf{C}$ as the clutter vector, $\mathbf{J}$ the jammer vector, and $\mathbf{N}$ as the background white

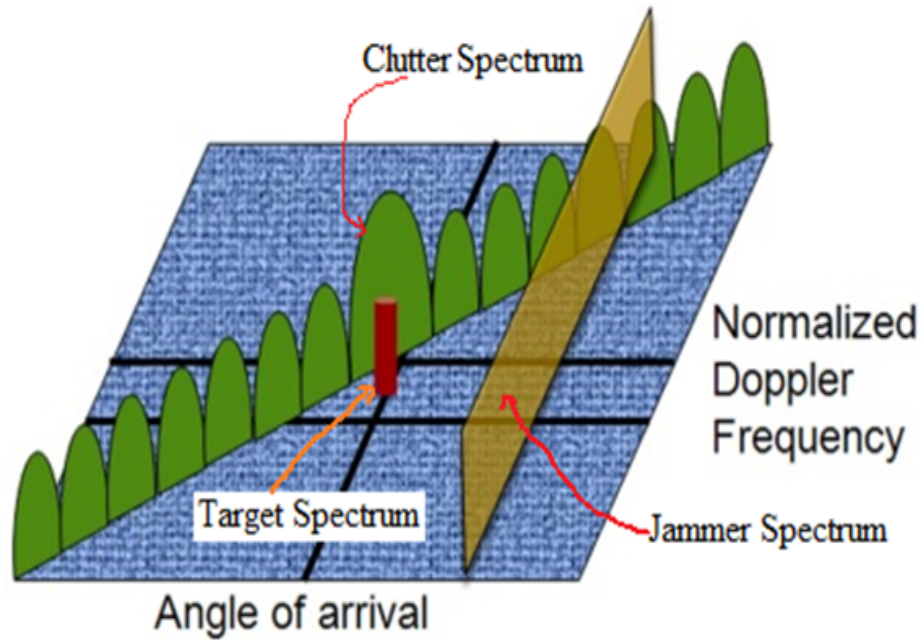


FIGURE 3.10: 2-D clutter distribution [60]

noise, $\mathbf{e}$ is a known spatio-temporal steering vector that models the target return for a particular angle-Doppler, and A is the unknown target complex amplitude. Assume the array is linear and all elements are equi-spaced. The following equations(3.28) to (3.32) indicate the steering vectors.

$$\mathbf{e} = \mathbf{e}_t \otimes \mathbf{e}_s \quad (3.28)$$

$$\mathbf{e}_t = \begin{bmatrix} 1 & z_t & z_t^2 & \dots & z_t^{M-1} \end{bmatrix} \quad (3.29)$$

$$\mathbf{e}_s = \begin{bmatrix} 1 & z_s & z_s^2 & \dots & z_s^{N-1} \end{bmatrix} \quad (3.30)$$

$$z_s = e^{j2\pi f_s} = e^{(j2\pi \frac{d}{\lambda} \sin \phi_t)} \quad (3.31)$$

$$z_t = e^{(j2\pi \frac{f_t}{f_s})} \quad (3.32)$$

Where ϕ_t is look-angle taken from broadside and f_t Doppler-frequency respectively. $\otimes$ indicates the Kronecker product for two steering vectors; f_s is the PRF; and λ represents wavelength of transmission frequency. The vectors involved in Kronecker product of temporal steering vector $\mathbf{e}_t$ and spatial steering vector $\mathbf{e}_s$.

3.6 Space time adaptive filter formulation for multi-dynamic radar scenario

The multi-dynamic radar configuration as shown in figure (3.11) is configured with various combinations of airborne Tx-Rx works in a coherent manner, it is assumed that all radars Tx-Rx shall be in synchronized. Each airborne Rx is synchronized and configured to receive the target echoes by any Tx to process the target of interest. As discussed in earlier section assuming that the radiation power by the target is isotropic.

Multi-dynamic configuration is considered for m Transmitters (Tx) and n Receivers (Rx) geographically apart but with shared coverage area, the $\text{SNR}_{\text{overall}}$ for multi-dynamic radar scenario is computed using equation (3.2) by coherent summing up all the partial SNR for each Tx-Rx pair.

The angle-Doppler image by using a two-dimensional, inverse FFT with the weight matrix applied on the signal for simulation for a given target in the presence of interference as seen by different radars. Figure (3.11) is the angle-Doppler plot for the target seen by one Rx in the presence of interference, figure (3.12) shows is the angle-Doppler plot for the case as seen by the receiver, which is kept off diagonal for multi-dynamic radar scenario. The simulation is made jammer at location $\theta = -34$ degree and 18 degree. Consider an example of 16 element linear array with uniformly distributed for S-Band frequency. The following assumptions are made in this example;

Per-element SNR of 0 dB.

Two jammers with JNRs of 50 dB.

Number of CPI pulses used for Doppler processing is 16 .

The target is located at an angle of zero and a Doppler of zero.

Half wavelength element spacing and a normalized PRI of unity

The single-pulse, per element SNR is 0 dB

The Doppler locations of the interference sources, corresponding to the above angles, are 0.217 Hz and 0.28 Hz

SCRs of the two clutter sources is 50 dB.

Maximizing the Signal to interference plus Noise ratio (SINR) output is the main objective of the STAP. The adaptive algorithms determine a weight vector $\mathbf{w}_k$ to obtain a test statistic threshold, Λ_k as shown in equation(3.34). Where k represents k^{th} radar of multi-dynamic scenario.

$$\Lambda_k = |\mathbf{w}_k^H \mathbf{z}|^2 \geq \lambda \quad (3.33)$$

For interference having Gaussian distribution, the optimum processing for interference covariance matrix $\mathbf{R}_k$, is the whiten-and-match filter for detecting a rank-1 signal given

by the equation(3.34) [79] [80]

$$w_k = \frac{R_k^{-1}e}{\sqrt{e^H R_k^{-1} e}} \Rightarrow \Lambda_{MF} = \frac{|e^H R_k^{-1} Y|^2}{e^H R_k^{-1} e} \quad (3.34)$$

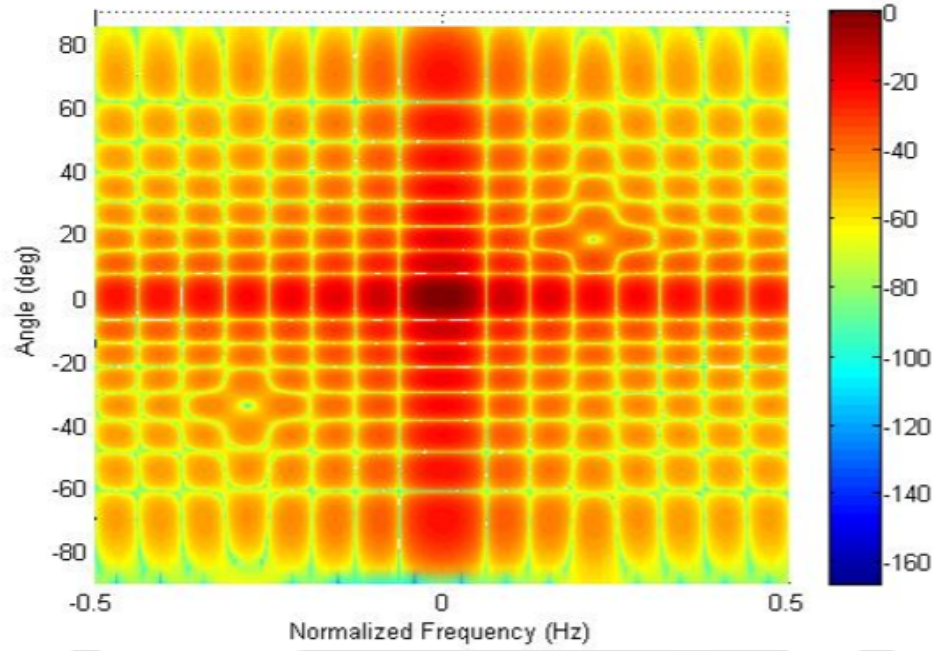


FIGURE 3.11: Angle-Doppler image for a target(with interference) for multi-dynamic radar

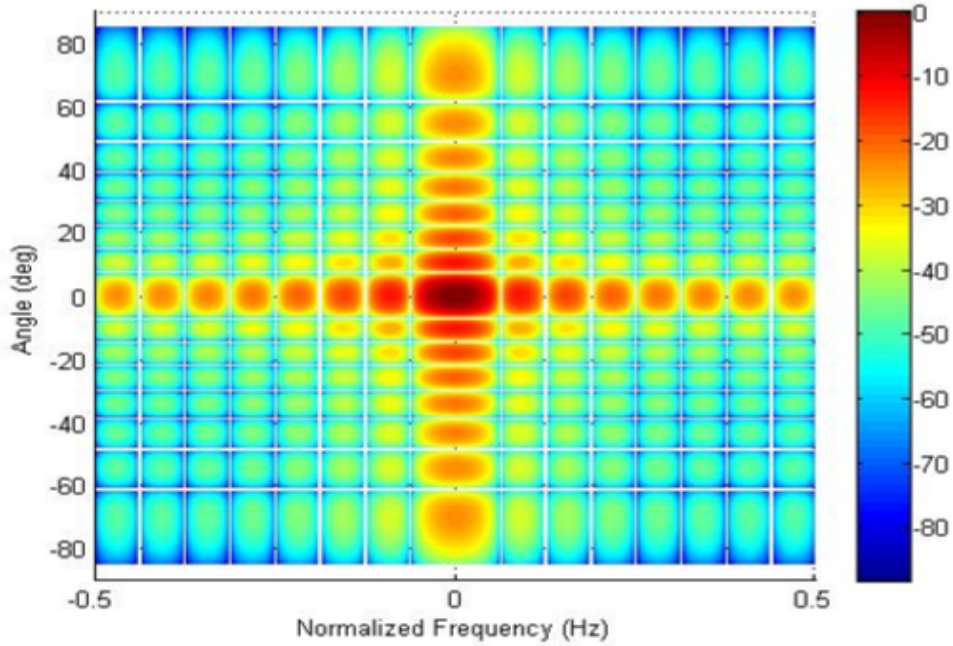


FIGURE 3.12: Angle-Doppler image for a target(with interference) for multi-dynamic radar(off diagonal Rx)

It can be readily shown that Λ_{MF} is SINR of the minimum variance distortion-less response (MVDR) beamformer, the maximum likelihood estimate of the target complex amplitude. In practice, the covariance matrix, $\mathbf{R}_k$, is unknown and must be estimated.

Reed, Mallet, and Brennan (RMB beam-former) [81] showed that the sample matrix inverse (SMI) method offered better convergence. In the SMI approach, the basis for most modern STAP algorithms, the interference covariance matrix is estimated using K data ranges for training, as shown in equation (3.35)

$$\hat{\mathbf{R}}_k = \frac{1}{K} \sum_{k=1}^K \mathbf{y}_k \mathbf{y}_k^H = \frac{1}{K} \mathbf{Y} \mathbf{Y}^H \quad (3.35)$$

where $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, \dots, \mathbf{y}_K]$ and the adaptive weights are obtained using $\mathbf{w} = \mathbf{R}_k^{-1} \mathbf{e}$. A drawback of the RMB approach is the lack of a CFAR [82] [83] [84] [85] [86] [87], i.e., the false alarm rate varies with the interference level.

3.7 Reduce Rank Transformation

3.7.1 Principal Components

PC uses the Eigen-value decomposition (EVD) to produce a low-rank estimate of the sampled covariance matrix $\hat{\mathbf{R}}_x$. This method is a good approximation to the original but it reduces the required computer processing power, which is acceptable trade-off (speed for accuracy) due to the benefit of less cost. Decomposing the eigen values [80] of $\hat{\mathbf{R}}_x$ as shown in equation(3.36)

$$\hat{\mathbf{R}}_x = \sum_{i=1}^N \lambda_i \mathbf{v}_i \mathbf{v}_i^H \quad (3.36)$$

where, λ_i and $\mathbf{v}_i$ represent the i^{th} eigenvalue and i^{th} eigenvector of $\hat{\mathbf{R}}_x$ and N is the total number of degree of freedoms. Retaining the r largest eigenvalues and their corresponding eigenvectors while rejecting the rest gives the best reduced r rank approximation of $\hat{\mathbf{R}}_x$ [88]. Therefore,

$$\hat{\mathbf{R}}_x = \sum_{i=1}^{r_{pc}} \lambda_i \mathbf{v}_i \mathbf{v}_i^H \quad (3.37)$$

where, $r_{pc} < N$, contains the principal components (components having the dominant signal power) of $\hat{\mathbf{R}}_{\mathbf{x}}$. Since, the data $\mathbf{X}$ is considered in weight vector calculation so it is data dependent and because the steering vector $\mathbf{e}$ is not considered so it is signal independent. Hence, it is called Principal-Component Signal Independent (PC-SI) algorithm [89]. This method is simple, but it degrades the performance as the steering vector $\mathbf{e}$ is not taken into account.

Another PC variation is PC-SD (signal dependent). This variation takes the main lobe responses (steering vector $\mathbf{e}$) into account for rank reduction. $\mathbf{B}$ symbolizes the orthogonal set of vectors to $\mathbf{e}$ (i.e. side lobe responses). The steering vector for PC-SD would be in equation (3.38) [78]:

$$\mathbf{e}_{\text{PC-SD}} = \mathbf{e} - \mathbf{B}\mathbf{w}_{\mathbf{a}} \quad (3.38)$$

where $\mathbf{w}_{\mathbf{a}}$ is weight computed for side-lobe response of signal.[78]:

3.7.2 Multi-stage Wiener Filter

It gives a weight vector that can yield the highest possible signal to interference and hence the maximum probability of target detection. The classic single stage Wiener filter equation (3.39) is [90]:

$$\mathbf{w}_{\text{opt}} = \mathbf{R}^{-1}_{\mathbf{x}_0} \mathbf{r}_{\mathbf{x}_0 \mathbf{d}_0} \quad (3.39)$$

It minimizes the mean square error by subtracting a weighted signal $\mathbf{x}_0(k)$ from the desired signal $\mathbf{d}_0(k)$. It shows that the weight vector depends on the given environment; hence it is adaptive [55] [91]. Both $\mathbf{R}$ and $\mathbf{r}$ are unknown and are estimated by acquiring the samples of the environment. Thus, the system performance is dependent on sample support. Reduction of rank for Wiener filter is heavily dependent on the cross-correlation vector. Figure (3.14) shows that MWF decomposes the sampled data $\mathbf{x}$ with a sequence of orthogonal projection blocks like $\mathbf{B}_0$. Rank reduction is done by converging these stages to the desired number r^{mwf} resulting into a reduced rank transformation basis that spans the Krylov subspace [90] instead of the eigenvector basis like PC [92].

Since it tailors its basis selection to the desired steering vector $\mathbf{e}$, thus MWF operates in a more closed subspace than PC. The total MWF weight vector is given by equation(3.40):

$$\mathbf{w}_{\text{mwf}} = \mathbf{e} - w_1 \mathbf{B}_0^H \mathbf{h}_1 + w_1 w_2 \mathbf{B}_0^H \mathbf{B}_1^H \mathbf{h}_2 - w_1 w_2 w_3 \mathbf{B}_0^H \mathbf{B}_1^H \mathbf{B}_2^H \mathbf{h}_3 + \dots \quad (3.40)$$

The main advantage of MWF is that its main computation operation is the vector cross-correlation and is hence simple. It does not require a covariance matrix which requires significant computation. EVD is not required, which are expensive operation. It has been found that that the MWF rank compression is lower than the number of interferers, but it does not give excellent sample compression [91].

Since, in MWF weight values depend on both eigenvalue and the cross-correlation coefficient, so its rank compression is always better than or equal to its PC counterpart. Thus, if $r_{\text{mwf}} < N$, then Krylov subspace dimension can be reduced based on low correlation, low eigenvalue, or a combination of both [93] [94].

3.8 Simulation and discussion

The following parameters are taken as mentioned in Table 3.5 for STAP based MDR simulation:

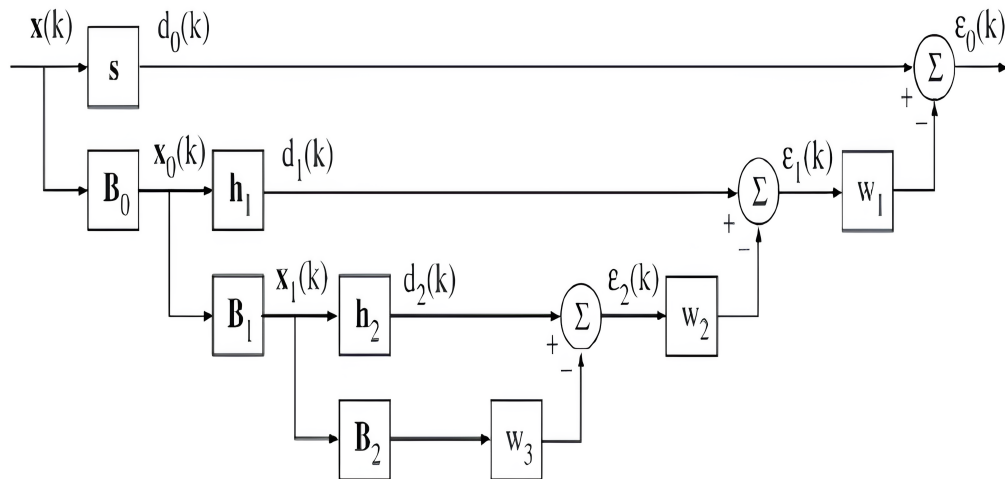


FIGURE 3.13: 3-stage Wiener filter

TABLE 3.5: STAP based MDR simulation parameters

Parameters	Value
N (No. of elements)	6
M (No. of pulses)	6
d (inter-element spacing)	$\lambda/2$
Clutter Power (CNR) dB	10 dB
Jammer Power (JNR) dB	50 dB
No. of Jammers	2
β	1
PRF	1 kHz
f_c Carrier Frequency	3 GHz
No. Tx-Rx Pair	3

3.8.1 Calculating eigen value and eigen vectors to estimate the rank

Figure(3.14) shows that the curve without the jammers is dominated by clutter. Brennan's rule estimates the number of significant eigenvalues created by clutter. From the simulation, we see that the addition of jammers results in strong eigenvalues and hence increases the rank. Therefore, it is highly desirable to implement RRT [93] to lower the processing cost in equation(3.41)

$$r_c = N + (M - 1)\beta \quad (3.41)$$

i.e $r_c = 11$ figure(3.14) shows that the simulated rank is 12.

3.8.2 Effect of number of training samples and jammers

As the number of training samples increases, SINR performance for MWF improves drastically as compared to PC-SI. MWF achieves desired range (within 3 dB) with approximately 25 samples as shown in figure (3.15). But as the number of jammers increases, more samples are required. Thus, MWF does not show much improvement with increasing interference as shown in figure (3.16).

3.8.3 MWF and PC-SD performance

Figure (3.12) shows the MWF and PC-SD performance against two jammers located at -28° and 55° . PC-SD needs 13 ADoF to suppress the interference to achieve MVDR. MWF only needs 5 ADoF to accomplish the same. As figure shows, MWF's MSE performance of ranks from 5 to 13 are all well within 3dB range of minimum mean

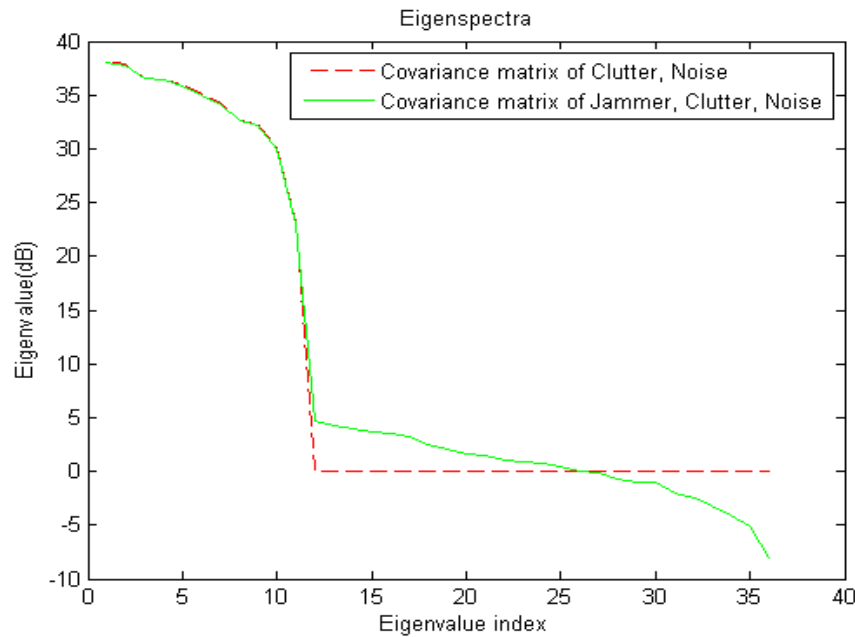


FIGURE 3.14: Eigen spectra

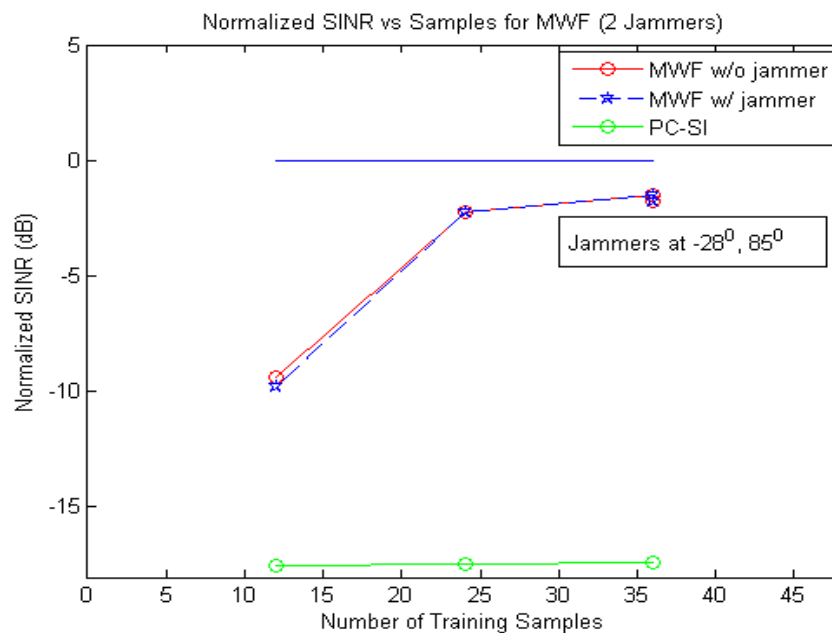


FIGURE 3.15: SINR v/s No. of training samples (2 Jammers)

square error (MMSE), so it can stop anywhere within stages 5 to 13 and still yield an acceptable result. Thus, MWF also offers more flexibility in rank selection as shown in figure (3.17).

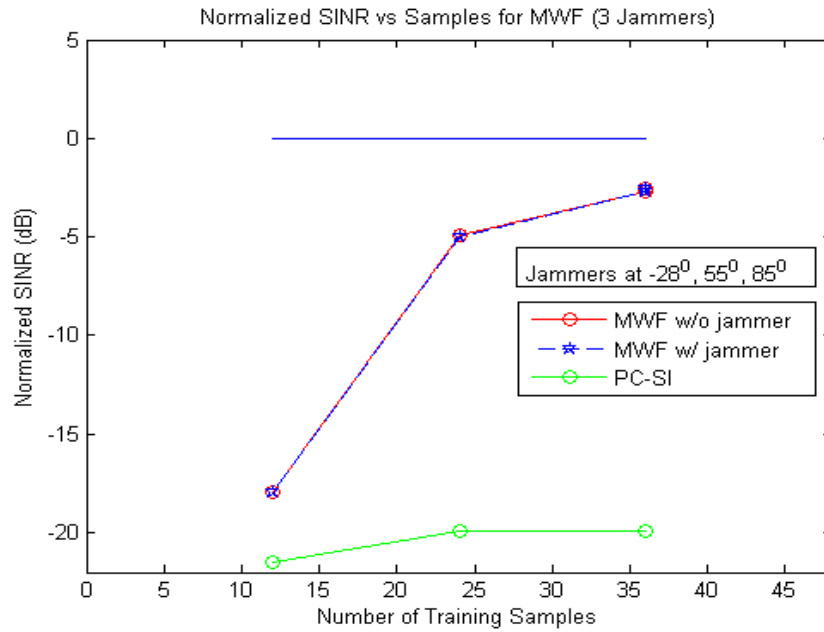


FIGURE 3.16: SINR v/s No. of training samples (3 Jammers)

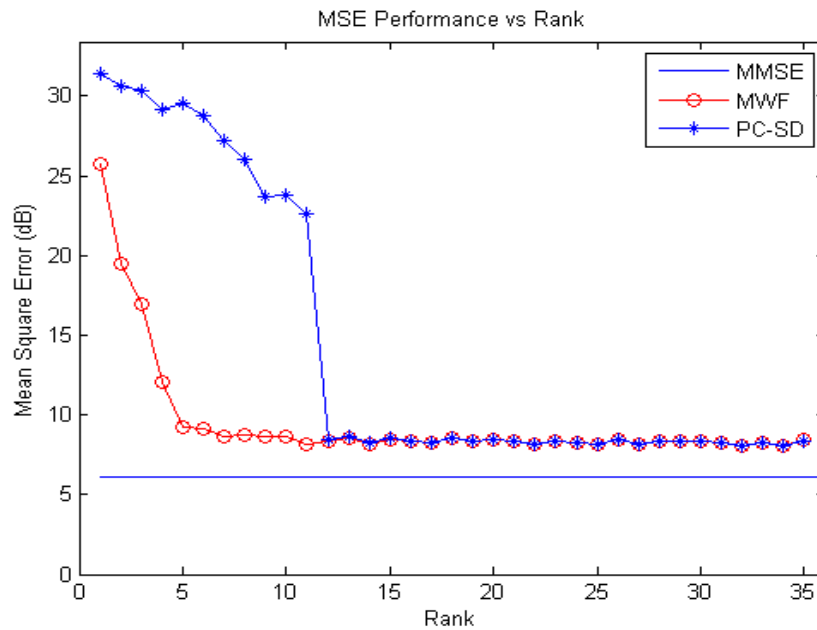


FIGURE 3.17: MWF v/s PC-SD with 50 dB JNR and 10 dB CNR

3.8.4 Beam-forming results

Figure (3.18) shows the beam-former pattern of an un-adapted phased array antenna. MWF and PC-SD array pattern shows the same array response. Both place the jammers at nulls with approximately the same accuracy (figure (3.19)). But MWF needs less DoF as compared to PC-SD (as shown in figure (3.19)).

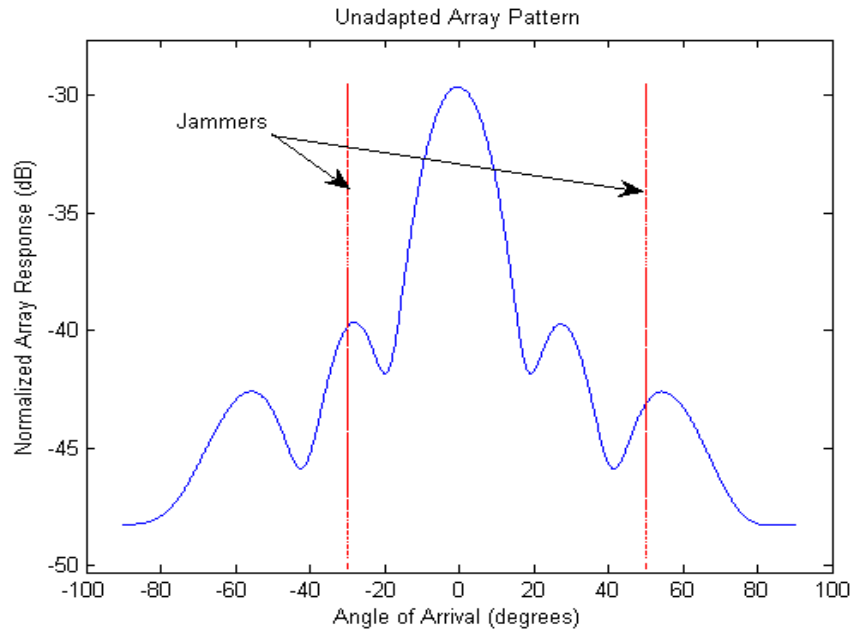


FIGURE 3.18: Unadapted array pattern (beamformer)

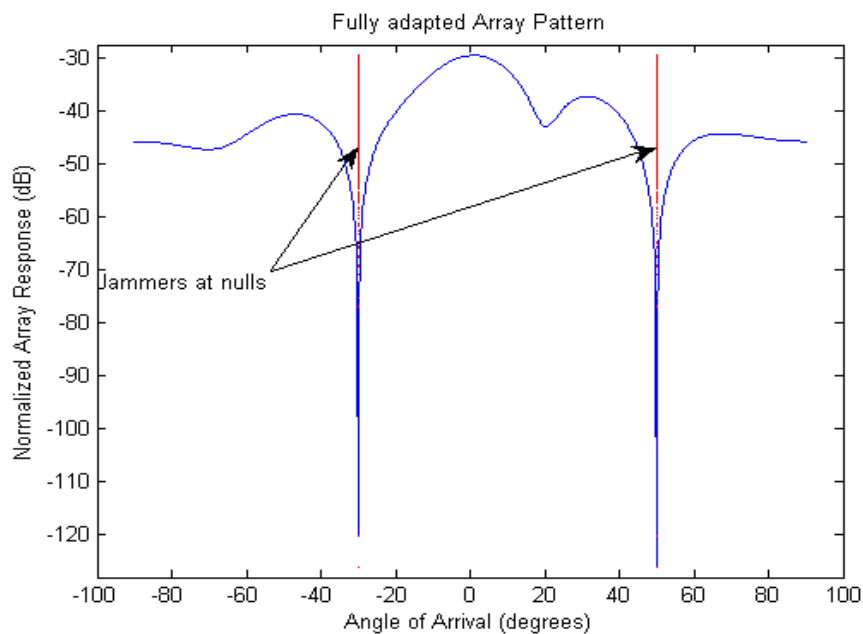


FIGURE 3.19: MWF(partially adaptive) array pattern

3.9 Summary

1. As All radars (Primary Radar) work in conjunction with IFF radar (Secondary Surveillance radar). The IFF radar works for cooperative targets, as it is the query (interrogator)-reply (transponder) based radar. The skin echo of the IFF radar waveform can be used as a waveform identifier of Primary radar. Hence in multi-dynamic radar case this IFF waveform ID processing can be useful for identifying the Primary radar waveform to process the radar signal. The technique of IFF interrogator Mode-S frames in the form of BPSK signal is introduced in this paper. The SNR estimation for various SNR scenarios is done for the synthesized AWGN added 56 bits. Since the hadamard decoder matrix size depends on the number of columns in the first order RM matrix, maximum message bits can be 5 (required 32x32 hadamard matrix). In future, higher order RM encoder and decoder can be implemented for any number of message bits.
2. The multi-dynamic radar scenario is proposed, the statistical nature of the scenario is dynamically changing. The analysis of the rank reducing transformation techniques namely principle component (PC) and multi-stage Wiener filter (MWF) for STAP simulated for multi-dynamic case. Both are reduced rank adaptive signal processing algorithms with several advantages. These include improved sample support (in MWF) with improved performance over their full rank counterpart [14]. The eigenvector and Krylov basis analysis tells us that MWF provides greater rank compression as compared to principal components. Thus, MWF RRT can be used efficiently to transform a higher rank noise covariance matrix to a much lower rank matrix [15].

4

Investigation on multi-dynamic radar waveform design

This chapter introduces the concept of multi-dynamic radar waveform design tool in the form of mathematical derivation of ambiguity function. How ambiguity function is having different configuration parameters in case of monostatic radar, bi-static radar, and further multi-dynamic radar case. It is shown the various investigation of OFDM-LFM waveform for multi-dynamic radar, this includes the trade-off of the number of OFDM-LFM waveform usage. Analysis for performance benchmarking of the waveform design and the derivation of Cramer-Rao Lower Bound (CRLB) is presented in this chapter.

4.1 Introduction

In a radar system, the proper choice of waveforms directly affects the performance of the system like the signal to noise ratio (SNR), range resolution, Doppler resolution, ambiguities in range and Doppler as well as side-lobe response levels of the processing radar signals. So the proper selection and analysis of the transmitted waveform is the first priority. To analyse the waveforms, ambiguity function [95] [96] is used as a waveform design tool. As mentioned earlier it gives the measure of time delay (τ) and a Doppler shift (ν) of a matched filter. Usually, target is detected when the optimal output obtained from matched filter crosses the threshold. Matched filters are also used to measure the time delay of an echo signal based on the position of the peak energy concentration of its output. The choice of radar waveform is concluded to be perfect if its matched filter response has a narrow main lobe and low side lobes. Ambiguity function is the essential tools for the waveform performance analysis of monostatic, bi-static, and multi-dynamic radar systems in the presence of additive noise and ambiguities in range and velocity measurements. The ambiguity function properties of a signal depending on the transmitted signal and the geometrical structure of the radar system. To obtain a high range resolution, the cross-correlation of the waveforms should be approximately an impulse function. In a multi-dynamic radar, for covering long range or coverage more average transmit power is required however, for echo processing it is required to compress the transmitted pulse width to achieve better range resolution. Hence, Linear Frequency-Modulated (LFM) based waveform is given preference. LFM based waveform used to have more time-bandwidth calculation product. Orthogonal Frequency Division Multiplexing (OFDM) based LFM waveform is projected to be one of the waveform for the multi-dynamic radar scenario. The simulation is focussed on the analysis of ambiguity function of OFDM based LFM waveform for multi-dynamic radar system.

4.2 Ambiguity function

Woodward developed [97] a tool to analyze the range and Doppler known as ambiguity function (AF) represented by $|\chi(\tau, \nu)|$. This describes the two dimensional response of

a matched filter [95] [98] [99]. It is an analytical tool for waveform design and the time response of a filter matched to a given finite energy signal when the target echo is received with a time-delay as well as Doppler-shifted. An ideal ambiguity function for a radar waveform, also known as thumb-tack ambiguity function, is shown in figure (4.1). This has a maximum impulsive energy peak at the origin and the remaining energy spreads uniformly throughout the range Doppler plane. The narrow central peak represents good resolution in both range, and Doppler [100], and the uniform plane shows there are no range and Doppler ambiguities. The uniform surface in the range Doppler plane signifies the presence of very low and uniform side lobes [101] with minimizing target masking effects. Assume $u(t)$ is the complex envelope of the transmitted signal, then it can be expressed as shown in equation (4.1):

$$s(t) = \text{Re}\{u(t) \exp(j \omega_c t)\} \quad (4.1)$$

The received signal will have a delay based on the travel path and the Doppler shift based on the target motion in the direction of the receiver. Ignoring the propagation losses, the received signal reflected from a slow-moving point target is given by equation (4.2),

$$s(t) = \text{Re}\{u(t - \tau) \exp(j (\omega_c + \omega_d) t)\} \quad (4.2)$$

where $\omega_d = 2 \pi \nu$

τ - time delay,

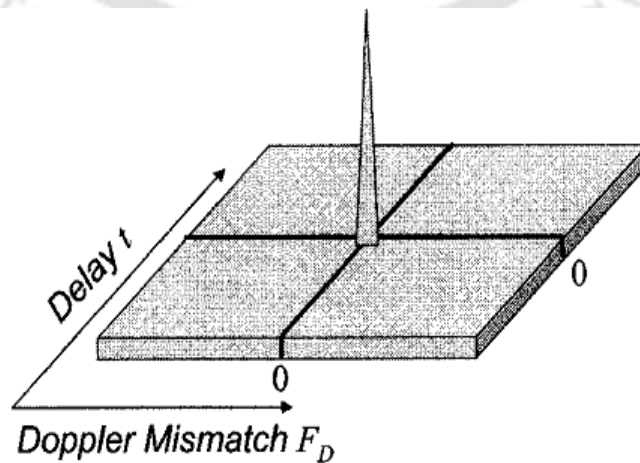


FIGURE 4.1: Ideal ambiguity function [97]

ω_c -carrier frequency,

ω_d - Doppler shift due to the target movement

ν - Doppler frequency.

By passing the received signal to a matched filter, the output (ambiguity function) mathematically represented by equation (4.3) [96]:

$$|\chi(\tau, \nu)| = \left| \int_{-\infty}^{\infty} u(t)u^*(t - \tau)\exp(j2\pi\nu t)dt \right| \quad (4.3)$$

Since the ambiguity function is two-dimensional, i.e. each function value is defined by two independent variables, it is helpful to analyse the ambiguity properties by looking at the cuts along two separate axes representing different independent variables, which by definition are delay and Doppler. The first cut is the ambiguity function cut along the delay axis by setting $\nu = 0$ in equation (4.4).

$$|\chi(\tau, 0)| = \left| \int_{-\infty}^{\infty} u(t)u^*(t - \tau)dt \right| = |R(\tau)| \quad (4.4)$$

$|R(\tau)|$ is the auto correlation function (ACF). This cut of ambiguity function, known as Doppler cut plot along the delay axis, represents the shape of “Range Window” of matched filter at zero Doppler. In other words, it shows the range resolution of the waveform.

The other cut of interest will be the cut along the frequency axis by setting $\tau = 0$ in equation (4.3). This cut of ambiguity function along the frequency axis tells us how the shape of the matched filter output will change with Doppler frequency, as shown in equation (4.5).

$$|\chi(0, \nu)| = \left| \int_{-\infty}^{\infty} |u(t)|^2 \exp(j2\pi\nu t)dt \right| \quad (4.5)$$

The Complex Ambiguity Function (CAF) is well known in the context of radar as a key tool for determining target resolution capability and is a consequence of the nature of the optimal detector, which involves decision-making based on the output of a matched filter determined from the transmitted waveform [95]. The CAF is the auto-correlation of the complex envelope of the waveform with shifted in time and frequency, and presents the point target response of the waveform as a two-dimensional function of range and

Doppler, showing the resolution, side-lobe structure, and ambiguities in the delay and Doppler domains. The mathematical definition of the CAF is given by equation (4.6) [95].

$$|\chi(\tau_H, \tau_a, \nu_H, \nu_a)| = \left| \int_{-\infty}^{\infty} u(t - \tau_a) u^*(t - \tau_H) \exp(-j2\pi(\nu_H - \nu_a)t) dt \right| \quad (4.6)$$

Where τ_a, ν_a the actual delay and Doppler frequency and τ_H, ν_H are the hypothesized delay and frequency.

4.2.1 LFM Pulses

Increasing the pulse energy improves radar detection performances. This can be achieved by a longer pulse but to get better range resolution a shorter pulse is required. To resolve this trade off the LFM is used where higher bandwidth can be achieved by a longer duration pulse by a technique known as pulse compression. Presently almost all the radar systems are using the LFM technique for signal transmission. The LFM signal mathematically represented by equation (4.7)

$$u(t) = \frac{1}{\sqrt{T}} \exp(j\pi k t^2) \text{rect}\left(\frac{(t - T/2)}{T}\right) \quad (4.7)$$

Where $kr^2 = TB$, the effective time bandwidth product of the signal and B is the signal bandwidth. The ambiguity function of the LFM pulse shown in figure (4.2), figure(4.3), the mathematical form is given by the equation (4.8):

$$|\chi(\tau, \nu)| = \left| \frac{\sin(\pi\nu - k\tau T - |\tau|)}{(\pi T\nu - k\tau)} \right| \quad (4.8)$$

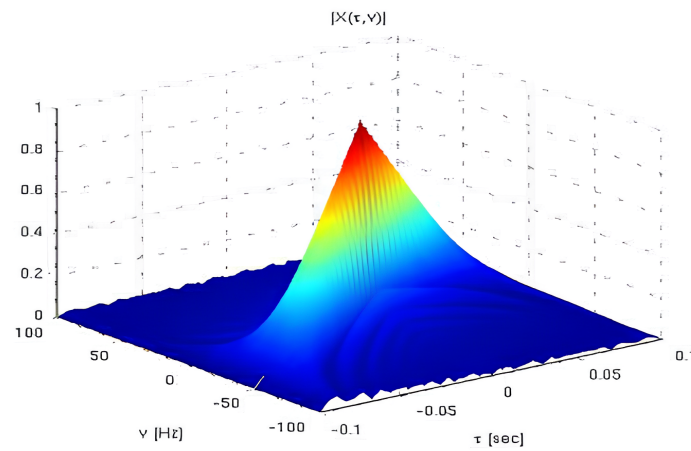


FIGURE 4.2: LFM ambiguity mesh

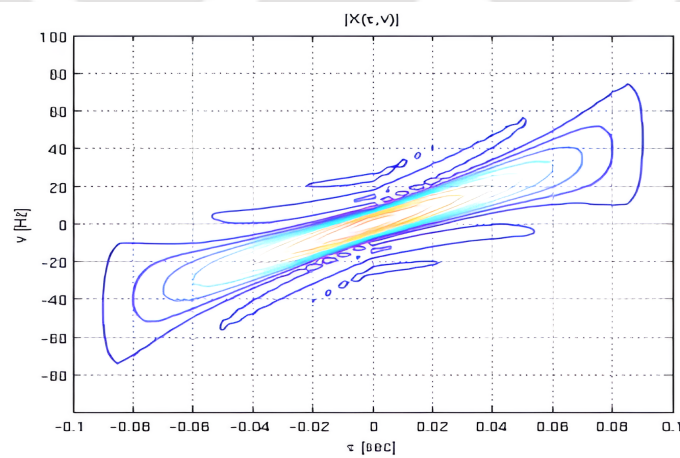


FIGURE 4.3: LFM ambiguity contour

4.3 Bi-static radar ambiguity function

In bi-static radar system transmitter and receiver are separated by a baseline distance L and the geometry between the transmitter, target, and receiver greatly affect the ambiguity function. A north-referenced coordinate system is used to represent bi-static geometry [27] [102] [103], which is shown in figure (4.4). This is a two-dimensional coordinate system confined to the bi-static plane formed by transmitter Tx, receiver Rx and target, with a north-up orientation. The look angle θ_T of the transmitter and the look angle θ_R of the receiver are measured positive clockwise from the north of the coordinate system. β represent the bi-static angle and ϕ angle is the angle formed by velocity vector and bi-static bisector. R_T and R_R are the range from transmitter to target and the range from receiver to target, respectively.

The delay-Doppler plane is not linear in bi-static case [104] [105]. This result in major differences between simple monostatic radar and bi-static radar ambiguity functions. From the bi-static geometry the total range R in terms of R_R , θ_R is given by equation (4.9) [3], the bi-static delay τ_b is given by equation (4.10) and equation (4.11) represents the computed Doppler (ν):

$$R = R_R + \sqrt{R_R^2 + L^2 - 2R_R L \sin \theta_R} \quad (4.9)$$

$$\tau_b = \frac{R_R + \sqrt{R_R^2 + L^2 - 2R_R L \sin \theta_R}}{c} \quad (4.10)$$

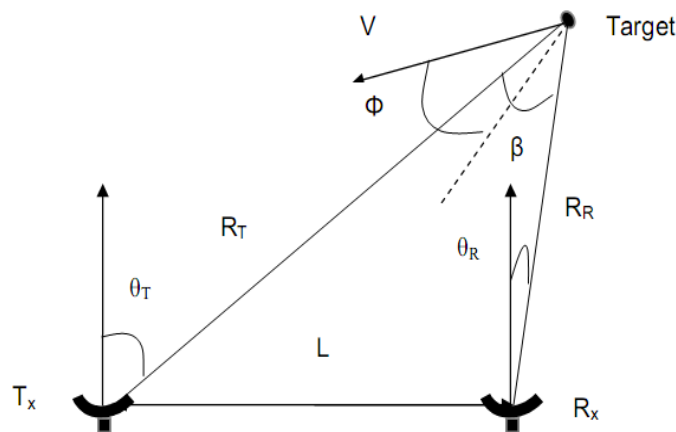


FIGURE 4.4: Bistatic system geometry

$$\nu = \frac{\omega_c V \cos \phi}{\pi c} \sqrt{\frac{1}{2} + \frac{R_R + L \sin \theta_R}{2\sqrt{R_R^2 + L^2 - 2R_R L \sin \theta_R}}} \quad (4.11)$$

where

ω_c is the carrier frequency,

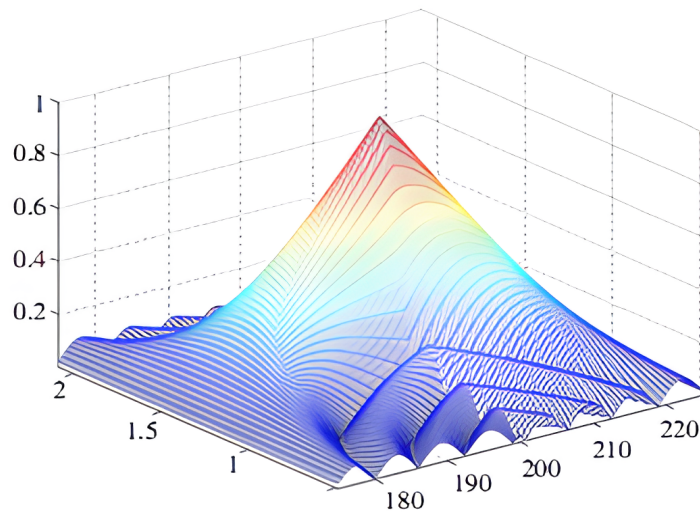
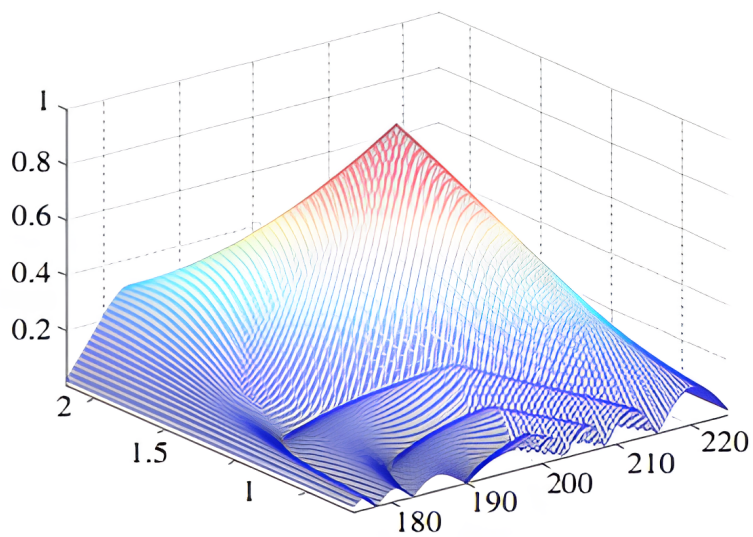
ν is bi-static radar Doppler shift, and

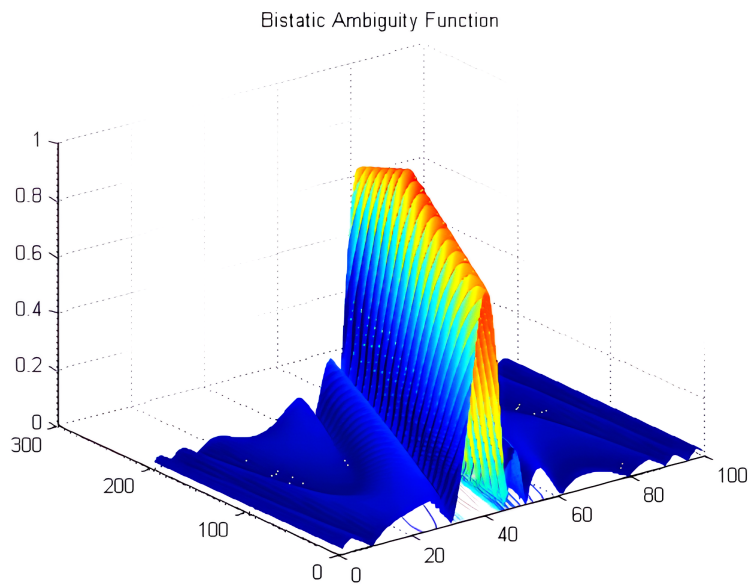
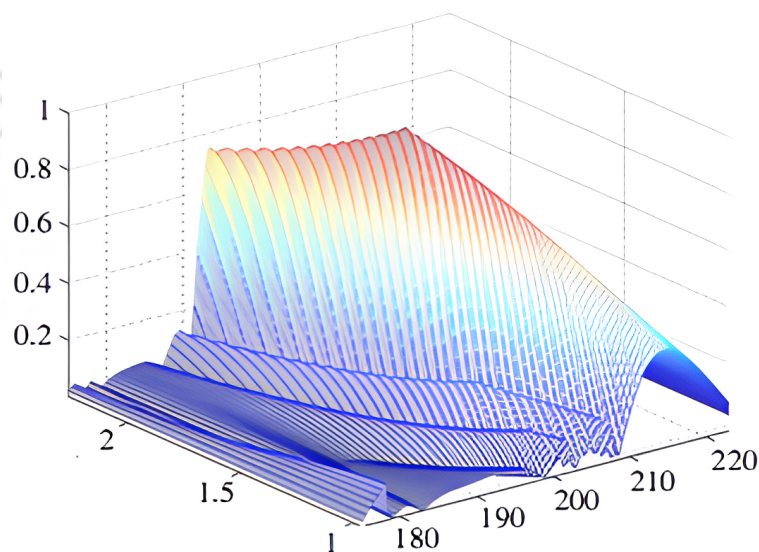
$V \cos \phi$ is the magnitude of the projection of velocity vector.

The bi-static ambiguity function is calculated by equation (4.12)

$$|\chi(R_R, \theta_R, L, V, \phi, \omega_c)| = \left| \int_{-\infty}^{\infty} u(t) u^*(t - \tau) \exp(j2\pi \nu t) dt \right| \quad (4.12)$$

From the above equation (4.12), the delay and Doppler in a bi-static radar system are defined by more factors, that relates to the bi-static radar system geometry. Therefore, it is expected that the shape of the bi-static radar ambiguity function will be depending on the bi-static radar system geometry. Despite the transmitted signal being used, as long as the target is close to the bi-static baseline, the ambiguity function deteriorates. Therefore it is the system geometry, instead of the transmitted signal, that plays an important role in determining the bi-static radar ambiguity properties. This makes a difference with the monostatic ambiguity function, where the shape of it only dependent on the signal is being used. Depending upon the geometrical configuration the ambiguity functions, are shown in figure (4.5), (4.6), (4.7) and figure (4.8) with different values of θ_R 45° , -45° , 0° and 85° .

FIGURE 4.5: Ambiguity function of bi-static radar at $\theta_R=45^\circ$ FIGURE 4.6: Ambiguity function of bistatic radar at $\theta_R=-45^\circ$

FIGURE 4.7: Ambiguity function of bistatic radar at different $\theta_R=0^\circ$ FIGURE 4.8: Ambiguity function of bistatic radar at different $\theta_R=85^\circ$

4.4 Waveform design and matched filter for MDR

Multi-dynamic radar (MDR) systems are required for the stealth aircraft detection and tracking. MDR more effectiveness is due to the dynamic geometry or coverage area which improves the detection capability for stealth aircraft. In this case multiple number of airborne transmitters (Tx) and multiple airborne receivers (Rxs) are deployed, each airborne Tx is capable of generating and transmitting arbitrary waveforms and each airborne Rx can receive the target echo from the respective transmitted waveforms. These waveforms design are required to provide the high performance of long range detection and resolution. Such waveform design have high challenge to define with respect to dependency and multi-dynamic radar geometry. In the design of a MDR system waveform it is required, better temporal resolution or range resolution as well as better Doppler resolution to achieve the effective performance. Hence, waveform design requires larger value of time bandwidth product value. Orthogonal frequency-division-multiplexing (OFDM) waveform with linear frequency modulated (LFM) signal is proposed for multi-dynamic scenario. The OFDM-LFM based waveform jointly make use of the achieved success of conventional LFM and the effectiveness of frequency orthogonality in OFDM signals. This is to ensure for achieving better spectral efficiency using multi carrier modulation as well as better range-Doppler resolution [106] [107]. As brought out in earlier chapter the multi-dynamic radar system is assumed to be a coherent airborne radar-network, i.e all airborne Tx and airborne Rx are synchronized in time and phase. The target returned is assumed to be isotropic for simulation. It is taken m number of airborne Txs and n number of airborne Rxs geographically apart in the MDR system network, the overall SNR is calculated based on partial SNR [4] of each Tx-Rx pairs. Here it is assumed that all are airborne Tx and Rx forming MDR system. Hence the $(\text{SNR})_{\text{overall}}$ is given by equation (4.13)

$$(\text{SNR})_{\text{overall}} = \sum_{i=1}^m \sum_{j=1}^n \frac{P_{ti} G_{ti} G_{rj} \lambda_i^2 \sigma_{tj}^2 F_{ti}^2 F_{rj}^2}{(4\pi)^3 k T_S B_{Rj} R_{ti}^2 R_{rj}^2 L_{ij}} \quad (4.13)$$

The scenario is more and more complex for detection and classification of the target when more and more targets come under MDR area coverage made by more and more

numbers of airborne radar Tx and Rx. There is no standard fixed signal model for multi-dynamic radar, it is due to the all dynamics nature of the platforms, the statistical characteristic of signals changes rapidly and randomly. Multi-dynamic radar is assumed to be similar to multi-static or multi-site Radar systems at a given instant of time. The advantages are to counter the detection loss of non-cooperative target, high accuracy of target localization, more target resolution capability, more jammer resistance, and more airborne clutter resistance [108]. In subsequent section the OFDM based LFM signal ambiguity function is analyzed for MDR waveform [109] [110] [111].

4.4.1 Waveform design for MDR

The OFDM-LFM (chirp-waveform), basically it is to utilize the property of orthogonality of discrete-frequency components. The orthogonality nature of the waveforms is used to be totally independent on the input-sequences. The concept for MDR scenario scheme, the utilization of frequency also increases as it exploits the entire bandwidth for every radar waveform [28] [112] [113].

Considering the OFDM modulation processing carried out with two OFDM and LFM radar-waveforms. Assuming the sequence input is $X(m)$ with N spectral discrete components, having separation by $2\Delta f$. For the first step, the input signal $X(m)$ is interleaved by N zeros, forming a new data sequence $X_1(m)$. Then, interleaved data-sequence $X_1(m)$ is shifted by Δf to get another data sequence $X_2(m)$. These input sequences are required to transform into the time-domain by taking $2N$ points inverse Discrete-Fourier Transform known as IDFT. This forms the OFDM modulation [32], resulting in obtaining two waveforms which are modulated by two orthogonal sub-carrier sets which are mutually shifted by Δf . Their demodulation is also computed by $2N$ -point DFT. Now both the data sets contain $2 - N$ sub-carriers but use only N sub-carriers to carry the input data-sequences.

4.4.2 Orthogonal frequency division multiplexing-LFM based signal model

The sequence as input is taken $X(m)$, it has N numbers of discrete spectral-components which are considered, separated by $K(\delta f)/(2)$. The considered $X(m)$ as input-sequence, is further interleaved by $K-1$ filled with zeros which gives rise to $X_1(m)$, and K can be $K = 1, 2, 3, 4, \dots, N-1$. Then $X_1(m)$ is shifted by $K(\delta f)/(2)$ (i.e. $\Delta f/2, \Delta f, 3\Delta f/2 \dots (N-1)\Delta f/2$) such that it generates the required signal sequences $X_2(m), X_3(m) \dots X_K(m)$ respectively. By taking IDFT operation for the KN -points sequence, all the K -sequences (data) are converted into temporal domain signal or time-domain. With this method to process, we achieve K orthogonal-modulated-waveforms by distinct sub-carrier sets which are having $K(\Delta f/2)$ shift. It is emphasized that all the required sets have KN sub-carriers however each uses N sub-carriers only. As mentioned previously in this simulation, it will be considered that the spectrum of linear-frequency-modulated signal for the OFDM-modulation. Equation (4.4) represents the discrete form of the LFM signal:

$$x(m) = \exp(j\pi k(mT_{ss})^2) \quad (4.14)$$

Where $m=0,1,2,\dots, N-1$. T_{ss} is the temporal sampling-interval, the chirp-rate is represented by k i.e. $k = \frac{B}{T_p}$, where bandwidth is B and T_p denotes the chirp-duration. The FT(Fourier-Transform) of the considered LFM signal is:

$$X(p) = FT[x(m)] = FT[\exp(j\pi k(mT_{ss})^2)] \quad (4.15)$$

Where FT in equation (4.15) denotes the Fourier-Transform operator, and $x(m)$ indicates the discrete-time sample of the chirp (LFM)-signal having N length sequence. Data sequences are subjected to interleave with zeros(0's) and shifted [114] [115]. Thus the K input data-sequences are produced as given in the equation (4.16) to (4.18):

$$X_1[p] = [x[0], 0_1, \dots, 0_K, x[1], 0_1, 0_2 \dots 0_K, x[2], 0_1 \dots 0_K] \quad (4.16)$$

$$X_2[p] = [0_1, x[0] \dots 0_K, 0_1, x[1], 0_2 \dots 0_K, 0_1, x[2], \dots 0_K] \quad (4.17)$$

Similarly

$$X_K[p] = [0_1 \dots 0_K, x[0], 0_1, \dots 0_K, x[1], 0_1, \dots 0_K, x[2]] \quad (4.18)$$

By taking the KN -point IDFT(Inverse-Discrete-Fourier-Transform) the temporal or time-domain sequence $x_1[n]$ is given by equation (4.19)

$$x_1(n) = \frac{1}{2} \left\{ x(n) \text{rect} \left[\frac{n}{N} \right] + x(n-N) \text{rect} \left[\frac{n-N}{N} \right] + \right. \\ \left. x(n-2N) \text{rect} \left[\frac{n-2N}{N} \right] + \dots + x(n-(K-1)N) \text{rect} \left[\frac{n-(K-1)N}{N} \right] \right\} \quad (4.19)$$

where $n=0,1,2,\dots KN-1$

it is known that the sequence of K orthogonal sub-carrier sets are altogether mutually shifted by $K\Delta(f)/2$, the other sequences are derived from the $x_1[n]$, that is represented as equation (4.20)

$$x_i[n] = x_1[n] \exp \left(\frac{j * (i-1) 2\pi n}{KN} \right) \quad (4.20)$$

where $i= 2, 3, 4 \dots K$,

The OFDM-LFM waveforms are generated by the equation(4.16)and (4.17), illustrated in figure(4.9). The continuous-time representation for the data sequences are indicated by taking into consideration $\Delta(f) = \frac{1}{NT_s}$. The final shifted signal is obtained using equation(4.21)

$$x_1(t) = \frac{1}{2} \left\{ x(t) \text{rect} \left[\frac{t}{T_p} \right] + x(t-T_p) \text{rect} \left[\frac{t-T_p}{T_p} \right] + \right. \\ \left. x(t-2T_p) \text{rect} \left[\frac{t-2T_p}{T_p} \right] + \dots + x(t-KT_p) \text{rect} \left[\frac{t-KT_p}{T_p} \right] \right\} \quad (4.21)$$

The other further signals are proceeded from the $x_1(t)$ as mentioned in equation(4.22)

$$x_i(t) = x_1(t) \exp \left(\frac{j\pi i t}{2T_p} \right) \quad (4.22)$$

The zero(0's) interpolation that reduces the peak-power of the waveform. The two adjacent waveforms are differently distinguish by $(K\Delta f)/2$ i.e the offset of sub-carrier frequency, and also has a $\pi/2$ phase difference at $t = T_p$. The temporal form or time-domain OFDM-LFM based waveform for the two (02) LFM waveform and six- LFM signal are shown in figure(4.10) and figure(4.11). All waveforms are showing similar type of frequency characteristics as these signals are chirp-like signals [116] [117].

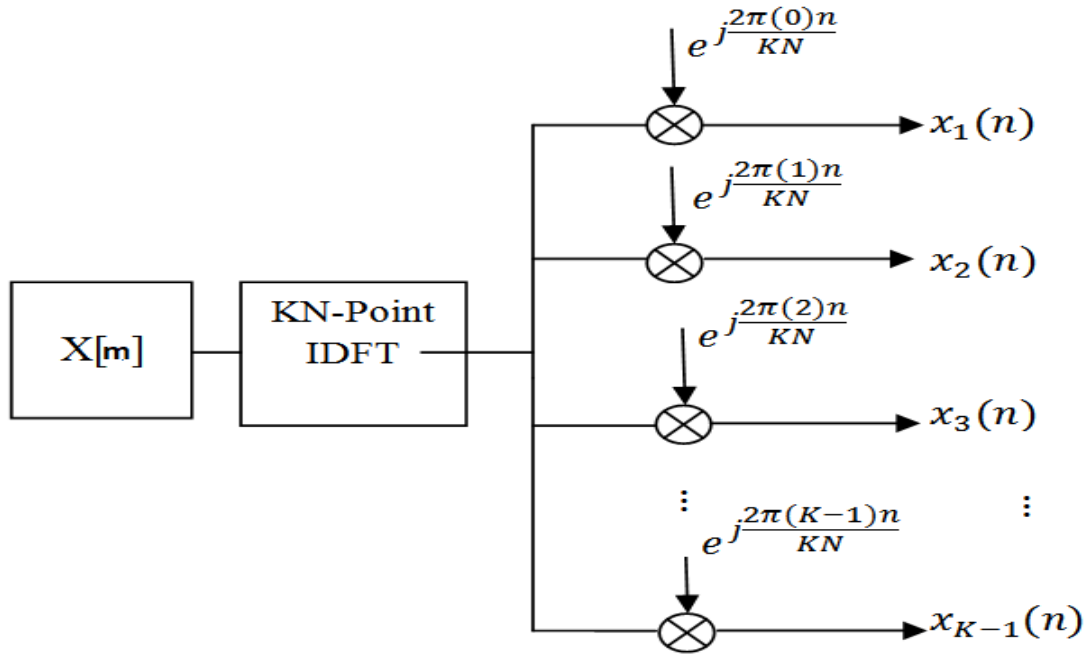


FIGURE 4.9: K-OFDM-LFM-waveform generation(block-diagram)

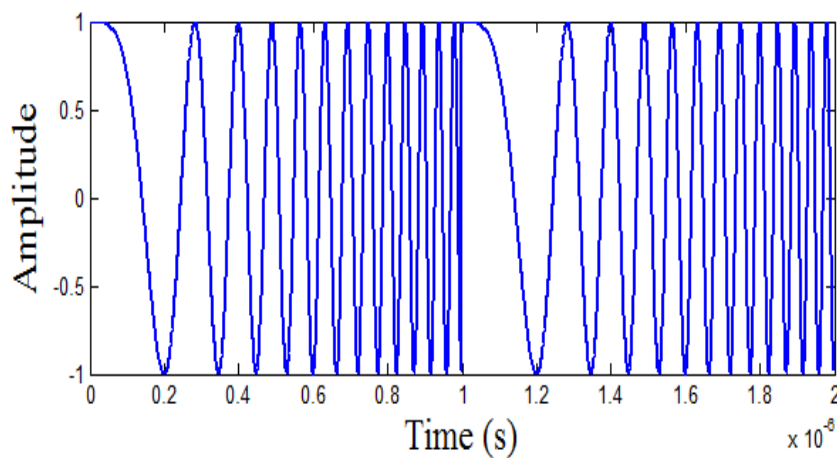


FIGURE 4.10: OFDM with two LFM signals

4.4.3 Ambiguity function analysis for multi-dynamic radar

Waveform diversity explains to use various waveforms (signals) for both transmitter (Tx) and receiver (Rx) design to improve the overall radar performance, such as detection and/or identification of target competing with interference and noise. The proposed orthogonal frequency-division-multiplexing (OFDM) with linear-frequency-modulated (LFM) waveform can be used in multi-dynamic scenarios. OFDM chirp-signal having N sub-carriers with number of K LFM signals having each T_w duration is given in equation (4.23) [26] [118] [119]

$$x(t) = \sum_{n=0}^N \sum_{k=0}^K u(t - kT_w) \exp[j2\pi f_{n,k}(t - kT_w)] \exp[j\pi k_{n,k}(t - kT_w)^2] \quad (4.23)$$

where $f_{n,k}$ and $k_{n,k}$ represents the frequency & the chirp-rate of sub carriers. Assumed the pulse duration of OFDM T_p shall satisfy such that $KT_w \ll T_p$. The derived required ambiguity-function is shown in equation (4.24)

$$|\chi(\tau, \nu)| = \left| \sum_{n=0}^N \sum_{k=0}^K u(t - kT_w) u(t + \tau - kT_w) \exp[j2\pi f_{n,k}(t - kT_w)] \exp[-j2\pi f_{n,k}(t + \tau - kT_w)] \exp[j\pi k_{n,k}(t - kT_w)^2] \exp[-j\pi k_{n,k}(t + \tau - kT_w)^2] \exp[j2\pi \nu t] dt \right| \quad (4.24)$$

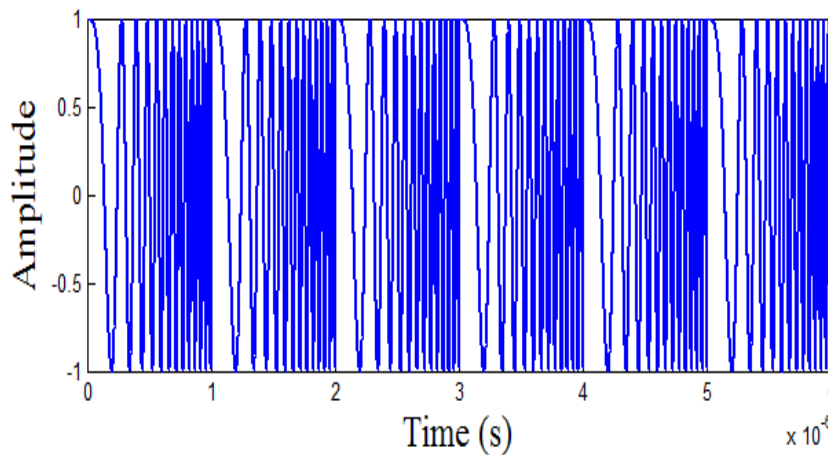


FIGURE 4.11: OFDM with six LFM signals

The details of OFDM-LFM for three LFM, five LFM and seven LFM as well as corresponding ambiguity function plot for multi-dynamic radar is shown in the following figure (4.12) to figure (4.17).

Each of the linear-frequency-modulated signal generates the corresponding side-lobes in the OFDM-LFM based waveform. Hence the number of K LFM signals generates $2K - 1$ numbers of side-lobes in ambiguity function response, the side-lobes are symmetric in nature with respect to main-lobe of response. It is to be noted that increasing in the number of LFM waveform, corresponding increasing of zero-padding by the same factor as side-lobes peak increases. The time-bandwidth product which is important parameter for the OFDM-LFM waveform also increases linearly with respect to increase the number of signals. But the side lobes also increase as shown in figure (4.20). Hence, it is to be noted

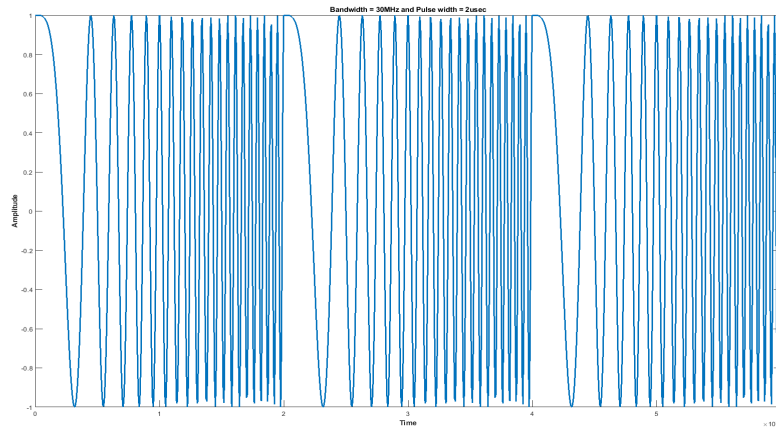


FIGURE 4.12: Three OFDM-LFM time-domain

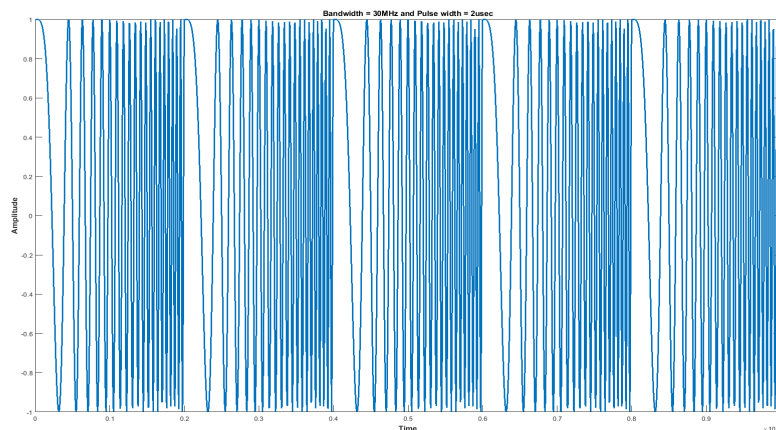


FIGURE 4.13: Five OFDM-LFM time-domain

there is a trade-off between, the time bandwidth product and the peak-side-lobe level for the waveform design to achieve better for both range as well as Doppler resolution radar waveform. The parameters have taken, for simulation purpose, which are the number of sub-carriers $N = 2048$, $T_w = 1\mu\text{sec}$ and bandwidth $B = 30\text{MHz}$. The matched-filter response of the waveform with different number of LFM signals are represented in figure (4.18) and figure (4.19). From the figure the conclusion can be made that the total number of side-lobes as illustrated and respective magnitude as indicated, increases with an increase in the number of LFM waveforms.

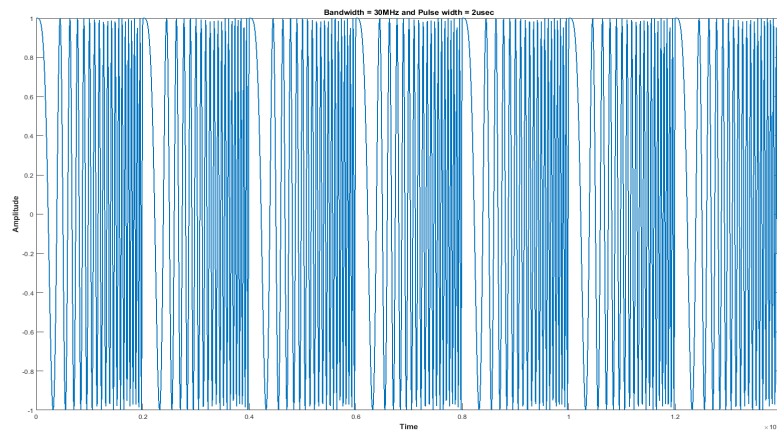


FIGURE 4.14: Seven OFDM-LFM time-domain

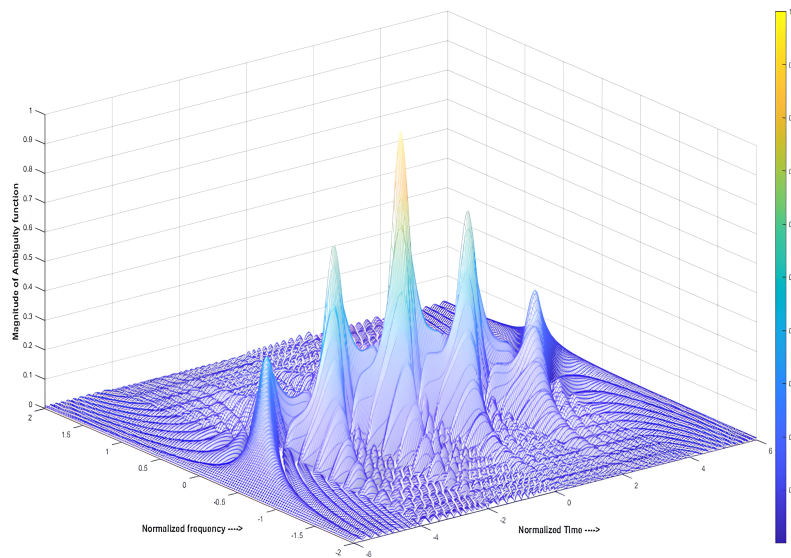


FIGURE 4.15: Three OFDM-LFM ambiguity function

The sensitivity-level for various application of radar, it is required to consider a different level of threshold. The first side-lobe as well as second side-lobe magnitudes for different number of LFM signals in OFDM-LFM based waveform is presented in figure (4.20). it is shown that the number of LFM increases in the waveform from two(02) number of LFM to six(06) number of LFM, the peak of the first side-lobe increases from -6dB to

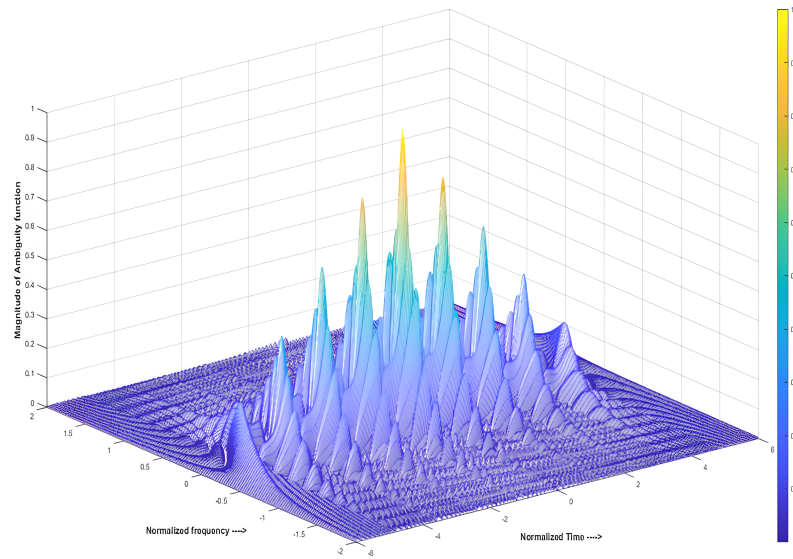


FIGURE 4.16: Five OFDM-LFM ambiguity function

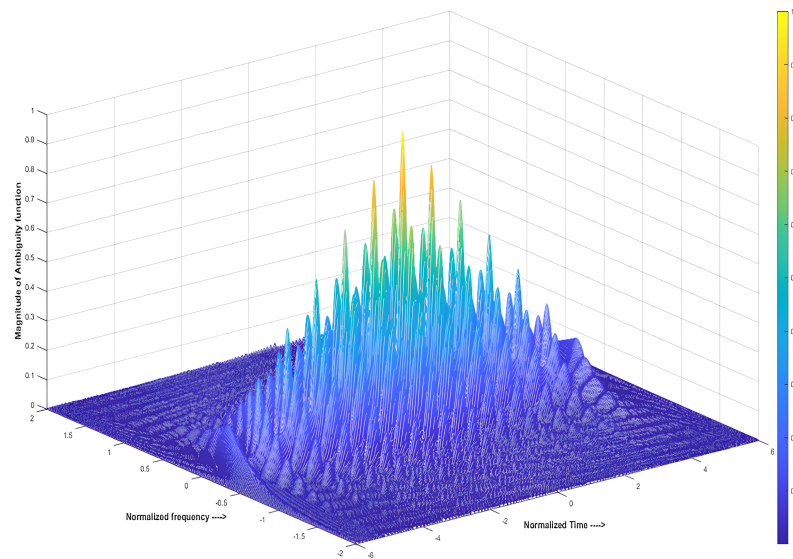


FIGURE 4.17: Seven OFDM-LFM ambiguity function

-1dB. As discussed the time-bandwidth product for the case for a various number of LFM signals indicate, more number of OFDM-LFM signals in the waveform and this will increase linearly the time-bandwidth product. But with an increase in the side-lobe level as presented in figure(4.20).

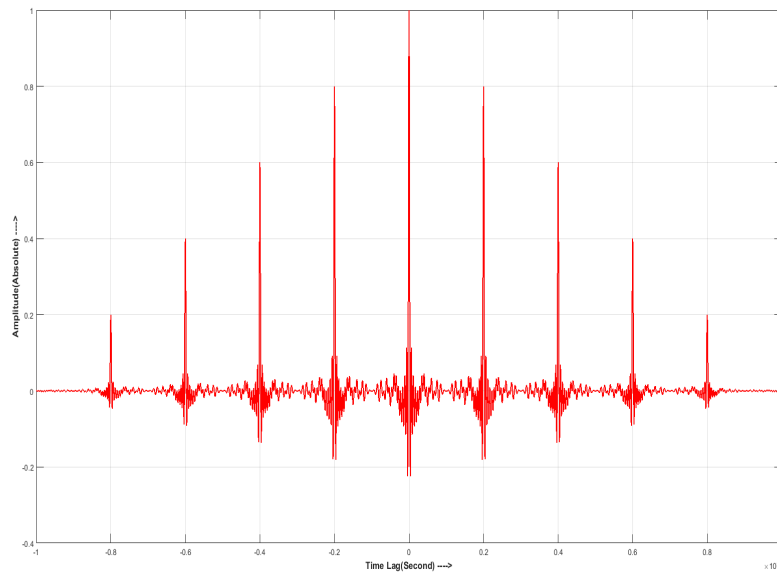


FIGURE 4.18: Five OFDM-LFM matched filter output

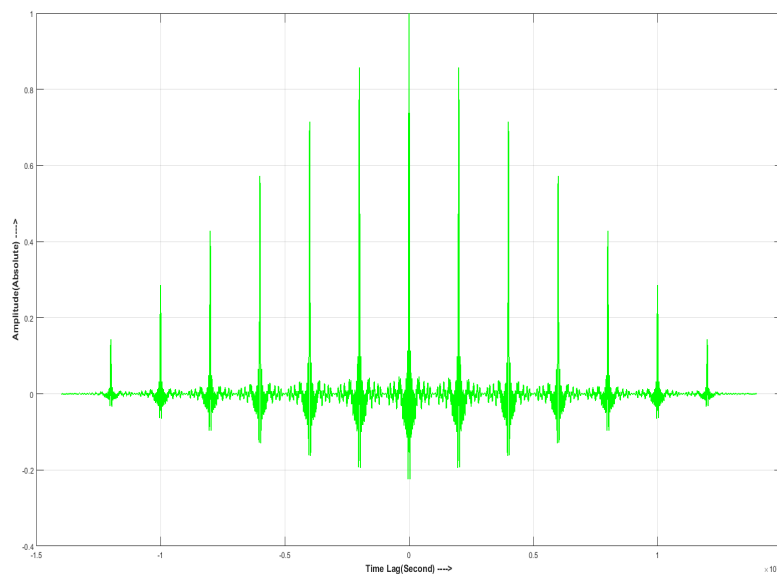


FIGURE 4.19: Seven OFDM-LFM matched filter output

4.5 Performance bench mark for waveform design for MDR

Considering the set bi-static geometry, unlike the ambiguity function, which provides information on the global resolution, the CRLBs [34] are a local measure of estimation accuracy. Both can be used to assess the error properties of the estimates of the signal parameters. In [34] [112] [113] the author derived a relationship between CRLB and AF, which has been successfully used in the analysis of passive and active arrays [34]. In monostatic configuration, [33] claims that for the Fisher Information Matrix (FIM) the following relationship holds:

$$\mathbf{J}_M(\tau_a, \nu_a) = -2\text{SNR}$$

$$\begin{bmatrix} \frac{\partial^2 |\chi(\tau, \nu)|}{\partial \tau^2} & \frac{\partial^2 |\chi(\tau, \nu)|}{\partial \tau \partial \nu} \\ \frac{\partial^2 |\chi(\tau, \nu)|}{\partial \nu \partial \tau} & \frac{\partial^2 |\chi(\tau, \nu)|}{\partial \nu^2} \end{bmatrix} \bigg|_{(\tau=0, \nu=0)} \quad (4.25)$$

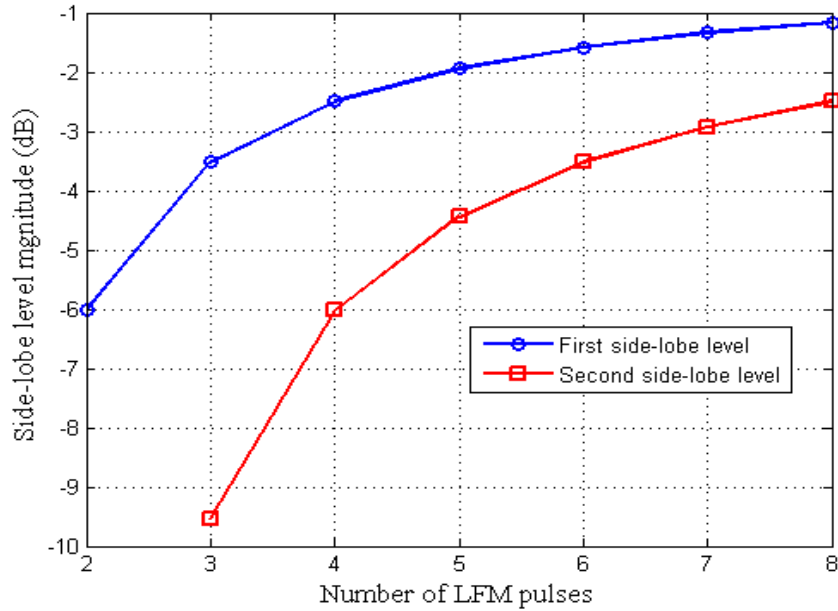


FIGURE 4.20: First and second side-lobe for various OFDM-LFM ambiguity function

The above FIM relation in equation (4.25) holds true for both monostatic, bi-static, as well as multi-dynamic cases it does not depend on the choice of the parameters of the ambiguity function. In a given time, multi-dynamics can be thought of set of bi-static configurations.

As discussed in the earlier section with simulation result, in a radar waveform as the number of OFDM-LFM signal increases, this will increase in the time-bandwidth product linearly. However, the side-lobes of ambiguity function, which are undesirable for radar performance also increase. It is also shown that there is a trade-off in such waveform design for the occurrence of side-lobe-level (SLL) of an ambiguity function response for more number of LFM pulses as shown in figure(4.20). Hence, waveform is required to be chosen according to the desired value of SLL.

It is required to have a waveform performance parametric benchmark for choosing the optimal waveform. MDR waveform design performance is determined by deducing CRLB for delay τ_a and Doppler ν_a of ambiguity function. This includes optimal Tx-Rx pair geometry for the estimation of range/Doppler for a particular target of interest. Fisher Information Matrix (FIM) gives a parametric benchmark to choose a radar waveform for optimal performance. FIM $\mathbf{J}_M$ for monostatic radar case is available in [34] as shown in equation (4.25). The CRLB for monostatic cases are derived in [33] for delay or range in equation (4.26) whereas for Doppler in equation (4.27). The corresponding root CRLB can be expressed as in equation (4.28).

$$\text{CRLB}(\tau_a) = [\mathbf{J}_M^{-1}(\tau_a, \nu_a)]_{(1,1)} \quad (4.26)$$

$$\text{CRLB}(\nu_a) = [\mathbf{J}_M^{-1}(\tau_a, \nu_a)]_{(2,2)} \quad (4.27)$$

$$\begin{aligned} \text{RCRLB}(\tau_a) &= \sqrt{\text{CRLB}(\tau_a)} \\ \text{RCRLB}(\nu_a) &= \sqrt{\text{CRLB}(\nu_a)} \end{aligned} \quad (4.28)$$

In MDR configuration, the ambiguity function for the estimation of true range of target, the range delay $\tau(R_R, \theta_R, L)$ is a function of R_R, θ_R, L and for estimating target velocity, the corresponding Doppler $f_d(R_R, V, \theta_R, L)$ is a function of R_R, V, θ_R and L . Where θ_R

is the angle of the target-receiver line with respect to an assumed north, V is the target velocity in the direction along the target-receiver line and R_R is the range from the target to the receiver. FIM for MDR $\mathbf{J}_{\text{MDR}}$ can be deduced with respect to $R_R^{(k)}$ and $V^{(k)}$, as performance parameter for range estimation and velocity estimation with respect to k th receiver and k th transmitter. The SNR for each bi-static combination depends on the effects of the path loss. $R_R^{(k)}$ and $V^{(k)}$ are denoted as range and Doppler of target with respect to k th receiver of MDR geometry configuration where as $R_a^{(k)}$ and $V_a^{(k)}$ are the actual range and velocity that would have been detected by monostatic radar.

The FIM is derived for MDR ($\mathbf{J}_{\text{MDR}}$), it is the function of $R_R^{(k)}$ and $V^{(k)}$ corresponding to k^{th} receiver, as shown in equation (4.29). The detailed mathematical derivation for $\mathbf{J}_{\text{MDR}}$ is shown below.

$$\mathbf{J}_{\text{MDR}}(R_R^{(k)}, V^{(k)}) = -2\text{SNR} \begin{bmatrix} \frac{\partial^2 |\chi(R_R^{(k)}, V^{(k)})|}{\partial R_R^{2(K)}} & \frac{\partial^2 |\chi(R_R^{(k)}, V^{(k)})|}{\partial R_R^{(K)} \partial V^{(K)}} \\ \frac{\partial^2 |\chi(R_R^{(k)}, V^{(k)})|}{\partial V^{(K)} \partial R_R^{(K)}} & \frac{\partial^2 |\chi(R_R^{(k)}, V^{(k)})|}{\partial V^{2(K)}} \end{bmatrix} \bigg|_{(R_R^{(k)} = R_a^{(k)}, V^{(K)} = V_a^{(K)})} \quad (4.29)$$

After approximation given [33] the FIM for multi-dynamic radar can be derived as given in equation(4.30) and equation(4.31) and CRLB can be computed from equation(4.32) and equation(4.33) for range and Doppler resolution respectively

$$\begin{aligned} & \left[\mathbf{J}_{\text{MDR}}(R_R^{(k)}, V^{(k)}) \right]_{(1,1)} = \\ & [\mathbf{J}_{\text{MDR}}]_{(1,1)} \left(\frac{\partial \tau}{\partial R_R^{(K)}} \right)^2 + 2 [\mathbf{J}_{\text{MDR}}]_{(1,2)} \frac{\partial \nu}{\partial R_R^{(K)}} \frac{\partial \tau}{\partial R_R^{(K)}} + [\mathbf{J}_{\text{MDR}}]_{(2,2)} \left(\frac{\partial \nu}{\partial R_R^{(K)}} \right)^2 \end{aligned} \quad (4.30)$$

$$\begin{aligned} & \left[\mathbf{J}_{\text{MDR}}(R_R^{(k)}, V^{(k)}) \right]_{(2,2)} = \\ & [\mathbf{J}_{\text{MDR}}]_{(1,1)} \left(\frac{\partial \tau}{\partial V^{(K)}} \right)^2 + 2 [\mathbf{J}_{\text{MDR}}]_{(1,2)} \frac{\partial \nu}{\partial V^{(K)}} \frac{\partial \tau}{\partial V^{(K)}} + [\mathbf{J}_{\text{MDR}}]_{(2,2)} \left(\frac{\partial \nu}{\partial V^{(K)}} \right)^2 \end{aligned} \quad (4.31)$$

$$\text{CRLB}_R = [J_{\text{MDR}}^{-1}(R_R^{(K)}, V^{(k)})]_{(1,1)} \quad (4.32)$$

$$\text{CRLB}_\nu = [J_{\text{MDR}}^{-1}(R_R^{(K)}, V^{(k)})]_{(2,2)} \quad (4.33)$$

4.6 Summary

In this chapter, the ambiguity function analysis of monostatic and bi-static radar system with LFM and rectangular pulses are analysed. The change in bi-static radar ambiguity function as the change in geometrical structure changes are described and shown in simulations results. An OFDM-LFM waveform design is discussed for the multi-dynamic radars in which a very high time bandwidth product can be achieved. But there is trade-off shown between the number of LFM signals and the side-lobe levels. When the number of LFM signals increases inside the OFDM-LFM waveform, time-bandwidth product increases. But this results increases the side-lobe level. This problem can be resolved by applying a better optimization algorithm.

5

Investigation on multi-dynamic radar operational requirement for airborne surveillance system

This chapter describes the various challenges of operational requirements of multi-dynamic radar. The challenges for phase and time synchronization across airborne transmitter and receiver. Synchronization is the first and fundamental requirement for radar signal processing, such as for matched filter. A complete design approach is presented in the form of a Phase and Time synchronization unit (PTSU). It also describes how phase sync can be achieved using high a precision stable oscillator with associated digital signal processing (DSP) circuitry.

5.1 Introduction

As explained earlier multi-dynamic radar(MDR) configuration is emerging as state of the art in the field of airborne radar, where several airborne radars work in a coordinated manner. MDR systems provide better spatial diversity and resistance to other vulnerabilities in combat scenario. MDR gives high performance to operational requirement compared to multi-static, bi-static, or monostatic radar for airborne surveillance. For the last few decades, stealth technology in aircraft design has been one of the most widely used approaches for target hide-out from radar systems. This technology uses the concept of diverging the electromagnetic energy away from the radar receiver hence it is based on low observable principles to reduce the radar cross section (RCS) in the direction of radar. Earlier, ground based radars like bi-static radar or multi-static radar were used to counter such target hideouts. This type of radar system has a number of transmitters as well as a number of receivers placed geographically apart at fixed locations. Airborne radar has its own advantages for better visibility and coverage for airborne surveillance. In multi-dynamic radar, the transmitters (Tx) and receivers (Rx) are placed on the airborne platform such as aircraft or un-manned air-vehicles (UAV). The operational challenge for MDR configuration is how to exploit the dynamics of Tx-Rx sensors for benchmarking the performance. The MDR system-geometry is considered as a set of bi-static-radar at a given time and simulated with all possible airborne Tx-Rx pair combinations for evaluating MDR performance. The simulation is also made in the earlier chapter for the proposed orthogonal frequency-division-multiplexing linear-frequency-modulated (OFDM-LFM) based waveform design using a derived ambiguity function for the proposed MDR scenario. However, the major challenges across the independent airborne Transmitter (Tx) and receiver (Rx) are to work in time and phase synchronization to process the target echo. GPS based local oscillator (LO) synchronization across TX-RX is available, but the inherent standard delay in GPS synchronization is much more than the required for any airborne radar waveform. In various scenarios as well as various situations, GPS signal may not be available to make phase-time synchronization across the platform. Hence a design methodology is required for an MDR to achieve phase synchronization across airborne Tx and Rx platforms.

5.2 Multi-dynamic radar coverage for airborne surveillance

The MDR scenario is more complex to model for the analysis of target detection and classification. An MDR at a given instant of time can be considered as a multi-static-radar system. The typical MDR scenario to detect a stealth target is shown in figure (5.1). The figure shows no target echo in the transmission direction (Tx/Rx). However, other receivers mounted over un-manned air-vehicles (UAVs) and positioned in different directions receive echo to process and detect the presence of the target. MDR coverage defines in total, the limits within which one or more radar systems can detect objects. It is determined by both radar sensitivity and propagation factors. The radar sensitivity is one of the parameters that indicate the ability performance of the radar system to detect the target by suppressing various interferences. It is defined in terms of signal-to-noise ratio (SNR), that depends on various governing factors such as transmitted power by antenna, gain of the antenna of Tx & beam, transmitted frequency etc. In multi-dynamic-radar maximum coverage area is computed by taking a set of Tx-Rx pair configurations. For simplification, here Tx-Rx pair is considered as bi-static pair, as shown in the earlier chapter. The bi-static angle (β) is taken between transmitter to target and target to the receiver. θ_R and θ_T angles are taken with respect to true north, these are the azimuth-angles (Az) of the receiver and transmitter, respectively. R_T and R_R are ranges of the target from transmitter and receiver, respectively. $R_T + R_R$ is computed from equation (2.1) [3], where τ is round trip time and c denotes the velocity of radar waves. The ranges R_T and R_R can be computed from equations (2) and (3), assuming known base-line distance L and beam-scheduling parameter θ_R and θ_T for transmitter-receiver pair. These angles can be computed from the beam scheduling parameters.

MDR system is assumed to be a coherent (time sync as well as phase sync) across all airborne Tx-Rx pairs, each receiver (Rx) is assumed to be capable of receiving target skin echoes from any transmitted waveform by any transmitter (Tx) as part of MDR network. The coverage area or detection range for MDR changes dynamically, however for benchmarking the performance, the temporal snapshot of MDR system is considered.

5.3 Time and Phase synchronization for multi-dynamic radar Doppler processing

MDR operates with different transmitting and receiving antennas mounted on separate airborne platforms. Time and phase synchronization is necessary and fore-most for MDR systems. The problem in MDR, the phase synchronization of the local oscillator (LO) across geographically apart airborne Tx-Rx. Since in both Tx and Rx independent unlocked local oscillators (LO) are used, and the common reference-time in such cases is not available for the airborne transmitter(Tx) and the airborne receiver(Rx). The lack of synchronization across the multi-airborne Tx-Rx units will not serve to exploit the advantages of MDR systems. The synchronization errors across Tx-Rx shall lead in range cell migration (RCM) error, also the distortion in azimuth dependent phase this will lead issue in mono-pulse tracking. Thus it implies that synchronization errors will degrade the MDR performance quality.

Temporal synchronization and phase synchronization of the transmit and receive systems gives true range and Doppler of target [120] [121] [122]. Global positioning system (GPS)-based technique for achieving successful time and phase synchronization for MDR is proposed as shown in figure(5.2). More importantly, residual time synchronization errors may be compensated with a high-precision range alignment method, and residual phase synchronization errors also may be compensated.

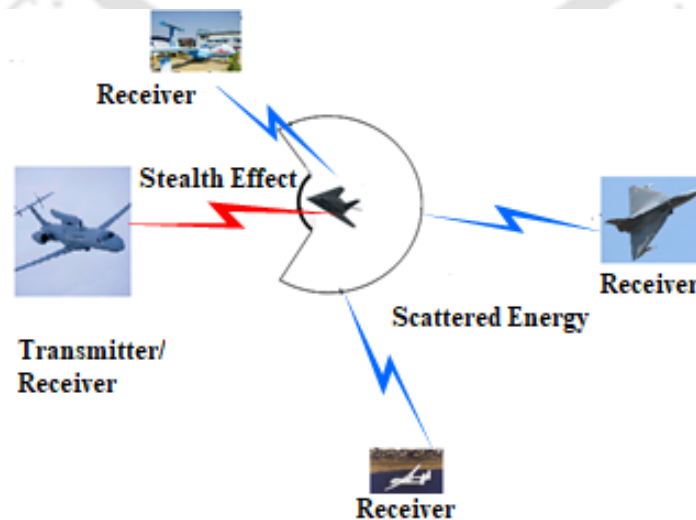


FIGURE 5.1: Multi-dynamic-radar scenario (RCS variation across Rx)

Time and phase synchronization is a significant challenge to process radar signals with phase-locking in RF up and down converter chain for pulse compression processing [123], digital beam-forming, digital mono-pulse tracking [124] [125], and subsequent Doppler processing. It is proposed to have a mechanism, GPS-based phase/time synchronization to the respective local oscillator (LO) for Tx and Rx units as shown in figure(5.3).

GPS will remove variation due to temperature, age, and random walk from the oscillator performance [126] [127]. The Allan variance for GPS will be a straight line that constantly decreases the variances, the variation is due to the white noise, but it does not have a random walk or flat characteristics. It is clear from figure (5.4), that GPS has better long term stability, where as oscillator has better short-term stability.

But the availability of a global positioning system (GPS), inherent GPS clock phase jitter, and phase stability affect the performance of MDR. Hence, the proposed mechanism is evolved and used in the phase/time synchronization unit (PTSU), which holds the phase stability for a sufficient long time. It includes maintaining the phase synchronization across PTSU. Multiple units of PTSU for each airborne platform (consisting of Tx / Rx radar) can be used for phase synchronization. PTSU can control and lock the phase

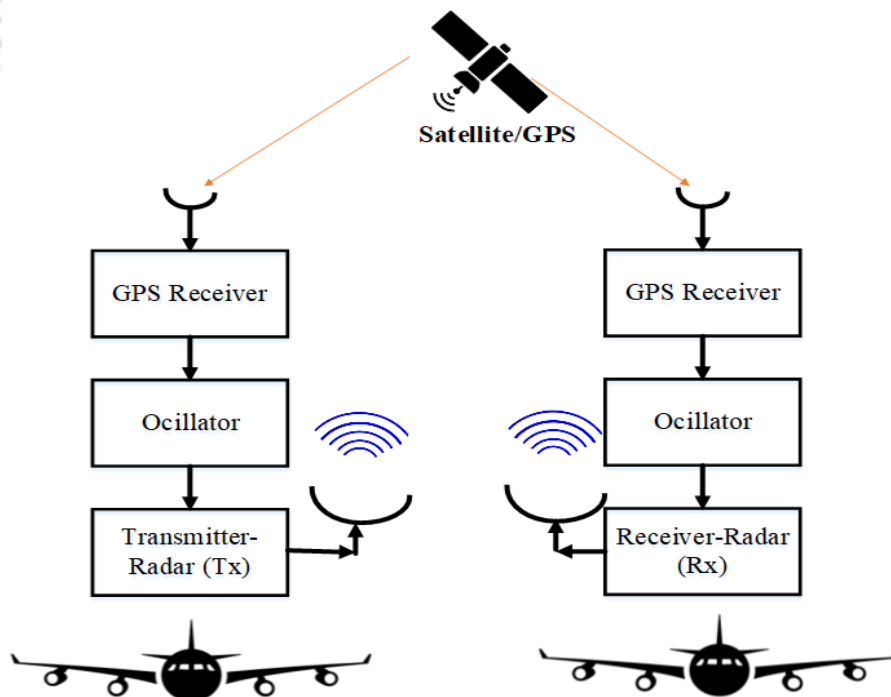


FIGURE 5.2: GPS based time-phase synchronization for MDR

of the local oscillator of the radar waveform generator [128][129]. Tx airborne platform can transmit the waveform, and airborne Rx can process the target echo as both Tx-Rx units are in phase sync with synchronized PTSU. The design can be made with rubidium like stable crystal oscillator-based circuitry in such a way that the phase synchronization

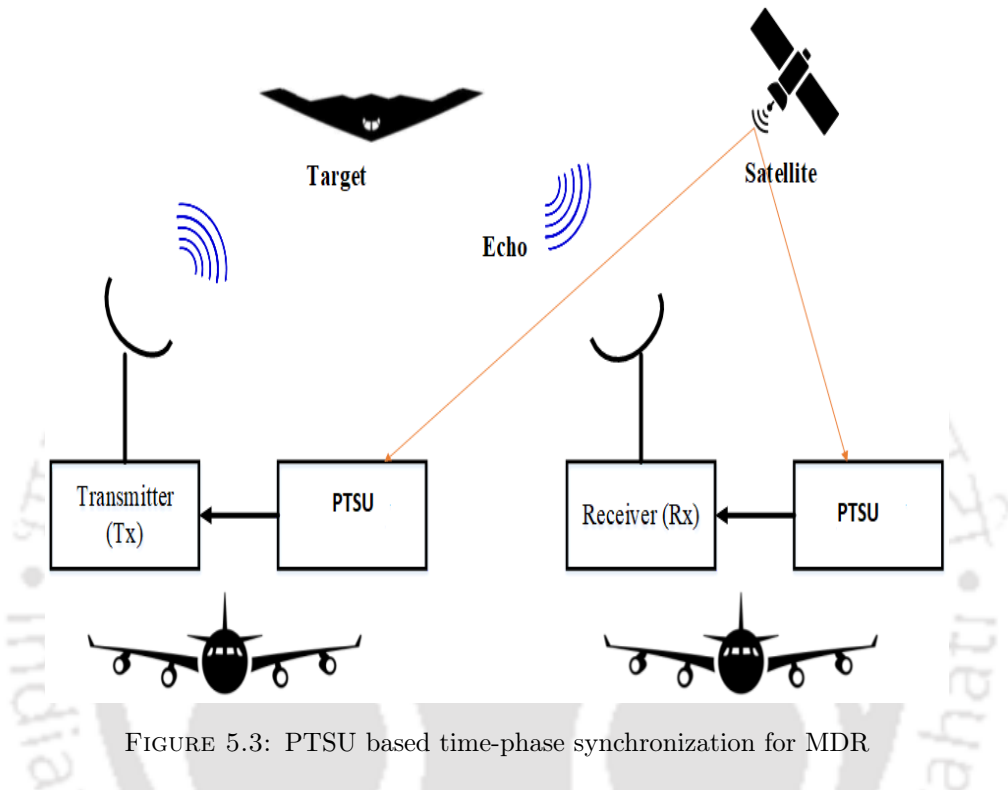


FIGURE 5.3: PTSU based time-phase synchronization for MDR

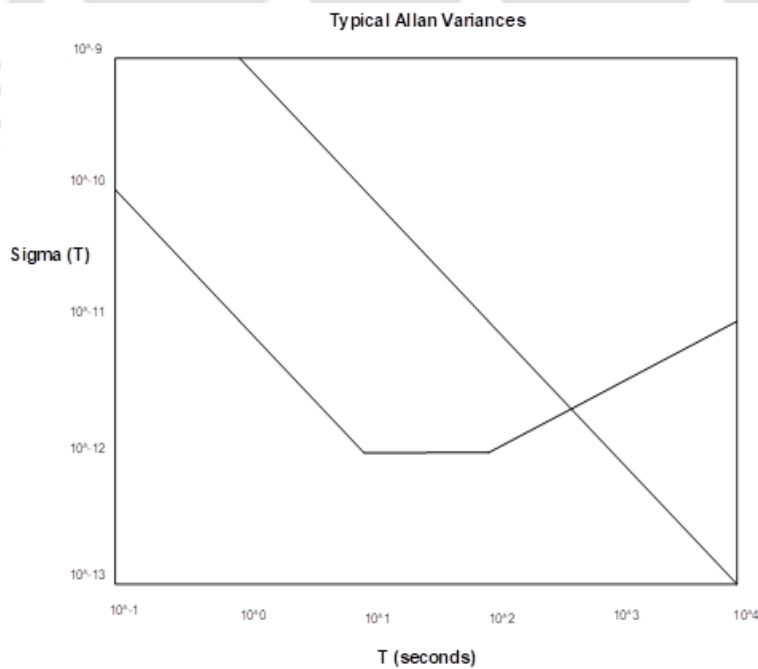


FIGURE 5.4: Typical Allan variance

across airborne radar Tx-Rx units can hold over a long time to exploit the MDR systems capability. The precise time and phase sync across systems is maintained by locking the crystal oscillator to GPS across multiple units on the ground for Tx-Rx platforms and the same phase lock hold while the system is flying over the test-bed. The stability of the oscillator's frequency and phase output is selected to meet the harmonics levels and power on warm-up conditions. The proposed system consists of a GPS receiver, which receives GPS data from GPS satellites and provides an accurate time and 1 PPS (Pulse Per Second) to the system as a reference for time roll-over. To maintain phase and time sync locally within the system, a crystal clock is used, which generates the required clock; for example, it is taken as 10 MHz clock signal, and 1 pulse per second is generated locally (LPPS) using a divider chain circuitry. The system synchronizes the 1 LPPS with the reference 1 PPS by disciplining the assumed frequency 10 MHz output of the crystal oscillator. This control and synchronization mechanism involves a processor and its associated devices. The PPS is disciplined with respect to GPS-PPS (GPPS) by the phase locking mechanism. The digital signal processing (DSP) circuitry continuously requires to compare the generated 1 PPS (LPPS) with respect to GPS-PPS and adjusts the oscillator frequency/phase through crystal control input such that both are perfectly matched in phase. The DSP firmware dynamically computes the required adjustment for frequency, which can be based on frequency measurements and oscillator characteristics. Based on this, DSP outputs data to digital to analog converter (D/A), which controls crystal's control input as shown in figure(5.5).

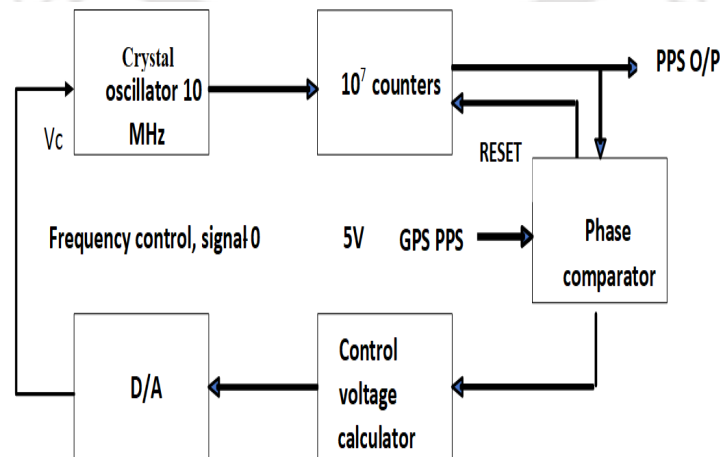


FIGURE 5.5: Phase locked loop for PTSU for MDR system

The clock/phase disciplining mechanism has several states. On switch on, GPS clock enters power-up state. In this state, the oscillator is allowed to warm up adequately to a stable state. After this, disciplining crystal oscillators with GPS begins by controlling the drift of generated local PPS with respect to GPS-PPS. Since the GPS itself has a jitter of more than ± 200 ns under good signal conditions (for time dilution of precision (TDOP) is 2), required to be filtered. However, when TDOP is more than 5, GPS jitter can be as high as ± 500 ns. Based on the drift measurement, the control-voltage of the crystal oscillator can be controlled. Variation in the control-voltage produces corresponding changes in the frequency of the oscillator. This process ensures that crystal output is locked to GPS-PPS. When phase locking is achieved, the divider chain, which generates 1 PPS, resets to indicate aligning of the PPS output perfectly with GPS-PPS. Then, the clock is considered to have attained synchronization. During this condition, when there is no GPS-PPS reference, the drift of the crystal oscillator is dependent on the crystal characteristics with respect to ambient temperature aging. So, it is very important that control-voltage is optimized, and the longer the averaging time, the better is the nullification of GPS drift, also control-voltage shall be better optimized with respect to temperature and aging. The parameter design for an MDR scenario as a case study is presented. The crystal oscillator clock frequency (f) is assumed 10 MHz. Crystal control frequency range is $\pm 2 \times 10^{-9} \times f$, corresponding to voltage level 0V to 5V. The clock period variation is computed as:

$$1/(10 \times 10^6 + 2 \times 10^{-9} \times 10^7) \text{ s to } 1/(10 \times 10^6 - 2 \times 10^{-9} \times 10^7) \text{ s}$$

The variation in 1PPS with respect to voltage is

$$\begin{aligned} &= [1/(10 \times 10^6 + 0.02) \text{ to } 1/(10 \times 10^6 - 0.02)] \times 10^7 \text{ ns} \\ &= 99999998.00 \text{ ns to } 1000000002.00 \text{ ns} = \pm 2 \text{ ns} \end{aligned}$$

The variation in PPS vs control-voltage provided for correction is shown in figure (5.6) to maintain PTSUs in sync across various transmitters and receivers on a different airborne platform.

The D/A used is of 16 bit for computation. Hence, maximum digital computed input shall be 65535. Assumed maximum control-voltage is 8 V. So, the minimum correction that can be applied to crystal 0.006104 ns/s. GPS jitter is considered ± 500 ns.

$$(500/65535) \times (8/10) = 0.006104 \text{ ns/s}$$

At any given time, the disciplining mechanism can be described as being in one of several states. The mechanism of transition from one state to the next in response to the condition is described in figure (5.7). From reset, the disciplining mechanism enters the warm-up or power-up state, in this state, the oscillator is allowed to warm up to an adequately stable state. At this point, the GPS receivers begin to track satellites, and time or position fixes become available. The initial synchronization process begins by adjusting the frequency of the oscillator based on the frequency offset information only. At the same time, the PPS output is constantly shifted into place. Once the frequency is close, the unit shifts into one of its phase-locking modes and the timing / phase output begin to settle in. After a period of settling, the PTSU transition to the synchronization state.

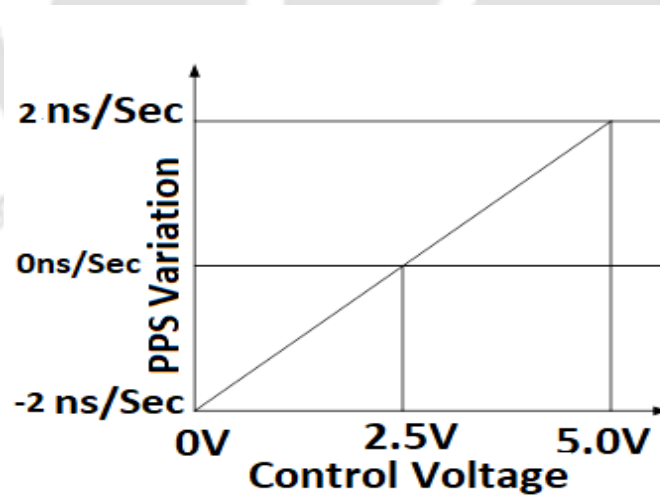


FIGURE 5.6: PPS variation vs control voltage

5.4 Summary

The state of art concept of PTSU is described to achieve phase synchronization across the airborne platform, also mentioned by the synchronization issues and reduction of time jitter. The design presented takes care the phase coherency across the same units installed to various airborne receivers. Only GPS based design cannot provide short-time stability, especially for the beam scheduling for medium PRF based radar. The design is made for specific airborne surveillance with respect to flying duration stability as well as specific to specific crystal used. For certain Airborne surveillance scenarios, the GPS links may not work as well GPS sync may lead to distorted incoherency. For MDR scenario, it is considered to be one Tx-Rx (Airborne Radar) and many receivers (Rxs) fitted over different UAVs. Unless and until time and phase are in sync across platforms the target data processing cannot be done. In such scenarios custom design made PTSU has to provide the Time/Phase coherency across Tx-Rx pairs over the required flying time. In multi-static radar case, phase can be locked via physical cable laying across the geographical location. But the same situation is not possible for multi-dynamic case, phase and time synchronization unit (PTSU) design is proposed, each unit can fly with each airborne Tx and airborne Rx. PTSU architecture is made such that once all

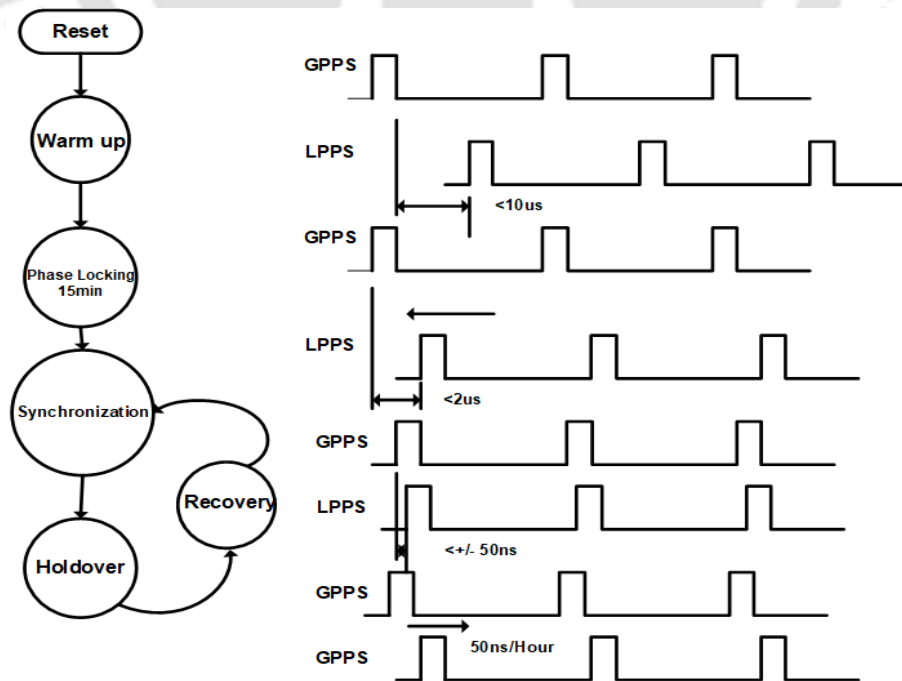


FIGURE 5.7: Disciplining of clock phase

PTSUs are made in sync in ground the after all PTSU will remain in sync upto many hours of flying. With this concept all radar Rxs can process the skin echo receive by the target from known airborne Tx transmitted waveform. Also indicated about MDR operation, however the complete signal and data processing chain for MDR with results on unclassified data is not shown here.





6

Summary and conclusion

This chapter is for the final summary of the research and proposed future work to enhance the effectiveness of multi-dynamic radar for airborne surveillance systems.

6.1 Summary of contribution and discussion

In this thesis we investigate the operational aspect of multi-dynamic radar scenario for airborne surveillance. We propose coded transmitter identity based tailored waveform for multi-static radar, which may not require all Tx-Rx to be physically connected. Studies and simulation shows IFF waveform usage for carrying out the primary radar waveform identity. In this thesis we have also presented the concept of STAP based MDR functionality. we propose ambiguity diagram for OFDM-LFM based waveform

and effect of side-lobe in choosing the waveform. Studies have been performed on how airborne radars with transmitter and receiver positions change dynamically and can work in a coordinated manner for radar signal processing this includes the requirement of time and phase synchronization using PTSU.

The main contributions of this thesis are summarized and listed below:

1. Mathematical derivation and simulation of Bi-static Radar Configuration, which includes the simulation of radar coverage for various SNR levels.
2. Study and simulation of multi-static radar, this includes the tailored waveform design with transmitter identity and signal processing for target detection and data fusion.
3. Study and simulation of multi-dynamic radar as an extended concept of multi-static radar. The mathematical derivation and simulation of the IFF Mod-S waveform based waveform identification requirement for primary radar. Additionally, simulations based on Space-Time Adaptive Processing (STAP) are shown for multi-dynamic radar systems, aiming to optimize the utilization of receiver sets in the presence of jamming signals.
4. Ambiguity function derivation for multi-dynamic radar, simulation of ambiguity function for the proposed OFDM-LFM waveform for medium PRF cases. Benchmarking the performance of the waveform by deriving the CRLB function for optimal results.
5. Details of design approach to achieve the Phase synchronization across airborne Tx and Rx by using Phase and Time Synchronization unit (PTSU) for Doppler processing.

6.2 Suggestion for future work

The approach for the following future work is shown in figure (6.1).

1. In the subsequent research, experiment and simulation based studies are planned to be performed, as well parametric performance based on CRLB derivation and evaluation to be done for target detection [130] [131] [132].
2. In parallel, the prospect of time reversal (TR-MIMO)[131][132][133][134] will also be explored. Subsequently, these two works on multi-dynamic and TR will be modified further to enhance the performance of the multi-dynamics scenario[21].

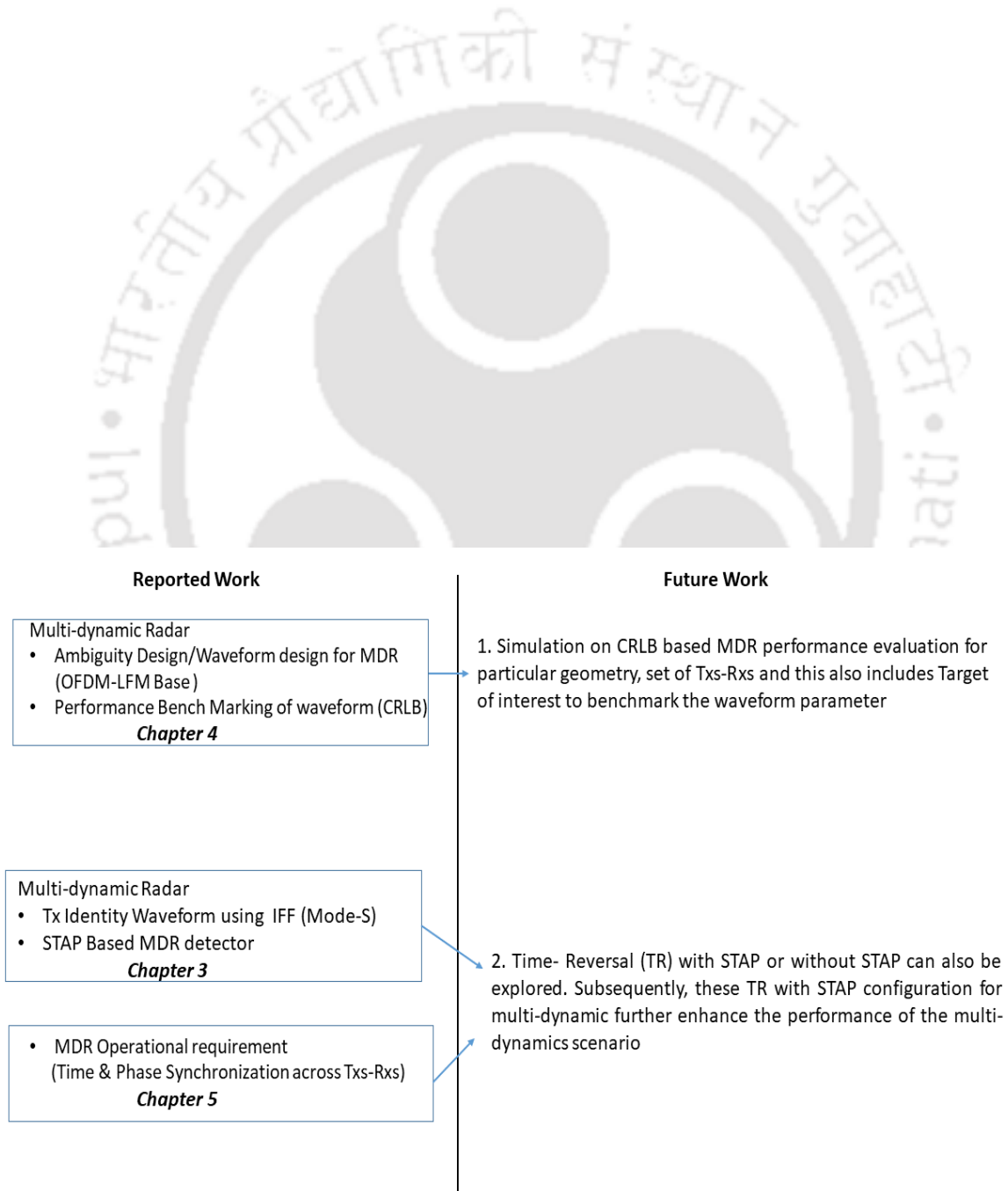


FIGURE 6.1: Future work



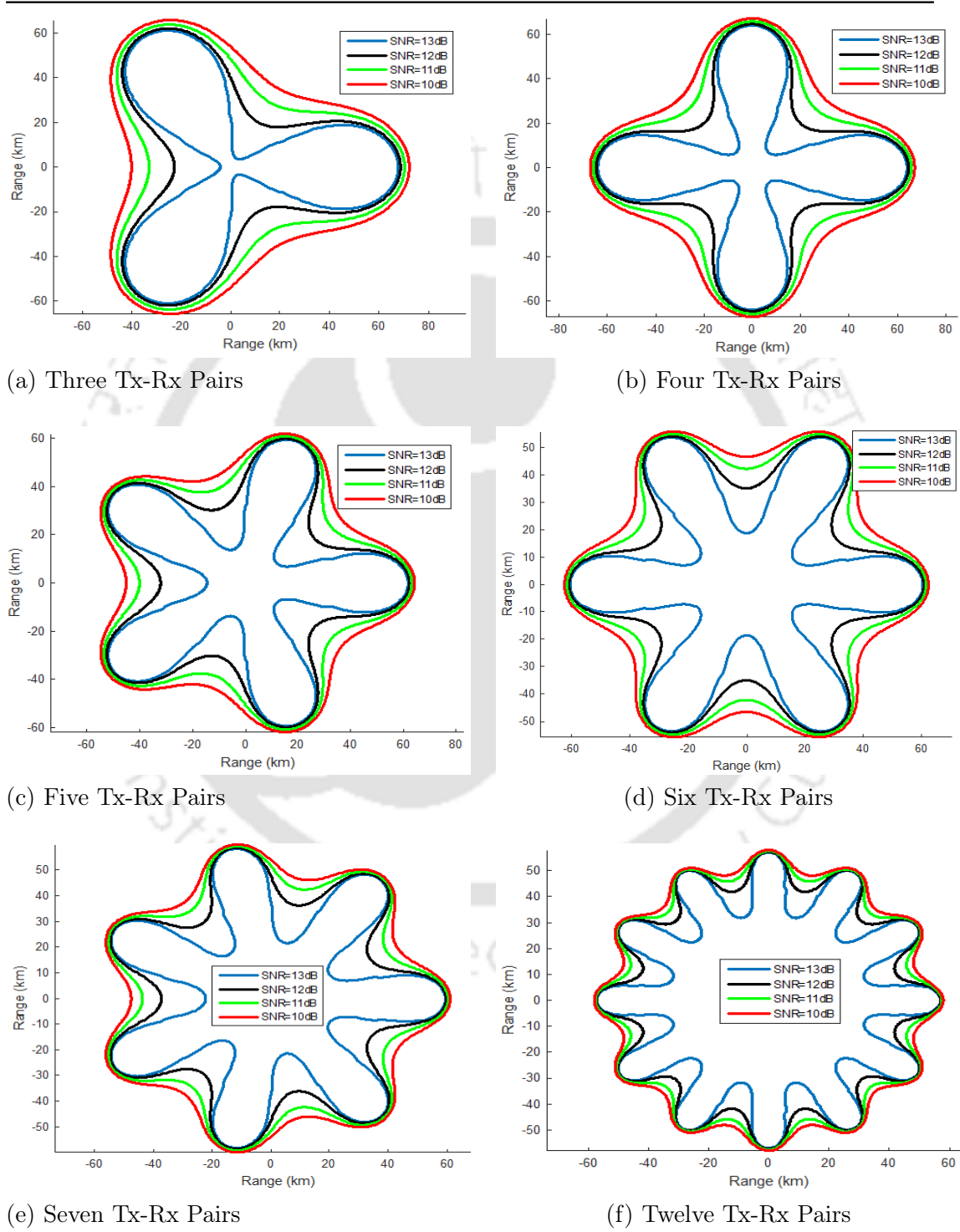
A

Appendix-A



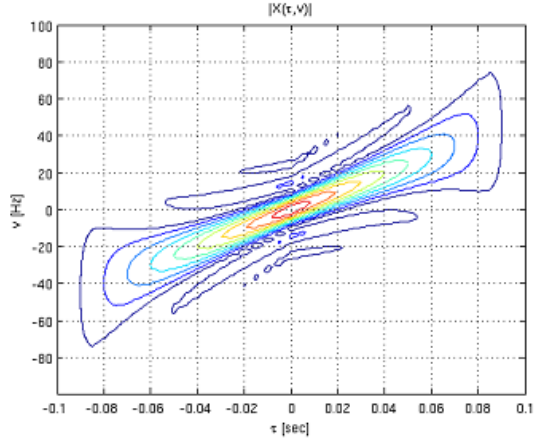
A.1 Multi-dynamic radar coverage for various Tx-Rx combination

TABLE A.1: Coverage area for Multi-dynamic radar scenario with various Tx-Rx pairs for different SNR

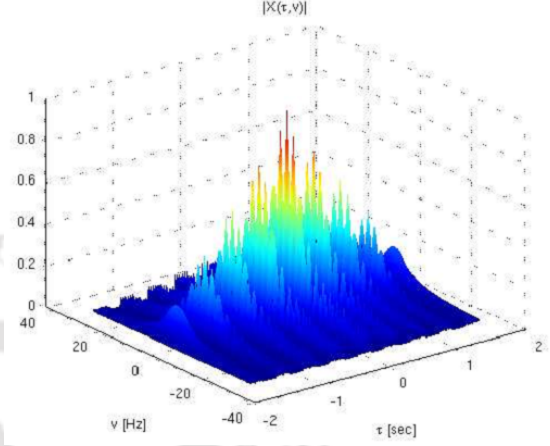


A.2 Ambiguity function analysis of MDR system (1)

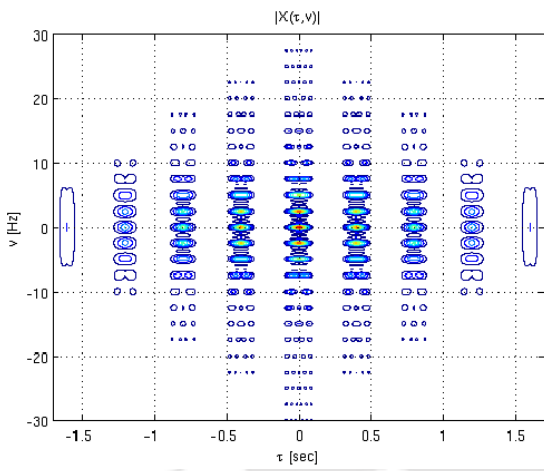
TABLE A.2: Ambiguity function representation for analysis MDR



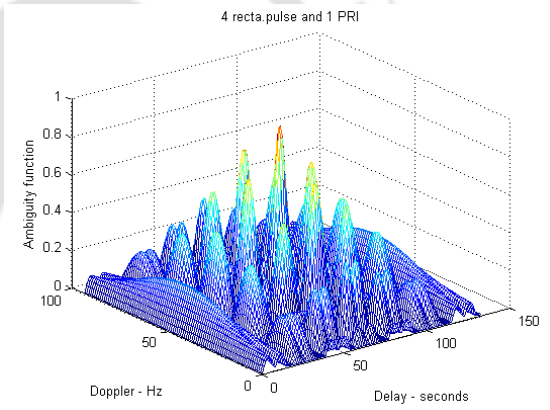
(a) Ambiguity contour for LFM



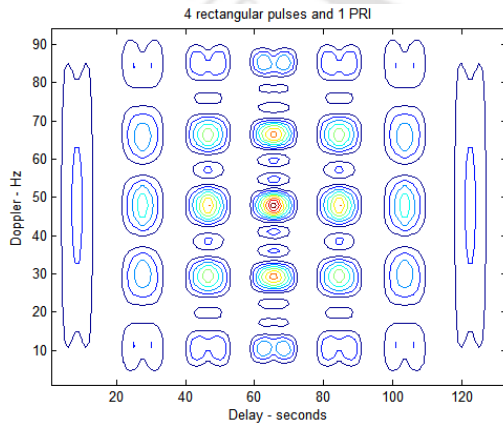
(b) 5 Burst Ambiguity plot



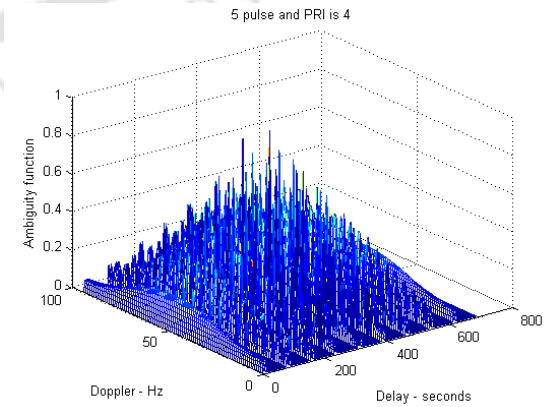
(c) 5 Burst Ambiguity contour



(d) Ambiguity plot complex Pulse PRI=1



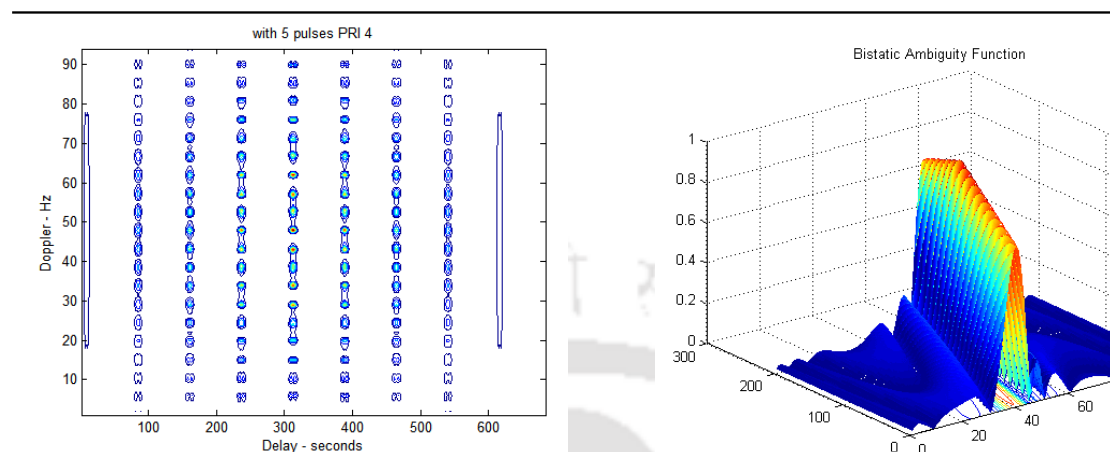
(e) Ambiguity contour complex Pulse=4 PRI=1



(f) Ambiguity plot complex Pulse=5 PRI=4

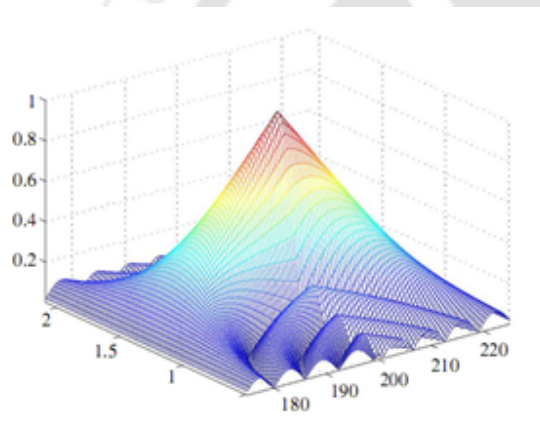
A.3 Ambiguity function analysis of MDR system(2)

TABLE A.3: Ambiguity function representation for analysis MDR(1)



(a) Ambiguity contour complex Pulse=5 PRI=4

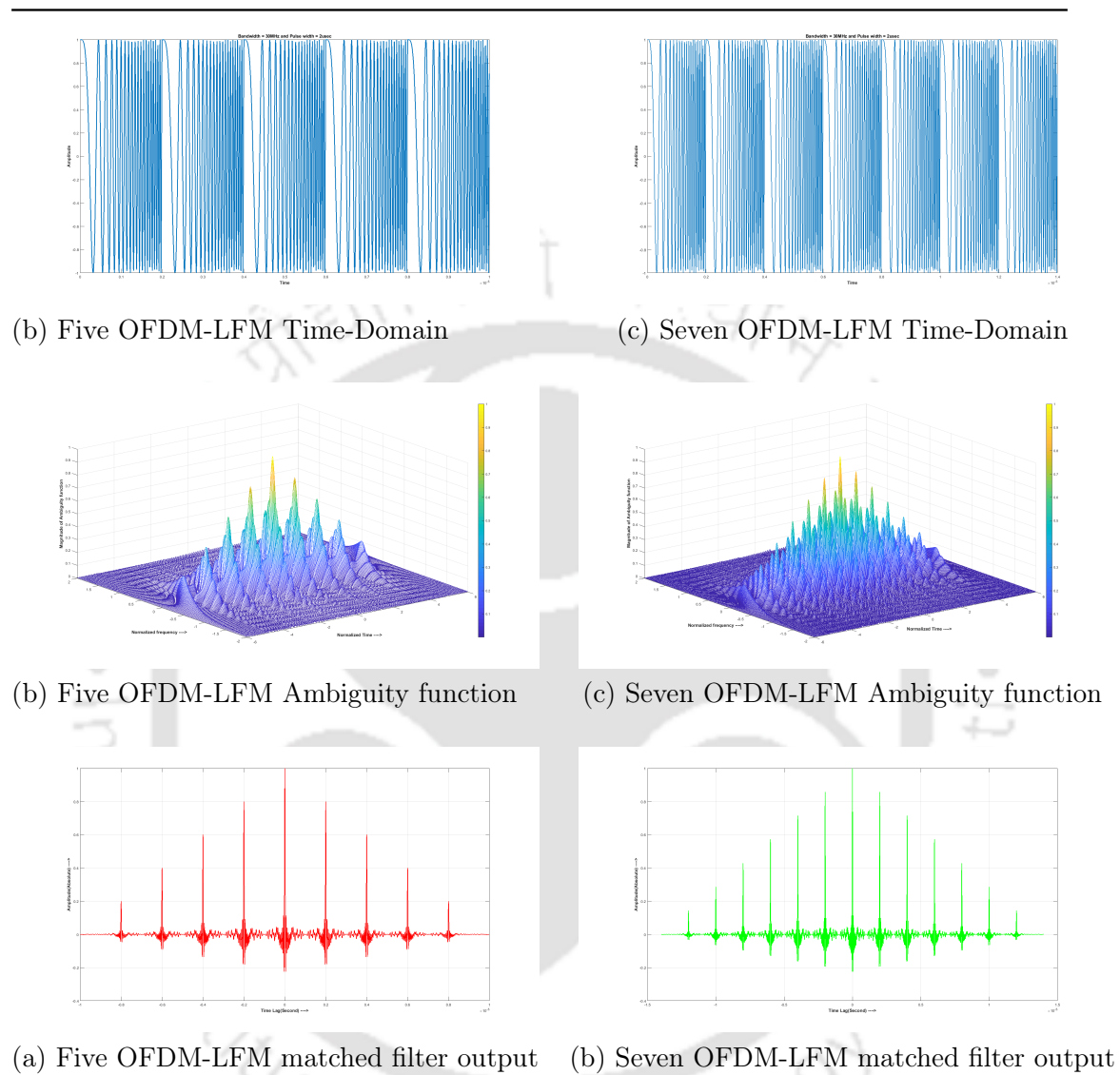
(b) AF Bistatic radar($\theta_R = 0$ degree)



(c) AF Bistatic radar($\theta_R = -45$ degree)

A.4 Ambiguity function analysis of MDR System(3)

TABLE A.4: Ambiguity function representation for analysis MDR(3)





B

Appendix-B



B.1 Issues and gap for research scope

TABLE B.1: The following tabular form brings the issues and gaps for research scope

SN	Issues or topics	State of Art (with Literature Reference)	Existing Research Gaps
1.	Challenges for Detection of Low Observable Targets for airborne surveillance system	- MIMO radar [135] There are two possibilities for MIMO configuration: MIMO radar with co-located antennas [23] [20] and MIMO radar with widely separated antennas [21], which is also referred to as statistical MIMO [136]. - MIMO radar: The transmission of multiple orthogonal waveforms from different antennas to exploit waveform diversity [137]. - The tradeoff for using MIMO versus Phased Array as it does not have coherent gain. The phased Array-MIMO concept is utilized where a group of antenna elements are used for TX and RX instead of single antenna elements [24]. This includes looking into MIMO virtual array concept [25] and side-lobes performance [138]. - MIMO Time reversal explores (TR-MIMO) the rich scattering in a multipath environment to achieve high target detect-ability [139][131]. Using orthogonal waveform signalling, the time reversal MIMO radar yields higher detection performance than conventional statistical MIMO radar in high clutter [140]. - Multi-static Radar, which exploits the concept of MIMO radar where Many Transmitter (TX) to Many Receiver (RX) enhances the detectability of the target [141][4][30][142]. In Multi-static where the platforms are assumed to be static but now a days most of the platforms are airborne platform. - Multi-dynamic radar waveform design and operational requirement.	-Multi-static/Bi-static, being Tx and Rx has limitations to position at fixed geographical locations, which may not be sufficient to detect all different dynamic aspect angles. -Developing the concept of a Multi-dynamic radar case study to exploit for the betterment of such low RCS target detection. - This includes the transmitter identity transmission, so that for wide spread in geographical locations where physical cable connection among Tx-Rx may not be required. -The concept of Airborne radar network working as MDR. What could be the waveform requirement or ambiguity diagram to define the characteristics of waveform. This includes how to benchmark the waveform performance. - How to synchronize Phase and Time among all airborne Tx-Rx, in order to work as fully coordinated and synchronized airborne radar as MDR.

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LIST OF PUBLICATIONS

International Journals

1. Jawed Qumar, S. Christopher, Ratnajit Bhattacharjee, "Multi-Dynamic Radar a Concept for Strategic Air Surveillance System" *IEEE Sensor Letter*, 2021

International Conferences

1. Jawed Qumar, S. Christopher, Ratnajit Bhattacharjee, "Multi-Static extended to Multi-Dynamic Radar for Airborne Surveillance Systems" *Proceedings of National Science day* , 2016
2. Jawed Qumar, S. Christopher, Ratnajit Bhattacharjee, "Novel techniques for target detection with waveform identity in multi-static radar using bi-static configuration" *Proceedings of IEEE Signal Processing and Communications (SPCOM)*, 2016
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7. Jawed Qumar, S. Christopher, Ratnajit Bhattacharjee, “Multi-dynamic Radar with Time Reversal Processing for better Clutter Mitigation” *Proceedings of Third International Conference on Smart Trends for computing and Communication (Smart-Com)*, 2019

Awards/Honor/Talks

1. DRDO National Science Day 2016

