Reduced complexity impulse radio ultra wide band direction finding systems

Lie, Joni Polili

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**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**REDUCED COMPLEXITY IMPULSE RADIO
ULTRA WIDE BAND DIRECTION FINDING
SYSTEMS**

**JONI POLILI LIE
SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING
2008**

**Reduced Complexity Impulse Radio
Ultra Wide Band Direction Finding Systems**

Joni Polili Lie

School of Electrical and Electronic Engineering

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Summary

Since FCC's approval of limited uses of UWB systems in the frequency band 3.1-10.6 GHz, research on UWB has proliferated extensively for the past few years. One area of growing interest in UWB impulse radio (IR) is localization. In realizing such a system, time-delay-based approaches enjoy high popularity due to its attractive centimeter-accuracy and low-power low-cost implementation, while direction-of-arrival (DOA) approach is of less interest due to an additional cost incurred for the use of antenna arrays. If compensated with a lower complexity processing, UWB IR direction finding (DF) offers more benefits as compared to a time-delay-based localization system. All of the contributions presented in this thesis aims to address the UWB IR DF system design using a reduced complexity receiver structure.

Several challenges emerge when considering low complexity UWB IR DF in a realistic channel. Detecting UWB pulse with a low complexity technique restricts the use of Nyquist rate sampling. Secondly, suppressing multipath optimally requires either a *prior* channel information or a high complexity structure. Thirdly, the well-known formulation of high resolution narrowband DOA estimation is no longer applicable. And lastly, extension of the aforementioned problems to a multiple-user case increases the difficulty of the problem.

In overcoming the above challenges, we propose several methods for UWB IR DOA estimation. The first contribution in this thesis is the DF receiver using digital

channelization receiver architecture (presented in Chapter 2), where the use of Nyquist rate sampling is no longer required. Suppressing multipath is achieved by a simple analog level threshold detector (LTD). Performance analysis reveals that the error in detection due to the LTD will give rise to DOA estimation errors. Therefore, this method is not robust enough, despite the advantage of feasible implementation.

An alternative means of finding the DOA using a simpler structure is the second contribution (presented in Chapter 3). In this contribution, the DF approach is motivated by the observation of a staircase-shape waveform whose slope indicates the DOA. We then design an analog differentiator to estimate the slope. The slope is detected at the output of the differentiator using a peak and hold detector. Although simpler than the channelization-based approach, this approach is also dependent on the detection accuracy of the LTD. Thus, it does not solve the problem posed to the channelization-based approach.

The third and most important contribution is the DF receiver based on the time delays estimated from channels of the array (presented in Chapter 4). The estimation can be realized using a simple analog circuitry and a low sampling rate analog-to-digital converter (ADC). Because of the presence of outliers in the time delay estimates, a fractional norm formulation is considered. The formulation serves to *de-emphasize* the impact of the large time delay estimation errors on the accuracy of the DOA estimation. In addition to the fractional norm metric, this technique also incorporates an error bound on the time delay estimation to ensure the reliability of the DOA estimator. Numerical analysis illustrates that this technique achieves good robustness and outperforms previous approaches.

In a time-hopping multiple access environment, the received signal originates from multiple transmissions of different sources coming from different directions. We address this problem in two ways. First is to estimate the DOA from one user-of-interest

and regard other active users as interferences. Second is to estimate the arriving directions from all active users. In the first problem, we shall assume that the receiver array has a *prior* knowledge of the hopping sequence assignment and the hopping sequence of the user-of-interest is required to be spectrally separable from other active users. That means the spectral information of the user-of-interest can be extracted from the spectral information of the received signal. The hopping sequence design that exhibits such characteristics is also described. The second problem is dealt with using a subspace-based DF method. The method effectively shows multiple peaks at the directions of all active users. Alternatively, the time delay based DF technique from the third contribution can also be used. This is because there is only single transmission at single chip duration. The works in the DOA estimation in a time-hopping multiple access environment are the last contribution in this thesis (presented in Chapter 5).

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List of Notations

$r(t)$	the received UWB signal in a multipath channel at single antenna .18
j	the frame index19
T_f	the frame interval19
τ_{toa}	the TOA of the received signal19
$\eta(t)$	zero-mean AWGN with power spectral density $N_o/2$19
$w_{mp}(t)$	the received multipath waveform at single antenna19
E_s	the pulse energy19
w_l	the pulse waveform of the l -th multipath component19
a_l	the fading level experience by the l -th multipath component19
τ_l	the delay experience by the l -th multipath component19
θ_l	the DOA of the l -th multipath component19
x_n, y_n	the position of n -th sensor in the two-dimensional space19
$\tau_{l,n}$	the differential delay due to the l -th multipath component19
n	the index of the sensor array19
$r_n(t)$	the received UWB signal in a multipath channel at n -th sensor21
c	the speed of light21
$w_{mp}^{(n)}(t)$	the received multipath waveform at n -th sensor21
L	the number of the received multipath components24
$R_n(\omega_k)$	the DFT of the received UWB signal at n -th sensor24
$W_l(\omega_k)$	the DFT of the l -th UWB signal waveform received24
$\aleph_n(\omega_k)$	the DFT of the received AWGN24
i	the imaginary unit, $\sqrt{-1}$24
$[\cdot]^T$	the matrix transpose operation25
$\mathbf{r}(\omega_k)$	the vector that contains $R_n(\omega_k)$25
$\mathbf{A}_f(\omega_k, \boldsymbol{\theta})$	the matrix that contains the directional information for the CSM method25
$\boldsymbol{\theta}$	the vector that contains θ_l25
$\mathbf{w}(\omega_k)$	the vector that contains $W_l(\omega_k)$25
$\mathbf{n}(\omega_k)$	the vector that contains $\aleph_n(\omega_k)$25
$\mathbf{P}_r$	the autocorrelation matrix of (r)25
$\mathbf{P}_w$	the autocorrelation matrix of $\mathbf{w}$25
$\mathbf{P}_n :$	the autocorrelation matrix of $\mathbf{n}$25
$[\cdot]^H$	the Hermitian matrix operation25

$E[\cdot]$	the expectation operation	25
$\mathbf{T}(\omega_k)$	the focusing matrix for the CSM algorithm	26
$\bar{\mathbf{P}}$	the coherent spectral matrix	26
θ_s	the scanning direction	27
$\mathbf{E}_n$	the matrix formed by the noise eigenvectors	27
$\mathbf{r}(j, \omega_k)$	the vector that contains $R_n(\omega_k)$ from the j -th frame	27
N_p	the number of pulses received (frames observed) for processing	27
$y(t)$	the sum of the received signals at the array	28
$Y(\omega_k)$	the CFT of $y(t)$ at ω_k	28
$R_o(\omega_k)$	the CFT of $r_o(t)$ at ω_k	28
k_m	the index of the frequency to be channelized	29
M	the number of the frequencies to be channelized	29
$\mathbf{y}$	the vector that contains $Y(\omega_{k_m})$	29
$\mathbf{r}_o$	the vector that contains $R_o(\omega_{k_m})$	29
$\text{diag}\{\cdot\}$	the diagonal matrix operation	29
$\mathbf{1}_{N \times 1}$	a column vector of all elements equal to 1	29
$\mathbf{A}(\theta)$	the matrix that contains the directional information formulated for the channelization-based DF method	29
$\check{\mathbf{A}}$	$\text{diag}\{\mathbf{r}_o\} \mathbf{A}$	29
$\mathbf{P}_A$	the projection of matrix $\check{\mathbf{A}}$	29
$\mathbf{P}_A^\perp$	the orthogonal projection of matrix $\check{\mathbf{A}}$	29
$\mathbf{I}$	the identity matrix	30
f_m	the m -th channel of the channelization structure	35
f_o	the fundamental frequency	35
$y_{xm}(t)$	the complex analog signal after the analog complex mixing	35
$H(i\omega)$	the frequency response of the analog filter	35
$h(t)$	the impulse response of the analog filter	35
$y_{fm}(t)$	the output of the analog filter in the channelization structure	35
$\otimes$	the convolution operation	35
f_s	the sampling frequency	35
y_m	the discrete sampled signal of $y_{fm}(t)$	35
N_{sh}	the number of samples of y_m used for the processing	36
$\check{r}_n(t)$	the output of the LTD at n -th antenna	38
$p(t)$	the rectangular pulse	38
τ_p	the rising time of $p(t)$	38
T_{latch}	the pulse width of $p(t)$ or the latch duration	38
V_p	the peak point voltage	42
V_v	the valley point voltage	42
V_{pp}	the projected peak point voltage	42
I_p	the peak point current	42
I_v	the valley point current	42
d	the inter-element spacing for ULA	50
$DF(\theta_s)$	the DF function whose peak indicates the estimated DOA	52

E_s/N_o	the ratio of the received pulse's energy and the noise PSD	60
f_L	the high frequency limit	76
f_H	the high frequency limit	76
$\lfloor \cdot \rfloor$	the floor function	96
N_b	the number of possible placements within the frame interval	96
$\check{R}_n(\omega_k)$	the CFT of $\check{r}_n(t)$	100
$\text{Re}\{\cdot\}$	the real part	100
τ_p^{max}	the maximum possible value of τ_p	100
$[t_a, t_b]$	the start and end time of the integration	102
x_n	the sampled integration's output of the n -th sensor	102
t_r	the resolution of the time delay estimation	106
$d_M(\theta_s)$	the Minkowski distance metric	110
ϵ	the threshold that limits the deviation of the TOA	111
ς	the threshold that limits the number of the good TOAs	111
q	the index of the user	139
c_j	the hopping sequence	139
$w_{tx}(t)$	the transmitted pulse waveform	139
N_f	the number of pulses in a period of the transmission	139
T_c	the chip interval	139
N_h	the number of the chips within the frame interval	139
$W(f)$	the CFT of $w_{tx}(t)$	140
δ_D	the Dirac delta function	140
$c_q(f)$	the sum of the code dependent terms in PSD of the TH-MA signal	140
ρ_q	the variable defined in $C_q(f)$	142
f_{nb}	the frequency of the narrowband system	144
$\mathbb{N}$	natural numbers, $\{0, 1, 2, 3, \dots\}$	144
Q	the number of active users	152
α_q	the path loss attenuation experienced by q -th user	153

List of Abbreviations

UWB	Ultra Wideband	1
FCC	Federal Communications Commission	1
IR	Impulse Radio	2
TOA	Time-of-Arrival	5
DOA	Direction-of-Arrival	5
GPS	Global Positioning System	6
DF	direction finding	6
ADC	Analog-to-Digital Converter	10
BPF	Band Pass Filter	16
CSM	Coherent Subspace Method	18
WCSM	Weighted Coherent Subspace Method	18
AWGN	Additive White Gaussian Noise	19
DFT	Discrete Fourier Transform	24
CFT	Continuous Fourier Transform	28
LPF	Low Pass Filter	32
LTD	Level Threshold Detector	37
pdf	probability density function	41
PSD	Power Spectral Density	50
SNR	signal-to-noise ratio	55
CIR	Channel Impulse Response	62
cdf	cumulative distribution function	68
CL	Confidence Level	66
DAC	Digital-to-Analog Converter	72
PRI	Pulse Repetition Interval	76
TR	Transmitted Reference	94
ED	Energy Detection	94
SFSBC	Single Frequency Subband Channelization	97
SR	Stored Reference	103
TH	Time Hopping	136
MAI	Multiple Access Interference	137

Chapter 1

Introduction

Research in wireless radio technology based on ultra wideband (UWB) radio has proliferated over the past few years. One of the research milestones in UWB radio, that could possibly be the trigger, is the release of “First Report and Order” by Federal Communications Commission (FCC) in 2002. In this report, FCC defines UWB transmissions as those having fractional bandwidth that is greater than 20% or occupy 500 MHz or more bandwidth [2]. As for UWB radio device operation, the permissible operational frequency range is limited to 3.1-10.6 GHz. Since other existing radio systems also operate within this band, FCC enforces a strict limitation on the transmission power. Different limits are imposed for different frequency ranges in order to have good coexistence with other existing radios. That means, the presence of UWB system’s transmission is very unlikely to interfere with other existing radios.

One possible means of operation for UWB radio devices is by employing a train of extremely narrow or short duration pulses. Each pulse instantly occupies a huge bandwidth because its pulse width is on the order of sub-nanoseconds. In between two pulses, there is a period of silence. Due to its unique discontinuous transmission, such a signalling scheme for UWB radio implementation is well-known as impulse radio

(IR). UWB IR enjoys great popularity among other signalling schemes. This is due to its potential capability for high data rate communications as well as high precision delay measurements.

1.1 Motivation

1.1.1 Unique Transmission of Ultra Wideband Impulse Radio (UWB IR)

As pointed out in Shannon's capacity equation [3], a bandwidth increment in a communication system is linearly proportional to a channel capacity increment. With higher channel capacity, the system is ideally capable of supporting a higher data rate communication. Nowadays, the demand for future communication has been increasing due to many potential applications including wireless transmission of audio-visual data. As a result, many researchers from industry and academic are motivated to exploit the potential of UWB IR unique transmission and propose solutions to answer for the demand.

Besides the ultra wide bandwidth potential, UWB IR systems can be realized using low complexity processing at both the transmitter and receiver end. The reason is that it requires no frequency-carrier modulation stage. The transmitted pulse is directly sent to the antenna from a pulse-generating circuit. Without the frequency-conversion stage that is commonly required for narrowband transmission, UWB IR transmitter is expected to be compact, lightweight and cheap. On the other side, the receiver's processing (after RF front-end) is also without a frequency-conversion stage. In the end, this potential might enable low cost realization of such a system. Figure 1.1 illustrates this potential as compared to a conventional communication system.

Figure 1.1: Comparison of conventional communication system and UWB communication system. Without the/any up conversion and down conversion stages, UWB system has the potential to provide a compact, lightweight and cheap system.

In addition to the advantage of low cost realization, UWB IR transmitter has the potential to be energy-savvy. This can be explained as follows. The pulse transmission is typically followed by a long silent period, that is much longer than the duration of the pulse itself (see Figure 1.2) and this silent period does not consume any power. If implemented for a localization system, the battery-powered transmitter will likely last longer and in the end reduces the maintenance cost of such a system.

Figure 1.2: A typical pulse train transmission of impulse radio UWB. The silent period duration is shown to be much longer than the pulse duration. This enables low power transmitter.

In a realistic multipath propagation, UWB IR system enjoys multipath diversity as a consequence of the short pulse transmission. It is able to temporally discriminate the direct and reflected UWB IR signals because multiple copies of the transmitted pulse from different propagation path are received. As a result, UWB IR waveforms can offer robust communication in urban and indoor environments [4,5]. Furthermore, when the transmitters and receivers are accurately synchronized, precise range measurements can be achieved by detecting and localizing the first arrival multipath component. Due to these advantages, systems employing UWB IR waveforms can simultaneously offer robust communication and positioning capability in a realistic multipath environment. In particular, applications such as combined wireless communications and Blue Force Tracking system [6], wireless sensor network [7] and real time location systems with wireless data exchange capability [8], will benefit from using such waveforms.

The high precision propagation-delay measurement capability, if translated into ranging, promises a centimeter-level ranging accuracy. This high precision ranging property makes UWB IR a viable candidate for localization enabling-technology. In fact, several commercial off-the-shelf localization systems are developed based on the UWB IR technology [9, 10]. The localization systems are based on the estimation of the time-of-arrival (TOA) of its emission to an array of distributed receivers. The TOA can be measured by detecting and localizing the first arrival of the received UWB IR signal with a leading edge detector [11].

1.1.2 Direction-of-Arrival (DOA) Estimation Based Localization

Apart from using the TOA as the distance-based location metric, alternatively, a localization system can be realized using a direction-based location metric, known as the direction-of-arrival (DOA) [12]. Since a TOA-based approach has been shown to deliver a high accuracy system, a DOA-based approach is rarely considered for UWB IR-based localization [13]. Besides, the increasing cost of using an antenna array is one of the main concerns for the DOA-based approach.

Principally, there is a distinct difference between locating a target based on TOA and DOA estimation. To estimate two-dimensional position of an object, one needs at least three good TOA estimates from three different receivers as references. Alternatively, one only needs two DOA estimates. Assuming that it requires the same processing complexity to calculate for the TOA and DOA and the cost of additional antennae is less significant as compared to the increasing cost due to higher complexity processing, locating a target based on DOA estimation is preferred over the TOA estimation. Figure 1.3 illustrates a TOA based system and Figure 1.4 shows a DOA

based system.

Secondly, the receivers must be accurately synchronized so that the TOA are estimated under a common time reference. Global Positioning System (GPS) may be used to synchronize the receivers. However, in practice, this might be difficult in urban and indoor environments due to their limited field of view. Alternatively, clock distribution networks or calibrated atomic clocks [14] may be used but at the expense of higher hardware complexity and system cost. In a DOA-based localization, locating the transmitter can be done in a non-cooperative environment and no synchronization is required. In addition to the advantages above, the DOA-based approach inherits the same advantages offered by UWB IR as the enabling-technology for localization system.

And lastly, taking into consideration the inaccuracy of the estimator, the TOA estimator may not be able to give the transmitter's location due to the ambiguity from the triangulation process. This case can be illustrated in Figure 1.5. In the presence of the DOA estimation error, locating the transmitter is unambiguous. The two lines drawn using the erroneous DOA estimates will definitely intersect at one point as shown in Figure 1.6. Another thing to note is that the smaller the tracking area in which the system is deployed, the higher the positioning accuracy given a fixed DOA estimation error.

The question raised now is whether UWB IR direction finding (DF) can be designed as a low complexity solution. This is important because the original purpose of UWB IR technology is to allow for a significantly simpler receiving structure as compared to a receiver for a sinusoidal-wave communication. It is only if the DF structure can be made as simple as possible, then the increasing cost of using an antenna array at the receiver is justifiable and it does not defeat the original purpose of UWB IR.

It is also important to note that the localization system is targeted for deployment

Figure 1.3: An illustration of a time-of-arrival (TOA) based localization system. At least three receivers that provide good TOA estimates are required to successfully locate the transmitter.

Figure 1.4: An illustration of a direction-of-arrival (DOA) based localization system. Only two DOA estimates from two arrays of receivers are required to successfully locate the transmitter.

Figure 1.5: An illustration of a time-of-arrival (TOA) based localization system when the TOA estimates are erroneous. The three arcs do not intersect, resulting in ambiguity of the transmitter's location. The ambiguity occurs when two out of three receivers report erroneous TOA estimates.

Figure 1.6: An illustration of a direction-of-arrival (DOA) based localization system when the DOA estimates are erroneous. Even if both receivers report erroneous estimates, the intersection of the lines will still be observed.

in a large indoor environment. Some of the possible applications are for locating the equipments in a hospital and locating the cargos in a large warehouse or a cargo area in an airport. The typical area of a hospital is approximately 15 by 20 meters. However, the typical area of such a large warehouse may be much larger.

The deployment area of the antenna array is critical for the far-field assumption to be valid. If the array is placed in the near-field of the propagating signal, the DOA estimation is no longer applicable for locating the target. For the purpose of illustration, the deployment scenario of the antenna array considered in this thesis, is shown in Figure 1.7. Based on this deployment scenario, the distance between the transmitter and receiver is approximately within 8 to 17 meters. Therefore, in this thesis, the channel model considered is an indoor dense multipath channel model with the distance between the transmitter and receiver assumed to be within 8 to 17 meters.

Figure 1.7: An illustration of the deployment area of the DOA based localization system.

1.2 Objectives

The problem of UWB IR DOA estimation is considered in this thesis. A number of delay-and-sum beamforming techniques has been proposed to estimate the DOA of a UWB IR signal in a multipath environment [15–18]. In [19], a weighted subspace wideband DF algorithm is proposed to estimate the DOA of the UWB IR signal. These proposed methods necessitate the receiver to sample the received signal at Nyquist's rate. Because of the large spectral support of the UWB IR signal, the receiver will need to sample at very high rates and handle a huge amount of data. For example, a UWB signal with a bandwidth of 4 GHz will require the analog-to-digital converter (ADC) to sample at 8 Giga-sample per second. This is a major drawback and may not be feasible where the localization systems are subjected to cost and size constraints [13].

A DF system can also be realized by using an array of TOA estimators [20, 21]. UWB IR TOA estimation techniques have been intensively discussed in the literature. Optimum estimator, using the matched filter, requires Nyquist rate sampling and a *prior* knowledge of the multipath propagation channel parameters while the suboptimal approaches limit the accuracy of the estimation. This limitation will result in a low angular resolution DF.

Therefore, the solution to the problem of UWB IR DOA estimation have to overcome the above mentioned limitations. To summarize, the challenges that need to be addressed are listed below.

- Detecting the received pulse of a sub-nanosecond-level duration as a low complexity solution requires a unique receiving technique. This is especially so, when the amplitude is very low, as to comply with the FCC emission mask. The low power transmission also causes the received pulse to appear as background noise.

- Acquiring the discrete representation of the UWB signal requires an extremely high sampling rate (typically in terms of GHz) and high dynamic range ADC. The computational resources to process such a huge amount of data, in turn, will be demanding and costly.
- Finding the direction of the first multipath component from the received signal that consists of hundreds of multipath replicas is a difficult problem. This is because the direction of each replica is unique and the directions from the rest of the multipath components are the nuisance parameters to be suppressed.
- Conventional narrowband DF methods are no longer applicable for UWB IR DOA estimation. This is because the conventional narrowband DF methods are modeled based on a continuous sinusoidal transmission. Likewise, wideband DOA estimation methods require a huge amount of array elements to estimate all directions from the multipath. In a typical indoor propagation environment, the number of multipath is more than 300. Such an array system is impossible to realize.
- The presence of other active UWB IR sources transmitting simultaneously introduces additional nuisance parameters. Otherwise, the direction of the other active sources has to be estimated as well. This means that the DF technique has to be extendable to a multiple access environment.

1.3 Major Contribution of the Thesis

The first major contribution of the thesis is a proposed DF technique using digital channelization receiver architecture. It does not require Nyquist rate sampling. The basic idea is to split the UWB spectrum into multiple single-frequency channels, which

can be arbitrarily selected within the UWB spectrum, and then the structure down-converts each channel into a much lower frequency, hence allowing a low sampling rate ADC to be used. The algorithm processes the output from these ADCs and estimates the DOA. This channelization process is applied to the sum of array received signal. In a multipath environment, a simple analog level threshold detector equipped with a latch circuitry is introduced to detect and localize the leading edge of the received UWB pulse before the summation. The output of the latch is a rectangular pulse, whose rise time indicates the TOA. The channelization is then applied to the sum of these rectangular pulses.

The second major contribution of the thesis is the analog-differentiation based DF approach. This approach is motivated from the observation of a staircase-shape waveform at the sum of the array's rectangular-pulse signal. The staircase-shape waveform is formed when the latch duration is longer than the inter-element propagation delay. Its slope indicates the DOA, theoretically. Thus an analog differentiator with a peak-and-hold detector can be used to estimate the slope.

The third major contribution of the thesis is the TOA-based DF approach. The received signal's TOA information is evidenced by the rise time of the rectangular pulse. A simple integrate-and-sum operation and a sampler can be used to calculate the TOA. Because of the presence of outliers in the time delay estimates, a fractional norm DF formulation is considered. The formulation serves to *de-emphasize* the impact of the large time delay estimation errors on the accuracy of the DOA estimation. In addition to the fractional norm metric, this technique also incorporates an error bound on the time delay estimation to ensure the reliability of the DOA estimator. Unlike previous approaches, no perfect detection is assumed prior to the estimation. Hence, this approach addresses both detection and estimation.

The last major contribution of the thesis is the extension of the DF approaches to

a time hopping multiple access environment. Multiple transmissions from active users are received at the antenna array. When one of the active users is regarded as the user-of-interest, the objective is to estimate the DOA of the user-of-interest in the presence of other users' transmissions. This can be done by utilizing the channelization-based approach, on the condition that the spectrum of the user-of-interest is separable from the rest. This spectral separation property is dependent on the hopping sequence assignment. Thus the hopping sequence has to be properly designed to have the spectral separation property. On the other hand, the problem can be defined as finding all active users' DOA. In this case, a new DF formulation based on a signal subspace transformation is proposed. Complexity-wise, this approach requires a larger channelization structure.

1.4 Organization of the Thesis

The rest of the thesis is organized as follows. In Chapter 2, the DF receiver using digital channelization receiver architecture is presented. Performance analysis reveals that this approach is not robust enough despite the advantage of feasible implementation.

Chapter 3 discusses an alternative means of UWB IR DOA estimation using a simpler structure. This approach is motivated from the observation of a staircase-shape waveform whose slope indicates the DOA. An analog differentiator and a peak-hold detector are utilized to estimate the slope. Although simpler than the approach in Chapter 2, this technique still does not solve the robustness problem.

The technique discussed in Chapter 4 is based on the time delay estimations at each antenna of the array. With a fractional norm formulation, the robustness of the estimated DOA can be achieved. Statistical analysis illustrates that this approach outperforms the previous approaches.

In a time-hopping multiple access environment, the received signal at the antenna originates from multiple transmissions of different sources coming from different directions. This problem is addressed in two ways. First is to estimate a single DOA from the user-of-interest and to regard other active users as interferences. Second is to estimate all active users' DOA. In the first problem, the hopping sequences assignment are required to be spectrally separable. To demonstrate the spectral separation, one example of a design for the hopping sequences is also proposed. The second problem is dealt with by using a subspace-based DF formulation that does not require any prior information on the hopping sequence assignment. The trade-off, however, is the increasing structure's complexity. The method effectively shows multiple peaks at the directions of all active users. Alternatively, the time delay based DF technique from the third contribution can also be used. This is because there is only a single transmission at a single chip duration. The work in the DOA last contribution in this thesis and presented in Chapter 5. Chapter 6 concludes this thesis.

Chapter 2

UWB IR Direction Finding using Digital Channelization Receiver Architecture

2.1 Introduction

UWB radio was originally utilized in military applications, especially in radar related applications. This is due to the fine-range resolution and large bandwidth characteristics of a UWB system. Recently, the focus research on the UWB radio has been diversified to the field of wireless communication and localization.

For narrowband and broadband array processing, DF or DOA estimation has been a major research topic, but this is not the case for UWB. Nevertheless, a number of delay-and-sum beamformers has been proposed to estimate the DOA of UWB IR signal [15, 16]. Although theoretically proven, these proposed approaches require the receiver to sample the received signal at Nyquist rate. Because of the extreme large bandwidth, the receiver will need an extremely high sampling rate. A structure

that utilizes an analog processing approach, as an alternative means of implementing a delay-and-sum beamformer, was proposed in [18]. However, the structure assumes multipath-free propagation and it requires a high resolution (sub-nanosecond) variable delay circuit. Therefore it may be difficult to be implemented.

Unlike narrowband signals, UWB signal covers a wide range of spectrum that causes many narrowband array processing methods to be ineffective. Moreover, the approach of broadband array processing faces difficulties in such a wide spectrum. Besides a high computational load, the practicality of sampling the signal with such spectrum can be questioned. Therefore, there is a need to develop a new approach for DOA estimation of UWB IR signal.

In this chapter, a novel DF technique for UWB IR using channelized digital receiver architecture is proposed. The basic idea of the proposed system is to split the array output into multiple frequency channels, which can be arbitrarily selected within the UWB spectrum, and then down-convert each channel to a much lower frequency for processing, hence allowing a low sampling rate ADC to be used. The algorithm processes the output from these ADCs and estimates the DOA.

The idea of channelized digital receiver architecture has been documented earlier in [22], which focuses on synthesizing the transmitted UWB signal using a single antenna. The architecture consists of multiple band pass filters (BPF) utilized to extract different sub-bands of the whole UWB spectrum. Because the sub-bands can be down-converted and they are narrower as compared to the whole UWB spectrum, a below-Nyquist rate ADC can be used for sampling. Note that to synthesize the UWB signal, the architecture has to be designed such that the combined sub-bands are required to cover the whole UWB spectrum. In addition, the BPF for each sub-band is required to have a sharp roll-off frequency response. Unlike that for synthesizing, the requirements of the BPF design for our proposed structure are greatly relaxed:

the combined sub-bands is not required to cover the whole UWB spectrum and the BPF's frequency response is not required to have a sharp roll-off.

In a multipath propagation environment, the received signal comprises of multiple impinging pulses from different directions. Resolving each DOA of the multipath components is difficult because the number of directions is of the order of hundreds. Since the objective is to estimate the impinging direction of only the first multipath component, the rest of the multipath components can be suppressed.

Because the first multipath component arrives first and the rest later, a simple analog level threshold detector (LTD) and a latch circuitry can be used to detect the leading edge of the direct path. When the leading edge exceeds a predetermined level, the latch is triggered and a rectangular pulse will be generated. The instant when the latch is triggered is then regarded as the estimated TOA of the first multipath component, in a form of the rising time of the rectangular pulse. This process is then introduced as a preprocessing step at each antenna.

To ensure that the DOA information of the direct path is not lost during the proposed preprocessing, the wavefront of the direct path is required to completely propagate across the array aperture (and its leading edge is detected by the thresholding) before the arrival of any subsequent multipath component. Unless the above propagation conditions are met, the proposed preprocessing will not be effective. Hence, this requires an investigation on the temporal and spatial information of a realistic UWB multipath channel model.

There has been many reported realistic UWB channel models in the literature. The work in [23] presents a detailed model on both the temporal and spatial characteristics of a realistic propagation medium. It is observed that some correlation exist between the temporal and spatial information. Further investigations suggest that large signals tend to arrive close to the direct path and the angular spread increases as a function

of the delay between the direct path and the multipath component. This means that the DOA of the direct path and the subsequent multipath are very close in value. This small deviation in the DOA will result in a deviation in the time required to propagate across the array aperture. If the deviation in the time required to propagate across the array aperture is smaller than the inter-arrival time between the direct path and the subsequent multipath, the wavefront of the direct path will propagate across the array aperture completely before the arrival of any subsequent multipath component. Thus, the propagation conditions for the proposed thresholding can be met.

The rest of this chapter is organized as follows. Systems model and preliminaries are presented next, followed by the multipath channel model and its spatio-temporal extension. A review of existing wideband DOA estimations, termed Coherent Subspace Method (CSM) and Weighted CSM (WCSM) , is given in Section 2.4. The approach for DF based on the sum of the received signals at the array is explained in Section 2.5. Section 2.6 describes the digital channelization structure that extracts the line spectra of the sum of the received signals at the array without Nyquist rate sampling. In a multipath propagation environment, the idea of using a level threshold detector is explained in Section 2.7. Numerical simulations of the system are discussed in Section 2.8 and Section 2.9. Finally, Section 2.10 concludes this chapter.

2.2 Systems Model and Preliminaries

Let the received UWB signal in a multipath channel be

$$r(t) = \sum_{j=-\infty}^{\infty} w_{mp}(t - jT_f - \tau_{toa}) + \eta(t) \quad (2.1)$$

where $w_{mp}(t)$ denotes the one-period received waveform capturing the effects of the multipath channel and is given by

$$w_{mp}(t) = \sqrt{E_s} \sum_{l=1}^L a_l w_l(t - \tau_l) \quad (2.2)$$

where E_s is the pulse energy, w_l is the pulse waveform of the l -th multipath component with unit energy and (a_l, τ_l) is the multipath fading coefficients and delays. Note that the multipath delays have been ordered such that $\tau_1 < \tau_2 < \dots < \tau_L$. Also, the first-arriving multipath delay is nil ($\tau_1 = 0$) because the delay due to the first-arriving multipath delay has been accounted for in the term τ_{toa} , the TOA of the received signal. The frame index and interval are denoted by j and T_f respectively. $\eta(t)$ is zero-mean additive white gaussian noise (AWGN) with power spectral density $N_o/2$ and $N_o/2$ also denotes the variance of the noise. Without loss of generality, the multiple access signalling and data modulation in the transmission are omitted. A multiple access signal model will be considered in Chapter 5.

The frame duration T_f is assumed to be sufficiently large such that the last-arriving multipath component does not overlap with the first-arriving multipath component of the subsequent pulse. That is,

$$T_f > \tau_L + \tau_{toa} \quad (2.3)$$

Each of the multipath component carries unique directional information denoted as $\{\theta_l\}_{l=1}^L$, measured with respect to the array axis (see Figure 2.1). If the propagation environment is an ideal multipath-free propagation, the array perceives only one impinging direction as shown in Figure 2.2.

To capture the directional information, an array of UWB sensors is used at the receiving-end. Let (x_n, y_n) denote the position of n -th sensor in the two-dimensional space. The differential delay due to the l -th multipath component can be expressed

Figure 2.1: An illustration of the multipath propagation environment. The transmitter is an far-field transmitter and the receiver is an array. Notice the presence of L multipath's DOAs.

Figure 2.2: An illustration of the multipath-free propagation environment. The transmitter is an far-field transmitter and the receiver is an array. The array observes a single DOA from the direct path.

as

$$\tau_{l,n} = \frac{x_n \sin(\theta_l) + y_n \cos(\theta_l)}{c} \quad (2.4)$$

where c is the speed of light. A simple illustration of the differential delay is given in Figure 2.3 assuming a uniform linear array (ULA) receiver.

The received signal at the n -th sensor is given by

$$r_n(t) = \sum_{j=-\infty}^{\infty} w_{mp}^{(n)}(t - jT_f - \tau_{toa}) + \eta(t) \quad (2.5)$$

where $w_{mp}^{(n)}(t)$ denotes the one-period received waveform at the n -th sensor, which is given by

$$w_{mp}^{(n)}(t) = \sqrt{E_s} \sum_{l=1}^L a_l w_l(t - \tau_l - \tau_{l,n}) \quad (2.6)$$

Observe that because each of the multipath component carries a unique directional

Figure 2.3: An illustration of the inter-element propagation delay due to propagating pulse from θ_l at the ULA receiver, where the inter-element spacing is denoted as d .

information, the received signal at n -th sensor is not the same as the rest of the sensors.

2.3 UWB Multipath Channel Model and Its Spatio-Temporal Extension

The indoor UWB multipath channel for a single antenna receiver has been examined intensively through a propagation measurement campaign and the model derived from the campaign, known as cluster-ray model, has also been reported in the literature [24–27]. Thus IEEE802.15.SG3a proposes to use this model as a common model to evaluate the performance of UWB IR systems in an indoor multipath propagation environment. In all the discussions presented in this thesis, this model is considered.

For cluster-ray model, each multipath component is grouped into different clusters. The multipath components that belongs to the same cluster, termed rays, arrive close to each other. Mathematically, the model can be expressed as a discrete time impulse

response (see [1] for its detailed expression). The impulse response is defined by the amplitude of each rays and the clusters' and rays' arrival times. The cluster's arrival is measured with respect to the arrival time of the direct path while the ray's arrival is measured with respect to the cluster it belongs to. To realize such a model, the simulation uses the Matlab code given in [1].

As for the arrival angle of the clusters and rays, neither [1] nor [26] proposes any model to follow. However, from the literature, [23] proposes the channel model for the angle of arrival based on the measurement campaign conducted in an office building. In [23], Cramer reports that the observation on the arrival angles reveals that the arrival angles and times are not totally independent. Taking the amplitude of each signal's arrival into consideration, Cramer shows that large signals tend to arrive at small deviations from the direct path arrival angle.

From the observations, Cramer proposes a channel model that includes the spatial information besides the temporal information, which agrees with the model proposed in [1]. Similar to the arrival time, the arrival angle is also grouped into different clusters. Each cluster's arrival angle follows the uniform distribution while the ray's arrival angle follows the Laplacian distribution centered at the cluster's arrival angle. For example, in the realization, the cluster's arrival angle is first generated, followed by the rays' arrival angle. For the purpose of illustration, if one of the cluster's arrival angle is 30° , its rays' arrival angle are Laplacian distributed random variable centered at 30° . To relate the amplitude and the angle of the signal's arrival, the generated arrival angles are sorted according to the amplitude of the signals: the larger the amplitude, the smaller the deviation with the direct path arrival angle.

2.4 Existing Wideband DOA Estimation – Coherent Subspace Method (CSM)

Here, the CSM proposed for wideband signal DOA estimation [28] is discussed and used as a benchmark for performance comparison analysis. Let L be the number of signals received at the antenna array. Note that this notation is the same as the notation used to represent the number of the received multipath components in Section 2.2. The one-period received signals at the n -th sensor, where $n = \{0, 1, \dots, N - 1\}$, can be simplified and expressed as

$$r_n(t) = \sum_{l=1}^L w_l(t - \tau_{l,n}) + \eta(t, n) \quad (2.7)$$

where $w_l(t)$ denotes the l -th signal waveform received and $\tau_{l,n}$ is the differential delay due to the propagation of $w_l(t)$ from θ_l .

Sampling the signal at Nyquist rate and applying the discrete Fourier transform (DFT) to the sampled signal, (2.7) becomes

$$R_n(\omega_k) = \sum_{l=1}^L W_l(\omega_k) \exp(-i\omega_k \tau_{l,n}) + \aleph_n(\omega_k) \quad (2.8)$$

where i is the imaginary unit ($i = \sqrt{-1}$). For N received signals, the above equation can be expressed in a matrix form

$$\mathbf{r}(\omega_k) = \mathbf{A}_f(\omega_k, \boldsymbol{\theta}) \mathbf{w}(\omega_k) + \mathbf{n}(\omega_k) \quad (2.9)$$

where the matrices are defined as follows

$$\mathbf{r}(\omega_k) = \begin{bmatrix} R_1(\omega_k) & \cdots & R_N(\omega_k) \end{bmatrix}^T \quad (2.10)$$

$$\mathbf{A}_f(\omega_k, \boldsymbol{\theta}) = \begin{bmatrix} \mathbf{a}(\omega_k, \theta_1) & \cdots & \mathbf{a}(\omega_k, \theta_L) \end{bmatrix} \quad (2.11)$$

$$\mathbf{a}(\omega_k, \theta_l) = \begin{bmatrix} \exp(-i\omega_k \tau_{l,0}) & \cdots & \exp(-i\omega_k \tau_{l,N-1}) \end{bmatrix}^T \quad (2.12)$$

$$\boldsymbol{\theta} = \begin{bmatrix} \theta_1 & \cdots & \theta_L \end{bmatrix}^T \quad (2.13)$$

$$\mathbf{w}(\omega_k) = \begin{bmatrix} W_1(\omega_k) & \cdots & W_L(\omega_k) \end{bmatrix}^T \quad (2.14)$$

$$\mathbf{n}(\omega_k) = \begin{bmatrix} \aleph_1(\omega_k) & \cdots & \aleph_N(\omega_k) \end{bmatrix}^T \quad (2.15)$$

where $[\cdot]^T$ denotes the matrix transpose operation. Assuming no correlation between the signal and noise, the correlation matrix is expressed as follows

$$\begin{aligned} \mathbf{P}_r(\omega_k) &= E[\mathbf{r}(\omega_k) \mathbf{r}^H(\omega_k)] \\ &= \mathbf{A}_f(\omega_k, \boldsymbol{\theta}) \mathbf{P}_w(\omega_k) \mathbf{A}_f^H(\omega_k, \boldsymbol{\theta}) + \mathbf{P}_n(\omega_k) \end{aligned} \quad (2.16)$$

where $\mathbf{P}_n(\omega_k)$ and $\mathbf{P}_w(\omega_k)$ are the autocorrelation matrix of $\mathbf{n}(\omega_k)$ and $\mathbf{w}(\omega_k)$, respectively, which are defined as

$$\mathbf{P}_w(\omega_k) = E[\mathbf{w}(\omega_k) \mathbf{w}^H(\omega_k)] \quad (2.17)$$

$$\mathbf{P}_n(\omega_k) = E[\mathbf{n}(\omega_k) \mathbf{n}^H(\omega_k)] \quad (2.18)$$

where $E[\cdot]$ symbolizes the expectation operation and the superscript H represents the Hermitian matrix operation. The size of the matrices are $N \times N$ and $L \times L$,

respectively.

Because the UWB signal spans over a wide range of frequencies, the steering vector is then spread over this range of frequencies. Let K denote the number of discrete frequencies of the received UWB signal after Nyquist-rate sampling. Thus, the expression in (2.9) is valid for $k = \{1, 2, \dots, K\}$ and the DOA information can be found in $\mathbf{A}_f(\omega_k, \boldsymbol{\theta})$, $\forall k$. The idea of the CSM is to focus all the steering vector at $\omega_k, \forall k$ into a reference frequency denoted as ω_o . In other words, the CSM uses a focusing matrix $\mathbf{T}(\omega_k)$ to transform the wideband DOA estimation problem to a narrowband problem. The focusing matrix has to satisfy

$$\mathbf{A}_f(\omega_o, \boldsymbol{\theta}) = \mathbf{T}(\omega_k) \mathbf{A}_f(\omega_k, \boldsymbol{\theta}) \quad \forall k \quad (2.19)$$

One possible solution is

$$\mathbf{T}(\omega_k) = \mathbf{V} \mathbf{U}^H \quad (2.20)$$

$$\mathbf{A}_f(\omega_k, \boldsymbol{\theta}) \mathbf{A}_f(\omega_o, \boldsymbol{\theta}) = \mathbf{U} \boldsymbol{\Sigma} \mathbf{V}^H \quad (2.21)$$

where $\mathbf{U}$ is an $N \times N$ unitary matrix, the matrix $\boldsymbol{\Sigma}$ is an $N \times L$ diagonal matrix with nonnegative diagonal terms and $\mathbf{V}$ is a $L \times L$ unitary matrix. They are the result of the singular value decomposition operation.

Thus the coherent spectral matrix can be formed as follows

$$\bar{\mathbf{P}} = \sum_k \mathbf{T}(\omega_k) \mathbf{P}_r(\omega_k) \mathbf{T}^H(\omega_k) \quad (2.22)$$

The coherent spectral matrix still maintains the model in (2.16). A small modification to the CSM is WCSM as proposed in [19]. The idea is to weight the coherent spectral matrix with the UWB power spectral density. This requires a *prior* knowledge of the

received waveform $w_l(t)$.

The model in (2.16) is very well-known in subspace-based DOA estimation. A common DOA estimation approach is to eigen-decompose the coherent spectral matrix $\bar{\mathbf{P}}$ into L signal subspaces and $N - L$ noise subspaces, followed by scanning the spatial spectrum for the peak. This method is well-known and has been documented in [28]. The scanning follows the formulation

$$\{\hat{\theta}_l; l = 1, \dots, L\} = \arg \max_{\theta_s} \frac{1}{\|\mathbf{a}^H(\omega_o, \theta_s) \mathbf{E}_n\|^2} \quad (2.23)$$

where θ_s symbolizes the scanning direction and $\mathbf{E}_n$ is the matrix formed by $N - L$ noise eigenvectors of the coherent spectral matrix. The correlation matrix $\mathbf{P}_r(\omega_k)$ in (2.16) is estimated as follows

$$\begin{aligned} \mathbf{P}_r(\omega_k) &= E[\mathbf{r}(\omega_k) \mathbf{r}^H(\omega_k)] \\ &\approx \frac{1}{N_p} \sum_{j=1}^{N_p} \mathbf{r}(j, \omega_k) \mathbf{r}^H(j, \omega_k) \end{aligned} \quad (2.24)$$

where $\mathbf{r}(j, \omega_k)$ is obtained from the observation at the j -th frame and N_p is the number of pulses received (frames observed) for averaging.

2.5 Directional Information Extraction with Frequency Domain Processing

The key idea of the proposed method is to extract the directional-information by using frequency-domain processing. Unlike CSM, the frequency-domain processing is formulated based on the sum of array received signals.

Considered the addition of the received signals from all N array channels. The

one-period expression of the sum of the received signals at the array can be written as

$$y(t) = \sum_{n=0}^{N-1} w_1(t - \tau_{toa} - \tau_{1,n}) + \eta(t, n) \quad (2.25)$$

where a multipath-free environment ($L = 1, a_1 = 1$) is assumed in deriving the DF technique.

The one-period expression given in (2.25) is a simplified expression. In a realistic case, $y(t)$ is observed over more than one period, thus it is a time-varying periodic signal and it can be decomposed into many frequency representations, known as Fourier series representations, given by [29]

$$y(t) = \sum_{k=-\infty}^{\infty} Y(\omega_k) \exp(i\omega_k t) \quad (2.26)$$

where $Y(\omega_k)$ represents the continuous Fourier transform (CFT) of $y(t)$, which is given by

$$Y(\omega_k) = W_1(\omega_k) \exp(-i\omega_k \tau_{toa}) \sum_{n=0}^{N-1} \exp(-i\omega_k \tau_{1,n}(\theta_1)) \quad (2.27)$$

where for simplicity, the energy and noise terms are ignored. By taking one of the sensors as a reference, the TOA term can be measured together with the term $W_1(\omega_k)$. Thus, the expression takes the form

$$Y(\omega_k) = R_o(\omega_k) \sum_{n=0}^{N-1} \exp(-i\omega_k \tau_{1,n}(\theta_1)) \quad (2.28)$$

where

$$R_o(\omega_k) = W_1(\omega_k) \exp(-i\omega_k \tau_{toa}) \quad (2.29)$$

Note that previously it was stated that K denotes the number of discrete frequen-

cies ω_k that cover the whole UWB spectrum. Now let ω_{k_m} denote a finite subset of ω_k where $m = \{1, \dots, M\}$ and $M < K$. Also, let the first sensor ($n = 0$) be taken as the reference. For an M-channel output, the output vector takes the following form

$$\mathbf{y} = \text{diag}\{\mathbf{r}_o\} \mathbf{A}(\theta_1) \mathbf{1}_{N \times 1} \quad (2.30)$$

where

$$\mathbf{y} = \begin{bmatrix} Y(\omega_{k_1}) & \cdots & Y(\omega_{k_M}) \end{bmatrix}^T \quad (2.31)$$

$$\mathbf{r}_o = \begin{bmatrix} R_o(\omega_{k_1}) & \cdots & R_o(\omega_{k_M}) \end{bmatrix}^T \quad (2.32)$$

$$\mathbf{A}(\theta_1) = \begin{bmatrix} \exp(-i\omega_{k_1}\tau_{1,0}(\theta_1)) & \cdots & \exp(-i\omega_{k_1}\tau_{1,N-1}(\theta_1)) \\ \vdots & \ddots & \vdots \\ \exp(-i\omega_{k_M}\tau_{1,0}(\theta_1)) & \cdots & \exp(-i\omega_{k_M}\tau_{1,N-1}(\theta_1)) \end{bmatrix} \quad (2.33)$$

where $\text{diag}\{\cdot\}$ denotes a diagonal matrix operation and $\mathbf{1}_{N \times 1}$ is a column vector with all elements equal to 1. Notice that the matrix $\mathbf{A}(\theta_1)$ is not the same as $\mathbf{A}_f(\omega_k, \boldsymbol{\theta})$ defined in Section 2.4.

Observe that matrix

$$\check{\mathbf{A}}(\theta_1) := \text{diag}\{\mathbf{r}_o\} \mathbf{A}(\theta_1) \quad (2.34)$$

is a unique matrix that characterizes the spatial and spectral information of the signal. Let $\mathbf{P}_A(\theta_1)$ denote the projection matrix and $\mathbf{P}_A^\perp(\theta_1)$ denote the orthogonal projection matrix of $\check{\mathbf{A}}(\theta_1)$. That is

$$\mathbf{P}_A(\theta_1) = \check{\mathbf{A}}(\theta_1) (\check{\mathbf{A}}(\theta_1) \check{\mathbf{A}}(\theta_1)^H)^{-1} \check{\mathbf{A}}(\theta_1)^H \quad (2.35)$$

and

$$\mathbf{P}_A^\perp(\theta_1) = \mathbf{I} - \mathbf{P}_A(\theta_1) \quad (2.36)$$

where $\mathbf{I}$ denotes an identity matrix. The space spanned by $\mathbf{P}_A(\theta_1)$ is termed the signal subspace and can be interpreted as the space that is uniquely formed by matching the accurate frequency channels and spatial parameters. Likewise, the space spanned by its orthogonal projection is called the noise subspace.

With the above definition, the DOA estimation can be realized by scanning the projection of matrix $\mathbf{y}$ onto the noise subspace that results in a minimum value. This is similar to the optimization problem where the projection onto the noise subspace is its cost function [30]

$$\hat{\theta}_1 = \arg \min_{\theta_s} \{ \mathbf{y}^H \mathbf{P}_A^\perp(\theta_s) \mathbf{y} \} \quad (2.37)$$

where θ_s is the scanning direction.

Alternatively, the problem of estimating θ_1 given the observed line spectra $Y(\omega_{k_m})$ can be solved by using the least squares method. First, the line-spectra of the sum of the received signals at the array is re-expressed as follows

$$\frac{Y(\omega_{k_m})}{R_o(\omega_{k_m})} = \sum_{n=0}^{N-1} b_n \exp(-i\omega_{k_m} \tau_{1,n}) \quad (2.38)$$

In the above expression, $R_o(\omega_{k_m})$ is known¹, $\exp(-i\omega_{k_m} \tau_{1,n})$ can be calculated if θ_1 is known and b_n is to be estimated. The variable b_n is introduced due to model inaccuracies. Given the observed line spectra $\{Y(\omega_{k_1}), Y(\omega_{k_2}), \dots, Y(\omega_{k_M})\}$, a linear matrix equation can be formed

$$\mathbf{x} = \mathbf{A}(\theta_1) \mathbf{b} \quad (2.39)$$

¹Generally speaking, $R_o(\omega_{k_m})$ is known only when τ_{toa} is known (see (2.29)). However, calibration can be performed to obtain the value of $R_o(\omega_{k_m})$ whereby the UWB signal source is positioned at $\theta_1 = 0^\circ$. Further explanation is given in Section 2.5.

$$\mathbf{x} = \left[\begin{array}{ccc} \frac{Y(\omega_{k_1})}{R_o(\omega_{k_1})} & \cdots & \frac{Y(\omega_{k_M})}{R_o(\omega_{k_M})} \end{array} \right]^T \quad (2.40)$$

$$\mathbf{b} = \left[\begin{array}{ccc} b_0 & \cdots & b_{N-1} \end{array} \right]^T \quad (2.41)$$

The least-squares solution for $\hat{\mathbf{b}}$ is straightforward if the matrix $\mathbf{A}(\theta_1)$ is known, that is

$$\hat{\mathbf{b}} = [\mathbf{A}(\theta_1)\mathbf{A}^H(\theta_1)]^{-1}\mathbf{A}^H(\theta_1)\mathbf{x} \quad (2.42)$$

With the expression of $\hat{\mathbf{b}}$, an expression for the minimum least-squares error can be obtained as

$$J_{min}(\theta_1) = \mathbf{x}^H\mathbf{x} - \mathbf{x}^H\mathbf{A}(\theta_1)[\mathbf{A}(\theta_1)\mathbf{A}^H(\theta_1)]^{-1}\mathbf{A}^H(\theta_1)\mathbf{x} \quad (2.43)$$

If θ_1 is unknown, the minimum least-squares error will be a function of the unknown direction. To solve for θ_1 , the minimum least-squares error can be expressed as a function of θ_s and the estimate of θ_1 is obtained by scanning for the direction that minimizes the error

$$\hat{\theta}_1 = \arg \min_{\theta_s} \{J_{min}(\theta_s)\} \quad (2.44)$$

Note that the cost function expression in (2.43) is positive definite. Thus, the solution of the optimization function is unique and global.

Modeling the problem as a linear system allows us to discover the minimum M required so that the linear system of equations gives a solution. The requirement depends on the number of unknowns in the system. In this case, the unknowns are $\tau_{1,n}$ and b_n . Since $\tau_{1,n}$ is a function of θ_1 , the number of unknowns are $N + 1$, which comprises of $\{b_0, b_1, \cdots, b_{N-1}\}$ and θ_1 . Therefore, the channelization structure designed is required to fulfil the following inequality

$$M \geq N + 1 \quad (2.45)$$

In the case when $M > N + 1$, the linear system is an over-determined system and the least-squares-error-based scanning is able to solve for θ_1 as expressed in (2.37).

2.6 Digital Channelization Receiver Architecture

The aforementioned technique requires Nyquist sampling. It is possible to get around this requirement by using the following proposed receiver architecture. In particular, the means to extract $Y(\omega_{k_m})$ without having to sample $y(t)$ is discussed and a channelization structure is proposed. This structure shares the same motivation as the structure reported in [22].

In general, the process of sub-dividing a wide-band spectrum into multiple sub-bands is known as channelization. In the audio equalizer application, the audio signal can be seen as a wide-band signal and the idea of the channelization is implemented by using a bank of digital BPFs. Each BPF passes the frequency components of the audio signal on a single subband. The audio equalizer then applies different attenuations to different sub-bands to adjust the frequency response of the audio signal. If translated into the case of UWB signal, this approach requires an extremely high sampling rate, as the whole range of spectrum for a typical UWB signal is in the order of 4 GHz. As the idea of having extremely high sampling rate ADC is not possible in practice, a slightly different approach that eases the requirements of the sampling rate is to implement analog band pass filter banks before the sampling. However, the same problem arises when it tries to sample the uppermost sub-bands.

The use of high sampling rate ADC can be avoided by shifting the spectrum of the wide-band signal to a lower frequency. This can be achieved by using an analog complex mixer followed by an analog low pass filter (LPF). With a complex mixer, the shifted wide-band signal will not contain the image of the spectrum. This property

ensures that no spectral overlapping occurs during the shifting and the spectrum of the sum of the received signals at the array is not changed during the mixing. Figure 2.4 illustrates the difference between complex mixing and real mixing in the frequency shifting process.

Figure 2.4: An illustration of frequency shifting process. The top figure shows the complex mixing while the bottom figures shows the real mixing. Real mixing changes the UWB spectrum.

When the channelization idea is applied to the line-spectra ($Y(\omega_{k_m})$) extraction to form the signal model in (2.30), the number of channels required is only $N+1$ (not the whole UWB spectrum) and the sampling rate can be further reduced by shifting the line spectra to lower frequencies. Because of this property, the proposed idea further relaxes the design requirements as compared to using it for signal reconstruction as proposed in [22]. Figure 2.5 illustrates the channelization process for the signal reconstruction and Figure 2.6 illustrates the process for the line-spectra extraction.

To extract the line-spectra $Y(\omega_{k_m})$, the channelization process can be summarized as follows:

Figure 2.5: An illustration of channelization process for signal reconstruction. For signal reconstruction, the whole UWB spectrum has to be included.

Figure 2.6: An illustration of channelization process for line-spectra extraction. The requirements are much relaxed as the objective is not to extract the whole UWB band.

Step 1 - Let f_m be the m -th channel's operating frequency of the channelizers and $f_o = 1/T_f$ be the fundamental frequency, the output of the analog mixer can be calculated from:

$$y_{xm}(t) = y(t) \times \exp(-i2\pi(f_m - f_o)t) \quad (2.46)$$

Note that $y_{xm}(t)$ is an analog complex signal.

Step 2 - Filtering with an analog filter $H(i2\pi f)$ that passes the frequency f_o and attenuates other frequencies

$$H(i2\pi f) = \begin{cases} 1 & f = f_o \\ 0 & \text{otherwise} \end{cases} \quad (2.47)$$

In practice, the above frequency response is not achievable. Instead, some frequencies around f_o are allowed to pass through the filter by having a slow roll-off response. Therefore, either a LPF or BPF can be used to implement the analog filter in a practical system². Let $h(t)$ denotes the filter's impulse response, the output of the filter can be expressed as

$$y_{fm}(t) = y_{xm}(t) \otimes h(t) \quad (2.48)$$

where $\otimes$ denotes the convolution operation. Since $y_{xm}(t)$ is a complex signal, the convolution of real and imaginary part of the signal is done separately.

Step 3 - Nyquist rate sampling ($f_s > 2f_o$) can be easily achieved. In fact, over sampling is also applicable. Let $f_s = 4f_o$, the sample data can be expressed as

$$y_m[n] = y_{fm}\left(\frac{n}{f_s}\right) \quad -\infty < n < \infty \quad (2.49)$$

The resulting samples form a complex discrete signal.

²A detailed discussion on the effect of the deviation of filter's response from the ideal response is given in Section 2.7.3

Step 4 - Taking N_{sh} samples and performing DFT at $N_{sh}/4$

$$Y(\omega_{k_m}) = \frac{1}{N_{sh}} \sum_{n=0}^{N_{sh}-1} y_m[n] \exp\left(-i\frac{\pi}{2}n\right) \quad (2.50)$$

The line-spectra of $R_o(\omega_{k_m})$ is extracted, following the same procedures described above, during a calibration stage whereby the UWB IR transmitter is positioned at $\theta_1 = 0^\circ$ with respect to the array broadside. Hence, $R_o(\omega_{k_m})$ can be obtained to form $\text{diag}\{\mathbf{r}_o\}$ in (2.30). Figure 2.7 shows the proposed structure for $Y(\omega_{k_m})$.

Figure 2.7: (a) Proposed receiver structure for line-spectra extraction of $Y(\omega_{k_m})$. After the ADC, the structure utilizes line-spectrum computation. (b) An illustration of line-spectrum computation of $Y(\omega_{k_m})$ from $y_m[n]$.

2.7 Level Threshold Detection for Multipath Suppression

Earlier, a multipath-free propagation was assumed in the derivation of the proposed DOA estimator. A realistic propagation environment of UWB IR signal is a dense multipath environment [31]. Unless the multipath effects are suppressed, the proposed method in Section 2.6 is expected to fail. To deal with multipath, a low complexity level threshold detection in analog manner is proposed (see [32]). The level threshold detector (LTD) decides the presence of the first multipath component based on the comparison between the amplitude of the received signal and a predetermined threshold level. When the amplitude exceeds γ , a latch will become active and generate a rectangular pulse. The rise time of the rectangular pulse indicates the delay of the first multipath component. Because the rectangular-pulse-generating process will last longer than the multipath delay spread, only one rectangular pulse will be generated by the latch circuit. The multipath delay spread is defined as the time interval between the arrival instant of the first multipath component and the last one. Figure 2.8 shows the structure of the proposed LTD.

Figure 2.8: The level threshold detection structure at one of the array channels. It consists of an analog comparator and a latch circuitry.

The LTD is implemented after the RF front-end at each antenna element that forms the array. Consider the received signal model at the n -th antenna expressed in

(2.5). Let $\check{r}_n(t)$ denote the output of the LTD at n -th antenna

$$\check{r}_n(t) = p(t - \tau_p^{(n)}) \quad (2.51)$$

where $p(t)$ denotes the rectangular pulse and $\tau_p^{(n)}$ denotes its rise time observed at the n -th antenna. The pulse can be modeled as

$$p(t) = \begin{cases} 1 & 0 \leq t \leq T_{latch} \\ 0 & \text{otherwise} \end{cases} \quad (2.52)$$

and T_{latch} denotes the pulse width (also known as the latch duration). In a practical latch circuitry, there is a delay between the instant when the latch is triggered and when the rectangular pulse is generated. This delay can be assumed to be calibrated before the estimation. Therefore it does not affect the estimation.

The rise time of the rectangular pulse at the n -th antenna is the instant when the observed amplitude exceeds the threshold, mathematically expressed as

$$\tau_p^{(n)} = \min\{t | r_n(t) > \gamma\} \quad (2.53)$$

An example of the operation of the proposed LTD is shown in Figure 2.9.

The rise time $\tau_p^{(n)}$ at *any* of the antenna is assumed to be a good estimate of $\{\tau_{toa} + \tau_1 + \tau_{1,n}\}$. If it is not the case, the resulting DOA estimation will not be accurate. To ensure that this assumption is valid, the following conditions have to be met:

- The threshold γ has to be set at the value smaller than the amplitude of the first arrival signal.
- The first arrival signal wavefront has to propagate completely across the array

Figure 2.9: The operation of the proposed LTD. (a) illustrates the received multipath signal and the threshold level (dotted line) and (b) illustrates the output of the latch: a rectangular pulse.

aperture before the arrival of the subsequent multipath signal at any element of the array. Mathematically, this can be expressed as

$$\tau_{1,n} < \tau_l + \tau_{l,n}, \quad \forall n \text{ and } l = \{2, 3, \dots, L\} \quad (2.54)$$

The sum of the received signals at the array is now the sum of the LTD's output signals (see (2.51)) at the array

$$y(t) = \sum_{n=0}^{N-1} \check{r}_n(t) \quad (2.55)$$

Subsequently, the sum of the LTD's output signals at the array is fed to the channelization structure described in Section 2.6 and the DOA of the direct path can be estimated accordingly. Note that this is possible because the channelization-based DF method is proposed without assuming any particular pulse waveform. Figure 2.10 depicts the proposed DF receiver structure.

Figure 2.10: Proposed DF receiver structure that comprises of the level threshold detector (LTD) at each antenna and the channelization structure. The LTD suppresses multipath and generates a multipath-free signal $\check{r}_n(t)$. The channelization structure (from Figure 2.7) extracts the line-spectra of the array output $y(t)$ that contains directional information. The line-spectrum computation is shown in Figure 2.7(b).

2.7.1 On The Use of Tunnel Diode in Level Threshold Detector

When designing the receiver structure to implement the LTD, [32] introduces a non-linear transformation in the structure prior to the thresholding operation (see Figure 2.11). The original motivation of introducing the transformation is to reduce the noise level such that the probability of false alarm due to a noise-exceeding-threshold event is also reduced. However, it will be shown later that the decrease in the probability of false alarm is accompanied by the decrease in the probability of detection. Hence, the probability of detection is unchanged given the required probability of false alarm.

To study the effect of the nonlinear transformation, the characteristics of the non-linear transformation are first described and then the comparison between the probability density function (pdf) before and after the transformation are presented. As a result of the non-linear transformation, the noise-only pdf is shifted away from the signal-plus-noise pdf. However, the probability of detection remains constant. Eventually, it can be concluded that the transformation is an analog means of implementing the LTD.

Figure 2.11: Level Thresholding Detector structure that utilizes a non-linear transformation prior to detection.

The non-linear transformation reported in [32] is due to a non-linear function of the voltage applied to a tunnel diode when observing the output current. This non-

linearity is depicted in Figure 2.12 as explained in [34]. An analytical expression of the non-linear function is formulated in [35] as

$$I(v) = I_p \left\{ e^{\left(\frac{-V_{pp}}{V_t}\right)} \left[e^{\left(\frac{v}{V_t}\right)} - 1 \right] + \frac{v}{V_p} e^{\left(1 - \frac{v}{V_p}\right)} \right\} + I_v e^{(v - V_v)} \quad (2.56)$$

where (V_p, I_p) , (V_v, I_v) and V_{pp} correspond to the peak point, valley point and projected peak, respectively (see Figure 2.12). V_t is 26 mV under the room temperature.

Figure 2.12: Analog behavioral model of a typical tunnel diode. The non-linear curve reflects the tunneling effect, whereby increasing the applied voltage results in decreasing the output current.

To help illustrate the non-linear transformation of the tunnel diode, a single realization of the received UWB signal, in a multipath environment before and after the tunnel diode, is generated as shown in Figures 2.13 and 2.14.

Figure 2.13: The received UWB signal before tunnel diode non-linear transformation. The simulation considers multipath environment with $E_s/N_o = 24\text{dB}$.

Figure 2.14: The received UWB signal after the tunnel diode non-linear transformation. The simulation considers multipath environment with $E_s/N_o = 24\text{dB}$. The tunnel diode realization uses the parameters listed in Table 2.1.

Parameter	Value	Unit
Peak voltage V_p	65	mV
Peak current I_p	22	mA
Valey voltage V_v	350	mV
Valey current I_v	3.1	mA
Projected peak voltage V_{pp}	510	mV

Table 2.1: Simulation parameters for tunnel diode.

Notice that the received signal amplitude that falls within the negative resistance region (between 0.1 and 0.3 volt) is nullified. This nullification process occurs when the received signal is applied to the non-linear region (see Figure 2.15).

Figure 2.15: Complete Voltage-current (VI) characteristics curve of tunnel diode that includes both non-linear and linear region.

Next, the changes in the noise-only and signal-plus-noise distribution when subjected to the tunnel diode's transformation are evaluated by comparing the pdf of the received signal before and after tunnel diode transformation. At any instant of time, the received signal can be modeled as a normally distributed random variable

with its mean defined by the amplitude of the received signal at that instant of time, and its standard deviation given by the standard deviation of the noise. When the amplitude of the received signal is zero (during the silent period), the random variable is a zero-mean normally-distributed random variable. The pdfs of the random variable for both cases are plotted in Figure 2.16. The corresponding pdfs after tunnel diode transformation are then plotted in Figure 2.17. It can be seen that when the threshold level is inside the non-linear region, the false alarm rate decreases. However, the detection rate decreases as well.

From this observation, it can be concluded that the reduction in false alarm rate is *counterbalanced* with the decrease in detection rate and the non-linear transformation is an analog means of implementing a level threshold detector because the two pdfs are shifted away from each other, so that the threshold level can be set anywhere inside the non-linear region in order to maintain the false alarm rate value.

Figure 2.16: The probability density function (pdf) of signal-plus-noise and noise-only random variables prior to tunnel diode transformation. The amplitude of the received signal before the tunnel diode nonlinear transformation follows this simple pdf.

Figure 2.17: The probability density function (pdf) of signal-plus-noise and noise-only random variables after tunnel diode transformation. The threshold is set within the nonlinear region to reduce the noise floor. However, this reduction is counterbalanced by the decrease in the detection rate.

2.7.2 On The Use of Envelope Detector In The Presence of a Dominant Jammer

A dominant jammer is defined here as a narrowband signal received together with the UWB signal whose spectrum is much larger in magnitude than the UWB spectrum (see Figure 2.18). In the presence of such a jammer, the use of LTD is expected to fail since the UWB pulses is buried within the narrowband signal. To illustrate this, a multi-tone representation of a narrowband signal is simulated and added to the received UWB pulse together with AWGN. Figure 2.19 illustrates this case. Because the pulse amplitude level is far below the jammer, it is not possible to detect it with a fixed threshold.

Figure 2.18: An illustration of the received UWB spectrum with and without the presence of a dominant jammer in AWGN channel. The spectrum of the dominant jammer is clearly much larger in magnitude than the UWB spectrum.

To address this issue, an envelope detector is proposed to provide an adaptive threshold to the comparator. It can also be seen as a rectifier that is subsequently low

Figure 2.19: An illustration of the received UWB signal that is jammed by high power narrowband interference. A multitone representation is used to generate the jammer, which is located in the GSM900 band (890-915 MHz). 10 tones are generated, each having a signal bandwidth of 200 kHz, occupying different bandwidth out of 125 possible bandwidths. The first and last tones are set at the first and last band respectively, while the rest are set in-between randomly. Each tone carries different random phase but equal amplitude. The level threshold detector fails to detect the presence of the pulse with a fixed threshold.

Figure 2.20: An illustration of the received UWB signal that is jammed by high power narrowband interference. The level threshold detector is trying to detect the presence of the pulse with the envelope of the jammer's signal. This shows successful detection.

pass filtered. However, the low pass filter has to be designed such that its response is slow enough in order not to follow the level of the received UWB pulse, but fast enough to follow the envelope of a narrowband signal. An illustration of how the adaptive threshold can be used to detect the weak pulse in the presence of strong jammer is given in Figure 2.20. By incorporating the envelope detector, the structure of the LTD is modified and changed to the structure given in Figure 2.21.

Figure 2.21: Level Threshold Detector structure that utilizes an envelope detector as the threshold to resolve the problem of pulse detection in the presence of a dominant jammer.

2.8 Numerical Analysis for Multipath-Free Channel

In the following simulations, a 2nd derivative Gaussian pulse is considered to be the received waveform. The mathematical expression of Gaussian 2nd derivative pulse can be found in [36] and is given by

$$w(t) = \frac{1}{\sqrt{2\pi}\sigma^3} \left(1 - \frac{t^2}{\sigma^2}\right) \exp\left(-\frac{t^2}{2\sigma^2}\right) \quad (2.57)$$

The parameter σ is defined such that the duration³ of the Gaussian pulse, $T_p \approx 1$ ns. As suggested in [36], σ is set such that 99.99% of the energy of the pulse is contained within the T_p duration. Therefore, $\sigma = 0.19$ ns in this simulation. Note that the proposed DF method is derived without assuming any particular waveform. Thus, other pulse waveforms can also be considered.

Unless otherwise stated, the receiver is a 7-element ULA with inter-element spacing $d = 50$ cm and the emitting source is from a direction of 56° . The source transmits a single pulse for every $T_f = 20$ ns interval. Only one transmitted pulse is received at the array (one frame duration is observed). The simulator generates a discrete signal with a very short sampling interval of 0.16 picosecond to represent the analog received signal. An example of the received UWB signal at the antenna array is shown in Figure 2.22.

The case of AWGN channel is first simulated, then followed by the case of multipath channel in the next section. In the AWGN case, the effectiveness of the proposed digital channelization DF receiver is investigated and compared with the WCSM. The array system does not include the LTD in a multipath-free propagation. The sum of the received signal at the array is fed to the channelization.

The frequency f_m in the channelization structure is arbitrarily chosen. As many as 15 frequencies are used for the channelization structure. These frequencies are chosen within the 1 to 4 GHz frequency range because the received power spectral density (PSD) is highest at this range. In order to speed-up the simulation, the line spectra of the sum received signal is extracted from the Fourier transform of the sum received signal. A more realistic simulation will be presented in Section 2.8.2, where the line spectra of the sum received signal is extracted from the output of the channelization

³Even though the duration of the Gaussian pulse and all of its derivatives is infinite, the simulation truncates the pulse so that the duration is T_p .

Figure 2.22: A realization of received UWB pulse at different antennae that forms the array for AWGN case.

architecture. The output is simulated by convolutions of the input signal and the impulse response of each analog component in the structure.

The estimated DOA is obtained by finding the scanning direction that maximizes the DF function defined as follows

$$DF(\theta_s) = \{\mathbf{y}^H \mathbf{P}_A^\perp(\theta_s) \mathbf{y}\}^{-1} \quad (2.58)$$

where $\mathbf{P}_A^\perp(\theta)$ has been defined in (2.36). To form this function, both the extracted line spectra and the knowledge of the array geometry are required, as described in Section 2.4.

2.8.1 Effect of Inter-Element Spacing

The effect of inter-element spacing in the DF function plot is discussed here. In narrowband array processing, increasing the inter-element spacing (stretching the array geometry) means increasing the resolution at the cost of increased array size. However, the stretching factor could only be increased up to a certain limit before grating lobes occurred. In the following simulations, no grating lobe is observed at the DF function when the inter-element spacing is increased. In fact, the wider the spacing the smaller the width of the main lobe that indicates the estimated DOA. This usually translates into better resolution. Figure 2.23 shows the DF function of various inter-element spacing d . The corresponding zoomed plots are shown in Figure 2.24.

Figure 2.23: Normalized DF function plot as a function of scanning direction under different inter-element spacing d . The simulation considers the emission of 5 pulses in AWGN environment with $E_s/N_o = 21\text{dB}$. The proposed system uses 7-element ULA receiver followed by the channelization structure with 15 channels.

Figure 2.24: The corresponding zoomed plots of Figure 2.23. As inter-element spacing d increases, the width of the main lobe reduces. The simulation considers the emission of 5 pulses in AWGN environment with $E_s/N_o = 21\text{dB}$. The proposed system uses 7-element ULA receiver followed by the channelization structure with 15 channels.

2.8.2 Effect of Channelization Structure

Recall that the DF function is formulated from the line spectra of $y(t)$, denoted as $Y(\omega_k)$. As a means of extracting $Y(\omega_k)$ without an extremely high rate ADC, the proposed channelization structure is utilized. With the aid of numerical simulations, it is shown that the proposed structure can be utilized to obtain the same DF function, as when the extreme rate ADC is used. When using the extreme rate ADC, $Y(\omega_{k_m})$ is extracted from the Fourier coefficients of the sampled $y(t)$ using a DFT operation. But if the proposed structure is used, $y(t)$ is passed through the structure and then sampled using low rate ADCs.

The proposed structure consists of the analog complex mixer, Bessel BPF, and sampler. The impulse response of the Bessel BPF is simulated using a 5th order analog filter with center frequency $f_o = 50\text{MHz}$ and 10 MHz bandwidth. Figure 2.25 shows the frequency response of the analog filter. The ADC samples the output of the filter at $4f_o$ sampling rate. A 15-channel structure with arbitrary frequencies chosen within the UWB spectrum is utilized.

Figure 2.25: Bode diagram of the designed analog Bessel band-pass filter with center frequency at $f_o = 50\text{MHz}$ and bandwidth 10 MHz.

The channelization process is illustrated in Figure 2.26 and Figure 2.27 which show the time domain and frequency domain plots, respectively. The plots shown are the sum of the received signal at the array, the real part of the mixer's output and the BPF's output.

The DF function calculated from the simulation is compared, between the first case when extremely high sampling rate is used and the second case when the channelization structure is used to extract the line spectra. The results are shown in Figure 2.28. The figure implies that the simulated channelizer performs as good as when an extremely high sampling rate is used.

2.8.3 Effect of Number of Channels

In a low signal-to-noise ratio (SNR) environment, the channelization-based DF system performs poorly. This can be observed from the result of the simulation shown in Figure 2.29(a). It shows several undesired peaks apart from the peak that indicates the true DOA. Increasing the number of channels at the channelizer can solve this problem at the cost of increasing the system's complexity as shown in Figure 2.29(b). It shows that those peaks are flatter hence the ambiguity can be removed. Eventually, there will be only single peak in the DF function plot.

Figure 2.30 plots the estimation bias and standard deviation curve as a function of the number of channels. In this simulation, E_s/N_o is fixed at 15 dB and the rest of the parameters are unchanged. As shown by both curves, the values decreases as the number of channels increases. Depending on what is an acceptable values for both the estimation bias and standard deviation, the channelization structure can be set accordingly.

Figure 2.26: A realization of the channelization process at one frequency channel looking at the time domain. (a) the array sum $y(t)$, (b) the complex mixer's output (the real part only) and (c) the filter's output (the real part only).

Figure 2.27: A realization of the channelization process at one frequency channel looking at the frequency domain. (a) the spectrum of $y(t)$, (b) the complex mixer's output and (c) the filter's output.

Figure 2.28: Normalized DF function plotted as a function of direction in AWGN environment. The estimated DOA is shown clearly as the peak. (a) is generated by using the proposed channelizer structure while (b) is generated by using the extremely high sampling.

Figure 2.29: Normalized DF function plotted as a function of DOA. The simulation considers the emission of 5 UWB pulses in AWGN environment with $E_s/N_o = 15\text{dB}$ where the true DOA is 56° . The channelization system uses 7-element ULA of $d=50$ cm with $M=5$ channels (a) and $M=15$ channels (b).

Figure 2.30: Estimation bias and standard deviation plot as a function of the number of channels used in the channelization structure. The simulation considers the emission of 5 UWB pulses in AWGN environment with $E_s/N_o = 15\text{dB}$. The channelization system uses 7-element ULA of $d=50\text{ cm}$.

2.8.4 Effect of Number of Pulses

The effect of increasing the number of received UWB pulses in a low SNR environment is shown in Figure 2.31. The figure compares the DF function obtained from taking different length of observation. As can be seen, the plot is restored after increasing the number of pulses observed. This leads to lower noise spectrum. Hence, increasing the number of pulses further will result in a better estimation performance.

Figure 2.32 plots the estimation bias and standard deviation curve as a function of the number of pulses received at the array. As shown by both curves, both the estimation bias and standard deviation decrease as the number of pulses received increases.

Figure 2.31: Normalized DF function plotted as a function of DOA. The channelization system uses 7-element ULA of $d=50$ cm with 15-channel channelizer. The simulation considers the emission of (a)50 pulses and (b)10 pulses in AWGN environment with $E_s/N_o = 15$ dB.

Figure 2.32: Estimation bias and standard deviation plot as a function of the number of pulses received at the receiver. The simulation considers the AWGN case with $E_s/N_o = 15$ dB, 7-element ULA of $d=50$ cm receiver array and 15-channel channelizer.

2.8.5 Performance Comparison with Weighted CSM

Here, the performance of the proposed channelization-based DF method is compared with the WCSM across various E_s/N_o levels in an AWGN environment. In the simulation, an extremely high sampling rate of 100 GHz is considered to implement WCSM. $K = 32$ frequencies centered around the maximum PSD are focused to the reference frequency at $\omega_o = 2\pi(100)$ MHz. $N_p = 50$ pulses from a single UWB IR source are received at the antenna array and used to estimate the signal-subspace $\bar{\mathbf{P}}$ (see (2.22)). The channelization-based DF method utilizes $M = 15$ -channel structure and observes only 5 UWB pulses (the observation time is $5T_f = 100\text{ns}$).

Based on the aforementioned considerations, 200 realizations for both methods are simulated and their performances are compared. The performances are evaluated in terms of the estimation bias and standard deviation in Figures 2.33 and 2.34 respectively. Both plots imply that as E_s/N_o increases, the performance of the channelization-based DF converges to that of the WCSM.

2.9 Numerical Analysis For Indoor Multipath Environment

The indoor multipath channel considered is simulated according to the proposed model in [1]. This model is based on the result from the measurement campaign of a dense multipath office/laboratory environment. Although different channel models are proposed for different propagation scenarios, only one model (CM1) is utilized because the essential difference between various models is the number of multipaths observed. Other multipath channel models that share the same basic principle have also been reported in [24–27].

Figure 2.33: Estimation bias plot as a function of E_s/N_o in a multipath-free environment. For low E_s/N_o cases, the weighted coherent subspace method (WCSM) achieves better performance. As E_s/N_o increases, the proposed channelization approach improves its performance significantly. And finally for high E_s/N_o , the channelization performs better than the WCSM.

Figure 2.34: Estimation standard deviation plot as a function of E_s/N_o in a multipath-free environment. For low E_s/N_o cases, the weighted coherent subspace method (WCSM) achieves better performance. As E_s/N_o increases, the channelization approach improves its performance significantly. And finally for high E_s/N_o , the channelization performs better than the WCSM..

To extent the temporal model into spatio-temporal model, the proposed cluster-ray model in [37, 38] is adopted and the model is modified based on the discussion given in Section 2.3. The clusters' DOAs are generated from a uniform distribution while the rays' DOAs follow a Laplacian distribution characterized by its standard deviation parameter and centered at the DOA of the cluster that the ray belongs to. For clusters' DOA, the possible value of DOA (measured with respect to the array axis) is limited within $[0^\circ, 180^\circ]$ since a ULA geometry is considered. In addition, for rays' DOA, they are generated as a random value following Laplacian distribution centered at the DOA of the cluster they belong to with a standard deviation set at 25.5° . Figure 2.35 shows one example of the simulated spatio-temporal channel impulse response (CIR). From the spatio-temporal CIR realization, the received signal at the antenna array can be simulated and is shown in Figure 2.36.

Figure 2.35: An example of the simulated spatio-temporal channel impulse response for indoor non-line-of-sight (NLOS) case. The multipath-delay values are generated following CM3 model proposed in [1]. The clusters' DOAs are uniformly distributed in $[0^\circ, 180^\circ]$ and the rays' DOA are sampled from a Laplacian distribution, centered at the DOA of the cluster that the ray belongs to, with 25.5° standard deviation.

Figure 2.36: A realization of received UWB pulse at (a) 0^{th} , (b) 3^{rd} and (c) 6^{th} antenna that forms the array for multipath case.

2.9.1 Multipath Suppression Using The Level Threshold Detector

In a multipath environment, the LTD at RF front-end is incorporated to suppress the multipath effects as shown in Figure 2.10. The threshold level for the LTD is $\gamma = 0.3$ for all antennae. The detection may introduce errors due to missed detection and inaccuracies of the estimated TOA. In a particular realization of the channelization-based DF, the first and 6th antennae miss the incoming UWB pulse while the rest of the antennae are subjected to some errors in the estimated TOA. The realization here considers a single UWB pulse transmission in multipath environment with $E_s/N_o = 17\text{dB}$. The pulse repetition interval is set at $T_f = 100\text{ns}$. The channelization structure utilizes 10 line spectra.

The channelization-based DF assumes the 2nd antenna to be the array phase center (in this case, the first antenna fails to detect the pulse). The estimated TOA, reflected in the rise time of the rectangular pulse, is normalized with respect to the array phase center. The observed values and true values of $\tau_{1,n}$ are shown in Table 2.2. The DF function based on the observed values is shown in Figure 2.37 regardless of the errors and missed detections. The DOA is still reflected at the peak of the DF function. The errors in the rise time of the rectangular pulses, up to a certain extent, does not affect the result of the estimation.

	Estimated delay [ns]	True delay [ns]	Remarks
$\tau_{1,-1}$	0	-1.443	no detection
$\tau_{1,0}$	0	0	reference
$\tau_{1,1}$	1.375	1.443	detected with error
$\tau_{1,2}$	2.875	2.887	good detection
$\tau_{1,3}$	4.250	4.3301	detected with error
$\tau_{1,4}$	0	5.774	miss detection
$\tau_{1,5}$	7.125	7.217	detected with error

Table 2.2: The observed and true delay estimates of $\tau_{1,n}$ in ns.

Figure 2.37: Normalized DF function plotted as a function of DOA in a multipath environment. Some of the LTDs incorporated at the RF front-end miss the incoming pulse and the rest reports erroneous TOA. In this harsh condition, the DOA is still reflected with the peak of the DF function.

2.9.2 Performance Analysis

Next, 1000 realizations of the system in a multipath environment are generated and the estimation's bias and standard deviation are calculated. The system considers a 7-element ULA with inter-element spacing $d = 50cm$ and the threshold for the LTD is set at $\gamma = 0.3$. The antenna array receives a *single* UWB pulse over $T_f = 100ns$ duration. The channelization structure utilizes a 10-channel structure. The frequencies are arbitrarily chosen within the UWB spectrum. Other simulation parameters follow the settings given in Section 2.8.

Figures 2.38 and Figure 2.39 show the estimation bias and standard deviation plots for various E_s/N_o . Both plots show values above 20° and these suggest that the estimator is statistically ineffective for the multipath case. As a comparison, the

estimation's bias and standard deviation are also shown for the time-delay-based DF approach (see Chapter 4).

To better understand the statistical characteristics of the estimation error, the cumulative distribution function curve is plotted for various E_s/N_o in Figure 2.40. The figure shows that the huge bias and standard deviation results are caused by the presence of outliers within the DOA estimates. The number of outliers increases as E_s/N_o increases.

To quantify the number of good estimates, a confidence level (CL) metric is used. CL is defined as the probability level that the estimation value lies within a predefined interval centered at the true DOA value. Without loss of generality, the CL of the estimator is calculated for interval $[\theta_1 - 5^\circ, \theta_1 + 5^\circ]$. Figure 2.41 depicts the results. The maximum confidence level achieved is approximately 45%. This is far below the acceptable level for a robust estimator.

The observation on Figures 2.40 and 2.41 reveals that the majority of the estimated values from the 1000 realizations is the true DOA. This means that it is possible to improve the estimation performance by collecting more estimation values and taking the mode of those values as the final estimate while keeping the system under consideration to be stationery when collecting these values. For example, if 30 estimation values are collected, the system is required to be stationery for $30T_f$ duration. For $T_f = 100$ ns, the required duration is only 3 microsecond. This is a very practical assumption.

The question now is how many values need to be collected so that the resulting CL is acceptable. To answer this, the following simulation considers collecting 5, 10, 20 and 30 estimation values and their performances are compared with the case when only *one* estimation value is considered. The simulation still considers 1000 realizations for each scenario where E_s/N_o is fixed at 23 dB. The resulting bias, standard deviation

Figure 2.38: Estimation bias plot as a function of E_s/N_o in a multipath environment. In all cases, the bias values are shown to be higher than 20° . This reflects that the estimator is not statistically robust.

Figure 2.39: Estimation standard deviation plot as a function of E_s/N_o in a multipath environment. In all cases, the standard deviation values are shown to be higher than 20° . This reflects that the estimator is not statistically robust.

Figure 2.40: Cumulative distribution function (cdf) plots of the estimation error for various E_s/N_o in multipath environment. The cdf plot at $E_s/N_o = 23\text{dB}$ implies that the errors are present as outliers. As E_s/N_o increases, the number of outliers decreases.

Figure 2.41: Confidence level plot of the estimation for various E_s/N_o in a multipath environment. The confidence level measures the probability of generating error less than $\pm 5^\circ$.

and CL are listed in Table 2.3. When 30 estimation values are used, the CL achieved can be as high as 99%. This result proves that the proposed DF method using digital channelization architecture is effective.

No. of initial estimates	1	5	10	20	30
bias [deg]	27.0710	2.1140	1.1100	0.1350	0.1010
standard deviation [deg]	42.3325	20.7547	12.7311	5.1598	3.6794
CL [%]	46	72	88.5	97.8	99

Table 2.3: The performance of the proposed estimator using the mode as the final estimate.

2.10 Conclusion

In this chapter, the use of digital channelization structure for UWB IR DF that leads to lower sampling rate requirement is explained. The DF approach is formulated based on the channelization of the sum of the received signals at the array. Existing wideband DOA estimation is presented in Section 2.3. The DF method is based on the focusing matrix, widely known as CSM. When applying CSM to pulse-based UWB, the coherent spectral matrix can be weighted with the UWB signal's PSD.

Section 2.5 discusses the digital channelization receiver structure. The structure is used to extract the line spectra of the sum of the received signals at the array. Since the set of line spectra extracted is not required to cover the whole UWB spectrum, the design requirements can be further relaxed. In a realistic multipath propagation environment, a simple analog level threshold detector can be incorporated. The received multipath signal is transformed into a clean rectangular pulse using a comparator-latch operation.

The AWGN numerical analysis in Section 2.7 demonstrates that the approach effectively provides good estimation of impinging UWB signal source, using a feasible receiver architecture with low sampling rate ADC. Comparison with the WCSM shows

that the approach may outperform the WCSM as SNR increases.

Section 2.8 provides the numerical analysis in a multipath case. When only some of the sensors successfully detect the pulse while the rest may miss it, the approach is still able to maintain the accuracy of the estimation. However, statistical analysis suggests that the confidence level of getting accurate estimation (absolute error to be less than 5°) is below 50%. This value is far below the acceptable level. To improve the performance, it is proposed that more estimation values are collected while keeping the system under consideration to be stationery when collecting these values. The final estimate will be the mode of these values. As high as 99% CL can be achieved when 30 estimation values are used.

When only one estimation value is available, the proposed method is non-realizable. This problem will be addressed in Chapter 4. As will be explained in Chapter 4, the estimation becomes non-robust because some of the LTDs detect noise that arrives prior to the UWB pulse and not the direct path signal.

Chapter 3

Analog Differentiation Based Impulse Radio UWB Direction Finding

3.1 Introduction

The receiving technique in the digital domain faces challenges due to the infeasibility of an extremely high sampling rate. On the other hand, an analog-receiving technique requires an exact time delay implementation.

In this chapter, an alternative approach for UWB IR DF using an analog differentiator (see Figure 3.4) is proposed. Unlike the approach explained in Chapter 2, this approach involves only simple analog time-domain processing by using an analog circuitry. Therefore, the receiver's complexity is lower than the previous approach.

The proposed system uses an array of level threshold detectors to suppress the multipath effects. The outputs of all array branches are then summed up to form a staircase-shape waveform, whereby the slope of the waveform indicates the direction

of the impinging UWB signal. In order to detect the slope and produce an estimation of the direction, an analog differentiator together with a peak hold detector is utilized. The proposed system is relatively simple and does not require high sampling rate or a high-speed digital-to-analog converter (DAC).

This chapter is organized as follows. Section 3.2 explains the formation of the staircase-shape-waveform. Simulation results reveal that there is a relationship between the slope of the waveform and the DOA. Thereafter, an estimation technique that is based on this relationship is derived in Section 3.3. Also, the analog differentiator design for slope detection is presented. Section 3.4 illustrates the expected operation of the proposed DF system via numerical simulations. The limitations of the proposed DF system are examined in Section 3.5 and the Monte Carlo simulation results is given in Section 3.6. Section 3.7 concludes the chapter.

3.2 Slope of the Staircase-Shape Waveform

Consider the ULA antenna array and the array output $y(t)$ as sum of the rectangular pulse from all N LTDs expressed in (2.55). If the latch circuitry is designed such that the pulse width T_{latch} (of the rectangular pulse $p(t)$) is large enough, the array output will form a staircase-shape waveform (see Figure 3.5). Then the DOA estimation can be achieved by looking at the slope of the staircase-shape waveform. Motivated by this graphical observation, an alternative of UWB IR DF technique is proposed.

Let the number of elements in the array be $N = 2$, for simplicity, and the first element is placed at the array phase center. The rectangular pulse at the output of the first element appears earlier as compared to the second element by $\tau_{1,1}$ seconds. This delay is termed the inter-element propagation delay. If T_{latch} is set to be equal to $\tau_{1,1}$, the resulting sum of the two rectangular pulses is another rectangular pulse with

$2T_{latch}$ pulse width. When $T_{latch} < \tau_{1,1}$, the resulting sum is two rectangular pulses separated by $\tau_{1,1} - T_{latch}$ interval. Only when $T_{latch} > \tau_{1,1}$, then the resulting sum will form a staircase-shape waveform. Figure 3.1 helps to illustrate the three cases.

Recall that $\tau_{1,1}$ is a function of IR source DOA θ_1

$$\tau_{1,1} = \frac{d}{c} \cos(\theta_1) \quad (3.1)$$

where d denotes the inter-element distance and c is the propagation speed. Therefore, the maximum possible value of $\tau_{1,1}$ is achieved when $\theta_1 = 0^\circ$. Hence, in general, to form a staircase-shape waveform, the pulse width has to satisfy the following inequality

$$T_{latch} > \frac{d}{c}(N - 1) \quad (3.2)$$

As $N \rightarrow \infty$, the resulting staircase-shape waveform can be seen as approximately a pair of upward and downward slopes. Both slopes reflect the source DOA. In fact, the slopes can be expressed as

$$y(t) \approx \begin{cases} \frac{1}{\tau_{1,1}}t, & \text{upward slope} \\ -\frac{1}{\tau_{1,1}}t, & \text{downward slope} \end{cases} \quad (3.3)$$

The upward slope occurs at

$$\tau + \tau_d \leq t \leq \tau + \tau_d + (N - 1)\tau_{1,1} \quad (3.4)$$

and the downward slope at

$$\tau + \tau_d + (N - 1)\tau_{1,1} + T_{latch} \leq t \leq \tau + \tau_d + T_{latch} + 2(N - 1)\tau_{1,1} \quad (3.5)$$

(a) $T_{latch} < \tau_{1,1}$.

(b) $T_{latch} = \tau_{1,1}$.

(c) $T_{latch} > \tau_{1,1}$.

Figure 3.1: An illustration of the sum of two rectangular pulses different settings of pulse width T_{latch} . Figure (a) shows the case when $T_{latch} < \tau_{1,1}$, (b) $T_{latch} = \tau_{1,1}$ and (c) $T_{latch} > \tau_{1,1}$. The staircase-shape waveform is formed only in figure (c).

where $\tau_{1,1}$ is the time delay between one element and the subsequent element.

3.3 DOA Estimation Based on Slope Calculation

In this section, the method for estimating the DOA from the slope of the staircase-shape waveform is described. In general, differentiation is a process to calculate the instantaneous rate-of-change of one measurement with respect to another. In this case, the interest is to calculate the slope of the staircase-shape waveform. From the calculated slope, the inter-element delay $\tau_{1,1}$ can be estimated and then the DOA is estimated based on the relationship given in (3.1) and (3.3). Mathematically, the relationship can be expressed as follows

$$\hat{\theta}_1 = \cos^{-1} \left\{ \frac{c}{d} \tau_{1,1} \right\} \quad (3.6)$$

$$\tau_{1,1} = \left[\frac{d}{dt} y(t) \right]^{-1} \quad (3.7)$$

Generally speaking, a ULA antenna array is unable to distinguish the signal wavefront impinging from the front and back of the array. In addition to such an ambiguity, the DOA estimation based on slope calculation suffers from ambiguity in differentiating the signal wavefront impinging from the left and right side of the array. To be precise, the resulting staircase-shape waveform is the same when the signal is impinging from θ and $(180 - \theta)$, where θ is measured with respect to the array axis.

3.3.1 Analog Differentiator Design

To calculate the slope of the staircase-shape waveform, it is proposed to use an analog differentiator to avoid high frequency sampling. In this subsection, the design of the

analog differentiator is discussed.

An op amp differentiator is chosen for the design. Figure 3.2 shows the circuitry of a general op amp differentiator. For a simple op amp, its transfer function is given by

$$H(i2\pi f) = -\frac{i2\pi f C_i R_o}{(1 + sC_o R_o)(1 + sC_i R_i)} \quad (3.8)$$

Figure 3.2: A typical op amp differentiator circuitry.

To determine the value of the passive components (resistors and capacitors) that are suitable for the design, the operating frequency range of the differentiator has to be defined. The frequency range is chosen in such a way that all the possible wave period of the staircase-shape waveform are within the frequency range. The wave period is in fact the pulse repetition interval (PRI) of the transmitted pulse. Let f_L denote the low frequency limit and f_H denote the high frequency limit of the operating frequency, the choice of the passive components is governed by the following equations¹

$$2\pi f_L = \frac{1}{C_i R_i} \quad (3.9)$$

¹This is true when both poles are assumed to be the dominant factors in the differentiator's frequency response.

$$2\pi f_H = \frac{1}{C_o R_o} \quad (3.10)$$

From the transfer function in (3.8), the impulse response of the differentiator can be derived by re-arranging the transfer function expression to a simplified expression

$$H(i2\pi f) = \frac{\alpha}{s + \frac{1}{C_o R_o}} + \frac{\beta}{s + \frac{1}{C_i R_i}} \quad (3.11)$$

where

$$\alpha = \frac{2\pi f_H^2 R_o}{R_o(f_H - f_L)} \quad (3.12)$$

$$\beta = \frac{2\pi f_L f_H R_o}{R_i(f_H - f_L)} \quad (3.13)$$

From the above transfer function expression, the impulse response can easily be written as

$$h(t) = \alpha e^{-\frac{1}{C_o R_o} t} + \beta e^{-\frac{1}{C_i R_i} t} \quad (3.14)$$

For the purpose of illustration, an impulse response of the analog differentiator design is simulated and shown in Figure 3.3. The parameters used for the design is listed in Table 3.1.

Parameter	Value	Unit
f_L	20	MHz
f_H	100	MHz
R_i	1	k Ω
R_o	4.3	k Ω

Table 3.1: Simulation parameters used to realize the analog op amp differentiator.

3.3.2 Peak Hold Detector

The output of the differentiator is a convolution between the impulse response and the input of the differentiator. Ideally, the peak of the output will indicate the slope

Figure 3.3: Impulse response of the designed op amp differentiator with parameters given in Table 3.1.

of the input. Therefore, the peak varies as the slope of the input changes. In our case, the slope of the staircase-shape waveform is observed within an extremely short duration (on the order of nanoseconds). To be exact, this duration can be calculated and expressed as $(N - 1)\tau_{1,1}$. Therefore, a peak hold detector is implemented to track-and-hold the peak.

Overall, Figure 3.4 shows the proposed UWB IR DF system comprising of an array of level threshold detectors, an analog differentiator and peak hold detector.

3.3.3 Receiver Architecture Comparison with Digital Channelization DF Receiver

Compared with the digital channelization receiver proposed in Chapter 2 (see Figure 2.10), the proposed structure based on the analog differentiation shown in Figure 3.4 requires less components. The channelization based receiver requires complex mixers,

Figure 3.4: Proposed UWB IR direction finding structure using analog differentiator to calculate the slope of the array output.

BPFs and ADCs while the differentiation based receiver requires only a differentiator and a peak hold detector.

To estimate the DOA, the channelization based receiver requires a digital processing unit to perform the complex additions and multiplications. The processing unit is also required to have some memory to store some complex values temporarily. On the other hand, the differentiation based receiver requires only a look-up table in order to associate the direction with its respective peak of the differentiator's output. No computing power is required for this process.

In summary, the differentiation based DF receiver requires less hardware than the channelization based one. Table 3.2 shows the receiver architecture comparison of the two DF receivers. Compared with a DF receiver using conventional Nyquist rate sampling, both approaches are more practical as they do not require an extremely high

sampling rate and tremendous computational resources to process a huge amount of data.

Hardware	Channelization	Differentiation	Quantity Required
Level Threshold Detector	Yes	Yes	N
Mixer	Yes	No	2M
BPF	Yes	No	2M
ADC	Yes	No	2M
Computing Power	Yes	No	Complex + and $\times$
Memory	Yes	No	Complex Values
Differentiator	No	Yes	1
Peak Hold Detector	No	Yes	1
Look-Up Table	No	Yes	1

Table 3.2: Comparison of the receiver's complexity between the channelization based DF receiver and the differentiation based one, where N represents the number of antennae and M represents the number of channels for the channelization architecture.

3.4 A Numerical Illustration

To help illustrate the operational theory of the proposed approach, a numerical realization of the system is utilized. The simulation considers a Gaussian 2nd derivative pulse with $T_p \approx 1\text{ns}$. The PRI is set at $T_f = 100\text{ns}$ and the latch duration is $T_{latch} = 15\text{ns}$. The antenna array is a 7-element array with $d = 30\text{cm}$. To form the staircase-shape waveform, the latch duration T_{latch} has to be greater than $\frac{d}{c}(N - 1) = 6\text{ns}$. Therefore, the latch duration is set to be 15 ns. The propagation environment is a multipath environment with $E_s/N_o = 25\text{dB}$. The threshold γ for the LD is set at 0.3. This realization results in the staircase-shape waveform shown in Figure 3.5.

In practice, the output of the analog differentiation can be modeled as a convolution between the analog differentiator's impulse response $h(t)$ and the staircase-shape-waveform $y(t)$. The differentiator's impulse response is given in (3.14) with the parameters set according to Table 3.1 and it has been shown previously in Figure 3.3.

Figure 3.5: Plots of the staircase-shape waveform for different propagation direction. The slope of the waveform increases as the DOA increases from 0° to 90° .

The simulation results of the differentiator's output is plotted in Figure 3.6.

The peaks indicate the slopes of the outputs as described in (3.3). Before reaching the peak, the output experiences a transition time, which is dependent on the duration of observing the slope of the staircase-shape waveform. Notice that this duration is smaller than 6ns. For different values of θ_1 , the curves reach different peaks' value. The highest slope is obtained when $\theta_1=90^\circ$.

Prior to estimation, the proposed structure has to be calibrated by measuring the differentiator's peaks for a few known directions. The calibration will produce a lookup table that maps the differentiator's peak and its corresponding direction. Given this lookup table, one is able to estimate the unknown DOA from the observed differentiator's peak using interpolation.

(a) The differentiator's output.

(b) Zoomed-in plot of the differentiator's output.

Figure 3.6: The output of the differentiator for different values of θ_1 plotted as a function of time. The peak indicates the slope of the staircase-shape waveform. As θ_1 increases, the slope also increases.

3.5 Limitations of the Differentiation Based DF

Ideally, the slope of the staircase-shape waveform is inversely proportional to the inter-element propagation delay $\tau_{1,1}$. In this section, it is shown that the peak of the differentiator's output does not match exactly with the inverse of the inter-element propagation delay.

Consider the realization of the proposed system for a propagation's direction from 0° to 90° with a 5° interval step-size. The peak of the differentiator's output is recorded and normalized. This peak should be equal to the slope or inversely proportional to the inter-element propagation delay $\tau_{1,1}(\theta_1)$. To compare the result with the ideal case, Figure 3.7 is plotted. The normalization is taken with respect to the highest peak, obtained when the signal propagates from the array broadside (90°). The ideal case (true slope) is plotted as the inverse of $\tau_{1,1}(\theta_1)$. As shown, the peak of the differentiator's output and the true slope are not an exact match. Although it shows a good match for directions below 45° , it starts to deviate when the direction increases further. The reason is due to the incapability of the differentiator to respond to very short-duration changes.

Apart from the above limitations, the proposed approach also suffers from the inaccuracy of the estimated TOA due to the LTD operation. This conclusion is supported by the observations on the cdf plots and the confidence level plots in the next section.

Figure 3.7: Plots of the slope as a function of scanning direction. The comparison shows that the peak of the differentiator starts to deviate from the true slope at higher scanning direction.

3.6 Performance Analysis in a Multipath Environment

Here, the effect of the system's parameters on the performance of the proposed system is investigated via numerical simulations. The effect of increasing the aperture of the array is discussed. Increasing the aperture can be done by either adding more sensors to the array or increasing the inter-element spacing. Both options will be considered and their effects will be investigated. In addition, the performance of the proposed analog differentiator's circuit may vary due to the tolerance of its passive components. This will also be investigated. A Monte Carlo simulation of 1000 realizations is utilized to calculate the estimation's bias and standard deviation. These results are then compared with those obtained using a digital channelization receiver discussed in Chapter 2. Besides the bias and standard deviation, the confidence level of the

proposed approach is also calculated and compared.

The simulations consider the same multipath propagation environment discussed earlier, except that the number of channels of the array and the inter-element spacing are varied. The first simulation considers a fixed inter-element spacing at $d = 50\text{cm}$ and varies the number of sensors N . The DOA is varied from 0° to 90° with a 5° interval step-size. The second simulation fixes the number of channels at $N = 7$ and varies the inter-element spacing d .

Generally, stretching the array geometry means increasing the resolution of the DOA estimation at the cost of an increase in the array aperture size. In this case, the peak of the differentiator's output indicates the slope. Figure 3.8 and Figure 3.9 show the normalized peak levels as a function of the scanning direction for different d and N respectively. The figure shows that increasing the array aperture leads to the larger difference between two peaks.

When the passive components (resistors and capacitors) in the analog differentiator's circuit are subjected to $\pm 5\%$ tolerance, the impulse response of the differentiator will also deviate as well. The deviation in the impulse response will then result in a different differentiator's output and peak value. To see how far the peak of the differentiator's output will deviate, the following simulations consider the case when the passive components are subjected to the tolerance and that of an ideal case. Figure 3.10 shows the resulting normalized peak of the differentiator's output as a function of the scanning direction. From the figure it can be concluded that no significant deviation is observed when the passive components are subjected to $\pm 5\%$ tolerance.

Next, the estimation bias and standard deviation are calculated and compared with the channelization based DF. The array under consideration is an ULA with $N = 7$ and $d = 50\text{cm}$. E_s/N_o is varied from 13 dB to 27 dB. Figures 3.11 and 3.12 show the estimation bias and standard deviation respectively. Although both the bias

Figure 3.8: Normalized peak of the differentiator output as a function of the scanning direction for different inter-element spacing d .

Figure 3.9: Normalized peak of the differentiator output as a function of the scanning direction for different number of array elements N .

Figure 3.10: Normalized peak of the differentiator output as a function of the scanning direction for the case when the passive components in the analog differentiator's circuit are subjected to $\pm 5\%$ tolerance and for an ideal case.

and standard deviation plots show that the differentiator method performs better than the channelization method, the following study proves otherwise when the statistics plot of the estimation error is observed.

Figure 3.13 depicts the cdf plots of the estimation error for different E_s/N_o . It is shown clearly that even when E_s/N_o is high, very few of the estimates are accurate. As a comparison with the channelization-based approach from previous chapter, the confidence level that the estimator reports accurate DOA (its absolute error is below 5°) is calculated. Figure 3.14 shows the confidence level plots. This figure reveals that the channelization based approach still outperforms the differentiation based approach in a high SNR environment (when E_s/N_o is higher than 19 dB).

Notice that this observation is inconsistent with Figures 3.11 and 3.12. This can be explained as follows. When a ULA geometry is considered, the differentiation based approach can only distinguish the direction of the signal wavefront impinging from

Figure 3.11: Estimation bias plots as a function of E_s/N_o . The plots compare the bias between the differentiator approach and channelization based approach. It shows that the bias of the differentiator approach is smaller than the channelization approach.

Figure 3.12: Estimation standard deviation plots as a function of E_s/N_o . The plots compare the standard deviation between the differentiator approach and channelization based approach. It shows that the standard deviation of the differentiator approach is smaller than the channelization approach.

Figure 3.13: Cumulative distribution function (cdf) plots of the estimation error for different E_s/N_o . The cdf plots show the ineffectiveness of the proposed estimation approach.

Figure 3.14: Comparison of confidence level plots of the estimation error between channelization and differentiation based approach for different E_s/N_o . The channelization based approach still outperforms the differentiation approach.

0° up to 90° while the channelization based approach can distinguish directions from 0° up to 180° . This means that the channelization based approach is subjected to a higher estimation error as compared to the differentiation based approach. This fact causes the estimation bias and standard deviation of the channelization based approach to be higher. Hence, a better means of comparing the performance of the differentiator based approach with that of the channelization based approach is by observing the the cdf plots (Figure 3.14).

Similar to the channelization based approach, it is possible to improve the estimation performance by collecting more estimation values and taking the mode of those values as the final estimate while keeping the system under consideration to be stationery when collecting these values. Therefore, the next simulation considers collecting 5, 10, 20 and 30 estimation values where E_s/N_o is fixed at 27 dB. The improvement can be observed from the increase in the confidence level, when more estimation values are used, as listed in Table 3.3.

No. of initial estimates	1	5	10	20	30
bias [deg]	4.5090	10.1730	4.5350	1.2820	0.2860
standard deviation [deg]	25.5328	11.7613	8.5631	5.4000	4.0527
CL [%]	41.8	45.4	65.9	82.4	89.4

Table 3.3: The performance of the proposed estimator using the mode as the final estimate.

Compared with the channelization based approach, the differentiation based approach performs worse under the same propagation condition. In other words, the differentiation based approach requires a higher SNR and more estimates to achieve the same CL that the channelization based approach is able to achieve. For example, the channelization based approach is able to achieve 99% CL when using 30 initial estimates while the differentiation based approach is only capable of achieving 89.4%.

Although it performs worse than the channelization approach, the differentiation based approach provides an alternative means of having a reduced complexity DF

system. With its hybrid digital-analog² structure, almost no computational resource is required to implement such system.

3.7 Conclusion

The DOA estimation approach based on the staircase-shape-waveform observation of the sum of the LTD's output signals is discussed in this chapter. The waveform, observed due to the sum of the rectangular pulses, contains the DOA information at its slope. By calculating the slope of the staircase-shape waveform, the DOA can be estimated (see Section 3.2). To do so, an analog differentiator is considered. The differentiator is realized by using a typical op amp differentiator circuit. Its impulse response can be derived and used in the simulation. The simulation shows that the peak of the analog differentiator output indicates the DOA. The implementation of the DF system requires the lesser components, and hence reduces the system's cost and complexity (Section 3.3).

In Section 3.4, the operation of the proposed structure is illustrated numerically. The staircase-shape waveform and the impulse response of the designed op amp analog differentiator designed are convolved to obtain the output of the differentiator. Different peaks' levels corresponding to different DOA are observed.

The major disadvantage is that the effectiveness of the approach is highly dependent on the accurate detection of the level threshold detectors at each antenna element. Otherwise, the staircase-shape waveform will be distorted and the slope is no longer an indication of the DOA. These are discussed in Section 3.5.

The last section investigates the performance of the proposed approach in a mul-

²Front-end systems (from antenna to level threshold detector) is an analog subsystem. Latch circuit produces digital output in the form of an rectangular pulse. The subsystem after latches is another analog subsystem.

tipath environment. The results imply that the presence of multipath and system's noise will render the level threshold detectors inaccurate. When comparing with the channelization based approach, no improvement is noticed.

Chapter 4

UWB IR Detection and Time Delay Based Direction Finding

4.1 Introduction

Reduced complexity UWB IR DF has been shown to be feasible in previous chapters. Two approaches are considered, both are based on processing the sum of all channels of the array¹. First approach, discussed in Chapter 2, utilizes a channelization structure. The channelization is used to extract the line-spectra information of the sum of all channels of the array. Then the DF function can be formulated from the line-spectra. Prior to the summation, the multipath effect due to signal propagation is suppressed with a simple level threshold detector. The detector transforms the multipath-contained received signal to a *clean* rectangular pulse, whose rise time reflects the propagation delay due to the first arrival pulse.

Second approach, discussed in Chapter 3, utilizes a simple analog differentiator to estimate the DOA. The use of the differentiator is motivated by the observation that

¹The term ‘sum of all channels of the array’ refers to the sum of all the LTDs’ output (each channel of the array implements the LTD).

the slope of a staircase-shape waveform (formed through summation of rectangular pulses) acts as an indicator of the DOA. Thus the analog differentiator is used to obtain the slope. Simulations show that the peak of the differentiator and the inverse of the inter-element propagation delay are a good match.

Both approaches are based on the processing of the sum of all the channels of the array and heavily dependent on the accuracy of the level threshold detector. Results presented in Chapters 2 and 3 imply that the level threshold detector is not reliable in providing a good TOA estimation. Thus, processing the sum of all channels of the array will lead to an unreliable DOA estimation. This drawback refrains both approaches from achieving robust DOA estimation.

An alternative approach for DF is time delay based DF. This approach estimates the DOA from the time delay estimated at each element of the antenna array [20]. However, there are some difficulties in estimating the time delay of the received UWB IR signal and extending it to DOA estimation. Although an optimal time delay estimation can be achieved with a matched filter receiver [39], it requires a *prior* knowledge of the multipath propagation channel parameters. Unfortunately this information is not available in many applications. Alternatively, suboptimal time delay estimators, such as the transmitted reference (TR) or energy detection (ED) approach [40], have also been proposed. These approaches perform an integrate-and-dump operation on the received signal before sampling. Besides reducing the sampling rate, it also reduces the angular resolution at the same time. This is because there exists an inter-dependency between the angular resolution and the sampling rate used in estimating the delay. As a result, the time delay DF approaches that utilize such techniques may not be suitable, especially when the aperture of the antenna array is small.

In this Chapter, a different approach is proposed. The approach achieves much higher angular resolution even when considering a small array aperture. In addition,

the approach overcomes the problem of the unreliable DOA estimation posed in Chapters 2 and 3. In other words, it is able to provide robust DOA estimation when the time delay estimation is not reliable. Rather than processing the sum of all channels of the array, the proposed approach estimates the rise time of the rectangular pulses from each channel, using either the proposed time-domain or frequency-domain processing. Both techniques utilize a low sampling rate, but high dynamic range ADC.

In some instances, the rise time of the rectangular pulse is not a good time delay estimate because the analog LTD operation may not successfully detect the leading edge of the received UWB IR signal. False alarms and missed detections may occur and cause huge estimation errors, especially at low SNR. To address this problem, the DF approach proposed here is based on fractional norm formulation. The fractional norm is used to de-emphasize the impact of the huge time delay estimation errors on the accuracy of the DOA estimation without having to identify which time delay estimates contain huge errors.

The DOA estimate is obtained by scanning for the direction that minimizes the fractional-norm cost function. This estimate is regarded as an initial estimate. Thereafter, an algorithm is proposed to compare the time delay estimates from each channel of the array with the time delay corresponding to the initial DOA estimate. If the difference is less than a preset error bound, the initial DOA estimate is regarded as a final estimate. It shall be illustrated, through numerical analysis, that a careful choice of the threshold level (of the LTD) and the error bound is critical to achieve a good DF performance. The results also show that the proposed approach is robust against multipath and outperform the conventional wideband DF method [19].

The rest of the chapter is organized as follows. In Section 4.2, an UWB IR array is described briefly and some preliminary definitions are discussed. The receiver structure operation is explained in Section 4.3. The structure comprises of the analog LTD,

the latch circuitry and the integrate-and-dump operation. For the frequency-domain processing, the single-frequency subband channelization operation is also presented. Then the proposed DF algorithm is given in Section 4.4. Finally, Section 4.5 provides the numerical analysis of the proposed technique and Section 4.6 concludes the chapter.

4.2 System Model and Preliminaries

This chapter considers the same problem as that discussed in Chapters 2 and 3. The received signal model used follows the same model described in (2.1). In addition to estimating the DOA of the received UWB IR signal, the receiver is required to detect the signal. The detection problem is explained as follows. The UWB pulse is transmitted every T_f duration. If the frame interval T_f is divided into multiple chip duration (assumed to be equal to pulse duration T_p), the number of possible placements is symbolized as

$$N_b := \left\lfloor \frac{T_f}{T_p} \right\rfloor \quad (4.1)$$

where $\lfloor \cdot \rfloor$ denotes the floor function, which returns the largest integer less than or equal to the operand of the function. The detection problem is thus finding, among N_b positions, where the pulse coming from the direct path is present. Only the pulse coming from the direct path or the first arrival pulse is counted as a detection. Figure 4.1 helps to illustrate the problem.

Consider a UWB antenna array receiver. Each element of the array implements the analog LTD. Recall that the output of the LTD is a rectangular pulse and its rise time indicates the time delay estimate. This estimate is needed for the DOA estimation, thus it shall be extracted. The extraction can be realized by using either an integrate-and-dump with a sampler (time-domain processing) or single frequency

Figure 4.1: An illustration of the detection problem. Within T_f interval, the observed signal comprises of noise-only observation E_o and signal-plus-noise observation E_1 . The detection problem is defined as finding among N_b observations that satisfies E_1 .

subband channelization structure (frequency-domain processing). Based on the time delay estimates, the receiver aims to calculate the DOA.

4.3 Time-Delay-Estimates Extraction

To extract the rise time of the rectangular pulse, two techniques are proposed. One utilizes frequency-domain processing while the other utilizes time-domain processing.

4.3.1 Single-Frequency Subband Channelization Structure

The frequency-domain technique utilizes a structure called single-frequency subband channelization (SFSBC). This structure adopts the idea of a channelization receiver. Instead of using a multiple-channel structure, this technique requires only a single-channel structure, but is implemented at each channel of the array (see Figure 4.2). Each structure consists of an analog mixer, a BPF, a low rate ADC and a simple multiply-and-sum to estimate the time delay. The analog mixer is used to shift the

frequency to be channelized f_m to a much lower frequency f_o . Each structure can be seen as an implementation of the line spectrum extraction as discussed in Chapter 2. For N antennae, the overall structure extracts N line spectra. As compared to the channelization-based DF approach that requires $N + 1$ line spectra extraction, this technique reduces the system's complexity.

Figure 4.2: Proposed UWB IR DF structured based on time delay estimation. The estimation is accomplished using SFSBC structure.

As the name implies, SFSBC is interested only in one line spectrum information from a single discrete frequency to calculate the time delay estimate. Figure 4.3 shows the structure of the SFSBC. From the line spectrum extracted, the time delay estimate can be calculated.

The following discussion uses the same notation as that in Chapter 2 to symbolize the output of the latch circuit, i.e. $\check{r}_n(t) = p(t - \tau_p^{(n)})$. Since $\check{r}_n(t)$ is periodic and *Dirichlet conditions* are satisfied [29], there exists a Fourier series representation

$$\check{r}_n(t) = \sum_{k=-\infty}^{\infty} \check{R}_n(\omega_k) \exp(i\omega_k t) \quad (4.2)$$

Figure 4.3: Single-Frequency Subband Channelizer Structure. It comprises of frequency shifting with analog mixing, filtering and sampling. After sampling, a discrete Fourier transform operation involving several sum and multiply is used to calculate the line spectrum.

where $\check{R}_n(\omega_k)$ represents the CFT of $\check{r}_n(t)$, which is given by

$$\check{R}_n(\omega_k) = P(\omega_k) \exp(-i\omega_k \tau_p^{(n)}) \quad (4.3)$$

The process of obtaining the line spectrum $\check{R}_n(\omega_k)$ from $\check{r}_n(t)$ has been explained in details in Section 2.5 and will not be repeated here.

The time delay term $\tau_p^{(n)}$ is calculated by assuming that $P(\omega_k)$ is known a priori. Thus $\tau_p^{(n)}$ can be calculated from the following formula

$$\tau_p^{(n)} = \text{Re} \left\{ \frac{\log(\check{R}_n(\omega_k)) - \log(P(\omega_k))}{-i\omega_k} \right\} \quad (4.4)$$

where $\text{Re}\{x\}$ symbolizes the real part of x .

In order to avoid the ambiguity when calculating $\tau_p^{(n)}$, the choice of the frequency to be channelized f_m has to follow a certain constraint. The constraint depends on the maximum possible value of the time delay term, denoted as $\max(\tau_p^{(n)})$. This can be explained as follows. From sampling theorem, a discrete sampled signal can uniquely represent an analog signal with frequencies less than half the sampling rate, otherwise ambiguity will occur. Likewise, $\tau_p^{(n)}$ is part of the discrete frequency term of the ADC output's signal. With simple manipulation, the constraint for f_m can be defined using the inequality

$$f_m < \frac{1}{4\pi \max(\tau_p^{(n)})} \quad (4.5)$$

The time delay term $\tau_p^{(n)}$ in (4.4) comprises of the inter-element propagation delay $\hat{\tau}_{1,n}$ and the propagation delay τ_{toa} (as described in the signal model explained in Section 2.1). The inter-element delay may vary depending on the error and the propagation delay τ_{toa} may vary depending on the distance between the transmitter and receiver. If the error is assumed to be limited within the frame interval T_f and

the maximum propagation delay can be calculated from the maximum range coverage intended for the receiver, $\max(\tau_p^{(n)})$ can be deduced.

Next, $\tau_p^{(n)}$ is observed at each antenna. To obtain the inter-element delay $\hat{\tau}_{1,n}$, the observed $\tau_p^{(n)}$ can be normalized with respect to the observed $\tau_p^{(0)}$ at the 0-th element as a reference. Note that for SFSBC implementation, the structure can be designed to use ADC with a sampling interval to be as large as the frame duration.

4.3.2 Integrate-and-Dump with Low Rate Sampling

A simpler method for time-delay-estimates extraction is the proposed method based on time-domain processing. This method utilizes a simple integrate-and-dump operation followed by an ADC. No high sampling rate requirement is needed for the ADC. In fact, the sampling rate requirement is only twice the frame rate. Figure 4.4 shows the proposed array structure.

Figure 4.4: Proposed UWB IR DF structure based on time delay estimation. The estimation is accomplished using a simple integrate-and-dump operation, followed by sampler.

It can be observed from Figure 4.4 that the proposed structure for the time-domain

technique is an extension of the basic LTD structure. The extension structure comprises of an integrate-and-dump operator and a low rate sampler. The latch output $\check{r}(t)$ is passed to the integrate-and-dump operation. The output of the integration is then sampled. The sampling interval t_s is equal to the integration interval and they are assumed to be aligned (thus, the term *integrate-and-dump* is used)

$$t_s = T_{latch} = t_b - t_a \quad (4.6)$$

where t_a and t_b are the start and end time of the integration. The rise time of the latch output may fall anywhere within the integration interval $[t_a, t_b]$. It is straight-forward to calculate $\hat{\tau}_{1,n}$ from the output of the integration.

The inter-element delay $\hat{\tau}_{1,n}$ is reflected in the rise time of the rectangular pulse and can be calculated accurately from the output of the low rate sampler. To show how this can be achieved, a few notations are first defined and then the calculation is demonstrated. The discrete signal at the sampler output is given by

$$x_n = \int_{t_a}^{t_b} \check{r}_n(t) dt \quad (4.7)$$

$$x_n = \begin{cases} i_n T_{latch} - \tau_p, & t_a < \tau_p < t_b \\ \tau_p - (i_n - 1) T_{latch} & t_a < T_{latch} + \tau_p < t_b \\ 0 & \text{otherwise} \end{cases} \quad (4.8)$$

$$i_n = \left\lceil \frac{\tau_p}{T_{latch}} \right\rceil \quad (4.9)$$

where $\lceil \cdot \rceil$ represents the ceiling function, which returns the smallest integer not less than the operand of the function. When the latch circuit outputs a rectangular pulse,

x_n is no longer zero. The ideal operation of the proposed structure is illustrated in Figure 4.5. Consider the 0-th antenna to be the array center, $\hat{\tau}_{1,n}$ is calculated as follows

$$\hat{\tau}_{1,n} = (i_n - i_0)T_{latch} + x_n - x_0 \quad (4.10)$$

Unlike others, this approach utilizes a very simple analog front-end operation. Furthermore, the estimation can be realized by extending the basic LTD structure with the simple integrate-and-dump operation with a very low rate ADC. The largest possible sampling interval is half the frame interval. Despite its simplicity, the estimation is highly accurate.

4.4 On the Angular Resolution of TOA Based DF in UWB IR

In most of the discussions in this thesis, a comparator-latch circuit has been used in the LTD structure frequently. In the literature of UWB IR TOA estimation, there are several approaches that do not require Nyquist rate sampling [41]. They include the stored reference (SR), TR and ED approaches. The common feature of these three approaches is that the sampling rate required is of the order of the chip rate. Before sampling, SR correlates the received signal with an internally generated pulse followed by an integrate-and-dump operation, while TR uses the transmitted reference pulse, however, ED simply squares the received signal before the integrate-and-dump operation to collect the energy. Figure 4.6 shows these structures. Other references that report similar structures are [42, 43]. [44–46] and [47, 48] report structures similar to TR and SR, respectively.

The TOA is estimated by finding the index at which the value is maximum.

Figure 4.5: Ideal operation of the proposed receiver structure. Figure (a) shows the received multipath signal $r_n(t)$. The dotted line shows the threshold level γ . Figure (b) shows the latch output $\tilde{r}_n(t)$. Its rising time is the first instant when the received multipath signal exceeds the threshold and it falls in-between the integration duration. Figure (c) shows the output of the sampler x_n . The amplitude is the result of the integrate-and-dump operation and it is a function of the rising time. Because the latch duration T_{latch} is known, the rising time can be calculated.

(a) Stored-Reference structure.

(b) Transmitted-Reference structure.

(c) Energy Detection structure.

Figure 4.6: TOA estimation structures based on chip-rate sampling.

Hence, the resolution is limited by the sampling interval. Consider using the ED structure for the time delay based DF. Assume a 10-element ULA geometry with 1 meter inter-element spacing. The impinging source is coming from 8° with respect to the array axis. This scenario is illustrated graphically in Figure 4.7. At the 10-th antenna, the inter-element delay with respect to the first antenna is only about 0.3 ns. Thus the maximum energy block at the 10-th antenna is the same as at the first antenna. However, if the impinging source is coming from 15° , the 10-th antenna perceives the inter-element delay to be 1 ns with respect to the first antenna. In addition, the maximum energy block at the 10-th antenna is now the next sampled data. This case is shown in Figure 4.8. For such an array, the angular resolution achieved is only 15° .

To formulate the angular resolution θ_r of a given time delay based DF system, the notation t_r is defined as the resolution of the time delay estimation. The relationship can then be expressed as

$$t_r = (N - 1) \frac{d}{c} [1 - \cos(\theta_r)] \quad (4.11)$$

Note that it is directly proportional to the array aperture length and the inter-element spacing. Figures 4.9 and 4.10 plot the time delay estimation resolution curve as a function of angular resolution for various number of array elements and inter-element spacing. This relationship is conceptually similar to the angular resolution defined in [49].

This limitation is not only faced by the ED structure but also other TOA estimation structures using sub-Nyquist rate sampling. Unlike these structures, the proposed LTD utilizes a latch circuit to produce a rectangular pulse. The duration of the rectangular pulse is known *prior* to the estimation, which is the latch duration. Given

Figure 4.7: A simple illustration of implementing TOA estimation structure with pulse-duration sampling to form a time delay based DF for UWB IR. No timing mismatch is considered. The pulse is arriving from 8° . Because all of the received signal falls within the same sampling duration, the system is unable to distinguish this case from around the 0° signal arrival case.

Figure 4.8: A simple illustration of implementing TOA estimation structure with pulse-duration sampling to form a time delay based DF for UWB IR. No timing mismatch is considered. The pulse is arriving from 15° . Observe that the received signal at the last element is one pulse-duration delayed from the received signal at the first element. If the TOA is estimated correctly, the system will be able to calculate the DOA.

Figure 4.9: Delay estimation resolution required as a function of angular resolution, from different values of d . The plot is generated from the theoretical derivation from (4.11).

Figure 4.10: Delay estimation resolution required as a function of angular resolution, from different values of N . The plot is generated from the theoretical derivation from (4.11).

the value of the sampled output, the time delay can then be calculated (see (4.8), the calculation of $\hat{\tau}_{1,n}$ will be discussed later in the next section). Theoretically, there is no limit on the accuracy of the time delay estimation. Implementation-wise, the time delay estimation will be limited by the dynamic range of the ADC used. This characteristic opens up the possibility of implementing a small aperture array for UWB IR.

4.5 Joint Detection and DF Technique

In this section, a DF function based on the estimation of the time delay is formulated. No assumption on the synchronization or sharing of the time reference between transmitter and receiver is required. In addition, the detection problem is jointly addressed with the estimation problem as no perfect detection is assumed prior to the DOA estimation.

4.5.1 Fractional-Norm Based Cost Function

A common approach used to solve for DOA estimation, given N time delays is to use the least squares (LS) approach. This is because the problem is an over-determined linear system. Similar work on applying LS to acoustic source DF based on time delay is reported in [20]. However, in this case, the time delay is an estimated value subjected to errors. These errors may be due to either false alarm or missed detection of the LTD.

As an example, 1000 realizations of the received multipath signal is generated and the estimation process is simulated as described in Section 4.3. The resulting errors are recorded and the cdf is plotted in Figure 4.11. As depicted, some outliers are observed in the error data. They do not appear frequently and may be very distant

from most of the error data. A similar observation has also been reported in [50]. Notice that most of the outliers are located in the negative TOA region. As E_s/N_o ratio increases, the number of outliers in the negative region reduces.

Figure 4.11: CDF plots of the erroneous time delay estimates with varying E_s/N_o parameters. The presence of outliers is clearly observed in the negative TOA region. This is due to the increasing false alarm rate. As E_s/N_o increases, the number of outliers reduces.

Hence, a fractional norm formulation is preferred over LS approach. The DOA is estimated by scanning for the direction that minimizes the fractional norm cost function $d_M(\theta_s)$

$$d_M(\theta_s) = \left\{ \sum_{n=0}^{N-1} \left| \hat{\tau}_{0,n} - s_n(\theta_s) \right|^p \right\}^{\frac{1}{p}} \quad (4.12)$$

$$s_n(\theta_s) = \frac{x_n \sin(\theta_s) + y_n \cos(\theta_s)}{c} \quad (4.13)$$

This cost function is generally defined as the Minkowski distance [51] of order p or p-norm distance between a point $(\hat{\tau}_{0,0}, \hat{\tau}_{0,1}, \dots, \hat{\tau}_{0,N})$ and $(s_0, s_1, \dots, s_N)$ in a N -dimensional space. The fractional-norm distance is a special case of Minkowski distance and it has been used to reduce the impact of less frequent extreme outliers as

reported in [52].

4.5.2 Ensuring The Reliability of DOA Estimation

When the majority of errors are due to the outliers, the minimization of the fractional-norm distance no longer gives a reliable DOA estimation. This is because the directional information has been lost during the analog front-end preprocessing. To identify such instances, the N data points from the inter-element delay estimates $\hat{\tau}_{1,n}$ are first partitioned into two subsets: good data points and outliers data. Then, the number of good data points is calculated and a proposed algorithm decides whether the initial DOA estimate $\hat{\theta}_1$ is reliable.

The partitioning is done by comparing the absolute difference between the inter-element delay estimates $\hat{\tau}_{0,n}$ and $s_n(\hat{\theta}_1)$, denoted as $d_n(\hat{\theta}_1)$, with a threshold ϵ . If $d_n(\hat{\theta}_1)$ is greater than ϵ , the data point is classified as an outlier and vice versa. $d_n(\hat{\theta}_1)$ can be calculated as follows

$$d_n(\hat{\theta}_1) = |\hat{\tau}_{0,n} - s_n(\hat{\theta}_1)| \quad (4.14)$$

If the number of good data points (denoted as N_ϵ) is greater than the threshold ς , the proposed algorithm decides that $\hat{\theta}_1$ is reliable. Otherwise, it is unreliable. The reliability of $\hat{\theta}_1$ determines the presence of the direct path's pulse. If $\hat{\theta}_1$ is reliable, the proposed algorithm decides that the direct-path's pulse is present in the observation. To summarize, the pseudo-code for the joint detection and DF is given in Table 4.1.

4.6 Numerical Results

Numerical simulations are used to evaluate the proposed joint detection and DF technique. A Gaussian 2^{nd} derivative pulse with $T_p \approx 1\text{ns}$ is considered. The mathematical

Joint Detection and DF Algorithm

1. Given: $x_n = \{x_0, x_1, \dots, x_{N-1}\}$ and $i_n = \{i_0, i_1, \dots, i_{N-1}\}$.
 2. Calculate for $n = \{1, 2, \dots, N-1\}$:

$$\widehat{\tau}_{0,n} = (i_n - i_0)T_{latch} + x_n - x_0$$

$$s_n(\theta_s) = n \frac{d}{c} \cos(\theta_s)$$
 3. Calculate the Minkowski distance metric:

$$d_M(\theta_s) = \left\{ \sum_{n=0}^{N-1} |\widehat{\tau}_{0,n} - s_n(\theta_s)|^p \right\}^{\frac{1}{p}}$$
 4. Find θ_s that minimizes the Minkowski distance metric:

$$\widehat{\theta}_1 = \arg \min_{\theta_s} d_M(\theta_s)$$
 5. Initialize: $N_\epsilon = 0$
 6. Evaluate for $n = \{1, 2, \dots, N-1\}$:

$$d_n(\widehat{\theta}_1) = |\widehat{\tau}_{0,n} - s_n(\widehat{\theta}_1)|$$
 If $d_n(\widehat{\theta}_1) < \epsilon$, then $N_\epsilon = N_\epsilon + 1$
 7. If $N_\epsilon < \varsigma$
 then the final DOA estimate is $(\widehat{\theta}_1)$ and the UWB signal is detected at x_0 .
-
-

Table 4.1: The proposed joint detection and DF algorithm.

expression of Gaussian 2nd derivative pulse is given by

$$w(t) = \frac{1}{\sqrt{2\pi}\sigma^3} \left(1 - \frac{t^2}{\sigma^2} \right) \exp \left(-\frac{t^2}{2\sigma^2} \right) \quad (4.15)$$

The parameter σ is defined such that the duration of the Gaussian pulse, $T_p \approx 1 \text{ ns}$.² As suggested in [36], σ is set such that 99.99% of the energy of the pulse is contained within the T_p duration. Therefore, $\sigma = 0.19 \text{ ns}$ is used in this simulation. The frame interval (PRI) is fixed at $T_f = 300 \text{ ns}$. At the analog front-end, the simulation parameters used are (unless otherwise stated) $\gamma = 0.6$ for the AWGN case and $\gamma = 0.3$ for the multipath case, $T_{latch} = 150 \text{ ns}$, and the integration duration is equal to T_{latch} . A uniform linear array (ULA) geometry is considered. The array considered has $N = 7$ antennae with $d = 50 \text{ cm}$ and the array phase center is located at the 4-th antenna (see Figure 4.12). The propagating IR source is assumed to be from 30° for all cases.

²Even though the duration of the Gaussian pulse and all of its derivatives is infinite, the simulation truncates the pulse so that the duration is T_p .

Figure 4.12: The considered array structure where the array phase center is located in the middle element.

The multipath propagation is modeled using the proposed channel model in [1] (indoor office/lab environment). To simulate the received signal at the array, this temporal model needs to be extended into a spatio-temporal model. To model the multipath's DOA, the proposed cluster-ray model in [37] is adopted. The clusters' DOAs are generated using a uniform distribution while the rays' DOAs are generated from a Laplacian distribution characterized by its standard deviation parameter and centered at the DOA of the cluster to which the ray belongs. For the clusters' DOA, the possible value of DOA is limited within $[0^\circ, 180^\circ]$ since a ULA geometry is considered. The standard deviation of the log-normal distribution is set at 25.5° in the simulations. This spatio-temporal model is similar to the model described in Section 2.8.

4.6.1 Effect of Using Fractional Norm

First, the effect of using the fractional-norm based cost function is investigated and compared against the least squares ($p = 2$) and L1-norm ($p = 1$) formulation. For the fractional-norm formulation, the simulation considers $p = 0.1$. The simulation also assumes an AWGN propagation channel with $E_s/N_o = 10dB$. All array channels employ $\gamma = 0.6$ for the level threshold. The proposed algorithm uses $\epsilon = 0.13ns$ and $\varsigma = 3$. The results of the TOA estimation from a single realization are as follows. The first array channel completely misses the pulse and reports no detection and no TOA estimation. The other channels, after normalizing the data with respect to the fourth antenna, report the following TOA estimation $\{-2.917, -1.417, 0, 1.542, -2.375, -2.750\}ns$. The

normalized cost function calculated with $p = 2$ (least squares), $p = 1$ (L_1 -norm) and $p = 0.1$ (fractional norm) are compared and depicted in Figure 4.13.

Figure 4.13: Plots of normalized cost function comparison with $p = 2$, $p = 1$ and $p = 0.1$. The use of fractional norm effectively maintains the robustness of the DOA estimator in the presence of outliers as compared to the other methods.

Observe that the estimated TOA estimations are erroneous. The correct estimates of the 2^{nd} – 7^{th} array channels should be $\{-2.8868, -1.4434, 0, 1.4434, 2.8868, 4.3301\}$ ns (see Table 4.2 for a better comparison). The 2^{nd} , 3^{rd} and 5^{th} channels give good TOA estimation. However, the last two estimates are due to false alarms. The Minkowski metrics based on the $p = 1$ and $p = 2$ formulations fail to estimate the DOA. Using the fractional norm ($p = 0.1$), the normalized cost function indicates two minimum points: at around 30° and 135° . The global minimum point is at 30° . This result implies that the use of fractional norm effectively provides good DOA estimator even when almost half of the TOA estimates are unreliable.

	Estimated delay [ns]	True delay [ns]	Remarks
$\tau_{0,-3}$	0	-4.3301	no detection
$\tau_{0,-2}$	-2.917	-2.8868	detected with error
$\tau_{0,-1}$	-1.417	-1.4434	detected with error
$\tau_{0,0}$	0	0	reference
$\tau_{0,1}$	1.542	1.4434	detected with error
$\tau_{0,2}$	-2.375	2.8868	false alarm
$\tau_{0,3}$	-2.750	4.3301	false alarm

Table 4.2: The estimated and true delays of $\tau_{0,n}$ in ns.

Figures 4.14 and 4.15 show the root-mean-square error (RMSE) plots and the detection rate plots as a function E_s/N_o , respectively. The results are obtained from 1000 different realizations. In this simulation, a multipath propagation channel is considered. The channel is assumed to be static for this Monte Carlo simulation. Due to the energy spreading effect of multipath, the minimum E_s/N_o evaluated is 15dB. Below 15 dB, the UWB IR signal will be buried below the noise level and no detection is possible in this case. The threshold level is set at $\gamma = 0.3$. In this figure, the fractional norm performance clearly surpasses the L1-norm in terms of RMSE. The detection performance is nearly unchanged.

Figure 4.14: Root mean square error (RMSE) plots as a function of E_s/N_o . The use of the fractional norm (1/10-norm) reduces the effect of outliers as compared to the commonly used L_1 -norm, thus reducing the RMSE.

Figure 4.15: Detection rate plots as a function of E_s/N_o . Comparing between the fractional norm and L_1 norm, no significant difference is observed.

4.6.2 Effect of Array Geometry

Next, the performance of the fractional norm is investigated while the inter-element spacing d and number of elements N are varied. The previous simulation settings are also used here. Figures 4.16 and 4.17 show the RMSE plots as well as the detection rate plots as a function E_s/N_o , respectively. Generally, both figures show better performance when the spacing is larger. A greater improvement is shown in the RMSE plot as compared to the detection rate plot. Figures 4.18 and 4.19 depict the performance comparison for different N . Better detection and estimation are observed for a higher N . On the whole, a larger aperture array leads to a better detection and estimation performance.

Figure 4.16: Root mean square error (RMSE) plots as a function of E_s/N_o for different inter-element spacing d . Better performance (lower RMSE) is achieved for larger d and the improvement is more significant as compared to the detection rate improvement shown in Figure 4.17.

Figure 4.17: Detection rate plots as a function of E_s/N_o for different inter-element spacing d . Better performance (higher detection rate) is achieved for larger d .

Figure 4.18: Root mean square error (RMSE) plots as a function of E_s/N_o for different number of array elements N . Better performance (lower RMSE) is achieved for higher N .

Figure 4.19: Detection rate plots as a function of E_s/N_o for different inter-element spacing N . Better performance (higher detection rate) is achieved for larger N .

4.6.3 Optimizing (γ, ϵ)

In this Section, the effects of the threshold level setting γ towards the performance of the proposed technique are investigated. It is clear that decreasing γ will increase the detection rate at the expense of a higher false alarm rate. However, the increase in the false alarm rate can be suppressed by the proposed technique. In addition, the effects of the time delay error bound ϵ towards the estimation performance will also be investigated. Thereafter, the optimum (γ, ϵ) can be evaluated.

1000 different realizations of AWGN and multipath cases are simulated with $E_s/N_o = 10dB$ and $E_s/N_o = 17dB$, respectively. For every (γ, ϵ) simulated, the detection rate, the false alarm rate and the RMSE are calculated. The proposed algorithm is set with $p = 0.1$ and $\varsigma = 3$ in this simulation. The 3D plots of the detection rate, false alarm rate and RMSE are shown in Figures 4.20, 4.22 and 4.24 respectively for the AWGN case. The corresponding 3D plots for the multipath case are shown in Figures 4.21, 4.23 and 4.25, respectively.

As shown for both the AWGN and multipath cases, decreasing γ results in the increase of the detection rate while the false alarm rate is suppressed. For the AWGN case, this occurs when γ is between 0.5 and 0.6. For the multipath case, γ is between 0.25 and 0.3. In fact, both cases show an optimum detection rate when γ is set at 0.6 for the AWGN case and 0.3 for the multipath case. The RMSE for both cases generally decreases as γ increases. In the multipath case, an increase of γ above 0.3 will result in an increase in RMSE but much slower than the decrease in RMSE when γ is below 0.3. This creates a minimum around $\gamma = 0.3$. This phenomenon is not observed for the AWGN case. The reason is because, in the multipath case, the detector misses the pulse due to the direct path but detects the DOA due to the multipath. As a result, the estimator reports a higher error.

Figure 4.20: Detection rate 3D plot as a function of (γ, ϵ) in AWGN case.

Figure 4.21: Detection rate 3D plot as a function of (γ, ϵ) in multipath case.

Figure 4.22: False alarm rate 3D plot as a function of (γ, ϵ) in AWGN case.

Figure 4.23: False alarm rate 3D plot as a function of (γ, ϵ) in multipath case.

Figure 4.24: RMSE 3D plot as a function of (γ, ϵ) in AWGN case.

Figure 4.25: RMSE 3D plot as a function of (γ, ϵ) in multipath case.

As for ϵ , a trade-off between detection rate and RMSE is observed for both the AWGN and multipath cases. The RMSE is reduced when ϵ is decreased but the detection rate is also reduced. Compared with the effects of γ , the decrease in detection rate is less significant.

Figure 4.26: CDF plot of the DOA estimation error using the optimum combination of (γ, ϵ) and a non-optimum one. With the optimum (γ, ϵ) , outliers due to TOA estimation are no longer existent in estimated DOA.

As a comparison, two different combinations of (γ, ϵ) are simulated and compared for the AWGN case at $E_s/N_o = 10\text{dB}$. When (γ, ϵ) are set at (0.6 volt, 0.13ns), the resulting DF will be optimized. To show this, the CDF curve is plotted in Figure 4.26 and compared against a non-optimum setting, $(\gamma = 0.4\text{volt}, \epsilon = 0.03\text{ns})$. The estimator is an unbiased estimator when the optimum setting is used. Otherwise, the estimator is not robust and is biased.

4.6.4 Effect of ς

In this subsection, the effects of varying ς are investigated. In the proposed algorithm, ς is used in testing the number of expected outliers among the estimated TOA reported from all sensors. In the simulation, 1000 realizations are generated for each different ς value. The multipath case is considered and E_s/N_o is varied from 15 dB up to 21 dB. Other parameters are set as follows: $\gamma = 0.3$, $\epsilon = 0.13\text{ns}$ and $p = 0.1$. Figure 4.27 depicts the detection rate curve as a function of E_s/N_o and Figure 4.28 shows the RMSE curve. Notice that there is a performance trade-off between the detection rate and RMSE. ς can be increased to have a better estimation performance, at the expense of a worsen detection performance.

4.6.5 Performance Comparison Against Focusing Matrix Approach (WCSM)

Here, the performance of the proposed DOA estimation is compared with the wideband DOA estimator based on focusing matrices proposed in [28] and [19], also known as weighted coherent signal-subspace method (WCSM). In addition, the detection performance of the proposed method is also compared with the energy detection (ED) method reported in [53].

This ED method is based on a simple structure that comprises of a squarer and an integrate-and-dump operation, followed by a sampler (see Figure 4.6(c)). The detection is based on the sampled integration-output. Unlike the proposed structure, the required sampling rate is of the order of one pulse duration (approximately 1 ns). The key idea is to collect the energy of the received pulse and to detect it by finding the maximum energy blocks at the output of the sampler. In the current literature, this is the simplest structure for detecting a UWB IR signal. The drawback is that

Figure 4.27: Detection rate plots as a function of E_s/N_o for different settings of ζ . Unlike the RMSE plot, a better performance (higher detection rate) is achieved for smaller value of ζ . Thus there is a trade-off between the RMSE and the detection rate.

Figure 4.28: Root mean square error (RMSE) plots as a function of E_s/N_o for different settings of ζ . A better performance (lower RMSE) is achieved for higher value of ζ .

the performance is degraded in low SNR environments [40].

Both the proposed structure and ED method are simulated in AWGN and multipath environments. The ED method uses 1 ns sampling rate while the proposed structure uses 30 ns sampling rate. Both detectors maintain a false alarm rate below 0.01 while varying E_s/N_o . For the AWGN case, Figure 4.29 depicts the detection rate performance comparison between the proposed technique and the ED method utilizing a square-law device. The corresponding plot for the multipath case is shown in Figure 4.30. It is clear that the proposed structure outperforms the ED method for both the AWGN and multipath environments most of the time. As E_s/N_o increases, the detection-rate-gap difference decreases.

For estimation performance comparison, the performance of WCSM is used as a benchmark. Although both WCSM and CSM are based on a subspace-search DF method, there is a slight difference between both methods. WCSM eigen-decomposes the focused signal-subspaces, which are weighted with UWB IR signal power spectral density (PSD) [19]. To know the PSD, weighted CSM assumes *prior* knowledge of the received pulse waveform.

The weighted CSM is simulated only in the AWGN channel because the number of elements required to resolve more than 100 multipaths and the processing power required are tremendous. To implement this method, a high sampling rate of 100 GHz is considered in the simulation (not possible practically). 32 frequencies centered around the maximum PSD are focused to the reference frequency at 100 MHz. 50 pulses from single UWB IR source are received at the antenna array and used for estimating the correlation matrix and forming the signal-subspace. Note that a perfect detection is considered for the weighted CSM. Based on the aforementioned considerations, 200 realizations are simulated and the performance is compared with the proposed DF method. The performance is evaluated in terms of the statistical bias and standard

Figure 4.29: Detection rate plots as a function of E_s/N_o in an AWGN propagation environment. The plots compare the proposed technique with the square-law device based detection. Most of the time, the proposed technique performs better. As E_s/N_o increases, the detection-rate-gap difference decreases.

Figure 4.30: Detection rate plots as a function of E_s/N_o in a multipath propagation environment. The plots compare the proposed technique with the square-law device based detection. Most of the time, the proposed technique performs better. As E_s/N_o increases, the detection-rate-gap difference decreases.

deviation.

Figures 4.31 and 4.32 show a performance comparison of the proposed DF approach and the weighted CSM in terms of bias and standard deviation in an AWGN propagation environment. The corresponding plots for the multipath case are shown in Figures 4.33 and 4.34, respectively. Since the WCSM is not able to resolve hundreds of multipath, the performance of the WCSM in the AWGN case is used for comparison with the proposed technique in multipath case. For both cases, the proposed DF approach maintains a constant bias below 0.5 degree, which is far below the simulated weighted CSM. The same observation is seen for the standard deviation curve. This demonstrates the superiority of the proposed DF over conventional wideband DOA estimators. Note that all the simulations presented consider the optimum (γ, ϵ) settings. With proper settings of the threshold γ and error bound ϵ , the multipath does not affect the detection and estimation performances of the proposed technique.

Figure 4.31: Estimation bias plots as a function of E_s/N_o in an AWGN environment. The plots compare the proposed technique with the WCSM. In all cases, the bias values are shown to be smaller than 0.5° . This reflects that the estimator is statistically unbiased.

Figure 4.32: Estimation standard deviation plots as a function of E_s/N_o in an AWGN environment. The plots compare the proposed technique with the WCSM. In all cases, the standard deviation values are shown to be smaller than the WCSM. This reflects that the proposed technique performs better than the WCSM.

Figure 4.33: Estimation bias plot of the proposed technique in a multipath propagation environment and the WCSM in an AWGN environment. The bias level of the proposed technique is shown to be lower than that of the WCSM.

Figure 4.34: Estimation standard deviation plot of the proposed technique in a multipath propagation environment and the WCSM in an AWGN environment. The proposed technique is better than the WCSM.

4.6.6 Estimator's Confidence Level

Lastly, the confidence level (CL) for the proposed time-delay based DF method is calculated. Without loss of generality, the CL is considered for estimation errors less than 5° . The simulation considers 1000 realization of both AWGN and multipath environments. Figure 4.35 depicts the results. Nearly 100% CL is observed for E_s/N_o higher than 9 dB in the AWGN environment and 15 dB in the multipath environment. Table 4.3 shows the CL together with the RMSE for both cases. Generally, the RMSE of the proposed DF method is below 2° in both cases. If the estimator is realized in $E_s/N_o > 15$ dB for the multipath case, the achievable CL is higher than 98% and the RMSE is below 1.5° . This ensures high reliability of the estimator.

Figure 4.35: Confidence level plot of the proposed DOA estimator as a function of E_s/N_o , generated using Monte Carlo simulation of 1000 realizations. The confidence level measures the probability of generating error less than $\pm 5^\circ$.

E_s/N_o [dB]	Confidence Level [%]	RMSE [degree]
7	81.5	1.8624
9	99.6	1.6057
11	99.5	1.5424
13	100	1.3649
15	100	1.2708
17	100	1.0858
19	100	1.0588
21	100	1.0909

E_s/N_o [dB]	Confidence Level [%]	RMSE [degree]
13	36	1.7127
15	98.2	1.3512
17	99.7	1.1606
19	100	1.1176
21	100	1
23	100	0.8802
25	100	0.8234
27	100	0.7622

Table 4.3: Performance of the proposed Joint Detection and DF method in (a) AWGN and (b) Multipath cases.

4.7 Conclusion

The DOA estimation approach based on time delay estimation is discussed in this chapter. The array system consists of the LTD at each antenna element, followed by time-delay-estimates extraction. The DF techniques in previous chapters are based on the sum of the LTD's output signals at the array whereas the proposed DF here is based on the TOA estimation of each sensor instead. This approach is driven by the presence of the non-idealities during the detection stage. Section 4.3 presents two different approaches of estimating the TOA without Nyquist-rate sampling. The first approach processes the received signal in the frequency domain while second approach process in the time domain.

Section 4.4 discusses the angular resolution limit of time delay based UWB IR DOA estimation. This limit is due to the accuracy bound of the time-delay-estimation. We show that the use of latch circuit in the LTD results in an improved angular resolution. With the latch operation, the angular resolution is determined by the ADC's dynamic range, not the sampling interval.

Section 4.5 describes the joint detection and DF technique. The DF formulation is based on a fractional norm metric in order to de-emphasize the impact of the large TOA estimation error. In addition, the technique also incorporates a TOA error threshold to ensure the reliability of the DOA estimates. The detection technique is motivated by the idea that the signal-plus-noise observation has spatial information while the noise-only observation does not. Whenever a reliable DOA estimation is reported, the detector decides that the signal is present. Together, it forms a joint detection and DF technique.

Section 4.6 presents the numerical analysis. Here it is shown that a careful choice of the LTD threshold level and the TOA error bound is critical to achieve good UWB-

IR signal detection and DF performances. The Monte Carlo simulation results suggest that this proposed approach outperforms the techniques discussed in previous chapters.

Chapter 5

UWB IR Direction Finding Extension to Time Hopping Multiple Access

5.1 Introduction

The fact that propagation space, time and frequency are limited has forced communication system designers to think of a means for channel-sharing. When the channel is shared, different transmitters emit different signals embedding different data. These signals will be received and processed at the receiver. To be able to decode the data, the receiver has to separate one user's transmission from the rest.

Time Hopping Multiple Access (TH-MA) is well-known for its UWB IR applications. In UWB-IR, the transmitted signal is a sequence of very short duration (sub-nanosecond) pulses. The TH-MA scheme dictates the instant when each pulse is transmitted. These instances are deduced from the hopping code uniquely assigned to different transmitters.

The spectral properties of TH-MA transmission has been widely analyzed in the literature. The derivation of the PSD expression of a TH signal is reported in [54]. The presence of timing jitter smoothes out the PSD, as reported in [55] and [56]. Their work also derive the PSD expression in the presence of timing jitter. Further research works suggest that, by designing a proper hopping code, the spectral properties can be adjusted. A means to control the spectrum with the TH code was proposed in [57] while [58] tries to minimize the PSD at some frequency band used for existing communication system to address the UWB co-existence issue. The pulse shape also determines the shape of the PSD as derived in [36].

Our research contribution presented in this chapter is in-line with the aforementioned efforts. As derived in [55], the PSD of a TH-IR signal is composed of multiple spectral lines. Here, the objective is to cancel the multiple-access interference (MAI) by isolating its spectral lines. This can be done by properly designing TH codes for all transmitters in the system such that the spectral lines of MAI is spectrally separable from those of the user-of-interest. In this chapter, a code design method to produce the codes with such spectral properties is also proposed. With the properly designed TH codes, it is possible to use the proposed channelization-based DF approach discussed in Chapter 2 to estimate the DOA from one of the users in the presence of the MAI transmissions from other users.

This method is based on the cancellation of code-dependent terms in the PSD expression. It can be used in different ways depending on applications

- In UWB-IR DF, several techniques are based on spectral lines extraction to estimate the DOA information [59], [19]. With our proposed code design, MAI can be cancelled by not selecting the spectral lines of MAI.
- For co-existence of UWB-IR system with other wireless system, our proposed

code design can be used to minimize the imposed UWB interference on a given narrowband system, without the need for a complex receiver design. For example, it is possible to design a hopping sequence with no spectral line contributions in the GSM-900 or GSM-1800 band.

- Other frequency-domain-processing based receiving techniques [22], [60] can also benefit from this scheme. The fact that they do not have to process at the frequencies, where the spectral lines of MAI are present, may help to increase their processing gain.

It is worth noting that our code design is based on a simple technique and the length of the code designed can be as short as four.

In addition, the problem of DOA estimation for multiple transmitting users of *unknown* TH codes is also addressed. The proposed solution is also based on the channelization structure, but requires a bigger structure with more frequencies to be channelized. The rest of the chapter is organized as follows. Section 5.2 introduces the TH signal model and formulates the received signal model at the antenna array, which is a continuation from the model derived in Chapter 2. In Section 5.3, the spectral separation characteristics of the TH code designed and its design method are discussed. The DOA estimation techniques, either with *prior* knowledge or an unknown hopping sequence, are presented in Section 5.4. Simulation results and concluding remarks are given in Section 5.5 and Section 5.6, respectively.

5.2 Spectral Properties of TH-MA Transmission

Earlier on, the received signal model in (2.1) assumes single-user transmission. In multiple-user transmission, the transmission of every user is governed by the MA

scheme. For TH-MA scheme, the transmitted signal model of the q -th user is given by [55]

$$s_q(t) = \sum_{j=-\infty}^{\infty} w_{tx}(t - jT_f - c_j^{(q)}T_c) \quad (5.1)$$

where $c_j^{(q)}$ symbolizes the hopping sequence of the q -th user. That is, the transmitted signal is a train of pulses positioned at different time instants. These instants are defined by the hopping sequence. Note that no data modulation is considered in the signal model since the system considered in this thesis is a localization system, where there is no exchange of information between the transmitter and receiver.

The hopping code is of a limited length (denoted as N_f), which is also the number of frames representing one information symbol. Since only one pulse is transmitted within one frame duration, the number of pulses used for representing one symbol is also N_f . Thus the periodicity of the transmission is N_fT_f . There are N_f pulses in one period of transmission. Each pulse is delay-shifted by c_jT_c where c_j is an integer value between 0 and $N_h - 1$

$$c_j = \{0, 1, \dots, N_h - 1\} \quad (5.2)$$

where T_c is the chip interval. Also, one frame interval is divided equally into N_h chip intervals

$$N_h = \frac{T_f}{T_c} \quad (5.3)$$

Figure 5.1 shows the single user transmission and q -th user transmission employing a TH-MA format. The hopping sequence of the q -th user in the illustration is $c_j^{(q)} = \{0, 3, 2, 1\}$ with $N_h = 4$. The length of the sequence, denoted by N_f , is 4 and the transmission period is N_fT_f .

The PSD of the TH signal has been derived and documented in [56], [55]. The

Figure 5.1: (a) An illustration of single user transmission and (b) the q -th user transmission employing the TH format. Single user transmission is a regular interval pulse train while the TH format transmission is irregular. The hopping sequence deduced from the illustration is $\{0, 3, 2, 1\}$ where $N_h = 4$.

PSD of the received signal due to the q -th transmission (assume no time jitter) can be expressed as [54]

$$S_q(f) = \frac{1}{(N_f T_f)^2} |W(f)|^2 \left| \sum_{j=0}^{N_f-1} \exp(-i2\pi f(jT_f + c_j^{(q)} T_c)) \right|^2 \sum_{k=-\infty}^{\infty} \delta_D \left(f - \frac{k}{N_f T_f} \right) \quad (5.4)$$

where $W(f)$ is the Fourier transform of the transmitted waveform $w_{tx}(t)$ and δ_D is the Dirac delta function. The PSD is composed of multiple line spectra separated by the reciprocal of $N_f T_f$ [61] (see Figure 5.3 for an illustration of the PSD plot and Figure 5.2 for its time domain plot). The PSD¹ of the TH signal is determined by both the transmitted pulse waveform and the TH code.

Notice that $C_q(f)$ is defined as the sum of the code-dependent terms in the PSD

¹Note that from this point onwards, the term “PSD” refers to the discrete PSD.

Figure 5.2: An illustration of a typical time hopping modulated UWB IR signal. The pulse waveform considered is a Gaussian 2nd derivative pulse (shown in inset figure).

Figure 5.3: The Power Spectral Density (PSD) plot of a time hopping modulated UWB IR signal. The PSD is composed of line spectra (shown in inset figure). The hopping sequence determines the magnitude of these line spectra.

expression

$$C_q(f) := \left| \sum_{j=0}^{N_f-1} \exp(-i2\pi f(jT_f + c_j^{(q)}T_c)) \right|^2 \quad (5.5)$$

It is possible to further simplify (5.5) by first expanding it using Euler's formula and then re-expressing it using trigonometric equalities. Thus the following expression is obtained (see Appendix A for a detailed derivation)

$$C_q(f) = N_f + \sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos[2\pi(\rho_q T_c)f], \quad j_1 \neq j_2 \quad (5.6)$$

where ρ_q is defined as a function of the frame index (j_1, j_2)

$$\rho_q = N_h(j_1 - j_2) + c_{j_1}^{(q)} - c_{j_2}^{(q)} \quad (5.7)$$

Recall that the PSD is a group of spectral lines at integer intervals of $1/(N_f T_f)$. This means that the PSD can be seen as a discrete signal and the continuous frequency index f can be replaced with the discrete frequency index k using the relationship $f = k/(N_f T_f)$. The expression for $C_q(f)$ is now $C_q(k)$

$$C_q(k) = N_f + \sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos \left[2\pi \frac{\rho_q}{N_f N_h} k \right] \quad j_1 \neq j_2 \quad (5.8)$$

From the above expression, the sum of the code-dependent terms is, in fact, a sum of cosine terms. It can be deduced that

- The cosine term is analogous to a discrete cosine signal

$$\cos(2\pi f k) \Leftrightarrow \cos \left(2\pi \frac{\rho_q}{N_f N_h} k \right) \quad (5.9)$$

- The cosine term is periodic in $N_f N_h$.
- The maximum value of $C_q(k)$ is N_f^2 , when all cosine terms are equal to 1.

5.3 Code Design for Spectral Separation in Time Hopping Multiple Access

By changing the TH code, the sum of the code-dependent terms will also change. Since this is a multiplicative factor of the PSD, nullifying this factor will result in nullifying the PSD. Our first objective is to develop a TH code design method for removing the imposed UWB-IR interference on a given narrowband system. To facilitate the discussion, a GSM-900 band system is considered. However, our design method is not limited to this system only. It is also applicable for any given narrowband systems (as will be explained in the following discussion).

The next objective is to remove the spectral line contributions from the MAI at a frequency subset of the whole UWB band. For simplicity of illustration, two transmitters are considered (the generalization to more transmitters is trivial), one being a user-of-interest and the other being the multiple-access interferer. The TH code employed for the interferer is designed such that no spectral line contributions are observed at the frequency subset, while the TH code employed for the user-of-interest is designed such that the spectral line contributions are maximized within the frequency subset. If such a condition is satisfied, the spectrum of the interferer is said to be separable from the user-of-interest.

Lastly, the methodology for the construction of the TH codes is extended for the case when the interferer now becomes the user-of-interest and vice versa. Within different frequency subset, the spectral line contributions of the interferer are maximized

while those of the user-of-interest are removed. If this condition is also satisfied, the spectrum of the two transmitters are said to be mutually separable. The extension of the design method for mutual spectral separation of more than two transmitters is also discussed in Section 3.

5.3.1 Removing the Imposed UWB-IR Interference on a Narrowband System

The discussion begins with a technique to null the sum of the code dependent terms at the GSM-900 band, and then to designing the code. It is assumed that the code length N_f and the parameter N_h are even numbers. Let f_{nb} denote the frequency of the GSM-900 band. The goal is to design the hopping sequence that results in $C_q(f_{nb}) = 0$. Since the PSD is a discretized spectrum, T_c has to be set such that f_{nb} is an integer multiple of $N_f T_f = N_f N_h T_c$. One possible way is to set $T_c = 1/(2f_{nb})$ or $k_{nb} = f_{nb}(N_f N_h T_c) = \frac{1}{2}N_f N_h$. Thereafter, the following expression for $C_q(k_{nb})$ is obtained

$$C_q(k_{nb}) = N_f + \sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos(\pi \rho_q) \quad j_1 \neq j_2 \quad (5.10)$$

where ρ_q has been defined in (5.7).

Because ρ_q is an integer value, the value of the cosine term is either 1 or -1 depending on whether ρ_q is even or odd in value

$$\cos(\pi \rho_q) = \begin{cases} 1, & \rho_q \in 2\mathbb{N} \\ -1 & \rho_q \in 2\mathbb{N} + 1 \end{cases} \quad (5.11)$$

where $\mathbb{N} = \{0, 1, 2, 3, \dots\}$ is a set of natural numbers or non-negative integers. To

nullify $C_q(k_{nb})$, the double sum of the cosine terms must be equal to $-N_f$. That is,

$$\sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos(\pi \rho_q) = -N_f \quad (5.12)$$

The number of cosine terms in the double summation is $N_f^2 - N_f$. The strategy is to have $\lceil \frac{1}{2}(N_f - 1) \rceil / (N_f - 1)$ cosine terms to be equal to -1 and the remaining $\lfloor \frac{1}{2}(N_f - 1) \rfloor / (N_f - 1)$ to be 1 (this is only true when N_f is even). As a result, the double summation is equal to $-N_f$. It can be proved that when N_h is even, it is possible to meet the requirement by setting half of the TH codes to be even and the other half to be odd. Hence, the TH code designed, using $N_h = 4$, is

$$c_j^{(q)} = \{2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N} + 1, 2\mathbb{N} + 1\} \quad (5.13)$$

The transmitted PSD of the TH signal with $T_c = 1/(2f_{nb})$ employing this code will not have any contribution at f_{nb} . Note that the order of the code does not affect the resulting PSD.

The GSM-900 system operates in the 890-915 MHz frequency range. To achieve good covertness, a UWB-IR device has to avoid using this band. Using the proposed TH code design method, it is possible to transmit a TH signal whose PSD does not occupy this band.

The following example shows the resulting PSD of the UWB-IR signal that employs the TH code design given in (5.13). The TH-MA parameters are: $T_c = 1/(2 \times 900 \text{ MHz}) = 0.555 \text{ ns}$, $N_f = 4$, $N_h = 8$ and $T_f = N_h T_c = 4.444 \text{ ns}$. The transmitted UWB pulse waveform is assumed to be the fourth derivative Gaussian pulse, which is given by [36]

$$w_{tx}(t) = \frac{A}{\sqrt{2\pi}\sigma^5} \exp\left(-\frac{t^2}{2\sigma^2}\right) \left(\frac{t^4}{\sigma^4} - \frac{6t^2}{\sigma^2} + 3\right) \quad (5.14)$$

The Fourier transform of the transmitted waveform $w_{tx}(t)$ is then given by

$$W_{tx}(f) = A(i2\pi f)^4 \exp\left(-\frac{(2\pi f\sigma)^2}{2}\right) \quad (5.15)$$

The pulse width parameter σ is set to 70 ps so that the pulse duration is kept within the T_c duration. The value of A is adjusted to ensure that the transmitted PSD does not exceed the FCC spectral mask [2]. The pulse waveform and its Fourier transform are shown in Figure 5.4. The TH code used is $c_j = \{0, 2, 1, 7\}$ and the transmitted PSD is shown in Figure 5.5. As can be seen, no contribution from the transmitted PSD falls at the frequency $f_{nb} = 900\text{MHz}$. If the GSM-1800 band is considered, the chip duration T_c can be simply changed to $T_c = 1/(2 \times 1800\text{MHz})$ and the other TH parameters are left unchanged.

Besides the null at 900 MHz, other PSD nulls also appear at the frequencies $\{2700, 5400, 7200, \dots\}$ MHz as shown in Figure 5.5. This means that by setting the chip duration $T_c = 0.555$ ns, the PSD is nullified at frequencies $\{900 + 1800\mathbb{N}\}$ MHz. This property is useful when the chip duration is required to follow a certain constraint. One possible scenario is explained as follows. Besides having a spectral null at the GSM-900 band, the design requires the chip duration to be larger than 100 ns. To satisfy both requirements, the following approach is considered. From the formula $T_c = 1/(2f_{nb})$, it can be deduced that $f_{nb} < 5$ MHz when $T_c > 100$ ns. Hence, if $f_{nb} = 2.4$ MHz, PSD nulls will appear at $\{2.4 + 4.8\mathbb{N}\}$ MHz. Notice that the frequency 900 MHz is also included. Therefore, both requirements are met.

5.3.2 Mutual Spectral Separation For Two Users

Now consider the case of two transmitters, one being the interferer and the other being the user-of-interest. For clarity, the interferer is termed user 1 and the user-of-interest

Figure 5.4: (a) Fourth derivative Gaussian pulse waveform, (b) Normalized absolute Fourier transform of the fourth derivative Gaussian pulse.

is user 2. To design the TH code for user 1, the TH code designed previously in (5.13) is utilized. Thus, user 1 will have no spectral contributions at k_{nb} . From here onwards, k_{nb} is denoted as k_2 . Because of the periodicity of $C_q(k)$, no spectral line contributions is observed within a frequency subset defined by $k_2 + \mathbb{N}N_fN_h$ or $f_2 + \frac{\mathbb{N}}{T_c}$. Within this frequency subset, the spectral line contributions of user 2 need to be maximized by designing its TH code. To do so, all the cosine terms in (5.10) have to be equal to 1. This can be achieved by setting the entire TH code to be comprised of only even numbers

$$c_j^{(2)} = \{2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N}\} \quad (5.16)$$

Figure 5.5: Normalized transmitted PSD of a UWB-IR TH signal utilizing the designed TH code. For the GSM-900 band, at 890-915 MHz, there is no contribution from the transmitted PSD.

At this state, the interferer (user 1), employing the TH code defined in (5.13), is spectrally separable from the user-of-interest (user 2), which employs the TH code defined in (5.16).

To achieve mutual spectral separation, an additional requirement is to find another frequency subset where the spectral line contributions of user 1 are maximized while those of user 2 are removed. This frequency subset is chosen at $k_1 + \mathbb{N}N_fN_h$ where $k_1 = \frac{1}{4}N_fN_h$. Thereafter, the following expression for $C_q(k_1)$ is obtained

$$C_q(k_1) = N_f + \sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos\left(\frac{\pi}{2}\rho_q\right) \quad j_1 \neq j_2 \quad (5.17)$$

The possible values of the cosine terms in the above expression are given by

$$\cos\left(\frac{\pi}{2}\rho_q\right) = \begin{cases} 1, & \rho_q \in 4\mathbb{N} \\ -1, & \rho_q \in 4\mathbb{N} + 2 \\ 0, & \rho_q \in 2\mathbb{N} + 1 \end{cases} \quad (5.18)$$

To nullify user 2, the TH code given in (5.16) is *refined* as follows

$$c_j^{(2)} = \{4\mathbb{N}, 4\mathbb{N}, 4\mathbb{N} + 2, 4\mathbb{N} + 2\} \quad (5.19)$$

Since the sets $4\mathbb{N}$ and $4\mathbb{N} + 2$ are subsets of $2\mathbb{N}$, the previous requirement for the TH code design of user 2 is still satisfied. This means that $C_2(k_2)$ is still maximized while $C_1(k_2)$ is nullified. Next, the TH code of user 1 given in (5.13) is also *refined* such that $C_1(k_1)$ is maximized. The resulting TH code is

$$c_j^{(1)} = \{4\mathbb{N}, 4\mathbb{N}, 2\mathbb{N} + 1, 2\mathbb{N} + 1\} \quad (5.20)$$

Notice that it is not possible to change the last two numbers to $4\mathbb{N}$, so as to have all cosine terms to be equal to 1, as it will invalidate the previous design ($4\mathbb{N}$ is not a subset of $2\mathbb{N} + 1$). At the current state, the mutual spectral separation has been achieved. Fig. 5.6 shows the sum of the code-dependent terms $C_q(k)$ as a function of the discrete frequency index k . At k_2 , user 1 is nullified while user 2 is maximized ($C_2(k_2) = N_f^2 = 16$). At k_1 , user 2 is nullified while user 1 is maximized ($C_1(k_1) = \frac{1}{2}N_f^2 = 8$).

Figure 5.6: Plot of the sum of the code-dependent terms $C_q(k)$ as a function of discrete frequency index k for two users. The TH codes employed are $\{5,5,0,4\}$ and $\{0,6,4,2\}$ for users 1 and 2, respectively. At $k_2 = 16$, $C_1(k_2)$ is zero and $C_2(k_2) = 16$. At $k_1 = 8$, $C_1(k_1) = 8$ while $C_2(k_1) = 0$.

5.3.3 Mutual Spectral Separation For Three Users

For the case of three users, the transmitted PSD of user 3 has to be nullified at the discrete frequencies $k_2 = \frac{1}{2}N_f N_h$ and $k_1 = \frac{1}{4}N_f N_h$. The resulting TH code design for user 3 can be expressed as

$$c_j^{(3)} = \{4\mathbb{N}, 4\mathbb{N} + 2, 2\mathbb{N} + 1, 2\mathbb{N} + 1\} \quad (5.21)$$

Hereafter, the design follow the same procedure as explained previously. Let k_3 denote the discrete frequency where user 3 is spectrally separable from users 1 and 2. The value of k_3 is chosen to be $\frac{1}{8}N_f N_h$ and the corresponding $C_q(k_3)$ can be expressed

as

$$C_q(k_3) = N_f + \sum_{j_1=0}^{N_f-1} \sum_{j_2=0}^{N_f-1} \cos\left(\frac{\pi}{4}\rho_q\right) \quad j_1 \neq j_2 \quad (5.22)$$

The possible values of the cosine terms in (5.22) are given by

$$\cos\left(\frac{\pi}{4}\rho_q\right) = \begin{cases} 1, & \rho_q \in 8\mathbb{N} \\ -1, & \rho_q \in 8\mathbb{N} + 4 \\ 0, & \rho_q \in 4\mathbb{N} + 2 \\ 1/\sqrt{2}, & \rho_q \in \{8\mathbb{N} + 1, 8\mathbb{N} + 7\} \\ -1/\sqrt{2}, & \rho_q \in \{8\mathbb{N} + 3, 8\mathbb{N} + 5\} \end{cases} \quad (5.23)$$

To nullify users 1 and 2 and to maximize user 3, the TH codes of user 3 given in (5.21) are *refined* as follows

$$\begin{aligned} c_j^{(1)} &= \{8\mathbb{N}, 8\mathbb{N} + 4, 4\mathbb{N} + 2, 4\mathbb{N} + 2\} \\ c_j^{(2)} &= \{8\mathbb{N}, 8\mathbb{N} + 4, 8\mathbb{N} + 1, 8\mathbb{N} + 5\} \\ c_j^{(3)} &= \{8\mathbb{N}, 4\mathbb{N} + 2, 8\mathbb{N} + 1, 8\mathbb{N} + 7\} \end{aligned} \quad (5.24)$$

Fig. 5.7 shows the sum of the code-dependent terms $C_q(k)$ as a function of discrete frequency index k for the three-user case. Similar to Fig. 5.6, it shows three indices k_1, k_2 and k_3 where only one of the users is not zero while the rest are null.

Table 5.1 summarizes the design process.

In the literature of TH-MA for UWB IR, [58] addresses the hopping sequence design that suppresses narrowband interference. The design technique used is to formulate a constrained optimization problem but the code length is significantly higher than that of the proposed design. Unlike in [58], the proposed design is based on a simple

Figure 5.7: Plot of the sum of the code-dependent terms $C_q(k)$ as a function of discrete frequency index k for the three-user case. The TH codes employed are $\{0,1,5,4\}$, $\{0,2,6,4\}$ and $\{0,1,7,2\}$ for users 1, 2 and 3, respectively. At $k_2 = 16$, $C_1(k_2) = C_3(k_2) = 0$ and $C_2(k_2) = 16$. At $k_1 = 8$, $C_1(k_1) = 8$ while $C_2(k_1) = C_3(k_1) = 0$. At $k_3 = 4$, $C_3(k_3) = 6.8284$ while $C_2(k_3) = C_1(k_3) = 0$.

manipulation of the PSD expression.

5.4 DOA Estimation in Time Hopping Multiple Access System

When Q users are active and transmit in a multipath environment, the total received signal can be seen as a summation of each user's received signal. The total received signal comprising of Q users' transmission can be modeled as

$$r(t) = \sum_{q=1}^Q \alpha_q s_q(t) + \eta(t) \quad (5.25)$$

User	Hopping Sequence	k	$C_q(k)$
User 1	$\{2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N}\}$	$\frac{1}{2}N_p N_h$	16
User 2	$\{2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N} + 1, 2\mathbb{N} + 1\}$		0
User 3	$\{2\mathbb{N}, 2\mathbb{N}, 2\mathbb{N} + 1, 2\mathbb{N} + 1\}$		0
User 1	$\{4\mathbb{N}, 4\mathbb{N}, 4\mathbb{N} + 2, 4\mathbb{N} + 2\}$	$\frac{1}{4}N_p N_h$	0
User 2	$\{4\mathbb{N}, 4\mathbb{N}, 2\mathbb{N} + 1, 2\mathbb{N} + 1\}$		8
User 3	$\{4\mathbb{N}, 4\mathbb{N} + 2, 2\mathbb{N} + 1, 2\mathbb{N} + 1\}$		0
User 1	$\{8\mathbb{N}, 8\mathbb{N} + 4, 4\mathbb{N} + 2, 4\mathbb{N} + 2\}$	$\frac{1}{8}N_p N_h$	0
User 2	$\{8\mathbb{N}, 8\mathbb{N} + 4, 8\mathbb{N} + 1, 8\mathbb{N} + 5\}$		0
User 3	$\{8\mathbb{N}, 4\mathbb{N} + 2, 8\mathbb{N} + 1, 8\mathbb{N} + 7\}$		6.8284

Table 5.1: A summary of the proposed design process for mutual-spectral-separation hopping sequences of 3 users.

where α_q is the path loss attenuation of q -th user.

$$s_q(t) = \sum_{j=-\infty}^{\infty} w_{mp}^{(q)}(t - jT_f - c_j^{(q)}T_c - \tau_{toa}^{(q)}) \quad (5.26)$$

$$w_{mp}^{(q)}(t) = \sqrt{E_s} \sum_{l=1}^{L_q} a_l^{(q)} w_l^{(q)}(t - \tau_l^{(q)}) \quad (5.27)$$

where each user's transmission goes through different paths.

When the receiver is an array (see Figure 5.8), the total received signal at the n -th element of the array is

$$r_n(t) = \sum_{q=1}^Q \alpha_q s_q^{(n)}(t) + \eta(t) \quad (5.28)$$

$$s_q^{(n)}(t) = \sum_{j=-\infty}^{\infty} w_{mp}^{(q,n)}(t - jT_f - c_j^{(q)}T_c - \tau_{toa}^{(q)}) \quad (5.29)$$

$$w_{mp}^{(q,n)}(t) = \sum_{l=1}^{L_q} a_l^{(q)} w_l^{(q)}(t - \tau_l^{(q)} - \tau_{l,n}^{(q)}) \quad (5.30)$$

where the transmitter's DOA $\theta_1^{(q)}$ is contained in the first-arriving multipath component's delay $\tau_{1,n}^{(q)}$. In this chapter, the DOA estimation of multiple transmitting users is addressed in two different ways: The first way is to assume any one specific user as

the user-of-interest and estimate its DOA in the presence of other users' transmission.

The second way is to estimate DOA of all transmitting users.

Figure 5.8: An illustration of multiple active users transmission in a multipath propagation environment. Each user goes through different paths. The receiver array is looking at multiple received signals from different path and different emitting sources.

In the course of developing the DOA estimation technique, it is assumed that:

- (i) The chip duration is long enough to let the multipath fades out within the same chip duration

$$T_p + \tau_{L_q} < T_c \quad \forall q \quad (5.31)$$

- (ii) The hopping sequence assigned to one user is unique to other active users. This means that within any chip duration, there will be only one pulse transmission.

5.4.1 With *Prior* Knowledge of The Hopping Sequences

Consider the case when more than one user is active and one of them is the user-of-interest. Given the knowledge of the hopping sequences of all active users, the DOA of the user-of-interest can be estimated accurately. The condition to be satisfied is that the user-of-interest is spectrally separable from other active users. When the TH codes of the active users are spectrally separable, the proposed channelization-based DF approach can be used to estimate the DOA of the user-of-interest. Thus, the line spectra, at the frequencies where the transmitted PSD from the user-of-interest is maximized while the MAI is nullified, can be extracted and fed to the channelization structure for DOA estimation.

Consider the three hopping sequences design presented in the previous section. First user's line spectra is spectrally separable at

$$f_1 = \left\{ \frac{k_1 + \mathbb{N}(N_p N_h)}{N_p T_f} \right\} = \left\{ \frac{2\mathbb{N} + 1}{2T_c} \right\} \quad (5.32)$$

Likewise, the second user's and third user's line spectra are spectrally separable respectively at

$$f_2 = \left\{ \frac{4\mathbb{N} + 1}{4T_c} \right\} \quad (5.33)$$

$$f_3 = \left\{ \frac{8\mathbb{N} + 1}{8T_c} \right\} \quad (5.34)$$

Depending on which user is the user-of-interest, the line spectra of that user can be extracted from the received PSD.

5.4.2 Unknown Hopping Sequences

When there is no *prior* knowledge of the hopping sequence, the DOA estimation formulation is no longer valid. Hence, a different estimation method is proposed in the following discussion.

Continuing from the expression of the sum of array channels' line spectrum in (2.28) for a single user's transmission, it can be re-expressed in the following way

$$Y(\omega_k + \Delta) = R_o(\omega_k + \Delta) \sum_{n=0}^{N-1} \exp(-i\omega_k \tau_{1,n}) \exp(-i2\pi \Delta \tau_{1,n}) \quad (5.35)$$

where $R_o(\omega_k + \Delta)$ is the $k + \Delta$ -th line spectrum of the rectangular pulse and Δ is an arbitrary value. In matrix form, a linear model is formed

$$\mathbf{y}(\Delta) = \mathbf{A}(\theta_1) \mathbf{e}(\Delta, \theta_1) \quad (5.36)$$

where

$$\mathbf{y}(\Delta) = \left[\begin{array}{ccc} \frac{Y(\omega_{k_1} + \Delta)}{R_o(\omega_{k_1} + \Delta)} & \dots & \frac{Y(\omega_{k_M} + \Delta)}{R_o(\omega_{k_M} + \Delta)} \end{array} \right]^T \quad (5.37)$$

$$\mathbf{e}(\Delta, \theta_1) = \left[\begin{array}{ccc} e^{-j2\pi \Delta \tau_{1,1}(\theta_1)} & \dots & e^{-j2\pi \Delta \tau_{1,N}(\theta_1)} \end{array} \right]^T \quad (5.38)$$

and the matrix $\mathbf{A}(\theta_1)$ has been defined in (2.33). For the case of multiple users' transmission, the expression now becomes

$$\mathbf{y}(\Delta) = \sum_{q=1}^Q \mathbf{A}(\theta_1^{(q)}) \mathbf{e}(\Delta, \theta_1^{(q)}) \quad (5.39)$$

which can also be linearized as

$$\mathbf{y}(\Delta) = \mathbf{\Phi} \mathbf{s}(\Delta) \quad (5.40)$$

$$\mathbf{\Phi} = \begin{bmatrix} \mathbf{A}(\theta_1^{(1)}) & \cdots & \mathbf{A}(\theta_1^{(Q)}) \end{bmatrix} \quad (5.41)$$

$$\mathbf{s}(\Delta) = \begin{bmatrix} \mathbf{e}^T(\Delta, \theta_1^{(1)}) & \cdots & \mathbf{e}^T(\Delta, \theta_1^{(Q)}) \end{bmatrix}^T \quad (5.42)$$

Notice that $\mathbf{y}(\Delta)$ is formed from the output of the M -th channel channelization structure. If the vector $\mathbf{y}(\Delta)$ is expanded into a matrix of $(P \times M)$ dimension, the following linear model can be formed

$$\mathbf{Y} = \mathbf{\Phi} \mathbf{S} \quad (5.43)$$

$$\mathbf{Y} = \begin{bmatrix} \mathbf{y}(\Delta_1) & \cdots & \mathbf{y}(\Delta_P) \end{bmatrix} \quad (5.44)$$

$$\mathbf{S} = \begin{bmatrix} \mathbf{s}(\Delta_1) & \cdots & \mathbf{s}(\Delta_P) \end{bmatrix} \quad (5.45)$$

The autocorrelation matrix of $\mathbf{Y}$ takes the form

$$\mathbf{P}_Y = E[\mathbf{Y}\mathbf{Y}^H] = \mathbf{\Phi} \underbrace{E[\mathbf{S}\mathbf{S}^H]}_{\mathbf{P}_S} \mathbf{\Phi}^H \quad (5.46)$$

This form is well-known in subspace-based DOA estimation. Therefore, $\{\widehat{\theta}_1^{(q)}; q = 1, \dots, Q\}$ can be obtained by doing a scan of the direction finding function

$$\widehat{\theta}_1^{(q)} = \arg \max_{\theta_s} \frac{1}{\|\mathbf{A}^H(\theta_s) \mathbf{E}_n\|^2} \quad (5.47)$$

where $\mathbf{E}_n$ is the matrix formed by $(M - Q)$ noise eigenvectors of the autocorrelation matrix $\mathbf{P}_Y$.

Alternatively, it is possible to re-use the time delay based direction finding technique discussed in Chapter 4. No changes or additional formulation are required to estimate the DOA of multiple active users. This is because the transmission of each user is confined to a single chip duration. In other words, the received signal at any

chip duration is due to the transmission from one active user. It can be shown that, by using numerical simulations, this technique estimates the DOA of all active users.

5.5 Simulation Results

The first simulation presented here considered the case of 2 users, with the hopping codes assigned as follows:

$$c_j^{(1)} = \{0, 6, 4, 2\} \quad (5.48)$$

$$c_j^{(2)} = \{5, 5, 0, 4\} \quad (5.49)$$

This assignment follows the design elaborated in Section 5.3.2. Since there are only 2 active users, the design requires 2 stages. And the line spectra of user 1 can be extracted from the total line spectra at f_1 and f_2 as given in (5.32) and (5.33), respectively. The other parameters for the simulation are listed in Table 5.2.

Parameter	Value
Baseband pulse	Gaussian 2 nd derivative
T_p	1 ns
Pulse 3-dB bandwidth	4 GHz
T_c	100 ns
$\theta_o^{(1)}$	30°
$\theta_o^{(2)}$	56°
$\theta_o^{(3)}$	65°
Array geometry	uniform linear array: $L = 7$ and $d = 50cm$
E_s/N_o	25 dB
Observation time	$N_f \times T_f = 3.2\mu s$
Number of channels M	15
f_o	50 MHz

Table 5.2: The simulation parameters used in Multiple Access DOA estimation.

Figures 5.9 and 5.10 show various plots of the received signal in the time domain and their PSD. It is shown clearly how line spectra of user 1 can be extracted from

the total PSD, and vice versa. The following simulation considers the case for 3 users and the hopping sequence assignment for the active users is given by

$$(5.50)$$

Figure 5.9: The received signal from (a) user 1 only, (b)user 2 only and (c) all users in AWGN channel.

Figure 5.10: The PSD of (a)user 1 only, (b) user 2 only and (c) the total PSD at f_1 and f_2 . The zeros PSD of user 1 is at f_2 and the zeros PSD of user 2 is at f_2 . Thus, the PSD of user 1 can be extracted from the total PSD at f_1 leaving out f_2 and vice versa.

5.5.1 Effects of Non-Ideality from Multipath Propagation

Next, the effectiveness of the estimator in a multipath propagation environment with 2 active users is investigated. The estimator aims to estimate one of the active users while the other user is regarded as MAI. The multipath propagation environment requires the implementation of LTD at the RF front-end. This may result in erroneous TOA estimates. It will be shown that the estimator is not affected by the error in the TOA estimates.

The simulation begins by generating a single realization of multiple-user transmission. The resulting TOA estimates are listed in Table 5.3. These estimates are obtained from the LTD operation at the chip duration that contains the transmitted pulse, which may be from either user 1 or user 2. To estimate the DOA of user 1, the frequencies of the channelization structure are chosen from the frequencies that are a subset of f_1 . In this realization, a $M = 15$ -channel structure is implemented for the channelization. These frequencies are given in Table 5.4. Next, the results of the estimation are compared with the ideal case when the TOA estimates are the true TOA. This is also the case for user 2.

TOA	$\tau_{0,0}$	$\tau_{0,1}$	$\tau_{0,2}$	$\tau_{0,3}$	$\tau_{0,4}$	$\tau_{0,5}$	$\tau_{0,6}$	Frame	Chip
Observed	0	5.333	2.895	4.340	5.763	7.201	8.673	1	1
True	0	1.443	2.886	4.330	5.773	7.216	8.660	1	1
Observed	250	250.888	251.833	252.729	253.659	254.611	255.541	1	6
True	250	250.932	251.864	252.796	253.728	254.659	255.591	1	6
Observed	650	650.937	651.826	652.791	653.736	654.645	655.583	2	6
True	650	650.932	651.864	652.796	653.728	654.659	655.591	2	6
Observed	700	701.347	698.902	700.347	701.805	703.208	704.666	2	7
True	700	701.443	702.886	704.330	705.773	707.216	708.660	2	7
Observed	800	800.930	801.847	802.806	803.694	804.625	805.562	3	1
True	800	800.932	801.864	802.796	803.728	804.659	805.591	3	1
Observed	1000	997.409	998.916	1000.819	1001.770	1006.986	1004.631	3	5
True	1000	1001.443	1002.886	1004.330	1005.773	1007.216	1008.660	3	5
Observed	1300	1301.395	1306.701	1304.291	1305.729	1310.944	1308.618	4	3
True	1300	1301.443	1302.886	1304.330	1305.773	1307.216	1308.660	4	3
Observed	1400	1400.930	1401.840	1402.805	1403.708	1404.652	1405.569	4	5
True	1400	1400.932	1401.864	1402.796	1403.728	1404.659	1405.591	4	5

Table 5.3: The observed and true values of $\tau_{0,n}$ in ns.

Frequency [GHz]	User 1	User 2
f_1	0.01	0.005
f_2	0.37	0.365
f_3	0.43	0.425
f_4	0.71	0.625
f_5	0.83	0.985
f_6	1.03	1.105
f_7	1.37	1.305
f_8	1.43	1.505
f_9	1.77	1.765
f_{10}	1.83	1.965
f_{11}	2.09	2.165
f_{12}	2.29	2.225
f_{13}	2.57	2.425
f_{14}	2.77	2.785
f_{15}	2.83	2.825

Table 5.4: The frequencies used in the channelization structure to estimate DOA of user 1 in the presence of user 2, and vice versa.

The normalized cost function (formulated in (2.43)) for the ideal case is shown in Figure 5.11. The level of the noise floor surrounding the peak is much lower in the ideal case, as compared to that for the non-ideal case, shown in Figure 5.12. This increased noise floor is due to the erroneous TOA estimates or jitter. The jitter smoothes the total PSD observed at the receiver [55]. This causes the PSD from user 2 to be present at k_1 frequencies.

Likewise, similar results are obtained when the channelizer's frequencies are chosen from a subset of f_2 in order to estimate the DOA of user 2 (shown in Figures 5.13 and 5.14). These results verify that by carefully choosing the frequency set for channelization, the estimator is still effective in the presence of multiple access interference. Also, the presence of the erroneous TOA estimates does not cause the estimator to fail. Instead, only the noise floor level of the cost function is increased while the peak is still observable.

Figure 5.11: The normalized cost function plot of user 1 from ideal TOA estimation in the presence of user 2. The channelization of the array system is configured such that only the PSD from user 1 is extracted.

Figure 5.12: The normalized cost function plot of user 1 from non-ideal TOA estimation (erroneous estimation) in the presence of user 2. Although the surrounding noise floor increases, the peak showing the DOA is still observable.

Figure 5.13: The normalized cost function plot of user 2 from ideal TOA estimation in the presence of user 1. The channelization of the array system is configured such that only the PSD from user 2 is extracted.

Figure 5.14: The normalized cost function plot of user 2 from non-ideal TOA estimation (erroneous estimation) in the presence of user 1. Although the surrounding noise floor increases, the peak showing the DOA is still observable. The presence of user 1 and erroneous TOA estimation does not render the estimator to be ineffective.

5.5.2 Effects of decreasing $C_q(k)$

Now, consider 3 active users transmission. Although the hopping sequence assignment ensures the PSD to be separable from one another, the resulting $C_q(k)$ is not maximized. Recall that the maximum value of $C_q(k)$ is N_f^2 . As shown in Table 5.1, $C_2(k_2) = 8$ and $C_3(k_3) = 6.8482$. As the number of active users that are required to be spectrally separable increase, $C_q(k)$ decreases. This also affects the noise floor level as demonstrated in the following discussion.

The simulation begins by generating a single realization of 3 active users transmission. In this realization, the simulation assumes that the LTDs successfully detect the direct path and the TOAs estimated are the true TOAs. The channelization frequency settings are given in Table 5.5 and the simulation results are shown in Figures 5.15-5.17. The results show that as $C_q(k)$ reduces, the surrounding noise floor increases. To ensure good spectral separation, the hopping sequence needs to be optimally designed.

No	User 1's Freq	User 2's Freq	User 3's Freq
1	0.0100GHz	0.0050GHz	0.0225GHz
2	0.2300GHz	0.3050GHz	0.2425GHz
3	0.5100GHz	0.5050GHz	0.5625GHz
4	0.6900GHz	0.7050GHz	0.7025GHz
5	0.9700GHz	0.9050GHz	0.8425GHz
6	1.0300GHz	1.1050GHz	1.0425GHz
7	1.2900GHz	1.3650GHz	1.3625GHz
8	1.4300GHz	1.4250GHz	1.4225GHz
9	1.7700GHz	1.7050GHz	1.6225GHz
10	1.9100GHz	1.9050GHz	1.9625GHz
11	2.1700GHz	2.1050GHz	2.1025GHz
12	2.2300GHz	2.3650GHz	2.2225GHz
13	2.5100GHz	2.5650GHz	2.4225GHz
14	2.6900GHz	2.7650GHz	2.7625GHz
15	2.9700GHz	2.8250GHz	2.8225GHz

Table 5.5: The frequencies set used for the channelization structure to estimate DOA of users 1, 2 and 3 in the presence of other active users.

Figure 5.15: The normalized cost function as a function of scanning direction. The system considers 3 active transmitting users and ideal TOA estimation. The first user is the user-of-interest while the rest are multiple access interferences. The designed hopping sequence results in maximum spectral separation ($C_1(k_1) = N_f^2 = 16$).

Figure 5.16: The normalized cost function as a function of scanning direction. The system considers 3 transmitting active users and ideal TOA estimation. The second user is the user-of-interest while the rest are multiple access interferences. The designed hopping sequence results in $C_2(k_2) = 8$.

Figure 5.17: The normalized cost function as a function of scanning direction. The system considers 3 transmitting active users and ideal TOA estimation. The third user is the user-of-interest while the rest are multiple access interferences. The designed hopping sequence results in $C_2(k_2) = 6.8482$.

5.5.3 Unknown Hopping Sequences

In the case when no *prior* knowledge is available at the receiver array, the technique described in Section 5.4.2 is utilized. The simulation considers a single realization of 3 active users, transmitting from $\{\theta_o^{(q)}\} = \{30^\circ, 56^\circ, 60^\circ\}$. The required channels for the channelization structure is $(M \times P)$ where we use $M = 10$ and $P = 9$ in this realization. The frequency settings are assumed to be arbitrary but follows the signal model described in Section 5.4.2. The transmission considers $T_f = T_c$ and each user transmits within a chip duration. The sum of the LTD's output signal at the array is plotted in Figure 5.18 and the normalized cost function plot is given in Figure 5.19. Each peak indicates one DOA of the active users.

Figure 5.18: The sum of the LTD's output signal at the array when three active users are transmitting. The hopping sequences governing the transmission are arbitrarily defined. The array is a ULA with $d = 1.2\text{m}$ and $N = 3$ elements. The chip duration is $T_c = 60\text{ ns}$.

Figure 5.19: The normalized cost function plot as a function of scanning direction. Three peaks are observed, where each peak indicates one DOA of the active users. The correlation matrix is formed with $M = 10$ and $P = 9$.

The number of active users is increased in the following simulation. 4 active users transmitting from $\{\theta_o^{(q)}\} = \{30^\circ, 45^\circ, 25^\circ, 60^\circ\}$, are considered. The method uses $M = 13$ and $P = 12$. The results are plotted in Figures 5.20 and 5.21. Both plots show that the proposed method successfully estimates all four directions, as indicated by four peaks observed at the normalized cost function plot shown in Figure 5.21.

Figure 5.20: The sum of the LTD's output signal at the array when four active users are transmitting. User 1 transmits in the first chip duration, followed in sequence by user 2, 3 and 4. The array is a ULA with $d = 1.2\text{m}$ and $N = 3$. The chip duration is $T_c = 60\text{ ns}$.

Figure 5.21: The normalized cost function plot as a function of scanning direction. Four peaks are observed, where each peak indicates one DOA of the active users. The correlation matrix is formed with $M = 13$ and $P = 12$.

Lastly, the performance of the time delay estimation based DF presented in Chapter 4 is evaluated in a multiple access, multipath propagation environment. The simulation parameters follow directly from the list in Table 5.2. Two users are considered in this simulation where $\{\theta_o^{(q)}\} = \{30^\circ, 56^\circ\}$. The transmission is arranged such that users 1 and 2 are transmitting at the chip durations with odd and even indices, respectively. Figure 5.22 shows the normalized cost function obtained from observing one chip duration due to the transmission from user 1 and Figure 5.23 shows the function obtained from observing the next chip duration due to user 2.

Figure 5.22: Normalized cost function plot of user 1 as a function of scanning direction. It is generated from the observation at the first chip duration. It shows the minimum point at the DOA of user 1.

Figure 5.23: Normalized cost function plot of user 2 as a function of scanning direction. It is generated from the observation at the second chip duration. The minimum point indicates the DOA of user 2.

Hereafter, this simulation is extended to a Monte Carlo simulation with 1000 realizations. Figures 5.24 and 5.25 show the pdf and the cdf, respectively. From these figures, it can be concluded that the technique is also reliable in a multiple access, multipath propagation environment.

5.6 Conclusion

The extension of the DF technique in TH-MA transmission for UWB IR system is considered in this chapter. Two different ways of looking into the DF problem are considered. For the first problem, the discussion considers the estimation of the user-of-interest DOA in the presence of multiple access interference. For the second problem, the discussion considers the estimation of all the active users' DOA. The hopping sequence assignment in the first problem is required to achieve spectral separation. This means that the PSD of the user-of-interest can be obtained from the total PSD. It has been shown that this is possible when the hopping sequence of the user-of-interest is properly designed. Examples of the hopping sequence design that satisfies the spectral separation requirement are also given. In the case when 3 active users are present and each user employs the hopping sequence from the design, the total PSD at frequencies $\{f_q\}$ contains only the PSD of the q -th user. The channelization structure can then be configured accordingly to estimate the DOA of the user-of-interest.

To solve the second problem, a new DF formulation is derived. This formulation resembles the conventional subspace-based approach. The difference is that this formulation operates in the frequency domain, rather than the time domain. To ensure good DOA estimation, the system is required to have a huge channelization structure to estimate the correlation matrix $\mathbf{P}_y$. Both problems consider the implementation of the LTD at the RF front-end in a multipath propagation environment. Alternatively,

Figure 5.24: The pdf plot generated from the Monte Carlo simulation of 1000 realizations in a multipath propagation environment. The system considers two users transmission at 30° and 56° .

Figure 5.25: The cdf plot generated from the Monte Carlo simulation of 1000 realizations in a multipath propagation environment. The system considers two users transmission at 30° and 56° .

the time delay based DF approach described in Chapter 4 can also be used in a multiple access environment. This is because at any chip duration interval, the received signal is coming from the transmission of one active user only.

Simulation results show that when the LTD does not provide good TOA estimates, the estimator still shows good DOA estimation. The condition, however, is that the error in the TOA estimates from the LTD should be small enough. This small variation is jitter and it smoothes the total PSD.

For the second problem, the simulation considers no TOA estimation error from the LTD. The simulation result shows that the technique successfully shows multiple peaks at the direction of the transmitting users. Up to 4 active users have been simulated and the technique has proven to be effective. The time delay estimation based direction finding is also extended to the case of multiple users transmission. Monte Carlo simulation show that the robustness of this technique is not affected by the multiple access interference.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

This thesis discusses DOA estimation of a UWB IR system with feasible approaches targeted for low data-rate localization. By using an antenna array for DOA estimation, the cost is increased. Therefore, the main concern, besides the ability to provide good DOA estimate, is to maintain a low complexity of the receiver array.

6.1.1 Channelization Based Direction Finding

The studies presented in this thesis look into the possibility of removing the Nyquist rate sampling requirement prior to any signal processing introduced. In turn, this will greatly reduce the complexity of the receiver. For illustration, a UWB signal covering a spectrum of 4 GHz wide will need a minimum of 8 GHz sampling rate. Not only it is not feasible, but the receiver's cost will also be an issue.

One possible solution to avoid Nyquist rate sampling is to utilize the idea of the channelized UWB receiver proposed in [22]. However, implementing the channelization in every antenna will increase the complexity. Instead, only the sum of the

received signals at the array is channelized. To estimate the DOA based on the channelization's output, the proposed method is based on a simple least squares scanning approach. From the derivation, the approach is motivated by the fact that the line spectra of the sum of the received signals at the array contain the directional information. These line spectra are within the UWB spectrum but not required to cover the whole spectrum. This will then simplify the channelization structure.

In a multipath propagation environment, the array sum observes multiple directions from both the direct path and the multipath. To suppress the multipath effects, each channel of the array implements a simple analog LTD with a latch circuitry. The combination of the LTD and the latch is an analog solution for the multipath problem.

6.1.2 Analog Differentiation Based Direction Finding

The output of the latch circuitry is in the form of a rectangular pulse. The pulse duration depends on the latch duration. When the pulse duration is larger than the propagation delay between the two ends of the array, the resulting sum of the LTD's output signals at the array is a staircase-shape waveform whose slope is inversely proportional to the DOA. This observation has motivated us to use the analog differentiation to extract the slope of the staircase-shape waveform. The DOA is then estimated from the slope. The proposed estimator is simple and requires minimum processing.

6.1.3 Time Delay Based Direction Finding

When the noise level increases, the detector may falsely identify the noise as the UWB pulse. Statistical simulations suggest that this is highly probable. Because the previous DOA estimation techniques are derived based on the sum of the LTD's output

signals at the array, a single false alarm error will cause the array system to fail. To overcome this, a time-delay based DF technique is proposed. The technique is based on the fractional norm formulation to reduce the impact of huge time delay estimation errors due to false alarm. In addition to the fractional norm metric, the technique also incorporates an algorithm to ensure the reliability of the DOA estimates. Through numerical analysis, the parameter settings of the algorithm and the threshold level used for the LTD has been shown to be critical to achieve good DF performance.

The erroneous time delay estimate is reflected in the rise time of the rectangular pulse at the output of the latch. To obtain the rise time with low complexity, a simple structure comprising of an integrate-and-dump operation followed by a sampler is proposed. The sampling interval required is equal to the integration interval, which is half the frame rate. Theoretically, angular resolution of time delay based DF is determined by the time delay resolution which is constrained by the sampling rate. This approach removes the inter-dependency of the angular resolution and the sampling rate because the time delay resolution is determined by the quantization of the integration output. Therefore, to achieve good angular resolution, it is possible to use very low sampling rate but required to have a high dynamic range ADC.

6.1.4 Time Hopping Multiple Access

In the presence of multiple access interference, the previously proposed time delay based DF technique can be utilized under certain assumptions. That is, within any chip duration, there is only one user transmitting. The received multipath spreads within the same chip duration. As a result, the time delay based DF will not have any problem in dealing with the multiple users transmission under such assumptions. Note that the MA scheme considered in this thesis is the TH MA scheme.

The channelization based DF is expected to fail even if the multipath is suppressed. This is because the spectrum of the sum of the received signals at the array contains all the DOA information of all transmitting users. Unless they are separated, it is impossible to estimate the DOA of any user. In view of this problem, a technique that alter the TH codes of all transmitting users is proposed to separate the spectrum of one user from the total spectrum. The alteration is motivated by the observation on the power spectral density (PSD) expression of a TH format signal as derived in [55]. The PSD expression is composed of line spectra as a function of the hopping sequence and the pulse waveform. The idea is to design a proper TH code such that at certain frequencies, the multiple access interference has no contribution. It has been shown that such a design is possible and a design technique is also proposed.

6.2 Recommendations for Further Research

Below is the list of the possible extensions of the work presented in this thesis.

- The level threshold detector is able to provide good UWB pulse detection only when the noise level is below the leading edge of the pulse. A better detection technique can be considered to extend the operational SNR of the proposed direction finder.
- The use of the envelope detector in the LTD structure to detect UWB pulse in the presence of dominant jammer can be further investigated. Different signal models to represent the dominant jammer can be considered. The robustness of the method needs to be addressed as well.
- The multipath suppression capability of the proposed threshold-latch structure has to be re-addressed in the presence of inter-symbol interference.

- The time delay estimation error resulting from the threshold-latch structure has a heavy-tailed distribution characteristics. Statistical modeling of this distribution can be considered to address the performance of the proposed time delay based DF. A comparison with the statistical lower bound performance will then be possible. A statistically optimum direction finder may also be available from this future research effort.
- To ensure the workability of the proposed technique, a proof-of-concept prototype can be built.

Appendix A: Derivation of $C_q(f)$

The definition of $C_q(f)$, given in (5.4), can be expressed as

$$C_q(f) = \left| \sum_{j=0}^{N_f-1} \exp[-i \underbrace{2\pi f(jT_f + c_j^{(q)}T_c)}_{\varphi_j}] \right|^2 \quad (6.1)$$

The exponent term inside the sum can be expanded using Euler's formula and further simplified as follows

$$C_q(f) = \left| \sum_j \{\cos(\varphi_j) + i \sin(\varphi_j)\} \right|^2 \quad (6.2)$$

$$C_q(f) = \left\{ \sum_j \cos(\varphi_j) \right\}^2 + \left\{ \sum_j \sin(\varphi_j) \right\}^2 \quad (6.3)$$

$$\begin{aligned} C_q(f) &= \sum_j \{\cos^2(\varphi_j) + \sin^2(\varphi_j)\} \\ &\quad + \sum_{j_1} \sum_{j_2} \{\cos(\varphi_{j_1}) \cos(\varphi_{j_2}) + \sin(\varphi_{j_1}) \sin(\varphi_{j_2})\} \\ &\quad j_1 \neq j_2 \end{aligned} \quad (6.4)$$

Utilizing the following two trigonometry identities

$$\begin{aligned} \cos^2(A) + \sin^2(A) &= 1 \\ \cos(A_1) \cos(A_2) + \sin(A_1) \sin(A_2) &= \cos(A_1 - A_2) \end{aligned} \quad (6.5)$$

we can now expressed $C_q(f)$ as

$$C_q(f) = N_f + \sum_{j_1} \sum_{j_2} \cos(\varphi_{j_1} - \varphi_{j_2}) \quad (6.6)$$

Using the definition of ρ_q in (5.7), we have

$$\varphi_{j_1} - \varphi_{j_2} = 2\pi f \rho_q T_c \quad (6.7)$$

With the above two equations, we will arrive at the simplified expression of $C_q(f)$ in (5.6).

Author's Publications

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