Robust ultra wideband time of arrival estimation for indoor localization applications

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**NANYANG
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UNIVERSITY**

**ROBUST ULTRA WIDEBAND TIME OF ARRIVAL
ESTIMATION FOR INDOOR LOCALIZATION
APPLICATIONS**

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SCHOOL OF ELECTRICAL AND ELECTRONIC ENGINEERING

2010

ROBUST ULTRA WIDEBAND TIME OF ARRIVAL ESTIMATION FOR INDOOR LOCALIZATION APPLICATIONS

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School of Electrical and Electronic Engineering

A thesis submitted to the Nanyang Technological University

in partial fulfillment of the requirement for the degree of

Doctor of Philosophy

2010

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

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Date

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Xu Chi

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To them I dedicate this thesis.

List of Abbreviations

ADC	analog-to-digital converter
AOA	angle of arrival
ASK	amplitude-shift keying
AWGN	additive white Gaussian noise
BPF	bandpass filter
BPSK	binary phase shift keying
CDF	cumulative distribution function
CIR	channel impulse response
CNR	clutter-noise region
DA	data-aided
DD	delay-dependent
DI	delay-independent

DP	dual pulse
DSO	digital sampling oscilloscope
ECC	European Commission Committee
ED	energy detection
FCC	Federal Communication Commission
FT	frame timing
GML	generalized maximum likelihood
GPS	global positioning system
IFI	inter-frame interference
IPI	inter-pulse interference
IR	impulse radio
ISI	inter-symbol interference
ITS	intelligent transportation system
LNA	low noise amplifier
LOS	line-of-sight
LS	least square

I&D	integrate-and-dump
MAE	mean absolute error
MCU	microcontroller unit
ML	maximum likelihood
MRO	maintenance, repair and overhaul
NBI	narrowband interference
NDA	non-data-aided
NLLS	nonlinear least square
NLOS	non-line-of-sight
NR	noise-only region
OWR	one-way ranging
PAM	pulse amplitude modulation
PDF	probability density function
PN	pseudo- noise
BPN	binary pseudo-noise
PPM	pulse position modulation

PRF	pulse repetition frequency
PSD	power spectral density
PWTC	positioning & wireless technology center
RF	radio frequency
RFID	radio frequency identification
RMS	root mean square
RMSE	root mean square error
RSSI	received signal strength indicator
Rx	receiver
SCR	signal to clutter ratio
SNR	signal to noise ratio
SR	signal region
TDOA	time difference of arrival
TDT	timing with dirty template
TOA	time of arrival estimation
TR	transmitted reference

TWR	two-way ranging
Tx	transmitter
Tx-Rx	transmitter-to-receiver
UWB	ultra wideband
WBI	wideband interference
WPAN	wireless personal area network
2D	two dimensional
2-PAM	2- pulse amplitude modulation

List of Symbols

a	design parameter for delay dependent threshold setting
$\{a_j\}$	binary bipolar sequence
b	design parameter for delay dependent threshold setting
$\{b_j\}$	binary bipolar sequence
b_{opt}	optimum value of threshold design parameter b
c	velocity of light
$\{c_{BPN_j}\}$	binary PN sequence
$\{c_j\}$	binary bipolar PN sequence
$C_{m,n}(u, v)$	current data bit modulating the front part of roundtrip tag response
$C_{m,n+1}(u, v)$	next data bit modulating the tail of roundtrip tag response
$\{d_j\}$	binary bipolar ranging sequence

D	degree of freedom of non-central Chi-square distribution
E_s	symbol energy
E_{tx}	energy of transmitted UWB pulse
f_s	sampling frequency
$F(\varepsilon)$	CDF function of ranging error
$\mathcal{F}_{pdp}^l$	power delay profile channel parameter set
$\mathcal{F}_{pl}^l$	path loss channel parameter set
$\mathcal{F}_{ss}^l$	small-scale fading channel parameter set
$g_{i,j}(t)$	overall roundtrip tag response to the UWB pulse $p_{tx}(t)$ transmitted at $t = iN_f T_f + jT_f$
$h(t)$	channel response
$h_d(t)$	overall downlink channel response to transmitted UWB pulse $p_{tx}(t)$
$h_{CIR}(t)$	overall channel impulse response
$h_{CIR_i}(t)$	channel impulse response of the i^{th} path
$I_l(x)$	the l^{th} order modified Bessel function of the first kind

$\mathbf{J}(\boldsymbol{\theta}_k)$	Jacobian matrix evaluated at $\boldsymbol{\theta}_k$
L	direct path propagation distance
L_c	number of clutter paths
L_d	number of paths in the downlink channel of backscattering RFID system
$L_{\max}$	maximum direct path propagation distance
L_{mean}	mean of a set of distances
L_d	number of paths in the downlink channel
L_u	number of paths in the uplink channel
L_{ref}	reference distance
L_0	calibration distance for path loss model
$\hat{L}$	estimated direct path propagation distance
$\hat{L}_{\text{ref}}$	estimated reference distance
N_a	code period of sequence $\{a_i\}$
N_{ch}	total number of channel realizations
N_{CNR}	number of samples within T_{CNR} duration in the clutter-noise region

N_d	code period of $\{d_j\}$
N_f	number of frames per symbol
N_s	number of symbols
N_{tot}	number of measured profiles used to generate RMSE
N_0	single-sided power spectral density
m_{nak}	small-scale fading parameter for Nakagami distribution
$p_d^{(l)}$	channel response of the l^{th} path to the transmitted UWB pulse $p_{tx}(t)$
$p_f(t)$	transmitted frame waveform with normalized energy
$p_i(t)$	the received pulse from i^{th} path with normalized energy.
$p_q^{(l)}(t)$	clutter channel response of the l^{th} path to $p_{tx}(t)$
$p_{tm}(t)$	pulse template
$p_{tx}(t)$	transmitted UWB pulse
$p_{u0}^{(l)}(t)$	uplink channel response of the l^{th} path to $\vartheta_0(t)$
$p_{u1}^{(l)}(t)$	uplink channel response of the l^{th} path to $\vartheta_1(t)$
$P(x)$	probability of event x

P_{FA}	false alarm probability
P_{loss}	path loss in the unit of dB
q	downsampling factor
$q(t)$	overall clutter channel response to transmitted UWB pulse $p_{tx}(t)$
q_j	the conditional false alarm probability when the j^{th} sample is the first sample containing the direct path sample
$\mathbf{q}$	sample vector of $q(t)$ within time interval $[0, T_f]$
$\tilde{\mathbf{q}}$	candidate value of $\mathbf{q}$
$Q_x(y_1, y_2)$	generalized Marcum's Q function
$r(t)$	received signal
$r_{jit}(t)$	received signal with jitter taken into account
$r_{tg}(t)$	noise-free signal input to the modulator of the RFID backscattering tag
$\mathbf{r}_{m,n}$	sample vector of $r(t)$ within time interval $[mN_fT_f + nT_f, mN_fT_f + (n+1)T_f]$
$R_{pp}(t)$	autocorrelation function of pulse template $p_{tm}(t)$

$R_{rp}(t)$	cross-correlation function between pulse template $p_{tm}(t)$ and received signal $r(t)$
$s(t)$	transmitted signal
$s_{jit}(t)$	transmitted signal with jitter taken into account
$s_{tg}(t)$	noise-free output of 2-PAM modulator
S	shadowing term for path loss channel model
S_{neg}	data set containing only negative ranging error
S_{sym}	data set constructed with the elements from S_{neg} and the inverse of them
t_{sta}	transmitting timing of Tx signal
t_{max}	arrival timing of largest path among the paths in $\mathbf{x}$
$\mathbf{t}$	vector of estimated path delay
t_{cal_ref}	arrival time of calibration signal in the reference measurement
t'_{cal_ref}	arrival time of calibration signal in the reference measurement with jitter taken into account
t_{toa}	time of arrival

$\hat{t}_{toa}$	estimated time of arrival
t'_{toa}	TOA of the signal with jitter taken into account
t'_{cal}	TOA of the calibration signal with jitter taken into account
t_{cal}	arrival time of the calibration signal in the absence of jitter
t_{toa_ref}	TOA of the signal in the reference measurement
t'_{toa_ref}	TOA of the signal in the reference measurement with jitter taken into account
T	integration interval of energy detection receiver
T_c	searching step size to adjust ζ in the first stage of a two-stage TOA estimation algorithm
T_{CNR}	expected duration of clutter-noise region
T_d	delay factor of the RFID tag with DP modulation
T_{ds}	presumed maximum channel delay spread used in TOA estimation
$T_{ds}^{(true)}$	true maximum channel delay spread
T_D	delay between the reference pulse and its following pulse in the signal

	with DP modulation
T_f	frame duration
T_g	maximum clutter duration
$T_{\max}$	maximum TOA
T_s	symbol duration
T_{tm}	Duration of the pulse template $p_{tm}(t)$
T_w	window length
T_0	integration time of the integrator in the first stage of a two-stage TOA estimation algorithm
T_1	integration time that is set in the second stage of a two-stage TOA estimation algorithm
u	PN code offset between RFID reader and tag
$\tilde{u}$	candidate value of u
$\mathcal{U}(x, y)$	uniform distribution with support on (x, y)
v	ranging symbol offset between RFID reader and tag

$\tilde{v}$	candidate value of v
$w(t)$	noise process
$\mathbf{w}_{m,n}$	sample vector of $w(t)$ within time interval $[mN_f T_f + nT_f, mN_f T_f + (n+1)T_f]$
W	signal bandwidth
$W(L)$	distance-dependent weighting factor for ranging error model
$x_{\max}$	amplitude of the largest path among the paths in $\mathbf{x}$
$x(t)$	overall uplink channel response to $\vartheta_0(t)$
$\mathbf{x}$	vector of estimated path amplitude
$y(t)$	overall uplink channel response to $\vartheta_1(t)$
$y_{atc}(t)$	autocorrelation output
$\mathbf{z}$	sample vector
$z[k]$	output sample of the receiver
z_w	arbitrary energy sample containing only noise component

α_i	instantaneous amplitude gain of the i^{th} path
$\mathcal{A}$	a constant for delay independent threshold setting
β	variable used in the adaptive threshold computation
$\beta_{i,j}$	jitter value of the signal in the $(iN_f + j)^{th}$ frame
$\beta_{0,0_ref}$	jitter value of the signal of the zero th frame in the reference measurement
γ	threshold
$\Gamma(\bullet)$	Gamma Function
$\gamma[k]$	threshold value associated with the k^{th} sample
δ	search duration prior to the direct path
$\delta_{i,j}$	relative jitter value defined by $\delta_{i,j} = \beta_{i,j} - \beta_{0,0}$
δ_{ns}	variable used set the value of design parameter a for delay dependent threshold
$\hat{\delta}_{i,j}$	estimated relative jitter value

ε	ranging error
ε_i	ranging error of the i^{th} profile
ζ	time factor introduced to control the transmitting time of Tx signal
$\eta(t)$	clutter channel response to $s(t)$
ϑ	factor for multi-trip distance computation
$\vartheta_0(t)$	front part of channel response $h_d(t)$ when the transition of two consecutive data bits falls within the time span of channel response $h_d(t)$
$\vartheta_1(t)$	tail of channel response $h_d(t)$ when the transition of two consecutive data bits falls within the time span of channel response $h_d(t)$
κ	kurtosis value
λ	mismatch between $h_d(t)$ and the opening window of the amplifier's enabling signal in the backscattering tag with DP modulation.
λ_k	noncentral parameter of non-central Chi-square distribution

$\lambda(L)$	distance-dependent exponential distribution factor for ranging error models
μ	mean of mixture distribution
μ_G	mean of Gaussian distribution
μ_{noise}	mean of noise samples
ξ_{floor}	computed noise floor
ξ_i	the polarity of the i^{th} path
ρ	the timing offset between the clock of RFID reader and tag
ρ_{pl}	path loss exponent
σ^2	variance of mixture distribution
σ_{CNR}	standard deviation of the samples in clutter-noise region
σ_G	standard deviation of Gaussian distribution
σ_{noise}	standard deviation of noise-only samples
σ_s	standard deviation of shadowing term S
σ_w^2	variance of noise process $w(t)$

τ_d	propagation delay of the direct path of the downlink channel
$\tau_d^{(l)}$	propagation delay associated with $p_d^{(l)}$
$\tau_q^{(l)}$	propagation delay of the l^{th} path in the clutter channel
τ_{tg}	total processing delay of the tag
τ_{tg}'	system processing delay from the tag's antenna to the modulator
τ_{tg}''	processing delay from modulator to the tag's antenna
τ_i	propagation delay of the i^{th} path
$\tau_{\max}$	maximum propagation delay of the direct path
τ_{sys}	delay introduced by Tx and Rx systems
$\tau_{u0}^{(l)}$	relative propagation delay of the l^{th} path associated with pulse $p_{u0}^{(l)}(t)$
$\tau_{u1}^{(l)}$	relative propagation delay of the l^{th} path associated with pulse $p_{u1}^{(l)}(t)$
τ_u	the propagation delay of the direct path in the uplink channel
$\varphi_{i,j}(t)$	received roundtrip tag response to the initial transmitted single UWB pulse $p_{tx}(t)$

$\Phi(t)$	modulation function consisting of a train of rectangular windows
$\Phi_{upper}(t)$	modulation of the modulator in the upper branch of the backscattering tag with DP modulation
$\Phi_{lower}(t)$	modulation of the modulator in the lower branch of the backscattering tag with DP modulation
$\omega(t)$	calibration signal
$\Omega(t)$	rectangular window with amplitude being one in $[0,1]$ and zero elsewhere
Ω_i	spatial averaged power gain of the i^{th} path
$\lceil \bullet \rceil$	ceiling operation
$ x _y$	modular operation $x \bmod y$
$\lfloor \bullet \rfloor$	integer floor operation
Δ	sampling interval
$\mathbf{0}_{x \times 1}$	represents a column vector of which the number of elements is x and all elements are zeros
$\mathbb{E}[\bullet]$	expectation operation

$\mathbb{E}_{ssa} [\cdot]$ operation of averaging the small-scale fading effect over a local area

Contents

Acknowledgements.....	ii
List of Abbreviations	iii
List of Symbols	viii
Contents	xxiii
List of Figures.....	xxvii
List of Tables	xxxii
Summary	xxxiii
Chapter 1 Introduction.....	1
1.1 Background Information	1
1.1.1 Impulse Radio Ultra Wideband Signal	2
1.1.2 Application of IR UWB signal for indoor localization	4
1.1.3 Indoor UWB Propagation Channel	9
1.1.4 UWB Receivers	11
1.1.5 RFID System	13
1.1.6 Pseudo-Noise Sequence	14

1.2 Literature Review and Motivation	15
1.3 Major Contributions of the Thesis	20
1.4 Organization of the Thesis	21
Chapter 2 Evaluation of UWB Ranging Performance in Realistic Indoor Environments	23
2.1 Introduction	23
2.2 Measurement Campaign	25
2.2.1 Measurement Setup	25
2.2.2 Measurement Environment	29
2.2.3 Measurement Procedures	36
2.3 Post Processing Algorithm.....	36
2.4 Measurement Results Analysis	42
2.4.1 Ranging Error Characterization.....	44
2.4.2 Effects of Parameter Settings on Ranging Performance	47
2.4.3 Ranging Error Modeling.....	52
2.5 Conclusion	60
Chapter 3 Delay-dependent Threshold Selection Technique.....	61
3.1 Introduction	61
3.2 Signal Model	62
3.3 Delay Independent Threshold Selection	66
3.4 Delay Dependent Threshold Selection.....	68
3.5 Simulation Results and Discussion	72

3.6 Conclusion	79
 Chapter 4 Least Square Time of Arrival Estimator for UWB Backscattering RFID	
System	80
4.1 Introduction	80
4.2 Signal Model and System Model	83
4.3 Least Square Time of Arrival Estimator	91
4.4 Immunity of the Estimator to Clutter	95
4.5 Simulation Results and Discussion	97
4.6 Experimental Evaluation	101
4.6.1 <i>Experimental Setup</i>	102
4.6.2 <i>Modified Signal Model</i>	104
4.6.3 <i>Methodology of Jitter Mitigation</i>	107
4.6.4 <i>Localization with TOA Measurements</i>	110
4.6.5 <i>Results and Discussion</i>	111
4.7 Conclusion	116
 Chapter 5 Clutter Suppression with Dual Pulse Modulation	
5.1 Introduction	117
5.2 Tag with DP Modulation	119
5.3 Application of the Proposed Tag	122
5.3.1 <i>Sequence Selection Criteria for Clutter Suppression</i>	127
5.3.2 <i>Two-Stage TOA Estimation</i>	129
5.4 Simulation Results and Discussion	132

5.5 Conclusion	137
Chapter 6 Conclusion and Future Works	138
6.1 Conclusion	138
6.2 Future Works	140
Author's Publications	142
Appendix A Derivation of LS Solution	144
Appendix B Transformation of $A_x(u,v)$, $A_y(u,v)$, and F	148
Appendix C Conditions for the Exceptional Case of LS Solution	151
Appendix D Transformation of Sample Model	153
Appendix E Variance of $\Psi_2(t)$	155
Reference.....	157

List of Figures

Figure 1.1. Illustration of unmodulated and uncoded IR-UWB pulse transmission with $N_f = 3$	4
Figure 1.2. Ranging scenarios for OWR and TWR: (a) OWR, (b) TWR.....	6
Figure 1.3. Illustration of the transmitted and received signals for OWR and TWR.	7
Figure 1.4. Block diagrams of different UWB receivers: (a). correlation receiver; (b) autocorrelation receiver; (c) energy detection receiver.....	12
Figure 2.1. Measurement System: (a). Measurement setup; (b) Block diagram of the system.	24
Figure 2.2. Normalized antenna patterns: (a). 3.1GHz, azimuth plane; (b). 10.6GHz, azimuth plane; (c). 3.1GHz, elevation plane; (d). 10.6GHz elevation plane.	26
Figure 2.3. Pulse characteristics of the measurement system: (a). The transmitted pulse shape (measured before transmitting antenna); (b). The PSD of the transmitted pulse (including transmitting antenna gain) vs. FCC spectrum mask; (c). Received monocycle at $L = 2\text{m}$	28
Figure 2.4. Four measurement environments: office (upper left), laboratory room (upper right), open hall (lower left), and corridor (lower right).	28
Figure 2.5. Floor plan of indoor office in PWTC.	30
Figure 2.6. Floor plan of laboratory. Notation of the objects follows Figure 2.5.	30

Figure 2.7. Floor plan of open hall. Notation of the objects follows Figure 2.5.....	31
Figure 2.8. Floor plan of corridor. Notation of the objects follows Figure 2.5.	31
Figure 2.9. Received channel profiles at $L = 5\text{m}$ in indoor office LOS, office NLOS, laboratory, open hall and corridor environments (from top to bottom). The vertical scale is in milivolts. Vertical dash line in each plot shows the true TOA of the direct path signal. .	32
Figure 2.10. Interference characterization.	33
Figure 2.11. Received interference with different times of averaging. In the five plots counting from top to bottom, averaging times are 1, 4, 16, 32, 64, and 512 respectively. The vertical scale is in volts.	35
Figure 2.12. Threshold selection in different SNR conditions: (a) High SNR; (b) Relatively low SNR; (c) Extremely low SNR.....	40
Figure 2.13. CDF of ranging errors with $\beta = 0.2$, $\delta = 50\text{ns}$, and $T = 1\text{ns}$	45
Figure 2.14. Scatter plot of ranging errors in NLOS environment for correlation receiver (left plot) and ED receiver (right plot) with $\beta = 0.2$, $\delta = 50\text{ns}$, and $T = 1\text{ns}$	46
Figure 2.15. RMSE versus sampling rate for correlation receiver in the office environment with $\beta = 0.2$, and $\delta = 50\text{ns}$	48
Figure 2.16. RMSE versus integration time for ED receiver in the office environment with $\beta =$ 0.2 , and $\delta = 50\text{ns}$	49
Figure 2.17. RMSE versus β in the office environment with $\delta = 50\text{ns}$ and $T = 1\text{ns}$	50
Figure 2.18. RMSE versus δ for ED receiver in office environment with $\beta = 0.2$ and $T = 1\text{ns}$	52
Figure 2.19. Error mean vs. distance for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta =$	

0.2, and $\delta = 50\text{ns}$.	55
Figure 2.20. Error variance vs. distance for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, β	
$= 0.2$, and $\delta = 50\text{ns}$.	55
Figure 2.21. Curve fitting for W for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$,	
and $\delta = 50\text{ns}$.	56
Figure 2.22. Curve fitting for λ for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$, and	
$\delta = 50\text{ns}$.	56
Figure 2.23. Comparison of CDFs generated from measurement data and the proposed model.	
.....	59
Figure 3.1. DD and DI threshold selections for ED receiver.	62
Figure 3.2. Block diagram of ED receiver employing DD technique.	68
Figure 3.3. Parameter a versus δ_{sn} for different degree of freedom.	69
Figure 3.4. P_{FA} vs. b for different a .	72
Figure 3.5. RMSE versus design parameter b at $E_{tx}/N_0 = 90\text{dB}$ in CM1.	73
Figure 3.6. The design parameter functions for different channel models.	74
Figure 3.7. RMSE performance of different threshold selection techniques in CM1 with $T =$	
1ns .	75
Figure 3.8. RMSE performance of different threshold selection techniques in CM2 with $T =$	
1ns .	76
Figure 4.1. System block diagram.	84
Figure 4.2. Transmitted/received signal present at various stages of the RFID system.	86
Figure 4.3. Comparison of the performance of the estimator under the scenarios with and	

without clutter for different number of sampled symbols in CM1 (left plot) and CM2	
(right plot) with $T_{ds} = 45\text{ns}$, $f_s = 8\text{GHz}$ and $\text{SCR} = -30\text{dB}$	98
Figure 4.4. MAE versus channel delay spread setting for different SNR values in CM1 with N_s	
$= 8$ and $f_s = 8\text{GHz}$	99
Figure 4.5. Effect of sampling frequency on MAE in CM1 with $N_s = 16$ for $T_{ds} = 40\text{ns}$ (dash	
line) and $T_{ds} = 50\text{ns}$ (solid line).....	101
Figure 4.6. Measurement setup.	102
Figure 4.7. Measurement setup layout. The measurement points are distributed in the shadow	
area.	103
Figure 4.8. Effect of jitter on the received signal ($N_f = 3$).	106
Figure 4.9. The received signal and the recovered tag response. The vertical scale is in	
millivolts.	112
Figure 4.10. The tag responses of 20 frames with different polarities.....	112
Figure 4.11. Jitter of different frames.	114
Figure 4.12. The scatter plot of roundtrip distance estimation error and its empirical PDF	
(curve fitting is shown as dash line).....	114
Figure 4.13. $ \varepsilon_x $ versus x and y coordinates.....	115
Figure 4.14. $ \varepsilon_y $ versus x and y coordinates.....	115
Figure 5.1. Tag structure.	119
Figure 5.2. Signal waveforms at different stages of signal reception.	120
Figure 5.3. Receiver block diagram.	126
Figure 5.4. MAE versus threshold for different N_s , SNR and SCR values in CM1 with $T_1 =$	
1ns . Dash line represents $\text{SNR} = 20\text{dB}$, solid line represents $\text{SNR} = 30\text{dB}$, dash-dotted	

lines (gray color) represents SNR = 40dB.	134
Figure 5.5. MAE versus SNR for different SCR values with $N_s = 64$, $T_1 = 1\text{ns}$ and $\theta' = 5\text{dB}$ in CM1.	135
Figure 5.6. MAE versus SNR for different N_s with $T_1 = 1\text{ns}$, $\theta' = 7\text{dB}$ and SCR = -10dB.	136
Figure 5.7. MAE versus SNR for different T_1 with $N_s = 128$, $\theta' = 9\text{dB}$ and SCR = 0dB.	136

List of Tables

Table 2.1. Performance of correlation and ED receivers in different indoor environments.....	45
Table 2.2. Fitting coefficients for ranging error model.....	59
Table 3.1. Fitting coefficients for DD.....	75

Summary

Ultra wideband (UWB) technology is particularly suitable for localization applications due to its fine time resolution, low power consumption, low complexity receiver design, and low probability of interception. The main goal of this thesis is to provide a good understanding of the challenges posed by UWB time of arrival (TOA) estimation in realistic indoor environments and develop robust TOA estimation techniques. To get insight into how the UWB signal performs in real indoor environments, experiments are carried out to collect a database of received signals captured in various indoor environments including both line-of-sight (LOS) and non-line-of-sight (NLOS) scenarios. With the measurement results, we perform the analysis and answer the following questions: How does the TOA estimation with different types of receivers perform in real indoor environment, what are the effects of system parameter settings on accuracy, and how do the ranging errors behave in real indoor environments? The study shows that the system parameter settings have a heavy impact on the performance especially in LOS environment. The analysis also allows us to build up a unified ranging error model that is applicable for both coherent and energy detection receivers.

To improve the accuracy while at the same time retain the low system complexity, we come up with several new techniques. The information of transmitted power combined with certain channel knowledge allows us to devise a delay-dependent threshold setting

strategy that helps the estimation overcome the excessive noise. The study also indicates that the threshold setting strategy is insensitive to possible variation or inaccuracy existing in the available power delay profile and small-scale fading channel parameters. To overcome the clutter problem, a novel least square (LS) TOA estimator is developed for UWB backscattering RFID system with tag implementing antipodal 2-Pulse-Amplitude-Modulation (2-PAM). We show that the proposed estimator is inherently immune to the clutter for arbitrary data sequence. The performance of the estimator is evaluated by both simulation and practical measurements. The above solution, however, requires the sampling rate of the order of GHz for UWB signal. To lower down the sampling rate requirement, we propose an RFID system implementing a dual pulse (DP) modulator in the tag which enables the use of the low complexity autocorrelation receiver in the reader. The polarity of modulated signal is jointly determined by a ranging sequence and a pseudorandom noise (PN) sequence. A low complexity two-stage TOA estimator is developed for the proposed system. When ranging sequence, PN sequence and sampling duration are properly selected, part of the clutter terms at the output of the autocorrelation receiver can be completely eliminated while the rest can be reduced by averaging over more symbols. The two solutions have their own advantages. The first one has better clutter suppression capability and does not require dedicated design for data sequence whereas the second one requires receiver with lower complexity and lower sampling rate. The performances of both solutions are validated by simulation results.

Chapter 1

Introduction

1.1 Background Information

Localization using radio frequency (RF) signals has a long history and nowadays becomes prevailing in many civilian and military applications such as intelligent transportation systems (ITS), public safety, automated billing, inventory management, and intrusion detection. In the United States, the Federal Communication Commission (FCC) has mandated cellular system operators to estimate the position of an emergency caller with accuracy of less than 125 meters in Enhanced 911 (E-911) service [1]. Localization systems designed for outdoor open space have been developed and deployed for many years. Some of these systems developed in earlier days including Loran C, Omega and Decca are terrestrial-based localization systems [2], [3]. The global positioning system (GPS) is able to achieve high accuracy using a constellation of 24 satellites [4], [5]. The aforementioned systems, however, encounter difficulty to achieve satisfactory localization accuracy in heavily built-up outdoor environment such as metropolis due to multipath fading. The localization service in cities may be provided by existing cellular network and the accuracy has been proven reasonable for some applications [2]. When the

scenario moves to indoor environments, the aforementioned systems with infrastructure designed for outdoor environments fail due to multipath impairments and/or incapability of penetrating the external wall of buildings. The indoor wireless network employing narrowband signal can realize indoor localization but its accuracy is limited by multipath fading [6]. Recently ultra wideband (UWB) emerges as a viable solution for indoor localization applications due to its excellent multipath mitigation capability [7]. To realize full potential of UWB signal, it is critical to design time-of-arrival (TOA) estimators able to produce accurate TOA metrics that can be incorporated into localization algorithms. In this thesis, we focus on designing robust UWB TOA estimators adaptive to the dynamic change of indoor environments and suppress various noise and interferences. Before proceed to give a thorough literature review, some essential basic concepts are introduced to facilitate the understanding of the rest of part of the thesis.

1.1.1 Impulse Radio Ultra Wideband Signal

The history of UWB signal can be traced back to the first oversea wireless communication experiment conducted by Guglielmo Marconi in 1897 in which the wireless signal generated by a spark radio transmitter covers very large bandwidth. The concept of IR was first introduced by Ross in his works describing the transient behavior of a certain class of microwave networks through their characteristic impulse response [8]. Although the first US patent was awarded for UWB communication applications [9] in 1973, the development of UWB systems focused more on radar applications in the earlier days due to the fine time resolution of UWB signal [10], [11]. The academic interest for UWB in communication started from the publications of some pioneer works of Scholtz and Win that demonstrated the

potential of UWB signals on systems with large user number and its robustness over fading in the multipath dense environment such as indoor office [12], [13], [14]. Due to the increasing interest in UWB in industry, FCC released the First Report and Order in 2002 to endorse the commercial use of UWB as unlicensed devices in overlapping frequency band with existing services. The power spectral density (PSD) mask was clearly defined to avoid potential interference of UWB on other systems [15]. In the report of FCC, UWB signal is defined as radio frequency signal having relative bandwidth larger than 20% or absolute bandwidth of more than 500MHz. In 2007, European Commission Committee (ECC) published the first draft of a European UWB spectral mask with stricter PSD mask than the mask released by FCC [16].

Generally, UWB signal can be classified as carrier-based and carrier-free types. The first type of UWB signal is generated by upconverting a baseband signal to the desired frequency band by mixing it with a narrowband carrier signal. The second type is produced by directly generating a series of extremely short duration pulses with central frequency located in the desired frequency band. The carrier-free UWB signal is usually referred as impulse radio (IR) UWB. In this thesis, only IR UWB is considered.

An uncoded transmitted IR UWB signal $s(t)$ consists of a series of symbols. Each symbol has a number of frames, i.e.,

$$s(t) = \sum_i \sum_{j=0}^{N_f-1} \sqrt{E_s / N_f} p_f(t - iT_s - jT_f) \quad (1.1)$$

where $\sqrt{E_s}$ is the symbol energy, $p_f(t)$ is the frame waveform with normalized energy

Figure 1.1. Illustration of unmodulated and uncoded IR-UWB pulse transmission with $N_f = 3$.

$\sqrt{E_s / N_f}$, $T_s = N_f T_f$ is the symbol duration, T_f is the frame duration and N_f is the number of frames per symbol. The detailed format of $p_f(t)$ is determined by modulation scheme. For instance, for an unmodulated signal, $p_f(t) = p_{tx}(t)$ where $p_{tx}(t)$ is a single IR UWB pulse while for a transmitted-reference (TR) signal, the frame waveform may be expressed as $p_f(t) = p_{tx}(t) + p_{tx}(t - T_f/2)$. Figure 1.1 shows a transmitted IR UWB signal with $p_f(t)$ being a single UWB pulse $p_{tx}(t)$ and $N_f = 3$.

1.1.2 Application of IR UWB signal for indoor localization

The reason why IR UWB is more suitable for indoor localization applications than those narrow band and conventional wideband signals is three fold. First, the enormous bandwidth greatly improves the multipath resolvability and hence alleviates the effects of multipath fading [14]. Also spreading information over large bandwidth reduces interference to other systems and lowers the probability of interception. In addition, UWB system is allowed to operate within certain frequency band for license free usage and the baseband nature of IR UWB signal eliminates the mixing-carrier stage required in the conventional narrow band system design, which imply that the UWB system is featured by low complexity and low cost.

Hence, the UWB system is particularly suitable for indoor localization applications that require high position accuracy and large scale device deployment such as wireless sensor networks. Several companies such as Multispectral and Time Domain have developed their UWB location systems that are now available in market [17], [18]. The Sapphire DART real-time location system developed by Multispectral achieves ranging accuracy better than 30cm and read ranges up to 200 meters in line-of-sight (LOS) scenario and 50 meters in non-line-of-sight (NLOS) with battery life longer than 7.5 years.

There are four types of measurement metrics available for UWB localizers: received signal strength indicator (RSSI), time of arrival (TOA), angle of arrival (AOA) and time difference of arrival (TDOA) [19]. Localization using TOA metric has the advantages of high accuracy and low implementation cost as compared to other metrics [20]. It was experimentally proven that UWB signal with TOA technique is able to achieve accuracy of the order of centimeter with low-power and low-cost implementation in indoor environment [21].

A localization system usually consists of many nodes that form a network. The coordinates of some nodes have been calibrated or estimated and therefore are available to the system. These nodes are here referred to as reference nodes while the rest of nodes in the network with unknown positions are known as blind nodes. To estimate the position of a blind node in the localization system using TOA technique, the range of the blind node to several different reference nodes has to be estimated. For a two-dimensional (2D) localization scenario, at least three such range estimations are needed to determine the position of the blind node. All the range estimations can then be fused and combined with additional information such as motion information to estimate the location of a blind node.

(a)

(b)

Figure 1.2. Ranging scenarios for OWR and TWR: (a) OWR, (b) TWR.

Range estimation between two nodes may be accomplished by one-way ranging (OWR) or two-way ranging (TWR) processes. Consider two nodes A and B in the localization system. Assume that there are two scatterers in the environments. As illustrated in Figure 1.2 and Figure 1.3, node A initializes the OWR or TWR processes by sending a signal at $t = t_{sta}$.

Figure 1.3. Illustration of the transmitted and received signals for OWR and TWR.

For OWR process, the signals propagating through two paths (P3, P5+P6) are received by node B. Among the received signals, the signal via path P3 is the direct path and the signal via path P6 is caused by the scatterer 2. Node B scans the received signal starting from $t = t_{sta} + \tau_{sys}$ where τ_{sys} is the system delay encompassing the delay of both node A and node B. If threshold detection technique is employed, the signal component first crossing a preset threshold is detected as the direct path signal. The timing associated with the direct path is taken as TOA t_{toa} .

For TWR process, the signals sent by node A propagating through the channel is received by node B. Node B may take two actions. The first possible action is that node B extracts useful information from the received signals and then generates a new signal to send back to node A. The second possible action is that node B modulates and reflects the received signal back to node A. Usually the processing delay of the first action is much longer than the second

one. Regardless of the action taken at node B, node A scans the received signal starting from timing $t = t_{sta} + \tau_{sys}$. The returned signal via path P4 is the desired direct path signal and the signal via P8 is the multi-path. Note that different from OWR, the signal path that is reflected by scatterer 1 and returns to node A via P2 comes into the picture in TWR. Unlike signals via P4 and P8, this signal does not interact with node B and is termed as clutter. This clutter suffers only system delay associated with node A. Recall that τ_{sys} includes system delays from both node A and node B. If node B takes the first aforesaid action generating new signal, the long delay at node B renders arrival of P2 clutter located prior to the starting time of receiver scanning. If node B takes the second action, due to the short delay at node B, clutter P2 will fall in the scanning region and may be falsely detected as the direct path if it arrives earlier than the direct path P4 and crosses the detection threshold. Detailed discussion on this problem will be presented in Chapter 4 and 5 of this thesis.

With estimated TOA $\hat{t}_{toa}$ by either OWR or TWR, the range may be computed as

$$\hat{L} = \vartheta c (\hat{t}_{toa} - t_{sta} - \tau_{sys}) \quad (1.2)$$

where $\hat{L}$ is the estimated range, c is velocity of light, and ϑ is a constant. Here $\vartheta = 1$ for OWR and $\vartheta = 1/2$ for TWR. In this thesis, we will focus on TOA estimation for ranging process although part of the studies can be applied to the synchronization of communication systems as well.

1.1.3 Indoor UWB Propagation Channel

The properties of UWB propagation channel together with transmitting and receiving antennas determine the distortion suffered by the received signal. According to the testing environment, UWB channel models may be classified as indoor and outdoor models. The propagating mechanism of the indoor environment is much more complicated than its outdoor counterpart due to numerous scatterers generating a large number of echoes. Various UWB channel models have been proposed in the literature such as [50], [51], [88] and [89]. IEEE 802.15.3a and IEEE 802.15.4a UWB channel models proposed in [88] and [50] are two standardized channel models commonly used for simulation studies. The former targets for short-range (up to 10 meters), high data rate (of the order of 500 Mb/s) Wireless Personal Area Network (WPANs), while the latter provides models for median-range (up to 20 meters), low data rate (hundreds of kb/s to a few Mb/s) WPANs, particularly for location-awareness applications. Since the focus of this thesis is TOA estimation for indoor localization applications, IEEE 802.15.4a channel models are adopted in this thesis.

The UWB channels models consist of channel impulse response (CIR) and path loss models. The general CIR $h_{CIR}(t)$ for IEEE 802.15.4a is a modified Saleh-Valenzuela model [90]

$$h_{CIR}(t) = \sum_{i=0}^{N_{mul}-1} \xi_i \alpha_i h_{CIR_i}(t - \tau_i) \quad (1.3)$$

where $h_{CIR_i}(t)$ is the CIR of the i^{th} path, $\xi_i \in \{\pm 1\}$, α_i and τ_i are the polarities, instantaneous amplitude gain and propagation delay of the i^{th} path respectively, and N_{mul} is

the number of multipaths. Denote the spatial average of α_i as $\Omega_i = \mathbb{E}_{ssa} [|\alpha_i|^2]$ where $\mathbb{E}_{ssa} [\cdot]$ is the operation of averaging the small-scale fading effect over a local area. Given the mean path power Ω_i , the path gain α_i follows Nakagami distribution (i.e. the small-scale fading is Nakagami distribution)

$$f_{\alpha_i}(\alpha_i | \Omega_i) = \frac{2m_{nak} m_{nak} \alpha_i^{2m_{nak}-1}}{\Omega_i^{m_{nak}} \Gamma(m_{nak})} \exp\left(-\frac{m_{nak} \alpha_i^2}{\Omega_i}\right). \quad (1.4)$$

The statistics of m_{nak} is a lognormally distributed random variable with its mean and standard deviation both having a delay dependence as specified in [50]. To take into account the fact that in some environment multipath caused by major scatterers tends to group together, the general CIR models in IEEE 802.15.4a are multi-cluster models with single cluster model being a special case. The arrival times of cluster follows a Poisson process. Within a cluster, the arrival time of multipath delays $\{\tau_i\}$ are mixtures of two Poisson processes. For details, interested readers may refer to the channel model report [50].

The path loss model is described as follows. The total path gain is defined by $\Omega_{tot} = \sum_{i=0}^{N_{mul}-1} \Omega_i$. The path loss P_{loss} in dB scale is related to Ω_{tot} by $P_{loss} = -10 \log_{10}(\Omega_{tot})$. P_{loss} can be modeled as a function of propagation distance L

$$P_{loss}(L) = P_{loss}(L_0) + 10\rho_{pl} \log_{10}\left(\frac{L}{L_0}\right) + S \quad (1.5)$$

where L_0 is a calibration distance, ρ_{pl} is path loss exponent, and S represents the shadowing (also referred as large-scale fading) term, which is a Gaussian-distributed random variable with zero mean and standard deviation σ_s . ρ_{pl} is equal to 2 for free space propagation, and

has a typical value of $1 \sim 2$ for LOS scenario and $2 \sim 8$ for NLOS scenario. The above discussions suggest that the path gain α_i is jointly determined by the characteristics of large-scale and small-scale channel fading with their statistics being distance-dependent.

IEEE 802.15.4a standard includes nine frequency selective fading UWB channel models for typical environments. In this thesis, the models CM1 and CM2 representing indoor residential LOS and NLOS environments are adopted to generate simulation results. For CM1, $P_{loss}(L_0) = 43.9$ dB, $\rho_{pl} = 1.79$, and $\sigma_s = 2.22$ dB; For CM2, $P_{loss}(L_0) = 43.9$ dB and $\rho_{pl} = 4.58$, $\sigma_s = 3.51$ dB.

1.1.4 UWB Receivers

UWB receivers may be classified as coherent or non-coherent. Coherent receiver requires the information of individual UWB pulses that may be estimated from received signal waveform. Such information may be pulse shape, path gain or pulse energy. Due to the unknown spreading and distortion effects of the channel, knowing the information of pulses propagating through different paths are generally difficult and a suboptimum solution is to assume that the channel effects of different paths are similar. For instance, we may presume a universal pulse shape for all paths. Such pulse shape may be extracted from the received pulse propagating through free space or even from the transmitted pulse. In contrast to coherent receiver, non-coherent receiver does not require the information of individual pulses and therefore its complexity is much lower than the coherent one.

Figure 1.4. Block diagrams of different UWB receivers: (a). correlation receiver; (b) autocorrelation receiver; (c) energy detection receiver.

Figure 1.4 depicts the block diagrams of three types of UWB receivers that will be discussed in this thesis. Correlation receiver shown in Figure 1.4 (a) falls into the category of coherent receiver. The received signal $r(t)$ is correlated with a pulse template $p_{tm}(t)$ locally generated at the receiver by analogue circuitry and then the output is sampled. Note that this type of receiver may also be implemented by sampling the received signal $r(t)$ and performing the correlation with the template stored as digital data. Figure 1.4 (b) and (c)

depict two types of non-coherent receivers: autocorrelation and energy detection (ED) receivers. Autocorrelation receiver multiplies the received signal with a delayed version of itself and the output is integrated and dumped (or may be directly sampled by analogue to digital circuitry). The ED receiver passes the received signal to non-linear devices so that the input signal is squared and then sampled.

1.1.5 RFID System

Radio frequency identification (RFID) is a generic term for technologies that use radio waves to identify people or objects automatically. An RFID system consists of a number of readers and tags. A simplest tag is formed by attached a microchip to an antenna. Serial numbers of the objects to be identified together with other useful information are stored in the writable memory of tags. A reader is a full functional transceiver initializing the identification process by interrogating the tag with a radio wave. It extracts the identity information from the radio wave returned from the tags. During this process, the tags and readers do not have physical contact and even the clear LOS path between them may not be present. Besides the identity information, the real-time information including stock level, process stages, shipping routing and geographical location can also be stored in the tag. Such information may be critical for the business management to make prompt decision on business activities, e.g., replenishment of stock. Due to the aforesaid advantages, RFID has wide usage in security monitoring and control, theft and counterfeiting prevention, healthcare monitoring, supply chain management, and airline maintenance, repair and overhaul (MRO), etc. [91]-[95].

A popular type of RFID system is backscattering RFID system. In this type of system, tags

do not generate their own signal but modulate and reflect the received signal back to the reader. Contrastingly, in other RFID systems, the tags receive the signal from the reader, process it and then generate new signals that is sent back to the readers. In these two kinds of of RFID systems, the tags of backscattering system usually consume much less power and have much lower structure complexity than those of the second type. In this thesis, TOA estimation for backscattering RFID system is our focus in Chapter 4 and 5.

1.1.6 Pseudo-Noise Sequence

A pseudo-noise (PN) sequence is a type of noise like sequence since its element appears to be randomly chosen from a set of numbers. However, the sequence is not really random since it is generated by a well-defined logic and repeats itself after a code period [96]. The generation can be realized by a feedback shift register and a combinational logic. If the elements of the sequence is confined to be 1 and 0, the sequence becomes a binary PN (BPN) sequence. Denote a BPN sequence as $\{c_{BPN_j}\}$. It can be transformed into a bipolar sequence by performing the conversion $c_j = 2c_{BPN_j} - 1$ on every element. In conventional UWB systems, PN sequence primarily has two functions. The first function is acting as unique identity of individual users. The second function is smoothing the signal spectrum shape so that the interference to other system caused by undesired spectral lines in the PSD of UWB signal can be mitigated [97]. In Chapter 5 of this thesis, PN sequence is assigned an additional function: mitigating undesired clutter for backscattering RFID system.

1.2 Literature Review and Motivation

To perform TOA estimation, the transmitted UWB signal usually consists of a number of data symbols. Each data symbol is associated with multiple frames. Each frame may contain one or more UWB pulses depending on the specific signaling format used. TOA estimation has to be performed not only at the frame level to find when the first frame in each symbol starts but also at the pulse level to find where a pulse is located within a frame.

Fully exploiting the potential of UWB and achieving accurate TOA estimation in the indoor environments is a challenging task. Due to the low PSD of UWB signal, the noise ahead of the direct path may trigger false detection that leads to high false alarm rate. Furthermore, the direct path may be overlapped and distorted by other multipath, which reduces the direct path energy captured by the correlation or matched filter receivers. Also due to the attenuation caused by blockages, the direct path may not be the strongest path. Hence, the conventional practice that picks the arrival time of the maximum sample as TOA is ineffective [22]. The situation is further complicated by the fact that the indoor environment is highly a dynamic environment. “Dynamic” here means that the strength of interference signals and/or the channel condition change dramatically within a short time as the mobile node moves around, thereby the signal to noise ratio (SNR) of the received signal varies in both spatial and temporal domains. Hence, the parameter setting should be robust to the change of channel condition. In addition, the narrow band and conventional wideband interference may also deteriorate the estimation accuracy if its bandwidth overlaps with that of UWB signal. For

some localization systems such as backscattering RFID system, besides multipath, interference and noise, the received signal also contains undesired clutter signal caused by the scatterers surrounding the mobile node or reference node. The clutter signal may overwhelm the desired response and hence severely degrade the estimation performance [23]. However, the fact that little attention has been dedicated to this issue motivates us to develop novel RFID systems with non-data-aided (NDA) TOA estimators to overcome clutter.

To improve the estimation accuracy, various estimators applicable for indoor environments have been proposed. In [24], under the assumption of unknown pulse shapes that differ from path to path, a maximum likelihood (ML) estimator is proposed to extract timing information when the uncertainty region is of pulse duration. In [25], [26], the CLEAN algorithm and the channel estimator developed based on ML criterion are able to estimate the amplitude and delay of the paths and hence can be implicitly used as TOA estimators. In [27], several suboptimal algorithms in principle similar to CLEAN are proposed and their performance are tested with real measurement data. In [28], a generalized maximum likelihood (GML) detection method is proposed. It performs the channel estimation in a predefined time interval prior to the largest sample and then picks the first channel tap crossing a preset threshold as direct path of which the timing is taken as TOA. In [29], a TOA estimator based on segmentation analysis for non-stationary process is introduced which does not require the details of received waveform. These estimators have to operate with at least subpulse sampling rate or even require Nyquist rate, and hence they are computationally prohibitive.

To lower down the sampling rate requirement and speed up the estimation, many estimation algorithms with low complexity and/or low sampling rate requirements are

proposed. In [30], [31], and [32], the timing acquisition is accomplished by using coarse bin search in conjunction with a correlator bank. In [33], the authors utilize a beacon/listener technique using Kasami sequences and divide-and-conquer algorithms to reduce coarse acquisition time. In [34], TOA is extracted by integrating the received signal energy within intervals of signal region (SR) which is assumed to be known but starts from different timings. The timing that gives the maximum integration output is taken as TOA. It is shown that sampling rate comparable to the inversion of the pulse width incurs small loss on performance as compared to the Nyquist rate. In [35], the received signal is correlated with a template estimated based on previous receiving frames and picks the maximum correlation output as the recovery timing.

In [36], it investigates channel estimation and TOA estimation with subspace-based methods. Using the concept of innovation rate, the received signal is projected onto a lower dimensional subspace via lowpass/bandpass-filtering and sampling below the Nyquist rate. The timing estimation problem is converted to a harmonic retrieval problem and is solved via a subspace method involving a Vandermonde equation system. However, the complexity is high and the Vandermonde systems are ill-conditioned if the delays of the paths are too closely spaced. In [37], an estimator consisting of coarse and fine timing estimation is proposed for UWB timing problem. The coarse timing phase adopts a fixed threshold while the fine timing phase uses an adaptive threshold based on noise statistics estimated from the first phase. In [38], the frame timing (FT) acquisition scheme operates in a non-data-aided (NDA) fashion making use of the cyclo-stationary nature of UWB signals. It only requires frame-rate sampling. In [39], a novel “timing with dirty template” (TDT) synchronization criterion is established

and its NDA and DA estimators are developed. TDT operates based on simple integrate-and-dump (I&D) operations over the symbol duration. In [40], the timing estimation problem is transformed into a ML amplitude estimation problem and a DA estimator is developed based on generalized likelihood ratio test. The symbol level timing offset is obtained through a linear search and a closed form solution is derived for the frame level timing offset. In [41], a DA TOA estimator is proposed for dual pulse (DP) signal structure. It correlates the received signal with its delayed version, and then makes use of threshold crossing to detect the direct path.

ED based estimators are gaining interest due to its low implementation complexity, low sampling rate requirement and robustness over pulse shape distortion. In [42], the authors propose to sample the received signal by sliding an integration window over the uncertainty region of TOA. The timing associated with the maximum output is estimated as TOA. And it shows that such integration makes the detection more robust in the scenario where the direct path is weaker than the later incoming paths. It also shows that the average power delay profiles (PDP) of propagation channels can be incorporated in TOA estimation to improve the accuracy. In [43], an ML estimator is derived under the assumption that the detailed waveform of the received signal path is unknown and the amplitude of channel taps follows Gaussian distribution. The estimator is also tested over non-Gaussian channel and its effectiveness is confirmed. In [83], assuming only an approximate knowledge of the received signal pulses' duration, a TOA estimator operates on the energy measurement is derived by making use of least mean square technique. Its performance under the sub-Nyquist sampling rates is investigated and the effects of analog-to-digital conversion with limited resolution on the

performance are explored.

In [44], a low complexity threshold-based ED estimator with threshold setting selected based on the maximum and minimum samples is proposed. In [45], the threshold of ED estimator is set based on Kurtosis value and the simulation result shows that such setting strategy is robust to channel condition variation. In [46], the performance of TOA estimation with different signaling formats is given. The conventional threshold-based estimator is sensitive to the SNR value and its performance deteriorates drastically when SNR falls below certain value. This is typically known in non-linear estimation as threshold phenomena [47]. Such sensitivity motivates us to find a better way to select the threshold value in order to improve its robustness over SNR variations. This leads to the delay-dependent (DD) threshold selection method that will be described in Chapter 2.

In [48], a practical ED based TOA estimation scheme is proposed which is robust to both narrow band interference (NBI) and wide band interference (WBI), and it shows superior performance as compared to other TOA estimation schemes in realistic multipath environments subject to NBI and WBI. For a particular case of simple TOA estimator based on threshold with averaging filter, an analytical criterion for threshold selection in the presence of NBI is proposed. In [49], nonlinear matrix filtering using minimum and median filters is applied to the energy samples of the received signal to mitigate WBI.

Many of the aforementioned estimators involve one or more parameter settings, for example, the threshold setting in the estimators proposed in [44] and [46], the length of signal region in the estimators developed in [34] and [35], and the integration length for those estimators used with autocorrelation or ED receivers. Among all the parameters, one of

the most critical parameter settings for any estimator that relies on digital samples is sampling rate. In order to adapt to the dynamic indoor propagation environment, the parameter settings have to be carefully determined. The effect of parameter settings on ranging accuracy for different TOA estimators may be carried out by simulation using existing statistical channel models such as [50] and [51]. However, the statistical channel model usually ignores some details to reduce implementation complexity and simulation time. This motivates us to perform an extensive measurement campaign to construct a database for real channel responses, with which the ranging errors in various indoor environments can be characterized and the effects of parameter setting, especially the sampling rate, can be modeled and studied.

1.3 Major Contributions of the Thesis

An extensive measurement campaign is carried out in different indoor environments. The captured channel responses are used to evaluate the ranging accuracy with TOA technique for different types of receivers, investigate the effects of critical parameter settings and build up a unified ranging error model. The results are published in [68] and [69].

With the understanding of the behavior of TOA estimation in real indoor channels, we devise a low-complexity DD threshold selection method to reduce the false alarm rate caused by excessive noise. The sensitivity of the estimator to the channel condition variation is also investigated. The results are published in [70].

To overcome the clutter problem for UWB RFID systems, two solutions are proposed. In

the first solution, a least square (LS) TOA estimator is derived for existing UWB RFID system with tag implementing antipodal 2-Pulse-Amplitude-Modulation (2-PAM). We prove that the estimator is inherently immune to the clutter regardless of SNR condition. The second solution consists of a novel RFID system with tags implemented DP modulation and a two-stage TOA estimator. The DP modulation enables low complexity autocorrelation detection at the receiver. The study indicates that it is able to mitigate the clutter effectively with low sampling rate and low hardware complexity. The results have been submitted to journals [71] and [72].

1.4 Organization of the Thesis

The remainder of the thesis is organized as follows. In Chapter 2, the measurement setup and procedure are described. An adaptive threshold method applicable for both coherent and noncoherent receivers is devised. With the threshold setting strategy, the ranging accuracies of both correlation and ED receivers are evaluated and compared. The effects of parameter settings such as sampling rate and integration length are analyzed. Moreover, a unified ranging error model applicable for different types of receivers is proposed.

In Chapter 3, a low complexity DD threshold selection method is proposed and simulations are conducted to evaluate its accuracy in both LOS and NLOS environments. The results indicate that the method can outperform the conventional threshold setting methods. Its sensitivity to the channel condition variation is studied as well.

In Chapter 4, an LS estimator with inherent immunity to clutter is developed for UWB

backscattering RFID system with its tags implementing 2-PAM modulation function. It is shown that the immunity of the estimator holds for arbitrary data sequence regardless of SNR condition. The effects of channel condition, sampling signal length and SNR on the accuracy are investigated via simulation. The estimator is also modified to accommodate the jitter and is implemented in a prototype of the UWB backscattering system. The experimental results validate the robustness of the estimator in the presence of strong clutter.

In Chapter 5, an RFID system with its tags having DP modulation capability is described. By putting some constraints on ranging sequence and interference suppression sequence, a two-stage estimator for the system is developed. The performance of the estimator under different signal-to-clutter conditions is evaluated. The effects of different system parameter settings on the accuracy are studied.

At last, Chapter 6 summarizes the thesis and recommends future research direction on the relevant topics.

Chapter 2

Evaluation of UWB Ranging Performance in Realistic Indoor Environments

2.1 Introduction

As elaborated by Equation (1.2), ranging accuracy is solely determined by the TOA estimation if the system delay is fixed and timers are perfectly synchronized. TOA estimation performance is in turn determined by the receiver type and the respective system parameters settings. UWB signal reception can be accomplished by either coherent or non-coherent receiver. Correlation receiver is a type of coherent receiver while ED receiver falls into the non-coherent category. To gain the insight of how different receivers behave in various indoor propagation environments, we resort to the experimental approach. A measurement campaign is conducted in several typical indoor environments to collect a database of real channel responses, based on which we perform the following analysis for both correlation and ED receivers: 1). the characteristic of ranging errors; 2). the effects of parameter settings on the ranging performance; 3) ranging error modeling. Since the power of the transmitted IR-UWB

(a)

(b)

Figure 2.1. Measurement System: (a). Measurement setup; (b) Block diagram of the system.

signal in the measurement system complied with the FCC PSD mask, the analysis gives a full picture of how a practical ranging system using TOA technique performs in real channels with pulse spreading, distortion and attenuation.

2.2 Measurement Campaign

2.2.1 Measurement Setup

The measurement setup and its block diagram are depicted in Figure 2.1. Agilent 8648B signal generator provides the clock for the entire measurement system. The kernel of the measurement setup is an in-house developed pulse generator generating UWB signal at a pulse repetition frequency (PRF) of 25MHz [52]. A bandpass filter (BPF) is then used to shape the spectrum of the UWB signal. The PRF of the UWB signal is lowered down to 2MHz by a HP11720 pulse modulator whose TTL control signal is provided by Tabor 8600 programmable pulse generator. The resultant UWB signal consists of repetitive monocycles evenly distributed with a time interval of 500ns which is sufficiently long to avoid inter-symbol interference (ISI) caused by multipaths. Tabor 8600 also provides a periodic rectangular waveform with period of 500ns to the receiver via a low loss coaxial cable. This signal acts as triggering signal for the sampling process. The received UWB signal is amplified by low noise amplifier (LNA) before being sampled by the Agilent 86100B digital sampling oscilloscope (DSO). Both transmitting and receiving antennas are asymmetrical biconical antennas [53] and have linear polarization. The radiation patterns of the antennas

(a)

(b)

(c)

(d)

Figure 2.2. Normalized antenna patterns: (a). 3.1GHz, azimuth plane; (b). 10.6GHz, azimuth plane; (c). 3.1GHz, elevation plane; (d). 10.6GHz elevation plane.

used in the experiment are examined in the anechoic chamber over frequency range [3.1GHz, 10.6GHz] and the results are shown in Figure 2.2. The antenna gains in each plot of Figure 2.2 are gains normalized to the maximum gain in that plot. It can be observed that the biconical antennas are omni-directional in azimuth plane and have around 30 degree beamwidth in elevation plane if -3dB from maximum gain is considered. The antennas have approximately 2dBi maximum gain in azimuth plane.

(a)

(b)

(c)

Figure 2.3. Pulse characteristics of the measurement system: (a). The transmitted pulse shape (measured before transmitting antenna); (b). The PSD of the transmitted pulse (including transmitting antenna gain) vs. FCC spectrum mask; (c). Received monocycle at $L = 2\text{m}$.

Figure 2.4. Four measurement environments: office (upper left), laboratory room (upper right), open hall (lower left), and corridor (lower right).

The dynamic range of the measurement system is $[-400\text{mV}, 400\text{mV}]$, the bandwidth is 2.9 GHz. The sensitivity of the system is 0.25mV. Figure 2.3 (a) shows the transmitted pulse shape measured before the transmitting antenna. Figure 2.3 (b) shows that the PSD of the transmitted pulse is compliant with the FCC PSD mask for indoor application. Hence, our study reflects the performance of a ranging system with realistic power constraint. Figure 2.3 (c) depicts the time domain waveform of the monocycle received at transmitter-to-receiver (Tx-Rx) distance $L = 2\text{ m}$.

2.2.2 Measurement Environment

Figure 2.4 shows four measurement environments, among which the office environment is the most complicated environment having both LOS and NLOS scenarios. The office measurements were performed at the Positioning & Wireless Technology Center (PWTC) which is located at the 4th floor of Research Techno Plaza in the campus of Nanyang Technological University, Singapore. PWTC is a typical modern office with width 27.3m, length 27.3m and height 2.8m. Its main area is soft partitioned and furnished with cupboards, metal cabinets, tables made of chipboard, plastic chairs, and computers, etc. The main area is either directly connected to other rooms or via very short corridors. Most of the walls inside the building are multilayer plasterboards. The external concrete walls of the building and the inner concrete walls surrounding the staircases and elevators cause great attenuation to the signal. The detailed layout of PWTC is shown in Figure 2.5. Besides the indoor office, the rest of measurement environments are the laboratory room, open hall and corridor. The floor plans of these measurement environments are shown in Figure 2.6 to Figure 2.8. These three environments only have LOS scenarios. The laboratory room is a single room with some

Figure 2.5. Floor plan of indoor office in PWTC.

Figure 2.6. Floor plan of laboratory. Notation of the objects follows Figure 2.5.

Figure 2.7. Floor plan of open hall. Notation of the objects follows Figure 2.5.

Figure 2.8. Floor plan of corridor. Notation of the objects follows Figure 2.5.

Figure 2.9. Received channel profiles at $L = 5\text{m}$ in indoor office LOS, office NLOS, laboratory, open hall and corridor environments (from top to bottom). The vertical scale is in millivolts.

Vertical dash line in each plot shows the true TOA of the direct path signal.

tables and measurement instruments placed against the walls. The door of the room is closed during measurement. Open hall is a large indoor open space where multipaths are primarily

Figure 2.10. Interference characterization.

caused by floor, walls and some supporting beams. Corridor is a typical waveguide-like environment. All the doors along the corridors were closed during measurements.

Figure 2.9 shows the received channel profiles of different environments. The office and the laboratory environments are featured by dense multipath while the multipath of open hall and corridor are relatively sparse. In the NLOS scenario shown in the second plot of Figure 2.9, the direct path in NLOS may be severely attenuated and thereby may be too weak to be detected.

Prior to the measurement, the noise and interference in the measurement sites were characterized by Agilent spectrum analyzer to ensure that proper steps were taken to

suppress them. As illustrated in Figure 2.10 , in the indoor office, there are three strong interference components located at 900MHz, 1.8GHz and 2.4GHz, corresponding to GSM 900, GSM 1800 and wireless LAN signals. The HP 86100 sampling oscilloscope employs equivalent time sampling that takes one sample from each pulse period and then compresses them to form a continuous waveform. Equivalent time sampling has the advantage of obtaining extremely high effective sampling rate using much lower physical sampling rate. However, it can only be applied to repetitive signals with repetition rate slightly deviated from the sampling clock signal. In the measurement campaign, the PRF is 2MHz and hence the rest of the signals that have different data rates are just random signals seen by the oscilloscope. Therefore, by averaging the captured waveforms of different frames, the interference can be suppressed.

Figure 2.11 shows the interference captured by the digital sampling oscilloscope in indoor office environment with different averaging times. The observation indicates that averaging is able to mitigate the interference significantly. It is found that 512 times averaging is sufficient to overcome the interference. Figure 2.11 also indicates that besides the interference, the noise floor is also lowered by the averaging operation. Note that even with 512 times averaging, the level of interference plus thermal noise still varies spatially as well as temporally and such variation can be as much as 10dB among the measured profiles due to the received signal strength from the aforementioned interference systems such as wireless LAN. In addition, the received UWB signal level also varies spatially due to the variation of channel condition as the receiver position varies. Hence, the measurement environments are dynamic environments.

Figure 2.11. Received interference with different times of averaging. In the five plots counting from top to bottom, averaging times are 1, 4, 16, 32, 64, and 512 respectively. The vertical scale is in volts.

2.2.3 Measurement Procedures

Both the transmitting and receiving antennas are placed at 1.65m above the floor. Every stored channel profile has duration of 400ns and is the result of 512 times averaging. During the measurement, the transmitter location was fixed and the receiver was placed at the spatial sample points uniformly distributed along some linear routes. The spacing between any two adjacent points was 0.5m for open hall, 1m for corridor and 0.2m for office and lab. Note that the spacing was chosen to be much larger than half wavelength to ensure independent sampling. For each measurement, both the transmitter (Tx) and the receiver (Rx) were kept stationary. The exact Tx-Rx distance was physically measured and recorded. The effective sampling rate of the oscilloscope was 40.5GHz that is much higher than the Nyquist rate of the signal. The measurement system was calibrated prior to the measurements in order to obtain the pulse template and system delay. The maximum measurement distances for both offices LOS and NLOS are 26m. For the laboratory, open hall and corridor, the maximum distances are 5, 30 and 30 meters respectively.

2.3 Post Processing Algorithm

The measurement system can be viewed as a single user UWB system with a repetitive coder described in [54]. The local clocks of the transmitter and the receiver are aligned and hence OWR can be performed [55]. The transmitter emits a signal at timing $t = t_{sta}$. After pre-filtering and averaging, the waveform of the received signal can be expressed as a

superposition of the distorted and delayed replicas of the transmitted signal

$$r(t) = \sqrt{E_{tx}} \sum_{l=0}^{N_{mul}-1} \xi_l \alpha_l p_l(t - \tau_l - t_{toa}) + w(t) \quad (2.1)$$

where $\sqrt{E_{tx}}$ is the energy of the transmitted pulse, N_{mul} is the number of multipath, $p_l(t)$ is the energy-normalized monocycle of the l^{th} path, α_l, ξ_l and τ_l are the path gain, path polarity and relative propagation delay of the l^{th} path, $t_{toa} = \tau_d + \tau_{sys} + t_{sta}$ is the true TOA, τ_d is the absolute propagation delay of the direct path, τ_{sys} is the delay introduced by the system hardware, $w(t)$ represents the aggregated effects of the thermal noise and the interference, and $r(t)$ is the raw waveform which will be further processed by different algorithms to estimate TOA.

For correlation receiver, $r(t)$ is first cross-correlated with the pulse template $p_{tm}(t)$ which has unit energy and finite time support on $[0, T_{tm}]$ where T_{tm} is defined as the duration in which 99% of the energy of the pulse template is concentrated. Here the finite time support means the finite time interval span by the pulse template. T_{tm} is 1.2ns in our experiment. The template is obtained by extracting the direct path waveform from the LOS channel profile measured at $L = 2$ m with a proper windowing function. Since the pulse template $p_{tm}(t)$ has very high SNR, it can be considered as quasi-noise free. The cross-correlation between the template and the received signal is given by

$$R_{rp}(t) = \int r(\xi) p_{tm}(\xi - t) d\xi. \quad (2.2)$$

Then the CLEAN algorithm is performed on $R_{rp}(t)$ to estimate the multipath amplitude

and arrival timings. The CLEAN algorithm was first introduced in [56] and has been widely used in radio astronomy and microwave communities. It was applied to UWB channel estimation in [25]. A variant of the CLEAN algorithm is introduced in [57] and is herein employed to extract path information. The conventional CLEAN algorithm manipulates on $r(t)$ and $p_{tm}(t)$ while the algorithm proposed in [57] manipulates on correlation functions $R_{pp}(t)$ and $R_{rp}(t)$. The steps of the algorithm are summarized below:

- 1). Calculate the autocorrelation function of the local pulse template and denote it as $R_{pp}(t)$. Normalize $R_{pp}(t)$ so that its peak is equal to 1.
- 2). $R_{rp}(t)$ is referred as a dirty map. Locate the global peak in the dirty map. Scale and align $R_{pp}(t)$ to this global peak and then subtract it from the dirty map. Record the amplitude and position of this global peak as the estimated amplitude and arrival timing of a multipath.
- 3). Repeat the above cleaning step until the preset number of iterations is reached or the global peak of the residual in the dirty map is below a certain value.

Suppose that the channel estimation using the above algorithm gives a vector of path amplitude $\mathbf{x} = [x_0, x_1, \dots, x_{M-1}]$ with respect to the delay $\mathbf{t} = [t_0, t_1, \dots, t_{M-1}]$ which have been sorted according to the ascending order.

For ED receiver, $r(t)$ passes through a square law device followed by a finite time integrator. Starting from timing t_{sta} , the integrator successively integrates the output of the square law device with time interval T . The sampling rate of ED receiver is $1/T$. The

sample vector is denoted as $\mathbf{z} = [z_0, z_1, \dots, z_{N-1}]$ with the k^{th} sample given by

$$z_k = \int_{t_{sta} + kT}^{t_{sta} + (k+1)T} |r(t)|^2 dt, k = 0, 1, \dots, N-1. \quad (2.3)$$

To fairly compare with the correlation receiver, the samples are converted to $x_k = \sqrt{z_k / T}, \forall k = 0, 1, \dots, N-1$ which are associated with timing $t_k = (k + 1/2)T + t_{sta}$.

With the amplitude vector $\mathbf{x}$ for correlation and ED receivers, we take the first path crossing a predefined threshold as the direct path and estimate its arrival time as TOA of the signal. Let the largest sample be $x_{\max}$ associated with timing $t_{\max}$. The estimated TOA is

$$\hat{t}_{toa} = \left(\arg \min_k [x_k > \gamma] + 1/2 \right) T + t_{sta}, t_{\max} - \delta \leq t_k \leq t_{\max} \quad (2.4)$$

where γ is the threshold and δ is the searching duration. To reduce the false alarm triggered by the noise-only samples prior to the direct path, the following empirical setting strategy is proposed:

$$\gamma = \begin{cases} \max[\beta x_{\max}, \xi_{floor}], & \text{if } \xi_{floor} \leq x_{\max} \\ x_{\max}, & \text{if } \xi_{floor} > x_{\max} \end{cases} \quad (2.5)$$

where $\xi_{floor} = \mu_{noise} + m\sigma_{noise}$ represents a computed noise floor, μ_{noise} and σ_{noise} are the mean and standard deviation of noise-only samples, m is a parameter setting the deviation of ξ_{floor} from μ_{noise} based on σ_{noise} and it is a positive real number, and β is a parameter setting a candidate threshold level proportional to the maximum sample and it is a real number satisfying $0 \leq \beta < 1$. For correlation receiver, μ_{noise} and σ_{noise} are the mean and standard deviation of $n(t)$. For ED receiver, μ_{noise} and σ_{noise} are the mean and standard deviation of the noise-only energy samples.

(a)

(b)

(c)

Figure 2.12. Threshold selection in different SNR conditions: (a) High SNR; (b) Relatively low SNR; (c) Extremely low SNR.

Equation (2.5) implies that the final threshold setting is determined by three candidate values: the computed noise floor ξ_{floor} , the candidate threshold level βx_{max} and the maximum sample amplitude x_{max} . Figure 2.12 illustrates the threshold selection among the three candidate values in different SNR conditions. When SNR is high, the threshold is dominated by βx_{max} which rises with higher SNR and hence the false alarm rate is reduced. When SNR is relatively low, βx_{max} is small and the threshold is likely to be set by ξ_{floor} which prevents the estimation from falling into noise-only region (NR) prior to the direct path. When SNR is extremely low, the computed ξ_{floor} may be even higher than x_{max} and detection based on ξ_{floor} miss entire received signal while detection based on βx_{max} will very likely fall into noise component. In such case, our best strategy to avoid false alarm event is to take the strongest sample x_{max} as the direct path, i.e., set threshold as x_{max} . According to the above discussion, we may conclude that the threshold setting in Equation (2.5) ensures that the false alarm rate has a ceiling regardless of SNR and decreases with increasing SNR. For the following numerical study, we set $m = 15$ dB, i.e., $m = 5.6234$. Note that $\hat{t}_{toa}$ includes the delay introduced by hardware as well as propagation delay.

To compute range L using estimated TOA $\hat{t}_{toa}$ and Equation (1.2), the unknown hardware delay τ_{sys} in Equation (1.2) should be calibrated out. To achieve that, we estimate the arrival time of signal captured in a LOS profile at a reference distance L_{ref} which is physically measured. Using Equation (1.2) and setting $\vartheta = 1$, $\hat{L}$ and $\hat{L}_{ref}$ are expressed as

$$\begin{cases} \hat{L} = c(\hat{t}_{toa} - t_{sta} - \tau_{sys}) \\ \hat{L}_{ref} = c(\hat{t}_{ref} - t_{sta} - \tau_{sys}) \end{cases} \quad (2.6)$$

Solving the two equations gives

$$\hat{L} = c(\hat{t}_{toa} - \hat{t}_{ref}) + \hat{L}_{ref}. \quad (2.7)$$

Since the captured reference profile can be considered as quasi noise free due to the high SNR condition, the estimated arrival time $\hat{t}_{ref}$ may be approximated as true timing t_{ref} leading to $\hat{t}_{ref} \approx t_{ref}$ and $\hat{L}_{ref} = L_{ref}$. By using these approximations, Equation (2.7) becomes

$$\hat{L} = c(\hat{t}_{toa} - t_{ref}) + L_{ref}. \quad (2.8)$$

In this chapter, L_{ref} is chosen as $2m$. The ranging error is defined as

$$\varepsilon = \hat{L} - L. \quad (2.9)$$

2.4 Measurement Results Analysis

There are four major sources causing TOA estimation errors in a UWB system when the signal propagates in indoor environments:

(1). Measurement imprecision. The measurement error is attributed to imprecise visual inspection of L or/and L_{ref} , the instability of the system caused by jitter and drift of clock, and the quantization error of analog-to-digital converters (ADCs) in the oscilloscope.

(2). Channel impairment. The blockages between reference node and mobile node may severely attenuate, distort and delay the direct path signal in NLOS scenario [58], [59]. Moreover, the pulse shape of the direct path is distorted if it is overlapped by the following multipath.

(3). Noise. If the response to a single transmitted UWB pulse is modeled as noise plus signal component impaired by the channel, it can be classified into two regions: NR and SR. The length of SR is equivalent to the channel delay spread. NR contains only noise while SR may contain signal components plus noise. A large noise sample in NR may be erroneously detected as direct path by threshold-based TOA estimators. For estimators like those reported in [34] and [42], the noise energy in the decisions variables which is accumulated over the integration length of SR also degrades the performance especially when the length of the integration is long.

(4). Interference. Besides the noise, various interference sources such as ISI, WBI and NBI contribute to the ranging errors as well.

In the above error sources, as indicated by Figure 2.11 and discussions in Section 2.2.2, the interference and noise can be effectively suppressed by averaging operation together with equivalent time sampling. As shown in the next section, using the proposed threshold setting strategy in (2.5), large negative range errors are absent in the ranging results. This observation indicates that the effects of the suppressed interference and noise are trivial since severe interference and noise can cause false alarm events leading to large negative ranging errors. In addition, due to careful geometry calibration performed during the measurement process, errors contributed by measurement imprecision can be considered as less than 1 cm which is small as compared to the error caused by the channel impairment. Therefore, the dominant error source for the ranging results present in the following discussion is channel impairment.

To gain insight into the behavior of ranging operation with TOA estimator, ranging errors should be properly characterized and modeled. In this section, the ranging error

of correlation and ED receivers are characterized and compared using measured profiles and algorithms with proposed threshold setting strategy in (2.5). This is followed by investigating the effects of system parameter settings on the ranging performance. Finally, the ranging error is statistically modeled.

2.4.1 Ranging Error Characterization

Both root mean square error (RMSE) and mean absolute error (MAE) can be used as ranging performance criteria [45], [60]. RMSE and MAE are defined as

$$RMSE = \sqrt{E[\varepsilon_i^2]} = \sqrt{\sum_{i=0}^{N_{tot}-1} (\varepsilon_i^2 / N_{tot})} \quad (2.10)$$

and

$$MAE = E[|\varepsilon_i|] = \sum_{i=0}^{N_{tot}-1} |\varepsilon_i| / N_{tot} \quad (2.11)$$

where N_{tot} is the total number of measured profiles used to generate the RMSE and ε_i is the ranging error of the i^{th} profile. With the Cauchy's inequality (see 3.2.9 in [61]), it can be readily shown that

$$\frac{1}{N_{tot}} (|\varepsilon_0| + |\varepsilon_1| + \dots + |\varepsilon_{N_{tot}-1}|)^2 \leq (\varepsilon_0^2 + \varepsilon_1^2 + \dots + \varepsilon_{N_{tot}-1}^2). \quad (2.12)$$

Equation (2.12) indicates that $MAE \leq RMSE$. Hence with the same set of $\{\varepsilon_i\}$, MAE is always smaller than RMSE.

Table 2.1 shows the RMSE and MAE evaluated with $\beta=0.2$, $T=1\text{ns}$ and $\delta = 50\text{ ns}$ for different environments. Both ED and correlation receivers are able to achieve centimeter

Figure 2.13. CDF of ranging errors with $\beta = 0.2$, $\delta = 50\text{ns}$, and $T = 1\text{ns}$.

Table 2.1. Performance of correlation and ED receivers in different indoor environments

Environment	Number of profiles	Maximum measurement distance (m)	Correlation with sampling rate 40.5 GHz		ED with sampling rate 1 GHz	
			RMSE (m)	MAE (m)	RMSE (m)	MAE (m)
Office LOS	326	26	0.023	0.018	0.101	0.084
Office NLOS	977	26	4.800	2.261	5.504	2.650
Laboratory	271	5	0.084	0.031	0.172	0.132
Open Hall	61	30	0.021	0.017	0.100	0.084
Corridor	31	30	0.019	0.015	0.102	0.085

Figure 2.14. Scatter plot of ranging errors in NLOS environment for correlation receiver (left plot) and ED receiver (right plot) with $\beta = 0.2$, $\delta = 50\text{ns}$, and $T = 1\text{ns}$.

accuracy in LOS scenarios regardless of type of environments. The estimated ranging performance of office NLOS is much worse than that of LOS environments. The reason is that the direct path in the NLOS may experience severe attenuation, distortion and delay effects due to the blockages.

Figure 2.13 shows the cumulative distribution functions (CDFs) $F(|\varepsilon|)$ of ranging errors for correlation and ED receivers with $\beta = 0.2$, $T = 1\text{ns}$ and $\delta = 50\text{ns}$. Since the width of the UWB pulse used in this measurement is approximately 1.3ns corresponding to an electrical length of 0.39m , we may declare the occurrence of an incorrect detection of direct path when $|\varepsilon| > 0.39\text{m}$. The correct detection probability $P(|\varepsilon| \leq 0.39)$ of both correlation and ED receivers in the LOS scenarios are very close to 1. In NLOS, the probabilities

for both correlation and ED receivers are only approximately 0.66. Within the interval $|\varepsilon| \in [0, 0.39]\text{m}$, for the same type of environment, the CDF curves of correlation receiver lies above those of ED receiver due to the fact that correlation receiver has higher time resolution than ED receiver.

Figure 2.14 shows the scatter plot of the ranging error ε in office NLOS environment. For both the correlation and ED receivers, the number of large errors increases as L increases. The absence of negative errors with large absolute values implies that the false alarm rate is negligible due to the threshold setting strategy in Equation (2.5). Thus the channel is the main source causing large errors.

2.4.2 Effects of Parameter Settings on Ranging Performance

The results in Table 2.1 and Figure 2.13 do not lead to the immediate conclusion that correlation receiver always outperforms ED receiver since those results are computed with certain system parameter settings. When the settings such as sampling rate are changed, the correlation receiver may have worse performance as shown in this section.

The performance of correlation receiver is heavily dependent on the sampling rate. In the experiment, the captured waveform is oversampled by the oscilloscope with sampling rate of 40.5GHz that is much higher than the Nyquist rate. To investigate the effect of sampling rate on correlation receiver, $r(t)$ and $p_{tm}(t)$ are downsampled by keeping every q^{th} sampling starting with the first sample. We consider $q = \{1, 2, \dots, 8, 9, 10, 20, \dots, 90, 100\}$. The signal frequency band of the UWB pulse used in the measurement is approximately [3.5GHz,

Figure 2.15. RMSE versus sampling rate for correlation receiver in the office environment with $\beta = 0.2$, and $\delta = 50\text{ns}$.

6.4GHz] if cutoff frequencies are defined as the frequencies where the power is -10dB lower than the peak power in PSD plot. Hence with sampling rate $f_s = (40.4/q)\text{GHz}$, $\forall q \geq 4$, the waveform is undersampled. Under-sampling induces distortion on the pulse shape of the template as well as the received signal. When the sampling is too low ($q \geq 30$), the pulse template becomes a delta function.

Figure 2.15 depicts the RMSE in office LOS environment as the sampling rate f_s varies. It is observed that the performance deteriorates as the sampling rate decreases. The reason is that decreasing of sampling rate makes the pulse distortion more severe and hence the correlation operation captures less energy of the direct path. In addition, lower sampling rate reduces the time resolution of the system and thereby reduces the path resolvability

Figure 2.16. RMSE versus integration time for ED receiver in the office environment with $\beta = 0.2$, and $\delta = 50\text{ns}$.

of the ranging algorithm. Note that for ED receiver, the sampling rate is $1/T$. If sampling rate is 1GHz, RMSE of the correlation receiver is 1.99m in LOS which is much worse than that of ED receiver which is only 0.10m as shown in Table 2.1. By curve fitting, the relationship between RMSE and sampling rate f_s for a correlation receiver is modeled as

$$RMSE = 10^{(b_0 \exp(b_1 f_s) + b_2)}. \quad (2.13)$$

The sampling rate in the above model has units of GHz. The model coefficients are obtained by curve fittings. For LOS, the coefficients are $b_0 = 2.386$, $b_1 = -0.262$ and $b_2 = -1.503$. For NLOS, the coefficients are $b_0 = 0.336$, $b_1 = -0.066$ and $b_2 = 0.662$.

Figure 2.16 shows the relationship between RMSE and integration time of ED receiver for

Figure 2.17. RMSE versus β in the office environment with $\delta = 50\text{ns}$ and $T = 1\text{ns}$.

office environment. In the LOS case, RMSE monotonically increases with integration time due to the reduction of time resolution of the system. In NLOS, surprisingly, RMSE improves with larger T within $[0\text{ns}, 5\text{ns}]$. A possible explanation for this counter-intuitive observation is given as follows. Longer integration time means the energy sample is expected to be larger. Such accumulative effect benefits more on the samples with smaller relative delays since earlier incoming paths are statistically stronger than the later incoming paths. Hence the probability of detecting the earlier incoming paths increases. Note that earlier incoming paths are closer to the direct path. When this accumulative effect has heavier impact on the ranging accuracy than the degradation effects caused by the reduced time resolution, the performance improves. At $T = 5\text{ns}$, RMSE curve hits a floor and remains unchanged until the integration time increases to $T = 11\text{ns}$, beyond which the curve goes up again due to the domination of

reduced time resolution effect on the ranging accuracy.

Figure 2.17 depicts the ranging performance as β varies. In LOS, when β is small, the variation of RMSE is negligible. The reason is that according to threshold setting strategy in (2.5), when β is too small, the threshold is determined by the parameter m which is unchanged. When β increases to 0.02, the threshold starts to be determined by $\beta x_{\max}$ which reduces the false alarm events which incurs large negative errors and hence dramatically reduces the RMSE. As β increase beyond 0.5, the performance starts to deteriorate drastically since $\beta x_{\max}$ becomes too large which likely misses the direct path and falsely detect the later coming path as direct path. In NLOS, RMSE monotonically increases with increasing of β . Note that in both LOS and NLOS scenarios, $\beta = 1$ is not the optimum threshold setting since the direct path may not be the strongest path even in LOS. The adjacent multipaths may constructively overlap with each other to create a path stronger than the direct path. It is also observed that the ranging accuracy is less sensitive to the threshold variation in NLOS than in LOS.

Figure 2.18 shows the impact of observation window length on the ranging performance. In both LOS and NLOS, increasing the window length lowers the probability of missing the direct path while increases false alarm rate. Since the threshold setting strategy in equation (2.5) effectively prevents the occurrence of false alarm event, RMSE is monotonically decreased with increasing window length. When the window length is widened beyond a transition point that is about $\delta = 30$ ns for LOS and $\delta = 100$ ns for NLOS, the performance reaches the saturation region. Generally, NLOS requires a wider observation window since the

Figure 2.18. RMSE versus δ for ED receiver in office environment with $\beta = 0.2$ and $T = 1$ ns.

probability that the direct path is further apart from the strongest path is higher in NLOS than in LOS.

2.4.3 Ranging Error Modeling

The ranging error model for LOS is relatively simple and usually modeled as a Gaussian distribution with all parameters independent of Tx-Rx distance. The ranging error models for NLOS are much more complicated. In this section, we will focus on the ranging error modeling for NLOS scenario and all the parameters extracted in the section uses the ranging results extracted from NLOS profiles obtained in the measurement campaign with the TOA estimation algorithm presented in Section 2.3. The algorithm parameters are set to be $T = 1$ ns, $m = 15$ dB, $\beta = 0.2$ and $\delta = 50$ ns.

Ranging error model can be used to simulate the performance of a ranging system or provide statistics to the positioning or tracking algorithm [63]. There are two methods to model the ranging error in NLOS environment. The first approach is to model the ranging error as combinations of some random variables with each assigned explicit physical meaning [62]. The second approach is to model the probability density function (PDF) of the ranging errors. In [63] and [64], the PDF of normalized ranging errors in NLOS are modeled as a mixture of Gaussian distribution and exponential distribution. The parameters in the mixture distribution are assumed to be distance independent. Here we use the second method. Unlike the model in [63] and [64], we model the PDF of the ranging error but not the normalized ranging error as a mixture distribution with the parameters being distance dependent. The motivation is to bypass the normalization process and explicitly relate the model coefficients to the distance L .

The absence of negative errors with large absolute values implies that the false alarm rate is negligible due to the threshold setting strategy given in (2.5). Thus the channel impairment is the main source causing large errors. The PDF of ε for a particular distance L is given by

$$f(\varepsilon; L) = (1 - W(L)) \frac{1}{\sigma_G \sqrt{2\pi}} e^{-\frac{\varepsilon^2}{2\sigma_G^2}} + \frac{\varepsilon + |\varepsilon|}{2} W(L) \lambda(L) e^{-\lambda(L)\varepsilon} \quad (2.14)$$

where $W(L)$ is a distance-dependent weighting factor, σ_G is the standard deviation of Gaussian distribution, and $\lambda(L)$ is the distance-dependent parameter for exponential distribution. σ_G is assumed to be independent of distance. The Gaussian distribution accounts for the effects of measurement noise and the exponential distribution is related to the impact of channel effects.

The mixture distribution with PDF expressed in (2.14) has mean

$$\mu = \frac{W(L)}{\lambda(L)} \quad (2.15)$$

and variance

$$\sigma^2 = (1 - W(L))\sigma_G^2 - \left(\frac{W(L)}{\lambda(L)} \right)^2 + \frac{2W(L)}{\lambda^2(L)}. \quad (2.16)$$

By solving (2.15) and (2.16), we have

$$\lambda(L) = \frac{-(\mu^2 + \sigma^2 - \sigma_G^2) + \sqrt{(\mu^2 + \sigma^2 - \sigma_G^2)^2 + 8\mu^2\sigma_G^2}}{2\mu\sigma_G^2}. \quad (2.17)$$

Substituting (2.17) into (2.15), $W(D)$ is computed as

$$W(L) = \frac{-(\mu^2 + \sigma^2 - \sigma_G^2) + \sqrt{(\mu^2 + \sigma^2 - \sigma_G^2)^2 + 8\mu^2\sigma_G^2}}{2\sigma_G^2}. \quad (2.18)$$

To estimate σ_G , we form a data set S_{neg} containing only the negative errors. We assume that the information of the Gaussian distribution can be fully obtained by exploiting the statistical property of the elements in S_{neg} . Using the symmetry property of Gaussian distribution, a new data set S_{sym} is constructed with the elements from S_{neg} and the inverse of them. The total number of points in S_{sym} is denoted by N_G . Denote the mean of the elements in S_{sym} as μ_G . Due to the symmetry property of those elements, $\mu_G = 0$. σ_G is then estimated as the variance of the elements in S_{sym} :

$$\sigma_G = \sqrt{\frac{1}{N_G - 1} \sum_i (\varepsilon_i - \mu_G)^2}, \varepsilon_i \in S_{sym}. \quad (2.19)$$

Figure 2.19. Error mean vs. distance for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$, and $\delta = 50\text{ns}$.

Figure 2.20. Error variance vs. distance for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$, and $\delta = 50\text{ns}$.

Figure 2.21. Curve fitting for W for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$, and $\delta = 50\text{ns}$.

Figure 2.22. Curve fitting for λ for NLOS measurements with $T = 1\text{ns}$, $m = 15\text{dB}$, $\beta = 0.2$, and $\delta = 50\text{ns}$.

An intuitive way to obtain $W(L)$ and $\lambda(L)$ is to estimate their function values from a large number of channel realizations for L . However, this method is impractical since the number of channel profiles associated with the same distance L is usually too few to generate statistics. To overcome this problem, we propose a low-complexity method to estimate $W(L)$ and $\lambda(L)$. First, the set of true distances $\{L_i\}$ is sorted and divided into six subsets $\mathcal{B}_j, j = 1, \dots, 6$ which are defined by $\mathcal{B}_1 = \{L_i \leq 5\}$, $\mathcal{B}_2 = \{5 \leq L_i \leq 7.5\}$, $\mathcal{B}_3 = \{7.5 \leq L_i \leq 10\}$, $\mathcal{B}_4 = \{10 \leq L_i \leq 12.5\}$, $\mathcal{B}_5 = \{12.5 \leq L_i \leq 15\}$, $\mathcal{B}_6 = \{15 \leq L_i \leq 17.5\}$. This is indeed a quantization process of distance axis. Every subset contains sufficient number of profiles (≥ 80) to generate statistics. Each subset is associated with a mean distance calculated by

$$L_{mean}^{(j)} = \frac{1}{M_j} \sum_i L_i, L_i \in \mathcal{B}_j \quad (2.20)$$

where M_j is the number of elements in the set $\mathcal{B}_j$. For each subset, its variance σ_j^2 and mean μ_j are then calculated from the ranging errors associated with the distance values in that subset. Figure 2.19 and Figure 2.20 suggest that the error mean μ_j and the error variance σ_j^2 generally increase as the distance L_{mean} increases. The computed μ_j and σ_j^2 are taken as the mean and variance associated with a Tx-Rx distance $L = L_{mean}^{(j)}$. By using (2.17) and (2.18), λ and W can be computed. As shown in Figure 2.21 and Figure 2.22, by curve fitting of λ and W values against their associated distance L , the relationship between λ and L for both correlation and ED receivers can be modeled as

$$\lambda(L) = B_1 / (\exp(B_2 L) + B_3), \quad (2.21)$$

and W and L are related by

$$W(L) = 1 / (C_1 \exp(-C_2 L) + 1). \quad (2.22)$$

As L increases, the direct path strength decreases due to longer traveling distance and more obstructions along the direct path. The probability that the multipath is detected as the direct path is increased and hence W increases. Moreover, as distance increases, chances that the earlier arrival paths are blocked or severely attenuated are increased as well. Hence, it is more likely that the multipath with larger relative delay will be detected as the direct path. This phenomenon is explained by the decrease of λ . By curve fitting, the coarse estimation of the parameters in (2.21) and (2.22) may be obtained. Due to the quantization of distance axis, the estimated model parameters are not very accurate. Hence, the fine estimation of the parameters is obtained by tuning the parameters to minimize the difference between the empirical CDF function (generated from empirical ranging error set $\{\varepsilon_i\}, 0 \leq i \leq N_{tot} - 1$) and the following model based CDF function

$$F(\varepsilon) = \frac{1}{N_{tot}} \sum_{i=0}^{N_{tot}-1} F(\varepsilon; L_i) \quad (2.23)$$

where $F(\varepsilon; L_i) = \int_{-\infty}^{\varepsilon} f(\xi; L_i) d\xi$. Figure 2.23 indicates the close match between the two functions after fine tuning of the parameters. The final estimated parameters for correlation and ED receivers are given in Table 2.2.

Figure 2.23. Comparison of CDFs generated from measurement data and the proposed model.

Table 2.2. Fitting coefficients for ranging error model

	B_1	B_2	B_3	C_1	C_2
Correlation	2.459×10^6	1.358	9.994×10^5	3.915×10^5	0.9701
ED	1.559×10^7	1.544	9.986×10^6	1.152×10^3	0.5544

2.5 Conclusion

In this Chapter, the ranging performance of correlation and ED receivers is evaluated and compared based on measured channel responses captured in four types of indoor environments. The results show that ranging performances of correlation and ED receivers are heavily dependent on the system parameters setting, especially in LOS environment. The correlation receiver outperforms ED receiver only if the sampling rate of correlation receiver is much higher than that of ED receiver. Further studies also reveal that the ranging error model of both types of receivers can be described by a unified model with distance-dependent parameters.

Chapter 3

Delay-dependent Threshold Selection Technique

3.1 Introduction

Achieving accurate estimation of TOA is critical to UWB ranging performance. The threshold-based TOA estimation method is attractive because of its simple hardware implementation. The threshold level may be set solely based on the noise level or it can be selected as a normalized value between maximum and minimum energy samples [44], [65]. To capture the information of the channel statistics and SNR, the kurtosis based normalized threshold selection proposed in [45] is proven to be robust over channel condition variation. The above methods are referred to as delay-independent (DI) threshold selection since the threshold value for individual channel realization is independent of propagation delay. In this chapter, we introduce a DD threshold selection method that is applicable to TOA estimation. As shown in Figure 3.1, DD sets the threshold as a function of propagation delay. The development of DD is motivated by the fact that the strength of the direct path increases with shorter Tx-Rx distance. Increasing the threshold values for smaller delays can significantly

Figure 3.1. DD and DI threshold selections for ED receiver.

lower the false alarm probability when Tx-Rx distance is relatively large. Simulation results show that significant performance improvement can be achieved with DD as compared to DI threshold. The proposed DD threshold can be applied to both OWR and TWR. The analysis in this chapter is based on OWR while the extension to TWR is straightforward.

3.2 Signal Model

Let us consider an OWR scenario in which both the transmitter and the receiver have been synchronized to a common clock so that OWR can be performed [55]. Let Tx-Rx distance be L and $L \sim \mathcal{U}(0, L_{\max})$ where $\mathcal{U}(\cdot)$ denotes uniform distribution and $L_{\max}$ is the maximum detectable distance. At timing $t = t_{sta}$, the transmitter sends out a ranging signal

$$s(t) = \sqrt{E_{tx}} p_{tx}(t) \quad (3.1)$$

where $p_{tx}(t)$ is the transmitted signal waveform with normalized energy and E_{tx} is the total energy of that signal. Although a single UWB pulse is adopted as $p_{tx}(t)$ here, the proposed method can be readily extended to the case when the transmitted signal is a pulse train.

Assume that the receiver has acquired the knowledge of E_{tx} by exchanging message with the transmitter [66].

The transmitted signal propagates through the channel via different paths and generates a large number of echoes. Among these signal paths, the first arriving path plays a crucial role in the TOA estimation process. A number of TOA estimators like the GML estimator and the threshold-based estimator reported in [28] and [45] estimate the arrival time of the first path as the TOA of the received signal. Similar first-path detection strategy is also adopted in this chapter.

The received channel response $h(t)$ is the convolution of transmitted signal $s(t)$ with the channel impulse response $h_{CIR}(t)$ given in Equation (1.3)

$$h(t) = s(t) \otimes h_{CIR}(t) = \sum_{i=0}^{N_{mul}-1} \sqrt{E_{tx}} \xi_i \alpha_i p_i(t - \tau_i) \quad (3.2)$$

where $\xi_i \in \{\pm 1\}$, α_i and τ_i are the path polarities, instantaneous path amplitude gain and propagation delay of the i^{th} path respectively, N_{mul} is the number of multipaths, and $p_i(t) = p_{tx}(t) \otimes h_{CIR-i}(t)$ is the pulse from the i^{th} path with normalized energy.

The received ranging signal $r(t)$ after a zonal bandpass filter with bandwidth W can be expressed in a real passband form

$$\begin{aligned} r(t) &= h(t) + w(t) \\ &= \sum_{i=0}^{N_{mul}-1} \sqrt{E_{tx}} \xi_i \alpha_i p_i(t - t_{sta} - \tau_i) + w(t) \end{aligned} \quad (3.3)$$

where $p_i(t)$ is the received monocycle waveform of the i^{th} path and has normalized energy,

and $w(t)$ is a zero mean Gaussian random process with variance $\sigma_w^2 = WN_0$ where N_0 is the single-sided power spectral density. Let the expected arrival time of the signal be $t_{toa} = t_{sta} + L/c$ where c is the velocity of light.

The low complexity ED receiver is considered here. It consists of a square law device followed by an integrator. The integrator performs the sampling operation by successively integrating the squared signal over time interval T . Hence T is not only the integration time but also the sampling period. Let M be the total number of samples and $\mathcal{B} = \{0, 1, \dots, M-1\}$ be the set of sample indices. The output samples can be expressed as

$$z[k] = \int_{t_{sta}+kT}^{t_{sta}+(k+1)T} |r(t)|^2 dt, k \in \mathcal{B}. \quad (3.4)$$

Let $N_w = \lfloor \tau_0 / T \rfloor$ where $\lfloor x \rfloor$ denotes the largest integer that is smaller than x . Let $N_{ds} = T_{ds}^{(true)} / T$ where $T_{ds}^{(true)}$ is the true length of maximum channel delay spread. Note that the samples with indices smaller than $N_w - 1$ or larger than $N_w + N_{ds}$ contain only noise energy while the rest of samples may contain signal component. Urkowitz has shown that if the sample contains only noise, the decision variables $2z[k]/N_0$ follow a central chi-square distribution with degree of freedom $D = 2TW$, while if the sample contains signal component, the decision variables $2z[k]/N_0$ follow non-central Chi-square distribution with degree of freedom $D = 2TW$ and noncentral parameter $\lambda_k = \sqrt{2E_k / N_0}$ where

$$E_k = \int_{t_{sta}+kT}^{t_{sta}+(k+1)T} |h(t)|^2 dt \text{ is the signal energy in the } k^{th} \text{ sample [67].}$$

The CDFs of the noise-only samples are

$$F_{z[k]}(z_k) = P(z[k] \leq z_k) = 1 - \exp\left(\frac{-z_k}{N_0}\right) \sum_{l=0}^{D/2-1} \frac{1}{l!} \left(\frac{z_k}{N_0}\right)^l. \quad (3.5)$$

The mean and variance of these noise-only samples are

$$\begin{cases} \mu_k = TWN_0 \\ \sigma_k^2 = TWN_0^2 \end{cases}. \quad (3.6)$$

The CDF of the samples that may contains signal energy is

$$F_{z[k]}(z_k) = 1 - Q_{(D/2)}\left(\lambda_k, \sqrt{2z_k/N_0}\right) \quad (3.7)$$

where $Q_x(y_1, y_2)$ is generalized Marcum's Q function, which is defined by

$$Q_m(x_0, x_1) = Q_1(x_0, x_1) + \exp\left(-\frac{x_0^2 + x_1^2}{2}\right) \sum_{l=1}^{m-1} \left(\frac{x_1}{x_0}\right)^l I_l(x_0 x_1). \quad (3.8)$$

Here $I_l(x)$ is the l^{th} order modified Bessel function of the first kind, which can be represented as

$$I_l(x) = \sum_{g=0}^{\infty} \frac{(x/2)^{l+2g}}{g! \Gamma(l+g+1)}, x \geq 0 \quad (3.9)$$

where $\Gamma(\bullet)$ is Gamma function. The mean and variance of these samples are

$$\begin{cases} \mu_k = TWN_0 + E_k \\ \sigma_k^2 = TWN_0 + 2N_0 E_k \end{cases}. \quad (3.10)$$

For a general threshold-based TOA estimator for the above ED receiver, starting from the first sample with index $k=0$, the samples are successively compared with the respective threshold values $\gamma[k]$, $\forall k \in \mathcal{B}$. The arrival time of the first sample crossing the respective threshold value within time interval $[t_{sta}, t_{sta} + L_{\max}/c]$ is estimated as TOA, i.e.,

$$\hat{t}_{toa} = t_{sta} + T \left(\arg \min_{k \in B_1} (z[k] \geq \gamma[k]) + 1/2 \right) \quad (3.11)$$

where $B_1 = B \cap \{kT + T/2 \leq L_{\max}/c\}$. If no crossing event occurs within $[t_{sta}, t_{sta} + L_{\max}/c]$, TOA is estimated as $t_{toa} = t_{sta} + 0.5L_{\max}/c$. The TOA estimation error is defined as

$$\varepsilon = \hat{t}_{toa} - t_{toa} \quad (3.12)$$

and the RMSE is used as performance criterion which is defined by Equation (2.10).

3.3 Delay Independent Threshold Selection

For DI threshold setting methods, all threshold values are equivalent to a value independent of delay, i.e., $\gamma[0] = \gamma[1] = \dots = \gamma[M-1] = \gamma$. The threshold value γ may be selected solely based on noise sample, i.e.,

$$\gamma = \mathcal{A}N_0 \quad (3.13)$$

where $\mathcal{A}$ is a constant for different channel realizations. This simple setting strategy is able to adapt to the variation of noise condition and accommodate the dynamic change of noise level to a certain extent. However, it could not adapt to the change of channel condition.

In [44], a normalized threshold method is proposed as

$$\gamma = \min \{z[k]\} + \vartheta_{trd} (\max \{z[k]\} - \min \{z[k]\}), \quad (3.14)$$

where $\vartheta_{trd} \in [0,1]$ is a constant, γ is selected to minimize the MAE defined by (2.11) for a particular received SNR value. The estimation of SNR is not trivial and the optimal threshold

setting may vary for different channel realizations at the same SNR. A normalized threshold selection based on kurtosis value is developed to overcome this problem [45]. Let z represent an arbitrary sample that may be modeled as a Gaussian variable with unknown mean and variance. The kurtosis of this channel realization relative to Gaussian distribution is defined as

$$\kappa = \frac{\mathbb{E}[z^4]}{\mathbb{E}^2[z^2]} - 3. \quad (3.15)$$

Let $f(z)$ denote function of z . The expectation $\mathbb{E}[f(z)]$ is performed over all possible samples. Practically, $\mathbb{E}[f(z)]$ may be replaced by an averaging operation, i.e.,

$$\mathbb{E}[f(z)] = \frac{1}{M} \sum_{k=0}^{M-1} f(z_k). \quad \text{The kurtosis values are computed over a number of channel}$$

realizations generated with discrete SNR values covering the dynamic range of the receiver under consideration, and the results are rounded to logarithmic integers. The channel realizations with the same logarithmic integer value are grouped into a set. The MAE of channel realization in each set is then evaluated with different normalized threshold settings. The threshold setting leading to the minimum MAE is referred as optimum normalized threshold. The relationship between the optimum normalized threshold and the logarithmic integers is modeled by mathematical equation. Any new channel realization just evaluates its own kurtosis value that is inserted into the equation to select the optimum normalized threshold. The TOA is then estimated with the selected threshold for the dedicated channel realization. The kurtosis based threshold selection is able to adapt to channel condition variation very well since it implicitly accounts for both the noise change and the individual channel realization.

Figure 3.2. Block diagram of ED receiver employing DD technique.

3.4 Delay Dependent Threshold Selection

Figure 3.2 shows the ED receiver employing DD technique. The threshold $\gamma(\tau)$ for DD is a function of delay which is defined as $\tau = t - t_{sta}$. The threshold values are defined as

$$\gamma[k] = \gamma(\tau = kT + T/2), k = 0, 1, \dots, M-1. \quad (3.16)$$

Note that DD indeed subsumes DI as a special case. Any function that monotonically decreases as delay increases can be used as DD threshold function. Here we heuristically choose an exponential function which is given by

$$\gamma(\tau) = aN_0 \exp(-b(\tau - \tau_{ref})), \tau \in [0, L_{\max}/c] \quad (3.17)$$

where a and b are design parameters, the unit of b is ns^{-1} , $\tau_{ref} = L_{ref}/c$ and L_{ref} is a reference distance which is set to $1m$ in the following analysis. Note that the above threshold functions are able to adapt to the noise level since they are proportional to N_0 .

Joint optimization of parameter a and b for all possible E_{tx} is computation intensive. Hence we propose the following setting strategy to reduce the computational complexity. Let

Figure 3.3. Parameter a versus δ_{sn} for different degree of freedom.

δ_{sn} be a value typically drawn from $10^{-5} \sim 10^{-10}$ which is small enough to ensure that the threshold function has sufficient noise suppression capability. The parameter a is obtained by numerically solving the equation

$$P(z_w \geq \gamma(\tau = \tau_{ref})) = P(z_w \geq aN_0) = \delta_{sn} \quad (3.18)$$

where z_w is an arbitrary energy sample containing only noise component. With (3.3),

$P(z_w \geq aN_0)$ can be derived as

$$P(z_w \geq aN_0) = 1 - P(z_w \leq aN_0) = \exp(-a) \sum_{l=0}^{D/2-1} \frac{1}{l!} a^l = \delta_{sn}. \quad (3.19)$$

Although no closed form solution for a in the above equation, a can be readily solved by numerical method due to the limited degree of freedom D . As demonstrated in Figure 3.3, for

a fixed degree of freedom, increasing δ_{sn} results in smaller a . Given any δ , larger degree of freedom leads to larger a .

After a is set, b is optimized via simulation during system design phase. Note that when $b = 0$, the DD threshold become a DI threshold solely based on noise level, i.e., $\gamma(\tau) = aN_0$. We also assume that the receiver knows if the channel is LOS or NLOS by applying the identification methods such as the one reported in [42] so that the predetermined design parameters for the respective scene can be applied. The sensitivity of the designed parameters on the variation of channel conditions will be discussed in the following section.

False alarm probability is one of the crucial performance metrics for TOA estimator. The false alarm event occurs when the noise-only sample prior to the direct path sample is erroneously chosen as direct path. The false alarm probability of DD setting method is derived as follows. The direct path signal component may be captured by a few samples. Let q_j be the conditional false alarm probability when the j^{th} sample is the first sample containing the direct path sample. For $j = 0$, the false alarm will never occur because no noise-only sample exists prior to the direct path sample and thereby $q_j = 0$. For $j \neq 0$, q_j can be expressed as

$$q_j = 1 - \prod_{k=0}^{j-1} F_{z[k]}(\gamma[k]), j = 1, 2, \dots, N_L \quad (3.20)$$

where $N_L = \lceil L_{\max} / (cT) \rceil$ and $\lceil \bullet \rceil$ denotes ceiling operation.

The mean false alarm probability is obtained by averaging over q_j for $j = 1, 2, \dots, N_L - 1$

$$P_{FA} = \frac{1}{N_L - 1} \sum_{j=1}^{N_L-1} q_j. \quad (3.21)$$

By substituting (3.3) into (3.21), we reach the mean false alarm probability for arbitrary DD threshold function

$$P_{FA} = 1 - \frac{1}{N_L - 1} \sum_{j=1}^{N_L-1} \left(\prod_{k=0}^{j-1} \left(1 - \exp \left(\frac{-\gamma[k]}{N_0} \right) \sum_{l=0}^{D/2-1} \frac{1}{l!} \left(\frac{\gamma[k]}{N_0} \right)^l \right) \right). \quad (3.22)$$

With (3.16), (3.17) and (3.22), the mean false alarm probability for the exponential DD threshold function is derived as

$$P_{FA} = 1 - \frac{1}{N_L - 1} \sum_{j=1}^{N_L-1} \left(\prod_{k=0}^{j-1} \left(1 - \exp(-A) \sum_{l=0}^{D/2-1} \frac{1}{l!} (A)^l \right) \right) \quad (3.23)$$

where $A = a \exp \left(-b \left((kT + T/2) - \tau_{ref} \right) \right)$. Formula (3.23) implies that P_{FA} for (3.17) is independent of noise spectral density N_0 and therefore the proposed threshold function is able to adapt to the variation of noise level. Figure 3.4 presents the false alarm rate P_{FA} versus b for different a when the system parameter settings are $N_L = 40$, $T = 1$ ns, and $W = 4$ GHz. As expected, larger value of a and smaller value of b result in lower P_{FA} . Given a particular b , the increment of P_{FA} in log scale is proportional to the decrement of a in linear scale. For instance, given $b = 0.01$ ns⁻¹, increasing a by 10 approximately gives $10 \log_{10}(P_{FA})$ a decrement of 29.3dB, and for $b = 0.04$ ns⁻¹ such decrement drops to 9.386dB.

Figure 3.4. P_{FA} vs. b for different a .

3.5 Simulation Results and Discussion

In this section, we evaluate the performance of DD threshold selection using threshold function given in (3.17) via simulation. Furthermore, its sensitivity to the channel parameters and threshold parameters are also investigated. The system parameters used in the simulations are $\delta_{sn} = 1 \times 10^{-9}$, $W = 4$ GHz, and $L_{\max} = 12$ m. The width of the monocycle $p_{rx}(t)$ is approximately 1 ns. For the results in this section, parameter a is computed based on the value of δ_{sn} .

Figure 3.5 presents the RMSE performance versus b for different integration time T at

Figure 3.5. RMSE versus design parameter b at $E_{tx}/N_0 = 90$ dB in CM1.

$E_{tx}/N_0 = 90$ dB in channel CM1 of IEEE 802.15.4a standard [50]. Each RMSE is generated from TOA estimation errors of 1000 channel realizations. For a given δ , there exists an optimum b_{opt} which gives the minimum RMSE for each performance curve. It is also observed that b_{opt} tends to decrease as T increases.

The optimal point b_{opt} is a function of E_{tx}/N_0 and such function is referred as design parameter function in the following discussion. This function is determined by the system parameters such as integration time as well as the channel parameters. Here we consider CM1 and CM2 corresponding to the channel models of indoor residential LOS and NLOS environments. As tabulated in [50], the channel parameters for CM1 and CM2 can be classified into path loss, power delay profile and small-scale fading parameter sets which are

Figure 3.6. The design parameter functions for different channel models.

denoted as $\mathcal{F}_{pl}^l$, $\mathcal{F}_{pdp}^l$ and $\mathcal{F}_{ss}^l$. Here the superscript l is the index of the channel, e.g., $l = 1$ for CM1. We also consider a modified CM1 channel parameter sets $\mathcal{F}_{pl}^1$, $\mathcal{F}_{pdp}^2$ and $\mathcal{F}_{ss}^2$, and a modified CM2 channel parameter sets $\mathcal{F}_{pl}^2$, $\mathcal{F}_{pdp}^1$ and $\mathcal{F}_{ss}^1$. Figure 3.6 depicts the design parameter functions of CM1, CM2, modified CM1 and modified CM2. For a particular T , the design parameter functions of the channels with identical path loss parameter sets are quite similar, while those with different path loss parameters are considerably different. Hence we may conclude that the design parameter function is primarily determined by the path loss parameters and is rather insensitive to the variation of other channel parameters. It can also be observed that for the same E_{tx}/N_0 and a , the channel with larger path loss parameters has smaller values of optimum setting b_{opt} . The reason is that when path loss increases, the

Figure 3.7. RMSE performance of different threshold selection techniques in CM1 with $T = 1\text{ns}$.

Table 3.1. Fitting coefficients for DD

Coefficient	CM1,	CM1	CM2	CM2,
	T=1ns	T=4ns	T=1ns	T=4ns
θ	0.527	0.372	0.341	0.256
β	0.0288	0.0286	0.0179	0.0177

Figure 3.8. RMSE performance of different threshold selection techniques in CM2 with $T = 1$ ns.

received direct path is weaker which means that the threshold values with larger delays have to be lowered down to reduce the missed detection probability. Hence, b_{opt} has to be lowered down accordingly.

By curve fitting, the design parameter functions of both CM1 and CM2 can be mathematically modeled as the following function

$$b_{opt} = \theta \exp(-\beta x) \quad (3.24)$$

where $x = 10 \log_{10}(E_{tx}/N_0)$ and the fitting coefficients are given in Table 3.1. It seems that θ decreases as T increases while β is rather insensitive to the change of T .

Figure 3.7. RMSE performances of different threshold selection techniques in CM1 with $T = 1$ ns.

Figure 3.7 and Figure 3.8 compare the exponential function DD set based on b_{opt} with kurtosis based normalized threshold selection and noise level based method (i.e., set $b = 0$ in Equation (3.14)) in CM1 and CM2 with $T = 1$ ns. The results show that the proposed methods are able to provide significant performance improvement in both CM1 and CM2. In addition, we also test the performance under the assumption of inaccurate parameters. It is assumed that the receiver operates in CM2 with its threshold set based on the channel parameters of modified CM2 channel. As shown in Figure 3.8, only marginal performance loss is incurred and DD still outperforms noise level based threshold selection and kurtosis based threshold selection that is optimized based on the correct parameter set. The observation implies that the performance of the proposed method is robust over the possible variation or

inaccuracy existing in the available power delay profile and small-scale fading channel parameters.

Finally, we give a practical implementation routine of DD:

1). The receiver gets the knowledge of the transmitted energy E_{tx} by exchanging message with the transmitter.

2). The receiver estimates N_0 by sampling the waveform in noise-only region. Denote these samples as $\{x_0, x_1, \dots, x_{N_x-1}\}$ where N_x is the total number of samples. The mean and

variance of these sample may be estimated as $\hat{\mu}_k = \frac{1}{N_x} \sum_{k=0}^{N_x-1} x_k$ and $\hat{\sigma}_k^2 = \frac{1}{N_x} \sum_{k=0}^{N_x-1} (x_k - \hat{\mu}_k)^2$.

The two equations in (3.4) lead to two possible simple estimators for N_0 , i.e.,

$$\hat{N}_0 = \frac{1}{TW} \hat{\mu}_k \quad (3.25)$$

and

$$\hat{N}_0 = \sqrt{\frac{1}{TW} \hat{\sigma}_k^2}. \quad (3.26)$$

3). With the knowledge of N_0 , E_{tx} and the type of channel by applying the method proposed in [42], the design parameter a and b are computed based on the design parameter function for the dedicated channel such as Equation (3.17).

4). Compare the ED output with the threshold function determined by chosen parameters.

3.6 Conclusion

In this chapter, we propose a technique to set the threshold as a function of delay instead of a single value as in the conventional methods. Simulations are performed using the IEEE 802.15.4a channel models CM1 and CM2 that are well-accepted and are devised from practical measurements as stated in the channel model reports [50]. Simulation results show that the proposed method is superior to those delay-independent methods as SNRs vary in both LOS and NLOS environments. A compact parameter setting strategy is described and the effects of system parameter settings on the system performance are investigated. The optimal parameter settings as a function of SNR are obtained by curve fitting. Simulation results show that the proposed method is insensitive to the possible variation or inaccuracy existing in the available power delay profile and small-scale fading channel parameters.

Chapter 4

Least Square Time of Arrival Estimator for UWB Backscattering RFID System

4.1 Introduction

Backscattering RFID system is a type of RFID system employing tags that do not generate their own signals but reflect the received signal back to the reader. As compared to the tags in other types of RFID systems, a tag of a backscattering RFID system has the merits of lower implementation complexity and less power consumption. Such systems have been widely applied to asset tracking, inventory management, health care and other fields [73]. Nowadays many RFID applications such as context-aware healthcare require location information of RFID tags with extended operating range. Such requirement, however, can not be fulfilled by the conventional backscattering RFID tag systems which use standard CW signaling and has poor distance estimation accuracy with limited operating range up to a few meters. Recently, UWB technology emerges as a viable candidate for the new generation of RFID systems. FCC defines UWB as a signal having a fractional bandwidth larger than 20% or absolute bandwidth of more than 500MHz [15]. Such enormous bandwidth provides high range accuracy and

robustness to multipath fading. In [74], the concept of UWB pseudo-random backscattering tag is introduced. The tag receives signal only in certain time intervals according to a pseudo-random time hopping sequence, slightly delays it and then reflects it back to the reader. In [75], a UWB backscattering RFID tag able to modulate the received signal by pulse amplitude modulation (PAM), pulse position modulation (PPM) or ON-OFF keying is proposed and its potential operating range/data rate trade-off in the presence of strong clutter is investigated.

Realizing accurate identification and localization function for UWB backscattering RFID systems requires TOA estimation of the tag response. However, recovering TOA information in multipath dense environment is challenging due to several reasons. First, the received signal may have low SNR due to the great path loss caused by the roundtrip propagation channel. A conventional method to improve SNR for ranging signal is to average the received signal over tens of frames for coherent receivers, and hundreds or even thousands of frames for noncoherent receivers [50]. Moreover, the TOA estimation for backscattering RFID systems is complicated by the undesired background clutter signals caused by the scatterers other than the RFID tags [23]. The clutter signals arriving earlier than the desired direct path may cause intolerable false alarm rate, while those overlapping with the direct path may severely distort its pulse shape and hence degrade the estimation accuracy. Averaging the signal over multiple frames is unable to suppress the clutter signal that has consistent waveform over different frames during sampling duration. To clean the received signal contaminated by the clutter before applying any TOA estimation, the empty-room technique and frame-to-frame techniques have been used in radar community [76]. The empty-room technique subtracts any received signal containing both target response and clutter from a channel response measured

in the absence of the target. This technique is sensitive to the environment change since any change renders the previously measured empty-room response inappropriate for further use. For the frame-to-frame technique, each measurement consists of two received signals captured at different timings. The technique subtracts one signal from the other to eliminate the clutter that is assumed to be static for each measurement but allows changes from measurement to measurement. This technique fails if the target does not move or moves a little between the received timings of the two received signals in the same measurement. Besides the empty-room and frame-to-frame techniques, it is also possible to mitigate the clutter by properly selecting the modulation sequence. In [75], it is shown that if the tag's modulation sequence used for antipodal 2-PAM signaling fulfils certain criteria, by averaging over multiple symbols, the received signal is immune to the clutter. The study, however, is carried out in the context of data communication under the assumption that perfect pulse synchronization to the first arriving path has been achieved, i.e., TOA information is known.

There are many existing TOA estimators that have been proposed for UWB systems as discussed in Chapter 1, for instance, the LS estimators proposed in [34] and [83]. These TOA estimators, however, are derived in the context of conventional UWB systems which usually involve one-way channel propagation and of which the modulations are done at the transmitter. Therefore the received channel response to a transmitted UWB pulse is associated with a single data bit. The UWB backscattering RFID system presents a different propagation scenario. It involves roundtrip channel propagation and modulation of the signal is performed at the tag that lies in the middle of the roundtrip channel [75]. Since the backscattering tag does not synchronize with the reader, such modulation incurs mismatch, i.e., the front part of a

channel response received at the tag may be modulated by one bit while its tail is modulated by the next data bit. Consequently, the received signal model of the UWB backscattering RFID systems is significantly different from that of conventional UWB systems. Furthermore, the aforementioned TOA estimators do not have clutter suppression capability that will incur severe accuracy degradation if the clutter overwhelms the desired tag response.

In this chapter, we derive an LS TOA estimator applicable for the UWB backscattering RFID system with antipodal 2-PAM based on the tag structure proposed in [75]. The estimator is able to recover the TOA information even when the reader and the tag are asynchronous. By examining the properties of the derived estimator, we find that it has inherent immunity to the clutter for arbitrary data sequence that is valuable for backscattering RFID systems operating in environments with dense scatterers. Unlike the empty-room and frame-to-frame techniques, such immunity holds without any empty-room response information whether the tag moves or keeps stationary. The simulation also indicates that the developed estimator is robust to undersampling estimation. Measurements are also conducted to test the performance of the estimator in indoor environment. To overcome the undesired jitter effects encountered in the implementation of the estimator for the practical measurement system, we also propose a robust jitter mitigating scheme and its performance is validated by the measurement.

4.2 Signal Model and System Model

Here we derive an LS TOA estimator for the UWB backscattering RFID system using tag implemented with antipodal 2-PAM modulation as in [75]. The RFID system

Figure 4.1. System block diagram.

consists of multiple readers and tags, and our object is to locate one of the tags. Each reader in the RFID system is a fully functional transceiver that transmits UWB pulses to tag and then estimates the TOA of the roundtrip channel response returned from the targeted tag. Note that the signal received by the reader contains both the signal reflected by the tags as well as the undesired clutter caused by other scatters in the surrounding environment. With the TOA metric recovered from the signal contaminated by both noise and clutter, the range between the reader and the tag can be calculated. Such range measurements from different readers can be combined to produce the estimation of the tag position. Without loss of generality, we will focus on the TOA estimation between one reader and one tag.

As illustrated in Figure 4.1, the tag under consideration consists of an antenna, an antipodal 2-PAM modulator and a bipolar sequence generator with output being $+1$ or -1 . The load condition of the tag antenna is determined by the 2-PAM modulator that is in turn controlled by the sequence generator. The polarity of the signal input to the 2-PAM modulator

is unchanged if the sequence generator outputs $+1$ and is inversed by the modulator if the sequence generator outputs -1 . The tag antenna acts like a scatterer and its scattering mechanism is composed of structural mode and antenna mode scatterings [77]. The structural mode scattering is solely determined by the physical properties of the antenna and is independent of the load condition of the antenna. Thus the signal caused by such scattering mechanism is not modulated by the tag's modulator and therefore does not carry any data information in the tag. In contrast, the signal caused by the antenna mode scattering is received by the antenna and reflected by the modulator. It is modulated by a bipolar sequence generated by spreading a ranging sequence $\{a_i = \pm 1\}$ with a PN code $\{c_j = \pm 1\}$ which has period N_f , i.e., $c_{j+N_f} = c_j, \forall j$. The PN code $\{c_j\}$ is used to suppress multi-user interference and is unique for each tag. The ranging sequence $\{a_i\}$ is periodic with period N_a . In the following discussion, the tag response is only referred to the signal produced by antenna mode scattering. The signal produced by the structural mode scattering is treated as part of clutter because similar to the environment clutter this part of signal is not modulated by the data sequence of the tag and has consistent waveform over different data frames.

To initialize the TOA estimation process, the transmitter of the reader repetitively sends UWB pulses starting at timing $t = 0$ with respect to its own clock. Every symbol consists of N_f frames with one pulse per frame. The transmitted signal is given by

$$s(t) = \sum_i \sum_{j=0}^{N_f-1} p_{tx} \left(t - (iN_f + j)T_f \right) \quad (4.1)$$

where $p_{tx}(t)$ is the transmitted UWB pulse and T_f is the frame duration.

Figure 4.2. Transmitted/received signal present at various stages of the RFID system.

After propagating through the downlink channel (from reader to tag), the transmitted signal impinges on the antenna of the backscattering tag. Letting the total number of paths in the downlink channel be L_d , the downlink channel response to $p_{tx}(t)$ is modeled as

$$h_d(t) = \sum_{l=0}^{L_d-1} p_d^{(l)}(t - \tau_d^{(l)}) \quad (4.2)$$

where $p_d^{(l)}(t)$ is the downlink channel response of the l^{th} path to the transmitted UWB pulse $p_{tx}(t)$ and $\tau_d^{(l)}$ is the associated relative propagation delay. $h_d(t)$ has a time support on $[0, T_{ds}^{(true)}]$, where $T_{ds}^{(true)}$ is the true maximum delay spread of one-way channel. Suppose that

τ'_{tg} is the processing delay from the tag's antenna to the modulator. The noise-free signal input to the modulator in the tag is expressed as

$$r_{tg}(t) = \sum_i \sum_{j=0}^{N_f-1} h_d \left(t - (iN_f + j)T_f - \tau_d - \tau'_{tg} \right) \quad (4.3)$$

where τ_d is the propagation delay of the direct path of the downlink channel. As illustrated in Figure 4.2 (b), the modulation function of the antipodal 2-PAM modulator is given by

$$\Phi(t) = \sum_i \sum_{j=0}^{N_f-1} a_i c_j \Omega \left(\frac{t - (iN_f + vN_f + j + u)T_f + (T_f - \rho)}{T_f} \right) \quad (4.4)$$

where v and u are integers reflecting the ranging symbol offset and PN code offset between the reader and the tag, $\rho \in [0, T_f)$ is the timing offset between the clock of the reader and the tag, and $\Omega(t)$ is a rectangular window with amplitude being one in $[0, 1]$ and zero elsewhere.

Equivalently, (4.4) may be also written as

$$\Phi(t) = \sum_i \sum_{j=0}^{N_f-1} \Lambda_{i,j} \left(t - (iN_f + vN_f + j + u)T_f \right) \quad (4.5)$$

with

$$\Lambda_{i,j}(t) = a_i c_j \Omega \left(\frac{t}{\rho} \right) + a_{i+[(j+1)/N_f]} c_{[j+1]_{N_f}} \Omega \left(\frac{t - \rho}{T_f - \rho} \right) \quad (4.6)$$

which has time support on $[0, T_f]$.

The output of 2-PAM modulator is expressed as

$$s_{tg}(t) = r_{tg}(t) \Phi(t) \quad (4.7)$$

Let $|x|_y$ be modular operation $(x) \bmod y$ and $\lfloor \cdot \rfloor$ be the integer floor operation. By substituting

(4.3) and (4.5) into (4.7) with some manipulations, the expression of $s_{tg}(t)$ can be rewritten as

$$s_{tg}(t) = \sum_i \sum_{j=0}^{N_f-1} \Psi_{i,j} \left(t - (iN_f + vN_f + j + u)T_f - \tau_d - \tau'_{tg} \right) \quad (4.8)$$

with

$$\Psi_{i,j}(t) = a_i c_j \vartheta_0(t) + a_{i+[(j+1)/N_f]} c_{[j+1]_{N_f}} \vartheta_1(t) \quad (4.9)$$

where

$$\vartheta_0(t) = h_d(t) \Omega \left(\frac{t + \tau_d + \tau'_{tg}}{\rho} \right) \quad (4.10)$$

and

$$\vartheta_1(t) = h_d(t) \Omega \left(\frac{t - \rho + \tau_d + \tau'_{tg}}{T_f - \rho} \right). \quad (4.11)$$

The physical explanation of $\vartheta_0(t)$ and $\vartheta_1(t)$ are given as follows. As shown in Figure 4.2 (b)-(d), due to the asynchronism between the clocks of the reader and the tag, for $\tau_d + \tau'_{tg} \leq \rho \leq \tau_d + \tau'_{tg} + T_{ds}^{(true)}$, the channel response $h_d(t)$ may be modulated by two consecutive data bits if the transition edge between the two data bits splits the channel response into two parts. Here $\vartheta_0(t)$ accounts for the front part of the channel response and $\vartheta_1(t)$ represents the tail. Note that for $0 \leq \rho < \tau_d + \tau''_{tg}$ we have $\vartheta_0(t) = 0$ since the support region of $h_d(t)$ does not overlap with that of $\Omega\left(\frac{t + \tau_d + \tau'_{tg}}{\rho}\right)$. Similarly, for $\tau_d + \tau'_{tg} + T_{ds}^{(true)} < \rho < T_f$, we have $\vartheta_1(t) = 0$.

The modulated signal $s_{tg}(t)$ is delayed by τ''_{tg} seconds and retransmitted back to the reader

via the uplink channel (from tag to reader). Assume that the uplink channel has L_u paths and its responses to $\vartheta_0(t)$ and $\vartheta_1(t)$ are given by

$$x(t) = \sum_{l=0}^{L_u-1} p_{u0}^{(l)}(t - \tau_{u0}^{(l)}) \quad (4.12)$$

and

$$y(t) = \sum_{l=0}^{L_u-1} p_{u1}^{(l)}(t - \tau_{u1}^{(l)}). \quad (4.13)$$

Here $p_{u0}^{(l)}(t)$ and $p_{u1}^{(l)}(t)$ are the responses of the l^{th} path to $\vartheta_0(t)$ and $\vartheta_1(t)$ respectively, and $\tau_{u0}^{(l)}$ and $\tau_{u1}^{(l)}$ are the associated relative propagation delays. Besides the returned tag response, the reader also inevitably receives clutter. Since the clutter does not pass through the modulation circuitry of the tag, it is not modulated by the tag's data information and hence has consistent waveform over different frames. Let the clutter channel response to a single transmitted UWB pulse $p_{tx}(t)$ be

$$q(t) = \sum_{l=0}^{L_c-1} p_q^{(l)}(t - \tau_q^{(l)}) \quad (4.14)$$

where $p_q^{(l)}(t)$ is the response of the l^{th} path in the clutter channel, $\tau_q^{(l)}$ is the propagation delay of that path and L_c is the number of clutter paths. $q(t)$ has time support on $[\tau_q^{(0)}, \tau_q^{(0)} + T_g]$ where T_g is the maximum clutter duration.

Let us define the true TOA t_{toa} as the arrival timing of the returned direct path corresponding to the pulse transmitted at $t = 0$. By this definition, t_{toa} is equal to the roundtrip propagation delay plus the processing delay of the tag, i.e., $t_{toa} = \tau_d + \tau_u + \tau_{tg}$ where τ_d and τ_u are the propagation delay of the direct path in the downlink and

uplink respectively, $\tau_{tg} = \tau'_{tg} + \tau''_{tg}$ is the total processing delay of the tag, and τ''_{tg} is the processing delay from modulator to the transmitting antenna of the tag. Assume that t_{toa} follows a uniform distribution, i.e., $t_{toa} \sim \mathcal{U}(0, T_{\max})$ where $T_{\max}$ is the maximum TOA. After passing through a zonal bandpass filter with bandwidth W , the overall signal received by the reader can be expressed as

$$r(t) = \sum_i \sum_{j=0}^{N_f-1} g_{i,j} \left(t - (iN_f + vN_f + j + u)T_f - t_{toa} \right) + \eta(t) + w(t) \quad (4.15)$$

where $w(t)$ accounts for thermal noise and multiple access interference and its double-sided power spectral density is assumed to be $N_0/2$,

$$\eta(t) = \sum_i \sum_{j=0}^{N_f-1} q \left(t - (iN_f + j)T_f \right) \quad (4.16)$$

is the overall clutter signal, $g_{i,j}(t)$ is the overall roundtrip tag response to the UWB pulse transmitted at $t = iN_fT_f + jT_f$ and has the expression

$$g_{i,j}(t) = \left(a_i c_j x(t) + a_{i+\lfloor (j+1)/N_f \rfloor} c_{\lfloor j+1 \rfloor_{N_f}} y(t) \right). \quad (4.17)$$

Suppose that the uplink channel has the same true maximum one-way channel delay spread $T_{ds}^{(true)}$ as the downlink channels. Assume that $T_g \leq T_f$ and $T_{\max} + 2T_{ds}^{(true)} \leq T_f$ so that the inter-frame interference (IFI) is avoided. As implied by (4.17), the energy of tag responses of different frames may be different since the data bits modulating $x(t)$ and $y(t)$ may be different. Define SNR as averaged symbol energy to noise ratio $SNR = E_s / N_0$ where E_s is the energy per symbol of the tag response averaged over all possible data bits, i.e.,

$E_s = N_f \mathbb{E} \left[\int_0^{T_f} g_{i,j}^2(t) dt \right] = N_f \int_0^{T_f} (x^2(t) + y^2(t)) dt$ and $\mathbb{E}[\bullet]$ is the expectation operator.

Also we define SCR as the signal to clutter ratio $SCR = E_s / (N_f E_q)$ where

$E_q = \int_0^{T_f} q^2(t) dt$ and $N_f E_q$ is the energy per symbol of the clutter.

4.3 Least Square Time of Arrival Estimator

In this section, we will first develop a general LS estimator for TOA estimation. The estimator is derived based on a presumed maximum delay spread T_{ds} of one-way channel. Generally T_{ds} is not equal to its true value $T_{ds}^{(true)}$. As we will show in Section 4.5, the discrepancy between T_{ds} and $T_{ds}^{(true)}$ does affect the estimation accuracy and its optimum value may be determined via simulation.

The received waveform $r(t)$ is sampled with period $\Delta = T_f / N$ where N is an integer. Define the sampling frequency as $f_s = 1 / \Delta$. Suppose that $r(t)$ is observed over N_s symbol duration and is sampled at $t = mN_f T_f + nT_f + k\Delta$ with $m = 0, 1, \dots, N_s - 1$, $n = 0, 1, \dots, N_f - 1$ and $k = 0, 1, \dots, N$. Assume that N_s / N_a is an integer, i.e., $N_s = N_a Q$ and Q is an integer. The channels for both the tag response and the clutter are assumed to be static during the sampling duration. The samples from the n^{th} frame of the m^{th} symbol interval are collected into vector $\mathbf{r}_{m,n} = [r_{m,n}[0], r_{m,n}[1], \dots, r_{m,n}[N-1]]^T$. Let $N_{ds} = \lfloor 2T_{ds} / \Delta \rfloor$ and $M = \lfloor t_{toa} / \Delta \rfloor$. Denote $\mathbf{x} = [x[0], x[1], \dots, x[N_{ds}-1]]^T$ and

$\mathbf{y} = [y[0], y[1], \dots, y[N_{ds} - 1]]^T$ as the sample vectors of $x(t)$ and $y(t)$. Define two column vectors $\mathbf{p}_x(M) = [\mathbf{0}_{M \times 1}^T, \mathbf{x}^T, \mathbf{0}_{(N-M-N_{ds}) \times 1}^T]^T$ and $\mathbf{p}_y(M) = [\mathbf{0}_{M \times 1}^T, \mathbf{y}^T, \mathbf{0}_{(N-M-N_{ds}) \times 1}^T]^T$ where $\mathbf{0}$ represents a vector with all elements being zeros. $\mathbf{r}_{m,n}$ may be expressed as;

$$\mathbf{r}_{m,n} = \mathbf{g}_{m,n}(u, v, M) + \mathbf{q} + \mathbf{w}_{m,n} \quad (4.18)$$

where $\mathbf{q}$ and $\mathbf{w}_{m,n}$ are $N \times 1$ column vectors containing samples of clutter and noise signal respectively, and $\mathbf{g}_{m,n}(u, v, M)$ which is the sample vector of $g_{i,j}(t)$ can be expressed as

$$\mathbf{g}_{m,n}(u, v, M) = C_{m,n}(u, v)\mathbf{p}_x(M) + C_{m,n+1}(u, v)\mathbf{p}_y(M). \quad (4.19)$$

Here $C_{m,n}(u, v) = a_{m-v+\lfloor (n-u)/N_f \rfloor} c_{m-u|N_f}$ is the current data bit modulating the front part of roundtrip tag response whereas $C_{m,n+1}(u, v) = a_{m-v+\lfloor (n+1-u)/N_f \rfloor} c_{m+1-u|N_f}$ is the next data bit modulating the tail of the response.

To proceed, let $\{\widetilde{M}, \widetilde{\mathbf{x}}, \widetilde{\mathbf{y}}, \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}\}$ be the candidate values of $\{M, \mathbf{x}, \mathbf{y}, u, v, \mathbf{q}\}$ among which M is the parameter of interest whereas the rest are nuisance parameters. $\{\widetilde{M}, \widetilde{\mathbf{x}}, \widetilde{\mathbf{y}}, \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}\}$ are chosen to minimize the LS function

$$\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}, \widetilde{\mathbf{y}}, \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \|\mathbf{r}_{m,n} - \mathbf{g}_{m,n}(\widetilde{u}, \widetilde{v}, \widetilde{M}) - \widetilde{\mathbf{q}}\|^2. \quad (4.20)$$

As shown in Appendix A, differentiating (4.20) with respect to $\widetilde{\mathbf{q}}$, $\widetilde{\mathbf{x}}$, and $\widetilde{\mathbf{y}}$ sequentially, setting the results to zero and solving for $\widetilde{\mathbf{q}}$, $\widetilde{\mathbf{x}}$, and $\widetilde{\mathbf{y}}$ yields the solution the expression for $\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}')$ depending only on $\{\widetilde{M}, \widetilde{u}, \widetilde{v}\}$

$$\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}') = \frac{\gamma_0 (\|\boldsymbol{\beta}_0\|^2 + \|\boldsymbol{\beta}_1\|^2) - 2\gamma_1 \boldsymbol{\beta}_1^T \boldsymbol{\beta}_0}{\gamma_1^2 - \gamma_0^2}. \quad (4.21)$$

where $\gamma_0 = N_s N_f - A^2 / (N_s N_f)$, $\gamma_1 = F - A^2 / (N_s N_f)$, $\boldsymbol{\beta}_0 = \boldsymbol{\theta}_B - A\boldsymbol{\theta}_c$, and $\boldsymbol{\beta}_1 = \boldsymbol{\theta}_D - A\boldsymbol{\theta}_c$, $\boldsymbol{\theta}_B$, $\boldsymbol{\theta}_D$ and $\boldsymbol{\theta}_c$ are $N_{ds} \times 1$ vectors with their respective k^{th} elements being $\theta_B[k] = B_{k+\widetilde{M}}(\widetilde{u}, \widetilde{v})$, $\theta_D[k] = D_{k+\widetilde{M}}(\widetilde{u}, \widetilde{v})$, and $\theta_c[k] = \theta[k + \widetilde{M}]$, $\forall 0 \leq k \leq N_{ds} - 1$,

$$B_k(\widetilde{u}, \widetilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} r_{m,n}[k] C_{m,n}(\widetilde{u}, \widetilde{v}), \quad (4.22)$$

$$D_k(\widetilde{u}, \widetilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} r_{m,n}[k] C_{m,n+1}(\widetilde{u}, \widetilde{v}), \quad (4.23)$$

$$F = Q \left(\sum_{i=0}^{N_a-1} a_i a_{i+1} \right) (c_{N_f-1} c_0) + Q N_a \left(\sum_{i=0}^{N_f-2} c_j c_{j+1} \right). \quad (4.24)$$

Note that F is a constant independent of $\widetilde{u}$ and $\widetilde{v}$.

Note that (4.21) is invalid when $\gamma_1^2 - \gamma_0^2 = 0$. To cover such exceptional situation, the following analysis is carried out. The equation $\gamma_1^2 - \gamma_0^2 = 0$ is first factorized as

$$\gamma_1^2 - \gamma_0^2 = (F - N_s N_f)(F + N_s N_f - 2A^2 / N_s N_f) = 0, \quad (4.25)$$

which leads to

$$F - N_s N_f = 0 \quad (4.26)$$

and

$$F + N_s N_f - 2A^2 / N_s N_f = 0. \quad (4.27)$$

In the Appendix C, we show that at least one of the equations (4.26) and (4.27) holds when the conditions $C_{m,n}(\widetilde{u}, \widetilde{v}) = C_{m,n+1}(\widetilde{u}, \widetilde{v}), \forall m, n$ or

$C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}) = 0, \forall m, n$ is met. Next, we will give some intuitions for these two conditions. The condition $C_{m,n}(\tilde{u}, \tilde{v}) = C_{m,n+1}(\tilde{u}, \tilde{v}), \forall m, n$ is met when every bit in the sequence $\{C_{m,n}(\tilde{u}, \tilde{v})\}$ has the same polarity, i.e., $C_{m,n}(\tilde{u}, \tilde{v}) = +1, \forall m, n$ or $C_{m,n}(\tilde{u}, \tilde{v}) = -1, \forall m, n$. In this case the tag response will retain polarity over different frames appearing as “unmodulated” signal like the clutter and therefore there is no way we can distinguish the tag response from the clutter. In the following discussion, we assume that such sequence is deliberately avoided in the system design so that $F \neq N_s N_f$ applies. The second condition $C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}) = 0, \forall m, n$ is fulfilled when the sequence of $\{C_{m,n}(\tilde{u}, \tilde{v})\}$ consisting of alternative $+1$ and -1 , i.e., $\{C_{m,n}(\tilde{u}, \tilde{v})\}$ is a sequence of $+1, -1, +1, -1, \dots$. With such sequence, we have $A = 0$, $F = -N_s N_f$ and $B_k(\tilde{u}, \tilde{v}) = -D_k(\tilde{u}, \tilde{v})$. As show in Appendix A, the LS solution to $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}})$ for $A = 0$, $F = -N_s N_f$ and $B_k(\tilde{u}, \tilde{v}) = -D_k(\tilde{u}, \tilde{v})$ leading to $\gamma_1^2 - \gamma_0^2 = 0$ is

$$\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}) = -\frac{1}{N_s N_f} \|\mathbf{0}_B\|^2. \quad (4.28)$$

According to the above discussion, we can conclude our final estimator. If $\gamma_1^2 = \gamma_0^2$, the decision function $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}})$ is directly computed based on (4.28); else $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}})$ is computed using (4.21). Subsequently, the candidate values $\{\tilde{M}, \tilde{u}, \tilde{v}\}$ minimizing the value of the decision function $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}})$ is adopted as final estimates,

$$\{\hat{M}, \hat{u}, \hat{v}\} = \arg \min_{\tilde{M}, \tilde{u}, \tilde{v}} [\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}})] \quad (4.29)$$

where $\{\widehat{M}, \widehat{u}, \widehat{v}\}$ are the final estimates of $\{M, u, v\}$. And the TOA is estimated as

$$\hat{t}_{toa} = \Delta \widehat{M}. \quad (4.30)$$

4.4 Immunity of the Estimator to Clutter

Clutter suppression capability is one of the most important functionalities to ensure that UWB backscattering RFID system is able to achieve accurate TOA estimation when it operates in a environment with dense scatterers. In [75], for data communication application, the clutter is suppressed via choosing the data sequence with zero mean (or quasi-zero mean). Intuitively we would have expected that similar sequence selection should also be imposed to assign the clutter immunity capability to the estimator derived in the last section. However, we are going to show that the proposed TOA estimator is indeed inherently immune to the clutter for arbitrary data sequence.

Note that the final decision function $\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}')$ is computed based on samples $\{r_{m,n}[k]\}$ which are a supposition of true tag responses $\{x[k], y[k]\}$, clutter $\{q[k]\}$ and noise $\{w_{m,n}[k]\}$. Recall that $\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}')$ is computed using (4.21) or (4.28). To explicitly reveal the possible effects of the clutter term $\{q[k]\}$ on $\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}')$, we first substitute the signal model $r_{m,n}[k]$ given in (4.18) into (4.21). After some derivations, we reach

$$\begin{aligned} & \Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}', \widetilde{\mathbf{y}}', \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}') \\ &= \frac{\gamma_0 \left\| (\gamma_0 - \gamma_1)(\mathbf{z}_x - \mathbf{z}_y) + \mathbf{f}_{w0} - \mathbf{f}_{w1} \right\|^2}{\gamma_1^2 - \gamma_0^2} + \frac{2(\gamma_0 \mathbf{z}_x + \gamma_1 \mathbf{z}_y + \mathbf{f}_{w0})^T (\gamma_1 \mathbf{z}_x + \gamma_0 \mathbf{z}_y + \mathbf{f}_{w1})}{(\gamma_0 + \gamma_1)} \end{aligned} \quad (4.31)$$

where $\mathbf{z}_x$ and $\mathbf{z}_y$, $\mathbf{f}_{w0}$ and $\mathbf{f}_{w1}$ are $(N_{ds} - 1) \times 1$ vectors with their k^{th} element given by

$$z_x[k] = p_x[k + \widetilde{M}], \quad z_y[k] = p_y[k + \widetilde{M}],$$

$$f_{w0}[k] = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} w_{m,n} [k + \widetilde{M}] \left(C_{m,n}(\widetilde{u}, \widetilde{v}) - \frac{A}{N_s N_f} \right), \quad (4.32)$$

and

$$f_{w1}[k] = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} w_{m,n} [k + \widetilde{M}] \left(C_{m,n+1}(\widetilde{u}, \widetilde{v}) - \frac{A}{N_s N_f} \right). \quad (4.33)$$

By observing (4.31)-(4.33), it is found that all terms having the clutter $\{q[k]\}$ have been completely cancelled out with each other during the derivation process, leaving no clutter term in the final expression. This finding suggests that clutter does not have any effect on the value of decision function. Therefore, the decision function (4.21) is immune to the clutter.

Similarly, to prove that (4.28) is immune to the clutter, we substitute the signal model $r_{m,n}[k]$ given in (4.18) into it and expand the resultant expression. Recall that (4.28) is used when $C_{m,n}(\widetilde{u}, \widetilde{v}) + C_{m,n+1}(\widetilde{u}, \widetilde{v}) = 0, \forall m, n$ occurs under which situation we have equations $A = 0$, $F = -N_s N_f$ and $B_k(\widetilde{u}, \widetilde{v}) = -D_k(\widetilde{u}, \widetilde{v})$ as discussed in the previous section. Using these equations, the expanded expression is derived as

$$\Upsilon(\tilde{M}, \tilde{\mathbf{x}}', \tilde{\mathbf{y}}', \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') = -\frac{1}{N_s N_f} \left\| N_s N_f (\mathbf{z}_x - \mathbf{z}_y) + \mathbf{f}_{w2} \right\|^2 \quad (4.34)$$

where $\mathbf{f}_{w2}$ is a $(N_{ds} - 1) \times 1$ vector with its k^{th} element given by

$$f_{w2}[k] = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} w_{m,n} [k + \tilde{M}] C_{m,n}(\tilde{u}, \tilde{v}). \quad (4.35)$$

The equation (4.34) and (4.35) implies that the decision function (4.28) is also immune to the clutter.

Consequently, based on the above discussions, we may conclude that the proposed estimator is immune to the clutter and this conclusion is supported by the simulation results presented in the next section.

4.5 Simulation Results and Discussion

For the tag response, the impulse responses of both downlink and uplink channels are generated from the channel models CM1 for residential LOS environment and CM2 for residential NLOS environment as described in IEEE802.15.4a standard [50]. The length of the generated one-way channel impulse responses are truncated beyond 60 ns, i.e., $T_{ds}^{(true)} = 60$ ns which is the same as the setting used in [34] and is sufficiently long to capture 99.97% and 90.87% total channel energy of CM1 and CM2. Since no UWB channel model for clutter has been reported in the literature, the roundtrip impulse response of the clutter channel is simply generated as the convolution of two impulse responses randomly generated from the channel model used for the respective tag response. The length of clutter responses are

Figure 4.3. Comparison of the performance of the estimator under the scenarios with and without clutter for different number of sampled symbols in CM1 (left plot) and CM2 (right plot) with $T_{ds} = 45\text{ns}$, $f_s = 8\text{GHz}$ and $\text{SCR} = -30\text{dB}$.

truncated beyond $T_g = 150\text{ ns}$ which is slightly longer than the true maximum roundtrip delay

spread of tag responses $2T_{ds}^{(true)} = 120\text{ ns}$.

The transmitted UWB pulse is shaped as the second derivative of Gaussian pulse with width of 1 ns . Without loss of generality, we set the tag processing delay to zero in the simulation, i.e., $\tau_{tg} = 0\text{ ns}$. The rest of the system settings are $W = 4\text{ GHz}$, $T_{\max} = 40\text{ ns}$, $T_f = 200\text{ ns}$, $N_f = 4$, $N_a = 8$. For each channel realization, the sequence $\{a_i\}$ and $\{c_j\}$ are randomly generated. The TOA estimation error for the κ^{th} channel realization is

$\varepsilon_\kappa = \hat{t}_{toa} - t_{toa}$ and the MAE given by $MAE = (1/N_{ch}) \sum_{\kappa=0}^{N_{ch}-1} |\varepsilon_\kappa|$ is used as performance

Figure 4.4. MAE versus channel delay spread setting for different SNR values in CM1 with $N_s = 8$ and $f_s = 8\text{GHz}$.

criterion where N_{ch} is the total number of channel realizations and is set to 1000 in the simulation.

Figure 4.3 compares the MAE of the LS TOA estimator in the scenarios with clutter and without clutter. The sampling rate is set at $f_s = 8\text{GHz}$. The SCR is set to -30dB representing the extreme environment where the clutter overwhelms the signal. It is found that regardless the variation of channel condition and the number of observed symbols, the estimator is immune to the clutter. Figure 4.3 also indicates that the MAE of CM1 is generally lower than that of CM2 especially in the high SNR region. The reason is that the direct path is statistically larger in CM1 than in CM2 due to the obstructions in CM2 environment that may severely attenuate the direct path. By comparing the MAE curves of different number of sampled

symbols, it can be concluded that using more symbols improves the performance of the estimator. In both CM1 and CM2, for a large range of SNR, doubling the number of sampled symbols results in 3 dB less of SNR to achieve a given MAE.

Setting of presumed maximum one-way channel delay spread T_{ds} is closely related to the performance of TOA estimation since it directly determines N_{ds} that is included in the discrete signal model used to derive LS estimator. Figure 4.4 presents the results of MAE versus T_{ds} for different SNR values in CM1 channels which have RMS delay spread 17ns. The optimal setting of T_{ds} ranges from 25 ns to 50 ns as SNR varies from 10 to 50 dB. The MAE is much more sensitive to the setting of T_{ds} for signal with high SNR.

Figure 4.5 shows the MAE of the estimator operating with various sampling rates for presumed delay spread $T_{ds} = 40$ and 50 ns. $f_s = 8$ GHz is the Nyquist rate while $f_s = 1$ GHz is the sampling rate equal to the inverse of pulse width. As expected, the MAE increases as the sampling rate decreases. The MAE performance difference between different sampling rates is reduced as SNR decreases. At moderate SNR level, for instance $SNR = 35$ dB, the MAE difference between $f_s = 8$ GHz and $f_s = 1$ GHz is only 0.1 ns for $T_{ds} = 40$ ns and 0.2 ns for $T_{ds} = 50$ ns. Therefore, we may conclude that the estimator is robust over undersampling operation.

Figure 4.5. Effect of sampling frequency on MAE in CM1 with $N_s=16$ for $T_{ds}=40\text{ns}$ (dash line) and $T_{ds}=50\text{ns}$ (solid line).

4.6 Experimental Evaluation

Implementing any algorithm in a real system requires more practical considerations such as taking the timing jitter into account [78]. To validate the effectiveness of the estimator proposed in Section 4.3 and discover the issues existing in its practical implementation, we conduct the experiments in a laboratory environment. The scenario under consideration is to localize a backscattering tag in a 2D space (x, y) by estimating the roundtrip distances from the signals received by two receivers using the TOA algorithms introduced in Section 4.3.

Figure 4.6. Measurement setup.

4.6.1 Experimental Setup

The experiment setup is shown in Figure 4.6. The reader consists of system clock, a transmitter and two receivers. Function generator HP3314A performs as system clock to synchronize the transmitter and receiver. The transmitter consists of one in-house designed UWB pulse generator, two power amplifiers with gain 10dB and 30dB respectively, and one transmitting antenna. The UWB pulse generator repetitively sends pulses with width 1.5ns at PRF of 125kHz, i.e., the nominal frame rate is $T_f = 1 / PRF = 8 \mu s$. The central frequency of the UWB pulses is 4 GHz and the total number of sampled frames is 50. The first and second channels of a real time DSO act as two receivers. The raw sampling rate of the DSO is set to

Figure 4.7. Measurement setup layout. The measurement points are distributed in the shadow area.

be 20 GHz. The data collected by the DSO is stored in a computer to be processed by the algorithm given in Section 4.3. The digital BPF used in the post processing is a digital Gaussian filter with bandwidth in the range of [2.5GHz, 5GHz].

The tag consists of a balloon planar antenna followed by a coaxial cable with one-way delay $T_D = 1.780$ ns. Microcontroller unit (MCU) MSP 430 F149 from Texas Instrument is programmed such that it repetitively generates a square waveform coded by a randomly generated bipolar code sequence with code length of 50. The antipodal 2-PAM modulator sets its terminal condition to be open or short according to the code sequence from MCU, so that the incoming signal is reflected back with inversed or original polarities. The above system settings lead to $N_f = 1$ and $N_a = N_s = 50$.

The measurements are performed in a laboratory room with length 8.35m and width 8.35m. Figure 4.7 illustrates the layout of the measurement setup. The Tx/Rx antennas and tag

antenna are placed in the same horizontal plane at a height of 1 m above the ground. The phase center of the transmitting antenna is located at (-94.5cm, 0cm) while the phase centers of the two receiving antennas are located at (-87.5cm, 24cm) and (-87.5cm, -24cm). There are a number of tables, metallic cabinets and measurement instruments placed in the room which are scatterers producing clutter signals. The tag and the reader are placed at the center of the room. The tag is placed at the points distributed along lines $y = \{10, 0, -6, -12, 18, -24\}$ cm with adjacent spatial points in each line being separated by 4cm. For each spatial point, the signals captured by the two receiving antennas are sampled by the two channels of DSO. Totally 104 spatial points are measured and the roundtrip distances range from 1.3m to 3.5m.

4.6.2 Modified Signal Model

Prior to going into the detailed signal modeling, we define a concept of calibration signal to facilitate the suppression of jitter. The selection criteria of calibration signal are given as follows. The calibration signals should have sufficient SNR and not overlap with tag response. In addition, the calibration signal should be independent of environment change. Finally, it should suffer the same amount of jitter as the tag response. The calibration signal may be chosen as early arrival part of the coupling signal between the transmitting and receiving antennas in the reader. It is also possible to tap some signal at the input to transmitting antenna as calibration signal. Compared to the second method, the first one requires simpler hardware implementation, but Tx and Rx antennas have to be oriented in the way that the coupling signal has sufficient SNR. In the experiment, we choose the first method due to its simplicity of implementation.

With the presence of jitter, the expression of the transmitted signal given in (4.1) should be rewritten as [79]

$$s_{jit}(t) = \sum_i \sum_{j=0}^{N_f-1} p_{tx} \left(t - (iN_f + j)T_f - \beta_{0,0} - \delta_{i,j} \right) \quad (4.36)$$

where $\beta_{i,j}$ is the jitter of the signal in the $(iN_f + j)^{th}$ frame while $\delta_{i,j} = \beta_{i,j} - \beta_{0,0}$ is the relative jitter value. By taking both the calibration signal and the jitter into consideration, the received signal accounting for jitter is

$$r_{jit}(t) = \lambda_{jit}(t) + \eta_{jit}(t) + \varpi(t) + w(t) \quad (4.37)$$

where

$$\lambda_{jit}(t) = \sum_i \sum_{j=0}^{N_f-1} g_{i,j} \left(t - (iN_f + j + v + u)T_f - t_{toa} - \beta_{0,0} - \delta_{i,j} \right) \quad (4.38)$$

is tag response, t_{toa} is the TOA of the tag response in the absence of jitter and

$$\eta_{jit}(t) = \sum_i \sum_{j=0}^{N_f-1} q \left(t - (iN_f + j)T_f - \beta_{0,0} - \delta_{i,j} \right) \quad (4.39)$$

is the clutter while

$$\varpi(t) = \sum_i \sum_{j=0}^{N_f-1} \omega \left(t - (iN_f + j)T_f - t_{cal} - \beta_{0,0} - \delta_{i,j} \right) \quad (4.40)$$

is the calibration signal and $\omega(t)$ has finite time support on $[t_{st}, t_{nd}]$, t_{cal} is the arrival time of the calibration signal in the absence of jitter which is solely determined by the system configuration of the reader but not that of the tag. Assume that the tag is located at the far field such that the propagation delay of the tag response is always larger than t_{nd} , i.e., $\tau_{\min} > t_{nd}$, which means the time region of $\varphi_{i,j}(t)$ does not overlap with that of $\omega(t)$. Similarly, we also assume that clutter always arrives later than t_{nd} such that $q(t)$ has finite time support

Figure 4.8. Effect of jitter on the received signal ($N_f = 3$).

on $(t_{nd}, T_g]$. As shown in Figure 4.8, jitter renders the leading edges of the received signal in different frames unevenly distributed in temporal domain. Besides the instability of system clock, other factors such as tracking and relative velocities between transmitter and receiver also contribute to total jitter budget [79]. In addition, sampling process will introduce additional jitter. All of these effects have been included in $\beta_{0,0}$ and $\{\delta_{i,j}\}$ for notation convenience. Without effective mitigation of jitter, the estimator in Section 4.3 would suffer intolerable performance loss.

$r_{jit}(t)$ is sampled by DSO and the discrete samples are upsampled by β times. Let $1/\Delta$ be the sampling frequency after the upsampling operation. The samples in $[mN_fT_f + nT_f, mN_fT_f + nT_f + \Delta N]$ are collected into sample vector $\mathbf{r}_{m,n} = [r_{m,n}[0], r_{m,n}[1], \dots, r_{m,n}[N-1]]^T$.

4.6.3 Methodology of Jitter Mitigation

The methodology to suppress the jitter in the process of TOA estimation is described as follows. Suppose that the roundtrip distance of the current measurement is L . In order to mitigate the jitter effect, a prior reference measurement is taken at a spatial position with a relatively short roundtrip distance L_{ref} to obtain a reference measurement profile with high SNR. Let c be the velocity of light. When the calibration signal is chosen according to the aforementioned condition, its delay and waveform shape remain unchanged. The situation holds even when the position of the tag varies as long as the system configuration does not change. The arrival times of the tag response and the calibration signal in the reference measurement and the current measurement are expressed as

$$t'_{toa} = t_{toa} + \beta_{0,0} \quad (4.41)$$

$$t'_{toa_ref} = t_{toa_ref} + \beta_{0,0_ref} \quad (4.42)$$

$$t'_{cal} = t_{cal} + \beta_{0,0} \quad (4.43)$$

$$t'_{cal_ref} = t_{cal} + \beta_{0,0_ref} . \quad (4.44)$$

Here t_{toa_ref} and t_{cal_ref} are the counterparts of t_{toa} and t_{cal} in the reference measurement while t'_{toa} and t'_{cal} are the counterparts with jitter taken into consideration, and lastly t_{cal} is the delay of calibration signal

Besides (4.41) - (4.44), the relationships between time of arrivals and ranges are given by

$$t_{toa_ref} = L_{ref} / c + \tau_{tg} \quad (4.45)$$

and

$$t_{toa} = L / c + \tau_{tg} . \quad (4.46)$$

Manipulating (4.41) - (4.46) yields the solutions for distance L ,

$$L = c(t'_{toa} - t'_{toa_ref} - \xi) + L_{ref} \quad (4.47)$$

where $\xi = t'_{cal} - t'_{cal_ref}$.

Equations (4.47) indicates that L is determined by the unknowns $\{t'_{toa}, t'_{toa_ref}, \xi\}$. To estimate these unknowns, $\{\delta_{m,n}\}$ needs to be determined first. Assume that the jitter $\delta_{m,n}$ is bounded within $[-M_\delta \Delta, M_\delta \Delta]$. Let $N_{st} = \lfloor t_{st} / \Delta \rfloor$ and $N_{nd} = \lfloor t_{nd} / \Delta \rfloor$. By definition, the time support region of the calibration signal does not overlap with that of clutter and tag response, and thus the samples $\{r_{m,n}[N_{st}], r_{m,n}[N_{st} + 1], \dots, r_{m,n}[N_{nd} - 1]\}$ contain only the component of the calibration signal $\omega(t)$. We form a calibration sample vector

$$\mathbf{r}_{m,n_cal} = [\mathbf{0}_{(N_{st}-1) \times 1}, r_{m,n}[N_{st}], r_{m,n}[N_{st} + 1], \dots, r_{m,n}[N_{nd} - 1], \mathbf{0}_{(N_{nd}-N_{st}) \times 1}]^T. \quad (4.48)$$

$\delta_{m,n}$ is estimated by correlating the calibration signal in the $(mN_f + n)^{th}$ frame with that in the 0^{th} frame and picking the location of the correlation peak,

$$\hat{\delta}_{m,n} = \Delta \hat{M}_{m,n} \quad (4.49)$$

where $\hat{M}_{m,n} = \left[\arg \min_{M_{m,n} \in \mathcal{A}_M} \left[\sum_{k=N_{st}}^{N_{nd}} r_{m,n_cal}[k + M_{m,n}] r_{0,0_cal}[k] \right] \right]$, $\mathcal{A}_M = \{-M, -M + 1, \dots, M\}$, and

$r_{m,n_cal}[k]$ is the k^{th} element of $\mathbf{r}_{m,n_cal}$. With the knowledge of $\hat{M}_{m,n}$, the samples of different frames can be aligned with the samples of the first frame so that the relative jitter $\delta_{m,n}$ can be eliminated. The sample vector after alignment is

$$\mathbf{r}'_{m,n} = \begin{cases} \left[r_{m,n} [\widehat{M}_{m,n}], r_{m,n} [\widehat{M}_{m,n} + 1], \dots, r_{m,n} [N - 1], \mathbf{0}_{\widehat{M}_{m,n} \times 1}^T \right]^T, & \text{if } \widehat{M}_{m,n} \geq 0 \\ \left[\mathbf{0}_{\widehat{M}_{m,n} \times 1}^T, r_{m,n} [0], r_{m,n} [1], \dots, r_{m,n} [N - \widehat{M}_{m,n} - 1] \right]^T, & \text{if } \widehat{M}_{m,n} < 0 \end{cases}. \quad (4.50)$$

By applying the LS TOA estimator described in Section 4.3 on $\mathbf{r}'_{m,n}$ instead of original sample vector $\mathbf{r}_{m,n}$, t'_{toa} can be estimated from (4.30). In the same way, t'_{toa_ref} can be estimated from the reference measurement profile. Let $\hat{t}'_{toa}$ and $\hat{t}'_{toa_ref}$ be the estimated versions of t'_{toa} and t'_{toa_ref} .

Suppose that ξ is bounded in $[-M_\xi \Delta, M_\xi \Delta]$. ξ is estimated by correlating the calibration waveform of the current measurement with that of reference measurement, i.e.,

$$\hat{\xi} = \Delta \widehat{M}_\xi \quad (4.51)$$

where

$$\widehat{M}_\beta = \left(\arg \min_{M_\beta \in \mathcal{B}} \left[\sum_{k=N_{st}}^{N_{nd}} r_{0,0_cal} [k + M_\beta] r_{0,0_cal_ref} [k] \right] \right). \quad (4.52)$$

Here $\mathcal{B} = \{-M_\xi, -M_\xi + 1, \dots, M_\xi\}$, and $r_{0,0_cal_ref} [k]$ is the counterpart of $r_{0,0_cal} [k]$ in the reference measurement.

Substituting the estimated $\hat{t}'_{toa}$, $\hat{t}'_{toa_ref}$ and $\hat{\xi}$ into (4.47) gives final expression for estimated distance $\hat{L}$,

$$\hat{L} = c \left(\hat{t}'_{toa} - \hat{t}'_{toa_ref} - \hat{\xi} \right) + L_{ref}. \quad (4.53)$$

4.6.4 Localization with TOA Measurements

Let the tag coordinates be (x, y) , the transmitter coordinates be (x_t, y_t) , the receiver coordinates be (x_i, y_i) , $\forall i = 0, 1, \dots, M_r$, where M_r is the number of receivers. To estimate the coordinate of the tag, we define a cost function

$$f(x, y) = \sum_{i=0}^{M-1} f_i^2(x, y) \quad (4.54)$$

where

$$f_i(x, y) = L_i - \sqrt{(x - x_i)^2 + (y - y_i)^2} - \sqrt{(x - x_i)^2 + (y - y_i)^2} . \quad (4.55)$$

Here L_i is the roundtrip distance for the i^{th} receiver that has been estimated in the previous section. Equation (4.54) is a nonlinear least square (NLLS) function about the unknown position vector $\boldsymbol{\theta} = [x, y]^T$. Newton-Raphson or Gauss-Newton can be used to find the global minima of such function by iterative estimation of the unknowns [80]. Here we use Gauss-Newton since it does not require the second order derivative of the NLLS function as Newton-Raphson does, and hence it is less computationally intensive. With Gauss-Newton method, the estimation update equation is given by

$$\boldsymbol{\theta}_{k+1} = \boldsymbol{\theta}_k - (\mathbf{J}^T(\boldsymbol{\theta}_k) \mathbf{J}(\boldsymbol{\theta}_k))^{-1} \mathbf{J}^T(\boldsymbol{\theta}_k) \mathbf{f} \quad (4.56)$$

where $\mathbf{f} = [f_0(x, y), f_1(x, y), \dots, f_{M_r-1}(x, y)]^T$, $\boldsymbol{\theta}_k$ is the estimation result for $\boldsymbol{\theta}$ at the k^{th} iteration, $\mathbf{J}(\boldsymbol{\theta}_k)$ is the Jacobian matrix evaluated at $\boldsymbol{\theta}_k$ and its element in the i^{th} row and the j^{th} column is defined by

$$[\mathbf{J}(\boldsymbol{\theta}_k)]_{ij} = \left. \frac{\partial f_i(x, y)}{\partial \theta[j]} \right|_{x=x_k, y=y_k} \quad (4.57)$$

where $\boldsymbol{\theta}[j]$ is the j^{th} element of $\boldsymbol{\theta}$.

In our experiment, $\mathbf{J}(\boldsymbol{\theta}_k)$ is

$$\begin{aligned} [\mathbf{J}(\boldsymbol{\theta}_k)]_{i0} &= \left. \frac{\partial f_i(x, y)}{\partial x} \right|_{x=x_k, y=y_k} \\ &= -\frac{(x - x_t)}{\sqrt{(x - x_t)^2 + (y - y_t)^2}} - \frac{(x - x_i)}{\sqrt{(x - x_i)^2 + (y - y_i)^2}} \Big|_{x=x_k, y=y_k}, i = 0, 1, \dots, M_r - 1 \end{aligned} \quad (4.58)$$

$$\begin{aligned} [\mathbf{J}(\boldsymbol{\theta}_k)]_{i1} &= \left. \frac{\partial f_i(x, y)}{\partial y} \right|_{x=x_k, y=y_k} \\ &= -\frac{(y - y_t)}{\sqrt{(x - x_t)^2 + (y - y_t)^2}} - \frac{(y - y_i)}{\sqrt{(x - x_i)^2 + (y - y_i)^2}} \Big|_{x=x_k, y=y_k}, i = 0, 1, \dots, M_r - 1 \end{aligned} \quad (4.59)$$

The iteration stops if the total number of iterations exceeds 1000 times or

$\|(\mathbf{J}^T(\boldsymbol{\theta}_k)\mathbf{J}(\boldsymbol{\theta}_k))^{-1}\mathbf{J}^T(\boldsymbol{\theta}_k)\mathbf{f}\| \leq 1 \times 10^{-12}$. And the $\boldsymbol{\theta}_k$ in the last iteration is assigned as the final

estimated location $\hat{\boldsymbol{\theta}} = [\hat{x}, \hat{y}]^T$.

4.6.5 Results and Discussion

Figure 4.9 depicts the received signal and recovered tag response $\hat{\mathbf{x}} + \hat{\mathbf{y}}$ in a received profile with roundtrip distance of $3.03m$. It can be observed that the true tag response is buried under the strong clutter signal before processing and it emerges after processing. Figure 4.10 shows the recovered tag responses of 20 frames, among which some have polarity of $+1$, while the polarity of the rest are -1 .

Figure 4.9. The received signal and the recovered tag response. The vertical scale is in millivolts.

Figure 4.10. The tag responses of 20 frames with different polarities.

Figure 4.11 shows the scatter plot of the jitter values $\{\beta_{m,n}\}$ of 512 frames. It is found that the variance of the jitter is 72.881ps which can severely affect the range estimation if not properly mitigated.

The left plot of Figure 4.12 depicts the estimation errors of roundtrip distance versus their respective true distances. The RMSE of roundtrip distance estimation computed by the 208 profiles is 1.5cm. Error slightly increases with the increasing of roundtrip distance due to lower SNR. The right plot shows the empirical PDF of errors. Using Kolmogorov-Smirnov test [81] with significance level 14%, we accept the hypothesis that distance estimation error is a Gaussian variable characterized by mean -0.26cm and standard deviation 1.55cm. The small error variance and symmetry of the PDF indicate that the effects of the clutter and jitter have been mitigated and the dominant error source is the noise.

Let the estimation error of the location along the x coordinate be $\varepsilon_x = \hat{x} - x$ and the error along y coordinate be $\varepsilon_y = \hat{y} - y$. Figure 4.13 and Figure 4.14 show the $|\varepsilon_x|$ and $|\varepsilon_y|$ versus x coordinate and y coordinate for all measurements. As it is expected, both $|\varepsilon_x|$ and $|\varepsilon_y|$ increase as the reader-tag distance increases while $|\varepsilon_y|$ is more sensitive to such increment. It is also found that ε_y has larger variation than ε_x . The reason is that the specific geometry of the receivers, the transmitter and the tag renders the range uncertainty of y axis to be larger than that of x axis due to the dilution of precision difference between y and x axis [82].

Figure 4.11. Jitter of different frames.

Figure 4.12. The scatter plot of roundtrip distance estimation error and its empirical PDF (curve fitting is shown as dash line).

Figure 4.13. $|\epsilon_x|$ versus x and y coordinates.

Figure 4.14. $|\epsilon_y|$ versus x and y coordinates.

4.7 Conclusion

A novel LS TOA estimator is derived for UWB backscattering RFID system. Both theoretical studies and simulation results indicate the estimator is inherently immune to clutter signal regardless of SNR variation and is applicable to arbitrary data sequence. It is shown that the performance of the estimator depends on the number of sampled symbols as well as the channel delay spread setting. The study also shows that the estimator is robust over undersampling operation. To validate the proposed estimator in practical implementation, a prototype of backscattering RFID system was setup with which measurements were conducted in a laboratory environment. To overcome the jitter, we propose a concept of calibration signal. A jitter mitigation technique is therefore proposed by using the information in the calibration signal and a reference measurement. Experimental results indicate the LS estimator is able to achieve high accuracy once the jitter is properly mitigated by the proposed technique. The experiments are conducted within relatively short roundtrip distance due to hardware constraint such as power limit. The possible effects of larger distances are discussed here. For larger distances, the received signal will be weaker due to larger path attenuation. Therefore, for larger distances, larger transmitting power is required to achieve the same ranging performance as compared to shorter distances. Larger distances also increase the possibility that the clutters caused by some scatterers arrive earlier than the direct path. Therefore, larger distance calls for better clutter suppress capability as compared to short distance. All these effects, however, do not affect the validity of the proposed estimator which is derived based on general scenarios.

Chapter 5

Clutter Suppression with Dual Pulse Modulation

5.1 Introduction

For backscattering RFID systems, extracting TOA information from the received signal contaminated by background clutter is challenging. In Chapter 4, we have derived an LS TOA estimator for the backscattering RFID system introduced in [75]. The solution works for arbitrary data sequence but the reader has to implement the receiver with ADCs that have high sampling rate close to the classic Nyquist rate. However, ADCs with high sampling rate are usually very expensive and their usage in low-power operations is problematic [83]. One of the possible ways to reduce the sampling requirement are to use non-coherent receivers such as ED receiver and autocorrelation receiver which only involve relatively cheaper I&D circuit and low sampling rate switch. Both transmitted-reference (TR) signaling and differential signaling enable the implementation of autocorrelation receivers. However, TR requires delay line that is longer than the channel delay spread, while differential receiver requires delay line in the order of symbol duration. Such long delay lines impose considerable implementation

difficulties, especially for UWB devices which have enormous bandwidth. Such drawback can be avoided by using dual pulse (DP) signaling [85]. DP was first introduced in [84] and is a modified form of the TR scheme described in [86]. The DP signal format consists of a reference pulse and a data pulse. The temporal spacing between the two pulses is of the order of a UWB pulse and hence the required delay line is much shorter than TR and differential schemes. Similar to TR and differential signaling formats, DP is also robust to the pulse distortion caused by the channel due to the autocorrelation operation adopted by its receiver [41].

To overcome the clutter problem while maintain low complexity of reader receiver, we propose a novel UWB RFID tag design implementing DP modulation scheme. The signal reflected by the tag consists of a channel response followed by its delayed version. The two responses are spaced with time interval of several nanoseconds and their polarities are determined by spreading a ranging data sequence with two different bipolar sequences. The processing delay of the tag is only up to tens of nanosecond that is much shorter than the delay of those conventional RFID systems that are of the order of milliseconds. Therefore the proposed RFID system is robust to clock drift between the reader and the tag. A low complexity two-stage TOA estimator is developed for the proposed system. In the first stage, the integration duration of the integrator following the autocorrelation receiver is set to a larger value which is of the order of the multipath delay spread length. Coarse synchronization between the tag and the reader is achieved in this stage. In the second stage, the integration duration is reduced to the time of the order of the UWB pulse width. Based on the energy measurement of the integrator, the leading edge of the received channel responses is then

Figure 5.1. Tag structure.

estimated using threshold-crossing technique. With properly designed ranging sequence, PN sequence, and sampling duration, the developed TOA estimator is able to eliminate part of the clutter effects while mitigate the rest by averaging energy samples over multiple data symbols. The clutter suppression capability of the estimator can be significantly improved by increasing the number of averaged data symbols. As compared to conventional clutter suppression techniques such as empty room and frame-to-frame techniques [76], the estimator does not require any empty room calibration and thereby is robust to the surrounding environment change. It is also insensitive to the motion state as long as the movement of the object is negligible during sampling duration.

5.2 Tag with DP Modulation

The structure of the proposed tag with DP modulation is depicted in Figure 5.1. The

Figure 5.2. Signal waveforms at different stages of signal reception.

antenna of such backscattering tag produces both antenna mode scattering and structural mode scattering responses. As discussed in Section 4.2 of Chapter 4, only antenna mode scattering response goes through the inner modulation circuitry of the tag whereas structural mode scattering response does not. Following the definitions and discussions in Section 4.2 of Chapter 4, the tag response in the sequel is only referred to the antenna mode scattering response whereas the structural mode scattering response is treated as part of the clutter.

As illustrated in Figure 5.2, any antenna mode signal received by the tag is amplified first, delayed by T_d and then passed through a switch. Define a function which is a train of rectangular windows each having length T_w

$$\Phi(t) = \sum_i \sum_{j=0}^{N_f-1} \Omega \left(\frac{t - (iN_f + j)T_f}{T_w} \right). \quad (5.1)$$

Let ρ be the clock offset between the reader and the tag. The amplifier is enabled or disabled by signal $\Phi(t - \rho)$ while the switch is controlled by $\Phi(t - \rho - T_d)$ where T_d is a delay factor. As illustrated in Figure 5.2, the control signals of both the amplifier and the switch consist of alternative ON and OFF states. Each ON state corresponds to an opening time window with length T_w , within which the amplifier is turned on and the switch is close to allow signal to pass through. Let $T_{ds}^{(true)}$ be the true maximum one-way channel delay spread. Suppose that the window length satisfies the condition $T_f - T_{ds}^{(true)} \geq T_w \geq T_{ds}^{(true)}$ so that the channel response induced by an arbitrary transmitted pulse should never appear in two adjacent opening windows. Assume that the reader transmits the signal at $t = \zeta$. The condition also implies that any channel response can be entirely confined within one opening window at tag when the transmitting timing ζ is properly set. Ideally, when $T_w \leq T_d$, no coupling occurs between the transmitting and receiving sections of the tag and thus in this case the transmitting and receiving antennas can be combined into one to simplify the hardware implementation. Else certain measures such as polarization mismatch should be taken to mitigate the coupling [87].

The signal passing the switch is equally split into upper and lower branches. The signal in the upper branch is fed into a binary phase shift keying (BPSK) modulator whereas the signal

in the lower branch passes through additional delay line T_D before being sent to the other BPSK modulator. The additional delay in the lower branch is of the order of UWB pulse width. The BPSK modulator in each branch inverses or retains the polarity of the incoming signals according to the output of a sequence generators. As shown in Figure 5.1, the two bipolar sequences adopted by the two sequence generators are obtained by spreading a ranging sequence $\{d_i\}$ with bipolar sequences $\{a_j\}$ and $\{b_j\}$ respectively. The sequences $\{a_j\}$ and $\{b_j\}$ are chosen in the way that their product is a PN sequence $\{c_j = a_j b_j\}$ which is unique to each tag to facilitate multi-user access functionality. Assume that the sequences $\{a_j\}$, $\{b_j\}$ and $\{c_j\}$ are periodic with code period N_f . The signals of the two modulators are then added up and reflected back to the reader.

5.3 Application of the Proposed Tag

The reader in the RFID system under consideration is a full functional transceiver. The transmitter and the receiver of the reader are synchronized to a common clock. The reader intends to accurately estimate the roundtrip distance between itself and the l^{th} tag by estimating the TOA of the received tag response.

To initialize the TOA estimation process, the reader transmits a sequence of identical UWB pulses

$$s(t) = \sum_i \sum_{j=0}^{N_f-1} p_{tx} \left(t - (iN_f + j)T_f - \zeta \right) \quad (5.2)$$

where $p_{tx}(t)$ is the transmitted UWB pulse, T_f is the frame duration, and $\zeta \in (0, T_f]$ is transmitting time of signal $s(t)$ and can also be interpreted as the clock offset of the reader relative to the zero timing point based on the time axis of itself. ζ is introduced to facilitate the TOA estimation and will be discussed later in this section. The number of frames per symbol is set to N_f which is also the code period of $\{a_j\}$ and $\{b_j\}$.

Suppose that the downlink channel (from reader to tag) has L_d path channel and its response to $p_{tx}(t)$ is $h_d(t)$ which includes the effects of path distortion and channel delay spreading. The signal model of $h_d(t)$ is given by (4.1) in Chapter 4. The transmitted signal impinges on the l^{th} tag with structure described in Section 5.2. After being amplified by the amplifier, delayed by T_d and then passing through the switch, the noise-free signal received by the l^{th} tag is given by

$$r_{tg}^{(l)}(t) = \sum_i \sum_{j=0}^{N_f-1} h_d'(t - (iN_f + j)T_f - \tau_d - T_d - \zeta) \quad (5.3)$$

where τ_d is the propagation delay of the downlink channel, and

$$h_d'(t) = h_d(t) \Omega\left(\frac{t + \lambda}{T_w}\right) \quad (5.4)$$

is the portion of $h_d(t)$ passing through the opening windows of both enabling signals of the amplifier and the switch. The variable $\lambda = \zeta + \tau_d - \rho$ can be interpreted as mismatch between $h_d(t)$ and the opening window of the amplifier's enabling signal. As illustrated in Figure 5.2, only when λ satisfies $T_w - T_{ds}^{(true)} + nT_f \geq \lambda > nT_f$ and n is an integer, the entire $h_d(t)$ passes through the opening windows, i.e., $h_d'(t) = h_d(t)$.

The output of the switch $r_{tg}^{(l)}(t)$ is split into two branches with equal power. The signal input to each branch is given by $r_{tg}^{(l)}(t)/\sqrt{2}$. In the upper branch, the split signal is directly fed to a BPSK modulator whereas in the lower branch it goes through an additional delay T_D . Define v and u as symbol and frame offsets respectively. As illustrated in Figure 5.2, the modulation functions of the upper and lower branches are

$$\Phi_{upper}(t) = \sum_i \sum_{j=0}^{N_f-1} d_i a_j \Omega \left(\frac{t - (iN_f + vN_f + j + u)T_f + (T_f - \rho - T_d)}{T_f} \right) \quad (5.5)$$

$$\Phi_{lower}(t) = \sum_i \sum_{j=0}^{N_f-1} d_i b_j \Omega \left(\frac{t - (iN_f + vN_f + j + u)T_f + (T_f - \rho - T_d) - T_D}{T_f} \right). \quad (5.6)$$

The output of the two modulators are then summed up and sent to the transmitting antenna of the tag. The signal retransmitted by the tag is given by

$$s_{tg}^{(l)}(t) = (r_{tg}^{(l)}(t)/\sqrt{2})\Phi_{upper}(t) + (r_{tg}^{(l)}(t - T_D)/\sqrt{2})\Phi_{lower}(t). \quad (5.7)$$

Note that (5.7) can be rewritten as

$$s_{tg}^{(l)}(t) = \sum_i \sum_{j=0}^{N_f-1} \varpi_{i,j} \left(t - (iN_f + j)T_f - \tau_d - T_d - \zeta \right) \quad (5.8)$$

where $\varpi_{i,j}(t)$ is given by

$$\varpi_{i,j}(t) = d_i a_j h_d'(t)/\sqrt{2} + d_i a_j h_d'(t - T_D)/\sqrt{2}. \quad (5.9)$$

Note that the $\varpi_{i,j}(t)$ in (5.8) is the superposition of two modulated signals spaced by T_D and thus takes the form of DP signaling format.

The signal $s_{tg}^{(l)}(t)$ is sent back to the reader. Since $s_{tg}^{(l)}(t)$ has DP format, the received tag response is the superposition of the two channel responses with one of them being the delayed

version of the other. The true TOA of the roundtrip signal is defined as the arrival time of that delayed channel response within $[\zeta, \zeta + T_f]$ which is induced by the UWB pulse transmitted at $t = \zeta$. Let τ_u be the propagation delay of the uplink channel. The true TOA can be expressed as $t_{toa} = \zeta + T_d + T_D + \tau$ where $\tau = \tau_d + \tau_u$ is the roundtrip propagation delay uniformly distributed within $[0, \tau_{\max}]$ and $\tau_{\max}$ is the maximum roundtrip propagation delay.

Note that the initial transmitted signal from the reader induces both the tag response and the clutter signal scattered by the scatterers in surrounding environment. Let $q(t)$ be the clutter channel response to the transmitted signal $p_{tx}(t)$ and its signal model is $q(t) = \sum_{l=0}^{L_c-1} p_q^{(l)}(t - \tau_q^{(l)} - \zeta)$ where $p_q^{(l)}(t)$ is the response of the l^{th} path in the clutter channel, $\tau_q^{(l)}$ is the propagation delay of that path and L_c is the number of clutter paths. The impulse response of the uplink channel consists of L_u paths. Denote $x(t)$ as the uplink channel response to the signal $(h_d'(t)/\sqrt{2})$ transmitted by the tag. $x(t)$ is given by

$$x(t) = \sum_{m=0}^{L_u-1} p_u^{(m)}(t - \tau_u^{(m)} - t_{toa}). \quad (5.10)$$

Here $p_u^{(m)}(t)$ is the response of the m^{th} channel path to $(h_d'(t)/\sqrt{2})$ and $\tau_u^{(m)}$ is the associated relative propagation delay. The received signal at the reader can be expressed as

$$r^{(l)}(t) = \sum_i \sum_{j=0}^{N_f-1} \varphi_{i,j} \left(t - (iN_f + j + vN_f + u)T_f \right) + \eta(t) + w(t) \quad (5.11)$$

where $\eta(t) = \sum_i \sum_{j=0}^{N_f-1} q(t - (iN_f + j + vN_f + u)T_f)$ is the overall clutter, $w(t)$ accounting for thermal noise and multiple access interference is modeled as additive white

Figure 5.3. Receiver block diagram.

Gaussian noise (AWGN) with zero mean and variance $\sigma_w^2 = N_0 W$, $N_0/2$ which is double-sided power spectral density,. $\varphi_{i,j}(t)$ is the received roundtrip tag response to the initial transmitted single UWB pulse $p_{tx}(t)$ and is given by

$$\varphi_{i,j}(t) = d_i a_j x(t) + d_i b_j x(t - T_D). \quad (5.12)$$

Since we will focus on the TOA estimation between one reader and one tag, the superscript l in (5.11) will be dropped in the following discussion for notation brevity.

Figure 5.3 shows the autocorrelation receiver under consideration. Multiplying the received signal with its delayed version yields a waveform $y(t)$ which is then integrated and sampled. Let $\tilde{u}$ be the trial value for u . The discrete samples are multiplied with coded sequence $\{c_{j+\tilde{u}}\}$ and then averaged over $N_s N_f$ frames, where N_s is the number of ranging symbols being sampled and averaged. The output after averaging operation is expressed as

$$\begin{aligned} & \Psi_k(t_{toa}, \zeta, \tilde{u}, v, T) \\ &= \frac{1}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \int_{t_0+kT+mN_f T_f+nT_f}^{t_0+(k+1)T+mN_f T_f+nT_f} y_{atc}(t) dt, \quad k \in \{0, 1, \dots, N-1\} \end{aligned} \quad (5.13)$$

where $t_0 = \zeta + T_d + T_D$, N is the total number of averaged samples, T is the length of the integration region, and $y_{atc}(t)$ is the autocorrelation output given by

$$y_{atc}(t) = r(t)r(t - T_D). \quad (5.14)$$

It is also assumed that $2T_{ds}^{(true)} + 2T_D + T_d + \tau_{\max} \leq T_f$ and $T_g + T_D \leq T_f$ so that no inter-frame interference (IFI) could occur at the output of autocorrelation receiver.

Define SNR as averaged symbol energy to noise ratio $SNR = E_s / N_0$ where E_s is the energy per symbol of the tag response averaged over all possible data bits. E_s may be expressed as $E_s = N_f \mathbb{E} \left[\int_0^{T_f} \varphi_{i,j}^2(t) dt \right] = N_f \int_0^{T_f} 2x^2(t) dt$. Also we define SCR as the signal to clutter ratio $SCR = E_s / (N_f E_q)$ where $E_q = \int_0^{T_f} q^2(t) dt$ and $N_f E_q$ is the energy per symbol of the clutter.

5.3.1 Sequence Selection Criteria for Clutter Suppression

Equation (5.13) can be equally transformed to

$$\Psi_k(t_{toa}, \zeta, \tilde{u}, v, T) = \int_{t_0 + kT}^{t_0 + (k+1)T} z(t) dt \quad (5.15)$$

where

$$z(t) = \frac{1}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \left(c_{n-\tilde{u}} y(t + mN_f T_f + nT_f) \right), \quad \forall t \in [t_0, t_0 + \tau_{\max} + 2T_{ds}^{(true)}]. \quad (5.16)$$

As shown in Appendix D, $z(t)$ can be expanded by substituting Equation (5.11) into (5.16)

which leads to

$$z(t) = \Upsilon_0(t) + \Upsilon_1(t) + \Upsilon_2(t) + \Upsilon_3(t) + \Upsilon_4(t), \quad (5.17)$$

where $\Upsilon_0(t)$ is the waveform solely dependent on tag response and is exploited to extract

TOA estimation, $\Upsilon_1(t)$ includes terms related to clutter only, $\Upsilon_2(t)$ includes terms related to both clutter and noise, $\Upsilon_3(t)$ is the inter-pulse interference (IPI), and $\Upsilon_4(t)$ subsumes the rest of terms related to noise. The expressions of Υ_0 to Υ_4 are given by

$$\Upsilon_0(t) = \left(\frac{1}{N_f} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} c_{n-u} \right) \Upsilon_{00}(t), \quad (5.18)$$

$$\begin{aligned} \Upsilon_1(t) &= \left(\frac{1}{N_s} \sum_{m=0}^{N_s-1} d_{m-v} \right) \left(\frac{\Upsilon_{10}(t)}{N_f} \sum_{n=0}^{N_f-1} a_{n-u} c_{n-\tilde{u}} + \frac{\Upsilon_{11}(t)}{N_f} \sum_{n=0}^{N_f-1} b_{n-u} c_{n-\tilde{u}} \right) + \frac{1}{N_f} \left(\sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \right) \Upsilon_{12}(t), \end{aligned} \quad (5.19)$$

$$\Upsilon_2(t) = \frac{q(t)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t - T_D) + \frac{q(t - T_D)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t), \quad (5.20)$$

$$\Upsilon_3(t) = \left(\frac{1}{N_f} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \right) \Upsilon_{30}(t) + \left(\frac{1}{N_f} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} c_{n-u} \right) \Upsilon_{31}(t), \quad (5.21)$$

$$\Upsilon_4(t) = \frac{1}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \Upsilon_{40}(t) \quad (5.22)$$

and the expressions of $\{\Upsilon_{00}(t), \Upsilon_{10}(t), \dots, \Upsilon_{40}(t)\}$ are given in Appendix D. To overcome the undesired clutter signal, the effects of $\Upsilon_1(t)$ and $\Upsilon_2(t)$ have to be mitigated, which will be discussed as follows. It is observed that if $(1/N_s) \sum_{m=0}^{N_s-1} d_{m-v} = 0$ and $(1/N_f) \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} = 0$ are satisfied simultaneously, we have $\Upsilon_1 = 0$ regardless of the clutter and the channel condition. These two conditions can be fulfilled by selecting $\{d_i\}$ and $\{c_j\}$ according to the following criteria:

1). $\{d_i\}$ and $\{c_j\}$ should be periodic. For $\{d_i\}$ or $\{c_j\}$, within each code period, the number of $+1$ and -1 should be equal.

2). The number of sampled symbols should satisfy $N_s = M_d N_d$, where N_d is the code period of $\{d_i\}$ and M_d is a non-zero integer number.

It is also interesting to note that when the above conditions are met, as a byproduct, the IPI term $\Upsilon_{30}(t)$ in (5.21) is suppressed. When T_D is large enough, the summands in the other clutter term $\Upsilon_2(t)$ can be treated as uncorrelated random variables and thus $\Upsilon_2(t)$ is approximately a Gaussian random variable with its mean and variance being zero and $\sigma_w^2(q^2(t) + q^2(t - T_D))/N_s N_f$ as proven in Appendix E. Therefore increasing N_s can reduce the variance of $\Upsilon_2(t)$ and hence mitigate the effect of $\Upsilon_2(t)$.

5.3.2 Two-Stage TOA Estimation

Assuming that the clutter effects have been effectively mitigated by the aforementioned method, we develop a two-stage TOA estimation algorithm. When $(1/N_s) \sum_{m=0}^{N_s-1} d_{m-v} = 0$, the algorithm treats the symbol offset v as a nuisance parameter which is not involved in TOA estimation. Thus $\Psi_k(t_{toa}, \zeta, \tilde{u}, v, T)$ can be abbreviated as $\Psi_k(t_{toa}, \zeta, \tilde{u}, T)$. In the first stage, by setting the integration duration T to a large value T_0 , $\{\zeta, \tilde{u}\}$ are jointly estimated. T_0 should be large enough to accommodate the roundtrip channel response and the uncertainty of TOA, i.e., $2T_{ds}^{(true)} + 2T_D + \tau_{\max} \leq T_0 \leq T_f$. In the second stage, T is set to a smaller value T_1 . The

fine TOA is then determined with the values of ζ and $\tilde{u}$ estimated in the first stage.

To start the first estimation stage, the receiver takes one energy sample per frame (i.e., $N = 1$). The samples are then multiplied with $\{c_{j+\tilde{u}}\}$ and averaged over $N_s N_f$ frames to produce an averaged sample $\Psi_0(\zeta, \tilde{u}, T = T_0)$ for each set of $\{\zeta, \tilde{u}\}$ values. In the absence of clutter and noise, $\Psi_0(\zeta, \tilde{u}, T = T_0)$ is indeed the energy of the captured roundtrip tag response accumulated over frames. $\{\zeta, \tilde{u}\}$ are selected to maximize that energy

$$\{\hat{\zeta}, \hat{u}\} = \arg \max_{\zeta, \tilde{u}} [\Psi_0(\zeta, \tilde{u}, T = T_0)] \quad (5.23)$$

where $\{\hat{\zeta}, \hat{u}\}$ are the estimated versions of $\{\zeta, u\}$. The maximization of equation (5.23) is achieved in the following way. First we note that when ζ is adjusted to the values leading to $T_w - T_{ds}^{(true)} + nT_f \geq \lambda > nT_f$ and $\tilde{u}$ is adjusted to the true value u , $\Psi_0(\zeta, \tilde{u}, T = T_0)$ reaches its maxima equal to $\int_{t_0}^{t_0+T_0} \Upsilon_{00}(t) dt$, since the channel responses of all frames pass through the opening window at tag with full length and their energy are added coherently. To realize the above maximization, the reader adjusted ζ for a set of candidate values $\vartheta = \{0, T_c, 2T_c, \dots, (M_c - 1)T_c\}$ where T_c is the searching step and M_c is an integer with $(M_c - 1)T_c + T_c = T_f$, i.e., $M_c = T_f / T_c$. For each value of ζ , $\Psi_0(\zeta, \tilde{u}, T = T_0)$ is computed for every possible $\tilde{u}$. To ensure that such maximization is realized by at least one trial value of ζ , the searching step should satisfy $T_c \leq (T_w - T_{ds}^{(true)})$. Recall that $T_w \leq T_f - T_{ds}^{(true)}$ and thereby T_c should satisfy $T_c \leq T_f - 2T_{ds}$. For instance, if $T_{ds} = 100ns$, $T_f = 200ns$, we may set $T_w = 150ns$ and $T_c = 100ns$, $M_c = 2$. In this case, we have $\vartheta = \{0ns, 100ns\}$,

which means that the reader only needs to adjust ζ twice. Note that although larger T_w may reduce the acquisition time, it also means more energy is consumed by the amplifier and the tag suffers worse Tx-Rx antenna coupling which needs to be taken into design consideration.

In the second stage of TOA estimation, $N = N'$ samples are taken

$$\beta_k = \Psi_k(\zeta = \hat{\zeta}, \tilde{u} = \hat{u}, T = T_1), k = 0, 1, \dots, N' - 1 \quad (5.24)$$

where $k = 0, 1, \dots, N' - 1$ with $N' = \left\lfloor (\tau_{\max} + 2T_{ds}^{(true)} + T_D) / T_1 \right\rfloor$. The integration time T_1 is of the order of UWB pulse width. TOA is then estimated as the arrival timing of the first sample crossing a preset threshold value γ ,

$$\hat{t}_{toa} = T_1 \left(\arg \min_{k \in \mathcal{A}} [\beta_k \geq \gamma] + 1/2 \right) + t_0, k \in \mathcal{A} \quad (5.25)$$

where $\mathcal{A} = \{kT_1 + T_1/2 \leq \tau_{\max}\}$. If no crossing event occurs for $k \in \mathcal{A}$, TOA is estimated as the timing of the center of the uncertain region, i.e., $\hat{t}_{toa} = t_0 + \tau_{\max}/2$.

The threshold setting is critical to the performance of the estimator. Define the sampling region that contains only clutter and noise components as clutter-noise region (CNR) and let σ_{CNR} be the standard deviation of the samples within that region. Note that CNR consists of a region prior to the direct path and a region after the delay spread of the channel echoes. Although the specific statistical properties of the clutter are unknown, it is generally true that increasing σ_{CNR} means the probability distribution function of the samples in CNR becomes flatter and more likely some samples in CNR will have large amplitude, which implies that the threshold should be increased in this case to lower the false alarm rate. Hence, the threshold is selected according to

$$\gamma = \theta \sigma_{CNR} \quad (5.26)$$

where θ is a positive constant. Define the decibel scale of θ as $\theta' = 10 \log_{10}(\theta)$.

The estimation method of σ_{CNR} in the sampled signal is discussed as follows. Denote $N_{CNR} = \lfloor T_{CNR} / T_1 \rfloor$ where T_{CNR} is the expected duration of CNR. First we form the following sample set $\mathcal{B}_g = \left\{ \Psi_k \left(\zeta = \hat{\zeta}, \tilde{u} = \hat{u}, T = T_1 \right), k = |g|_{N'}, |g+1|_{N'}, \dots, |g+N_{CNR}-1|_{N'} \right\}$, $g = 0, 1, \dots, N' - 1$ where $|x|_y$ is the modular operation $(x) \bmod (y)$. Let the standard deviation of the samples in $\mathcal{B}_g$ be σ_g . It is expected that the set with all samples in CNR gives smallest standard deviation since any random signal component will increase the standard deviation. Therefore, σ_{CNR} is estimated as

$$\sigma_{CNR} = \min \left\{ \sigma_g \right\}, g = 0, 1, \dots, N'. \quad (5.27)$$

5.4 Simulation Results and Discussion

In the simulation, $p_{rx}(t)$ is the second derivative of Gaussian pulse and has the width of approximately 1ns. For the tag response, the impulse responses of both downlink and uplink channels are generated from the channel models CM1 for residential LOS environment and CM2 for residential NLOS environment as described in IEEE802.15.4a standard [50]. The length of the generated one-way channel impulse response is truncated beyond 60ns, i.e., $T_{ds}^{(true)} = 60$ ns. Following the simulation settings in Chapter 4, the roundtrip impulse response

of the clutter channel is simply generated as the convolution of two impulse responses randomly generated from the channel model used for the respective tag response. The length of clutter responses are truncated beyond $T_g = 150$ ns.

For each channel realization, the ranging sequence $\{d_i\}$ is randomly chosen as one of the permutation of a bipolar sequence with period $N_d = 8$ and equal numbers of $+1$ and -1 . Similarly $\{c_j\}$ is chosen as one of the permutations of a bipolar sequence with period $N_f = 4$ and the sum of the code elements in the sequence is zero. The rest of system settings are $T_d = 0$, $T_D = 2$ ns, $W = 4$ GHz, $\tau_{\max} = 40$ ns, $T_0 = 145$ ns, $T_{CNR} = 40$ ns, $T_w = 150$ ns, $T_c = 100$ ns, $\vartheta = \{0, 100\}$ ns, and $N_{ct} = 100$. The TOA estimation error for the κ^{th} channel realization is $\varepsilon_\kappa = \hat{t}_{toa} - t_{toa}$ and the MAE given by $MAE = (1/N_{ch}) \sum_{\kappa=0}^{N_{ch}-1} |\varepsilon_\kappa|$ is used as performance criterion where N_{ch} is the total number of channel realizations.

Figure 5.4 shows an analysis for threshold setting with different N_s , signal to clutter ratio SCR and SNR values in CM 1. For each SNR, there is an optimal threshold point for each SNR. For each SNR, increasing threshold will reduce false alarm probability while at the same time increase the probability of missed detection. When the threshold is too low, the false alarm probability is intolerable while when the threshold is too high, the missed detection probability is too high. In either situation, the performance is poor. The optimal point is achieved where the combination of the positive and negative effects minimizes the MAE. The optimal threshold point increases with increasing SNR. However, such increment in threshold value is only about 2dB when SNR changes from 20dB to 40dB. It is also interesting to note

Figure 5.4. MAE versus threshold for different N_s , SNR and SCR values in CM1 with $T_1 = 1\text{ns}$. Dash line represents SNR = 20dB, solid line represents SNR = 30dB, dash-dotted lines (gray color) represents SNR = 40dB.

that the optimal point is also insensitive to the variation of N_s and SCR. The reason is that the threshold setting strategy in (5.26) adapts to the CNR level of each channel realization which is determined by noise level, clutter strength and N_s .

Figure 5.5 shows the MAE performance of the system in different SCR conditions in CM1. Due to the clutter suppression capability of the system, the estimator suffers little performance loss in moderate SCR condition (SCR=10dB) as compared to the scenario in the absence of clutter, for instance, only 0.18ns accuracy loss occurs at SNR=30dB.

As expected, the performance degrades with the increasing of SCR as illustrated in Figure 5.5. However, such degradation can be compensated by increasing the number of averaged

Figure 5.5. MAE versus SNR for different SCR values with $N_s = 64$, $T_1 = 1\text{ns}$ and $\theta' = 5\text{dB}$ in CM1.

symbols N_s as demonstrated in Figure 5.6. Given a fixed threshold, doubling N_s has the effects of reducing 2 to 3 dB of SNR required to achieve a particular MAE value. When the number of N_s is sufficiently large, significant performance improvement can be achieved even in very poor SCR condition. Figure 5.6 also suggests that the performance of CM1 is generally much better than that of CM2. The reason is that the direct path in CM2 may be severely attenuated by the blockages and hence its amplitude is statistically lower than that of CM1.

Figure 5.7 illustrates the effect of sampling interval T_1 on MAE with $N_s = 128$, $\theta' = 9\text{dB}$ and $SCR = 0\text{dB}$. Generally, smaller time interval leads to higher estimation accuracy due to better time resolution. The estimation in CM1 is more sensitive to the

Figure 5.6. MAE versus SNR for different N_s with $T_1 = 1\text{ns}$, $\theta' = 7\text{dB}$ and $\text{SCR} = -10\text{dB}$.

Figure 5.7. MAE versus SNR for different T_1 with $N_s = 128$, $\theta' = 9\text{dB}$ and $\text{SCR} = 0\text{dB}$.

variation of T_1 . For instance, at $SNR = 35$ dB, the MAE difference between $T_1 = 1$ ns and $T_1 = 2$ ns is 0.3 ns for CM1 and is only 0.179 ns for CM2.

5.5 Conclusion

A novel backscattering RFID system with tag implemented with DP modulation is proposed for ranging and positioning applications. A low-complexity two-stage TOA estimator is developed for the reader. Results show that the background clutter can be effectively mitigated by the algorithm when the ranging sequence, PN sequence and the sampling duration are properly designed. It is found that increasing the number of averaged symbols can effectively improve clutter suppression capability and therefore enhance the estimation accuracy. The impact of system parameter settings including threshold and sampling interval are also investigated.

Chapter 6

Conclusion and Future Works

6.1 Conclusion

This thesis mainly concentrates on characterizing ranging errors and developing robust TOA estimators for UWB indoor multipath fading channel with noise and clutter. To get the insight into TOA estimation in the real indoor environments, a measurement campaign is carried out to collect a database of received signals captured in four typical indoor environments including indoor office, open hall, laboratory, and corridor, where LOS and NLOS scenarios are present. Based on the measurement results, the ranging errors by TOA technique are evaluated for both correlation and ED receivers. From the performance point of view, correlation receiver is preferable since it can achieve better accuracy than ED receiver. Such superiority, however, is achieved at the expense of higher sampling rate that increases the cost of the receiver and complicates the system design. The impact of system parameter settings on the ranging performance is investigated and it is found that ranging accuracy in LOS is more sensitive to the parameter settings than that in NLOS. Based on measurement results, a unified ranging error model applicable for both correlation and ED receivers is

developed.

To overcome the excessive noise and improve the probability of correctly detecting the direct path, a novel threshold selection strategy that sets the threshold as a function of delay is proposed. The performance improvement of DD approach over the conventional DI approach is demonstrated via simulation. The study indicates that such DD setting strategy is insensitive to the possible variation or inaccuracy existing in the available power delay profile and small-scale fading channel parameters.

To overcome the clutter problem for TOA estimation in backscattering RFID system, we propose two solutions. The first solution is an LS estimator applicable to the UWB backscattering RFID system with tag implementing with antipodal 2-PAM modulation. The estimator is proven to be inherently immune to the clutter for arbitrary data sequence. Both simulation and measurement results show that the estimator is able to suppress severe clutter and accurately recover TOA information. The second solution is a novel RFID system with tag implementing a DP modulator. The DP modulation enables the low complexity autocorrelation receiver to be used in the reader. The polarity of modulated signal is jointly determined by a ranging sequence and a PN sequence. A low complexity two-stage TOA estimator is then developed for this system. When ranging sequence, PN sequence and sampling duration are properly designed, part of the clutter terms at the output of the autocorrelation receiver can be eliminated while the rest can be reduced by averaging over multiple symbols. The performance of this solution is tested by intensive simulation and proven to be robust over clutter. These two solutions have their own advantages. The first one has better clutter suppression capability and faster signal acquisition, while the second system

has lower receiver complexity and requires lower sampling rate. The effectiveness of both systems is validated by simulation results.

6.2 Future Works

In this thesis, the clock drift effect is neglected. In practice, the clock instabilities will affect the TOA estimation accuracy, especially for coherent reception. Future work should investigate the effects of the clock drift on the TOA estimation and develop robust TOA estimator when clock drift is present.

The study of this work focuses on the transmission link for single user. In the real UWB localization systems such as sensor network, there may be a large number of users in the same field which suggests that multi-user interference should be taken into consideration for TOA estimation in the further work. Furthermore, the wideband and narrowband interference are assumed to be eliminated by proper filtering in this work. In practice, the filter might not be able to completely null off the interference due to hardware design complexity. Therefore, developing TOA estimator to suppress wideband and narrowband interference can be included in further work.

Only single transmitting and single receiving antennas are considered for the receivers throughout this work. Multiple receiving antennas can provide the array gain and diversity gain and hence improve the performance. In future work, TOA estimators for multiple antennas could be developed.

Finally, certain location or environment information can facilitate the parameter settings for TOA estimation. The location and environment information may be provided by feedbacking the location estimation results to the TOA estimation stage. For instance, via interpreting the location results, the LOS or NLOS may be differentiated and this information can be feedback to TOA estimator to adjust the parameters accordingly. How to design the feedback mechanism remains a topic to be investigated in future research work.

Author's Publications

Journal Paper

(i) Chi Xu and Choi L. Law, "Delay-dependent threshold selection for UWB TOA estimation," *IEEE Commun. Lett.* vol. 12, no. 5, pp. 380-382, May. 2008.

(ii) Chi Xu and Choi L. Law, "Experimental evaluation of UWB ranging performance for correlation and ED receiver in indoor environments," *International Journal of Hybrid Information Technology*, vol. 2, no. 2, pp. 37-53, April 2009.

(iii) Chi Xu and Choi L. Law, "TOA estimation for UWB backscattering RFID system in the presence of background clutter." Accepted for publication by *EURASIP Journal on Wireless Communications and Networking*.

(iv) Chi Xu nad Choi L. Law, "TOA estimation for UWB backscattering RFID system with Dual Pulse Modulation for Clutter Suppression," Accepted for publication by *Wireless Communications and Mobile Computing*.

Conference Paper

(i) Chi Xu and Choi L. Law, “Experimental evaluation of UWB ranging performance for correlation and ED receivers in dense multipath environment,” in *Proc. Future Generation Communication and Networking* (FGCN 2007), vol. 2, Jeju Island, Korea, 6-8 Dec. 2007, pp. 186-192.

Appendix A

Derivation of LS Solution

The LS function is expressed as

$$\Upsilon(\widetilde{M}, \widetilde{\mathbf{x}}, \widetilde{\mathbf{y}}, \widetilde{u}, \widetilde{v}, \widetilde{\mathbf{q}}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \left\| \mathbf{r}_{m,n} - \widetilde{\mathbf{g}}_{m,n}(\widetilde{u}, \widetilde{v}, \widetilde{M}) - \widetilde{\mathbf{q}} \right\|^2. \quad (\text{A.1})$$

Differentiating (A.1) with respect to $\widetilde{\mathbf{q}}$, setting the result to zero and solving for $\widetilde{\mathbf{q}}$ yields the intermediate expression

$$\widetilde{\mathbf{q}}' = \boldsymbol{\theta} - \frac{1}{N_s N_f} A_x(\widetilde{u}, \widetilde{v}) \widetilde{\mathbf{p}}_x(\widetilde{M}) - \frac{1}{N_s N_f} A_y(\widetilde{u}, \widetilde{v}) \widetilde{\mathbf{p}}_y(\widetilde{M}) \quad (\text{A.2})$$

where $\widetilde{\mathbf{p}}_x(\widetilde{M}) = \left[\mathbf{0}_{\widetilde{M} \times 1}^T, \widetilde{\mathbf{x}}^T, \mathbf{0}_{(N-\widetilde{M}-N_{ds}) \times 1}^T \right]^T$, $\widetilde{\mathbf{p}}_y(\widetilde{M}) = \left[\mathbf{0}_{\widetilde{M} \times 1}^T, \widetilde{\mathbf{y}}^T, \mathbf{0}_{(N-\widetilde{M}-N_{ds}) \times 1}^T \right]^T$, $\boldsymbol{\theta}$ is a $N \times 1$

vector collecting the samples averaged over $N_s N_f$ frames and given by

$$\boldsymbol{\theta} = \left(\frac{1}{N_s N_f} \right) \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \mathbf{r}_{m,n}, \quad (\text{A.3})$$

$A_x(\widetilde{u}, \widetilde{v})$ and $A_y(\widetilde{u}, \widetilde{v})$ are defined as

$$A_x(\widetilde{u}, \widetilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} C_{m,n}(\widetilde{u}, \widetilde{v}), \quad (\text{A.4})$$

$$A_y(\tilde{u}, \tilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} C_{m,n+1}(\tilde{u}, \tilde{v}). \quad (\text{A.5})$$

As shown in the Appendix B, it can be proven that

$$A_x(\tilde{u}, \tilde{v}) = A_y(\tilde{u}, \tilde{v}) = Q \left(\sum_{j=0}^{N_f-1} c_j \right) \left(\sum_{i=0}^{N_s-1} a_i \right) \triangleq A \quad (\text{A.6})$$

which is indeed a constant and is irrelevant to $\{\tilde{u}, \tilde{v}\}$. Therefore, (A.2) is rewritten as

$$\tilde{\mathbf{q}}' = \mathbf{0} - \frac{1}{N_s N_f} A \left(\tilde{\mathbf{p}}_x(\tilde{M}) + \tilde{\mathbf{p}}_y(\tilde{M}) \right). \quad (\text{A.7})$$

With (A.7), substituting $\tilde{\mathbf{q}} = \tilde{\mathbf{q}}'$ into (A.1) and dropping the factor $\left(\sum_m \sum_n \|\mathbf{r}_{m,n}\|^2 \right) - \|\mathbf{0}\|^2$

which is irrelevant to the decision making process, the expression for $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}')$, we

have

$$\begin{aligned} & \Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') \\ &= \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \left\| \mathbf{r}_{m,n} - C_{m,n}(\tilde{u}, \tilde{v}) \tilde{\mathbf{p}}_x(\tilde{M}) - C_{m,n+1}(\tilde{u}, \tilde{v}) \tilde{\mathbf{p}}_y(\tilde{M}) - \tilde{\mathbf{q}}' \right\|^2 \\ &= \gamma_0 \|\tilde{\mathbf{x}}\|^2 + \gamma_0 \|\tilde{\mathbf{y}}\|^2 - 2\boldsymbol{\beta}_0^T \tilde{\mathbf{x}} - 2\boldsymbol{\beta}_1^T \tilde{\mathbf{y}} + 2\gamma_1 (\tilde{\mathbf{x}})^T \tilde{\mathbf{y}} \end{aligned} \quad (\text{A.8})$$

where $\gamma_0 = N_s N_f - A^2 / (N_s N_f)$, $\gamma_1 = F - A^2 / (N_s N_f)$, $\boldsymbol{\beta}_0 = \mathbf{0}_B - A\mathbf{0}_c$, and

$\boldsymbol{\beta}_1 = \mathbf{0}_D - A\mathbf{0}_c$, $\mathbf{0}_B$, $\mathbf{0}_D$ and $\mathbf{0}_c$ are $N_{ds} \times 1$ vectors with their respective k^{th} elements being

$\theta_B[k] = B_{k+\tilde{M}}(\tilde{u}, \tilde{v})$, $\theta_D[k] = D_{k+\tilde{M}}(\tilde{u}, \tilde{v})$, and $\theta_c[k] = \theta[k + \tilde{M}]$, $\forall 0 \leq k \leq N_{ds} - 1$.

$B_{k+\tilde{M}}(\tilde{u}, \tilde{v})$, $D_{k+\tilde{M}}(\tilde{u}, \tilde{v})$ and F are expressed as

$$B_k(\tilde{u}, \tilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} r_{m,n}[k] C_{m,n}(\tilde{u}, \tilde{v}), \quad (\text{A.9})$$

$$D_k(\tilde{u}, \tilde{v}) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} r_{m,n}[k] C_{m,n+1}(\tilde{u}, \tilde{v}), \quad (\text{A.10})$$

$$F = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} C_{m,n}(\tilde{u}, \tilde{v}) C_{m,n+1}(\tilde{u}, \tilde{v}). \quad (\text{A.11})$$

It is shown in the Appendix B that F can be transformed to

$$F = Q \left(\sum_{i=0}^{N_a-1} a_i a_{i+1} \right) (c_{N_f-1} c_0) + Q N_a \left(\sum_{i=0}^{N_f-2} c_j c_{j+1} \right) \quad (\text{A.12})$$

which is a constant.

Armed with the definitions (A.9) to (A.12), by solving $\partial \Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') / \partial \tilde{\mathbf{x}} = 0$ and $\partial \Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') / \partial \tilde{\mathbf{y}} = 0$ for $\tilde{\mathbf{x}}$ and $\tilde{\mathbf{y}}$, we obtain

$$\gamma_0 \tilde{\mathbf{x}} + \gamma_1 \tilde{\mathbf{y}} = \boldsymbol{\beta}_0, \quad (\text{A.13})$$

$$\gamma_1 \tilde{\mathbf{x}} + \gamma_0 \tilde{\mathbf{y}} = \boldsymbol{\beta}_1. \quad (\text{A.14})$$

Solving (A.13) and (A.14) gives the intermediate expression for tag response

$$\tilde{\mathbf{x}}' = \frac{\gamma_1 \boldsymbol{\beta}_1 - \gamma_0 \boldsymbol{\beta}_0}{\gamma_1^2 - \gamma_0^2}, \quad (\text{A.15})$$

$$\tilde{\mathbf{y}}' = \frac{\gamma_1 \boldsymbol{\beta}_0 - \gamma_0 \boldsymbol{\beta}_1}{\gamma_1^2 - \gamma_0^2}. \quad (\text{A.16})$$

With (A.15) and (A.16), substituting $\tilde{\mathbf{x}} = \tilde{\mathbf{x}}'$ and $\tilde{\mathbf{y}} = \tilde{\mathbf{y}}'$ into (A.8) and simplifying the resultant equation, we have the expression for $\Upsilon(\tilde{M}, \tilde{\mathbf{x}}', \tilde{\mathbf{y}}', \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}')$ depending only on $\{\tilde{M}, \tilde{u}, \tilde{v}\}$

$$\Upsilon(\tilde{M}, \tilde{\mathbf{x}}', \tilde{\mathbf{y}}', \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') = \frac{\gamma_0 (\|\boldsymbol{\beta}_0\|^2 + \|\boldsymbol{\beta}_1\|^2) - 2\gamma_1 \boldsymbol{\beta}_1^T \boldsymbol{\beta}_0}{\gamma_1^2 - \gamma_0^2}. \quad (\text{A.17})$$

Equation (A.17) is only valid for $\gamma_1^2 \neq \gamma_0^2$. The discussion in Section 4.3 indicates that

three conditions $A = 0$, $F = -N_s N_f$ and $B_k(\tilde{u}, \tilde{v}) = -D_k(\tilde{u}, \tilde{v})$ leads to $\gamma_1^2 = \gamma_0^2$. To derive the solution for this exceptional case, using the three conditions and with (A.8), (A.13), (A.14), it can be derived that

$$\begin{aligned}
 & \Upsilon(\tilde{M}, \tilde{\mathbf{x}}, \tilde{\mathbf{y}}, \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') \\
 &= \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \left\| \mathbf{r}_{m,n} - C_{m,n}(\tilde{u}, \tilde{v}) \tilde{\mathbf{p}}_x(\tilde{M}) - C_{m,n+1}(\tilde{u}, \tilde{v}) \tilde{\mathbf{p}}_y(\tilde{M}) - \tilde{\mathbf{q}}' \right\|^2 \\
 &= \gamma_0 \|\tilde{\mathbf{x}}\|^2 + \gamma_0 \|\tilde{\mathbf{y}}\|^2 - 2\boldsymbol{\beta}_0^T \tilde{\mathbf{x}} - 2\boldsymbol{\beta}_1^T \tilde{\mathbf{y}} + 2\gamma_1 (\tilde{\mathbf{x}})^T \tilde{\mathbf{y}} \\
 &= N_s N_f \|\tilde{\mathbf{x}} - \tilde{\mathbf{y}}\|^2 - 2\boldsymbol{\theta}_B^T (\tilde{\mathbf{x}} - \tilde{\mathbf{y}})
 \end{aligned} \tag{A.18}$$

and

$$\tilde{\mathbf{x}}' - \tilde{\mathbf{y}}' = \frac{1}{N_s N_f} \boldsymbol{\theta}_B. \tag{A.19}$$

Substituting (A.19) into (A.18) yields the solution for $\gamma_1^2 = \gamma_0^2$

$$\Upsilon(\tilde{M}, \tilde{\mathbf{x}}', \tilde{\mathbf{y}}', \tilde{u}, \tilde{v}, \tilde{\mathbf{q}}') = -\frac{1}{N_s N_f} \|\boldsymbol{\theta}_B\|^2. \tag{A.20}$$

Equations (A.17) and (A.20) form a complete LS solution covering all possible values for γ_0 and γ_1 .

Appendix B

Transformation of $A_x(u, v)$, $A_y(u, v)$, and F

Since $\{a_i\}$ is periodic with period N_a , we have $a_m = a_{m+N_a}$. Together with $0 \leq u < N_f - 1$ and $N_s = M_a N_a$, it can be shown that

$$A_x(u, v) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} a_{m-v+\lfloor (n-u)/N_f \rfloor} c_{n-u|N_f} = Q \sum_{n=0}^{N_f-1} (c_{n-u|N_f} \varphi) \quad (\text{B.1})$$

where $\varphi = \sum_{m=0}^{N_a-1} a_{m-v+\lfloor (n-u)/N_f \rfloor}$. Setting $i = m - v + \lfloor (n - u) / N_f \rfloor$, we have

$$\varphi = \sum_{m=0}^{N_a-1} a_{m-v+\lfloor (n-u)/N_f \rfloor} = \sum_{i=-v+\lfloor (n-u)/N_f \rfloor}^{N_a-1-v+\lfloor (n-u)/N_f \rfloor} a_i. \quad (\text{B.2})$$

If $-v + \lfloor (n - u) / N_f \rfloor = 0$, we have $\varphi = \sum_{i=0}^{N_a-1} a_i$. If $-v + \lfloor (n - u) / N_f \rfloor \neq 0$, using

$a_m = a_{m+N_a}$ again, same conclusion can be drawn since

$$\begin{aligned} \varphi &= \sum_{i=-v+\lfloor (n-u)/N_f \rfloor}^{-1} a_i + \sum_{i=0}^{N_a-1-v+\lfloor (n-u)/N_f \rfloor} a_i = \\ &= \sum_{i=N_a-v+\lfloor (n-u)/N_f \rfloor}^{N_a-1} a_i + \sum_{i=0}^{N_a-1-v+\lfloor (n-u)/N_f \rfloor} a_i = \sum_{i=0}^{N_a-1} a_i. \end{aligned} \quad (\text{B.3})$$

Substituting $\varphi = \sum_{i=0}^{N_a-1} a_i$ into (B.1) yields

$$A_x(u, v) = \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} a_{m-v+\lfloor (n-u)/N_f \rfloor} c_{\lfloor n-u \rfloor_{N_f}} = Q \varpi \left(\sum_{i=0}^{N_a-1} a_i \right) \quad (\text{B.4})$$

where $\varpi = \sum_{n=0}^{N_f-1} c_{\lfloor n-u \rfloor_{N_f}}$. Denote $j = \lfloor n-u \rfloor_{N_f}$. If $u = 0$, we have $\varpi = \sum_{j=0}^{N_f-1} c_j$ which is

also true for $u \neq 0$ as shown below. Noting that $\lfloor -x \rfloor_{N_f} = N_f - x$ if $0 < x < N_f - 1$, for

$n = 0, 1, \dots, u-1, u, u+1, \dots, N_f-1$, j has respective values

$j = N_f - u, N_f - (u-1), \dots, N_f - 1, 0, 1, \dots, N_f - u - 1$. Thus

$$\varpi = \sum_{j=N_f-u}^{N_f-1} c_j + \sum_{j=0}^{N_f-u-1} c_j = \sum_{j=0}^{N_f-1} c_j. \quad (\text{B.5})$$

Substituting (B.5) into (B.4) leading to

$$A_x(u, v) = Q \left(\sum_{j=0}^{N_f-1} c_j \right) \left(\sum_{i=0}^{N_a-1} a_i \right). \quad (\text{B.6})$$

Analogous to the proof for $A_x(u, v)$, it can also show that $A_y(u, v) = Q \left(\sum_{j=0}^{N_f-1} c_j \right) \left(\sum_{i=0}^{N_a-1} a_i \right)$.

Therefore, we have

$$A_x(u, v) = A_y(u, v) \triangleq A \quad (\text{B.7})$$

where A is a constant.

To prove (A.12), we express $F(u, v)$ as

$$\begin{aligned}
 F(u, v) &= \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} C_{m,n} C_{m,n+1} \\
 &= Q \sum_{n=0}^{N_f-1} c_{|n-u|_{N_f}} c_{|n+1-u|_{N_f}} \left(\sum_{m=0}^{N_a-1} a_{m-v+\lfloor (n-u)/N_f \rfloor} a_{m-v+\lfloor (n+1-u)/N_f \rfloor} \right) \\
 &= Q \sum_{\substack{n=0,1,\dots,N_f-1, \\ n \neq u-1}} c_{|n-u|_{N_f}} c_{|n+1-u|_{N_f}} \left(\sum_{m=0}^{N_a-1} a_{m-v+\lfloor (n-u)/N_f \rfloor} a_{m-v+\lfloor (n+1-u)/N_f \rfloor} \right) \\
 &\quad + Q \left(\sum_{m=0}^{N_a-1} a_{m-v-1} a_{m-v} \right) (c_{N_f-1} c_0)
 \end{aligned} \tag{B.8}$$

For $n \neq u-1$, we have $|n+1-u|_{N_f} = 1 + |n-u|_{N_f}$ and

$m-v+\lfloor (n-u)/N_f \rfloor = m-v+\lfloor (n+1-u)/N_f \rfloor$. Therefore, (B.8) may be rewritten as

$$F(u, v) = Q N_a \sum_{\substack{n=0,1,\dots,N_f-1, \\ n \neq u-1}} c_{|n-u|_{N_f}} c_{|n-u|_{N_f}+1} + Q \left(\sum_{m=0}^{N_a-1} a_{m-v-1} a_{m-v} \right) (c_{N_f-1} c_0). \tag{B.9}$$

Finally, letting $j = |n-u|_{N_f}$, and $i = m-v-1$, we have

$$\begin{aligned}
 F(u, v) &= Q N_a \left(\sum_{j=N_f-u}^{N_f-2} c_j c_{j+1} + \sum_{j=0}^{N_f-u-1} c_j c_{j+1} \right) + Q \left(\sum_{i=-v-1}^{-1} a_i a_{i+1} + \sum_{i=0}^{N_a-v-2} a_i a_{i+1} \right) (c_{N_f-1} c_0) \\
 &= Q N_a \sum_{j=0}^{N_f-2} c_j c_{j+1} + Q \left(\sum_{i=0}^{N_a-1} a_i a_{i+1} \right) (c_{N_f-1} c_0) \\
 &\triangleq F
 \end{aligned} \tag{B.10}$$

Appendix C

Conditions for the Exceptional Case of LS Solution

Using equation (A.11) and the fact that $C_{m,n}^2(\tilde{u}, \tilde{v}) = C_{m,n+1}^2(\tilde{u}, \tilde{v}) = 1$, the first condition (4.26) can be transformed to

$$F - N_s N_f = 0 \Leftrightarrow \frac{1}{2} \sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) - C_{m,n+1}(\tilde{u}, \tilde{v}))^2 = 0 \quad (\text{C.1})$$

which implies that the first condition (4.26) is true only when $C_{m,n}(\tilde{u}, \tilde{v}) = C_{m,n+1}(\tilde{u}, \tilde{v}), \forall m, n$ holds, .

Using (A.11), (A.4) - (A.6), the second condition (4.27) may be rewritten as

$$\begin{aligned} (F + N_s N_f) - \frac{2A^2}{N_s N_f} &= \frac{1}{2} \sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}))^2 - \frac{(A_x + A_y)^2}{2N_s N_f} = 0 \\ \Leftrightarrow \frac{1}{2} \sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}))^2 &= \frac{\left(\sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v})) \right)^2}{2N_s N_f} . \\ \Leftrightarrow N_s N_f \sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}))^2 &= \left(\sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v})) \right)^2 \end{aligned}$$

(C.2)

To get more insight into (C.2), we invoke the Cauchy-Schwarz inequality which states that

$$\left| \sum_i x_i y_i \right|^2 \leq \left(\sum_i |x_i|^2 \right) \left(\sum_i |y_i|^2 \right) \quad (\text{C.3})$$

is always true and the equality holds if and only if $x_i / y_i = x_j / y_j \forall i, j$ [61]. Applying the inequality to the right side of (C.2) yields

$$\left(\sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v})) \right)^2 \leq N_s N_f \left(\sum_m \sum_n (C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}))^2 \right) \quad (\text{C.4})$$

and the equality holds if $C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}) = G, \forall m, n$ and G is a constant. Note that both $C_{m,n}$ and $C_{m,n+1}$ can only be $+1$ or -1 and hence G can only be ± 2 or 0 . $G = 0$ is true when $C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}) = 0$ whereas $G = \pm 2$ is equivalent to $C_{m,n}(\tilde{u}, \tilde{v}) = C_{m,n+1}(\tilde{u}, \tilde{v})$.

In conclusion, when the conditions $C_{m,n}(\tilde{u}, \tilde{v}) = C_{m,n+1}(\tilde{u}, \tilde{v}), \forall m, n$ or $C_{m,n}(\tilde{u}, \tilde{v}) + C_{m,n+1}(\tilde{u}, \tilde{v}) = 0, \forall m, n$ are fulfilled, at least one of Equations (4.26) and (4.27) holds.

Appendix D

Transformation of Sample Model

With the following shorthand notations:

$$x_{m,n}(t) = x(t + mN_fT_f + nT_f), \quad (\text{D.1})$$

$$q_{m,n}(t) = q(t + mN_fT_f + nT_f), \quad (\text{D.2})$$

$$w_{m,n}(t) = w(t + mN_fT_f + nT_f), \quad (\text{D.3})$$

we have

$$\begin{aligned} & r(t + mN_fT_f + nT_f) \\ &= \sum_i \sum_{j=0}^{N_f-1} d_i a_j x_{i+v-m, j+u-n}(t) + \sum_i \sum_{j=0}^{N_f-1} d_i b_j x_{i+v-m, j+u-n}(t - T_D). \quad (\text{D.4}) \\ &+ \sum_i \sum_{j=0}^{N_f-1} q_{i+v-m, j+u-n} + w_{m,n}(t) \end{aligned}$$

Since $z(t)$ has finite time support on $t \in [t_0, t_0 + 2T_{ds} + \tau_{\max}]$ and T_f satisfies

$2T_{ds} + 2T_D + T_d + \tau_{\max} \leq T_f$ and $T_g + T_D \leq T_f$. The support regions of $x_{i+v-m, j+u-n}(t)$,

$x_{i+v-m, j+u-n}(t - T_D)$ and $q_{i+v-m, j+u-n}(t)$ are overlapped with that of $z(t)$ only if

$i + v - m = 0$ and $j + u - n = 0$. Hence, by omitting those summands with support regions

outside $t \in [t_0, t_0 + T_f]$ and noting that $x_{0,0}(t) = x(t)$ and $q_{0,0}(t) = q(t)$, Equation (D.4) may be rewritten as

$$r(t + mN_f T_f + nT_f) = d_{m-v} a_{n-u} x(t) + d_{m-v} b_{n-u} x(t - T_D) + q(t) + w_{m,n}(t) \quad (D.5)$$

$$, t \in [t_0, t_0 + 2T_{ds} + \tau_{\max}]$$

Substituting Equation (D.5) into (5.16) yields

$$z(t) = \Upsilon_0(t) + \Upsilon_1(t) + \Upsilon_2(t) + \Upsilon_3(t) + \Upsilon_4(t), \quad (D.6)$$

where the expressions of first three terms in the above equation are given in Equations (5.18) – (5.22) and the expressions of the rest of terms are

$$\Upsilon_{00}(t) = x^2(t - T_D), \quad (D.7)$$

$$\Upsilon_{10}(t) = x(t)q(t - T_D) + x(t - T_D)q(t) \quad (D.8)$$

$$\Upsilon_{11}(t) = x(t - T_D)q(t - T_D) + x(t - 2T_D)q(t) \quad (D.9)$$

$$\Upsilon_{12}(t) = q(t)q(t - T_D) \quad (D.10)$$

$$\Upsilon_{30}(t) = x(t)x(t - T_D) + x(t - T_D)x(t - 2T_D) \quad (D.11)$$

$$\Upsilon_{31}(t) = x(t)x(t - 2T_D) \quad (D.12)$$

$$\begin{aligned} \Upsilon_{40}(t) = & d_{m-v} a_{n-u} x(t) w_{m,n}(t - T_D) + d_{m-v} b_{n-u} x(t - T_D) w_{m,n}(t - T_D) \\ & + d_{m-v} a_{n-u} w_{m,n}(t) x(t - T_D) + d_{m-v} b_{n-u} w_{m,n}(t) x(t - 2T_D) + w_{m,n}(t) w_{m,n}(t - T_D) \end{aligned} \quad (D.13)$$

Appendix E

Variance of $\Psi_2(t)$

The expression of $\Upsilon_2(t)$ is

$$\Upsilon_2(t) = \frac{q(t)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t - T_D) + \frac{q(t - T_D)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t). \quad (\text{E.1})$$

Given that $\mathbb{E}[w_{m,n}(t - T_D)] = 0$ and $\mathbb{E}[w_{m,n}(t)] = 0$, the mean of $\Upsilon_2(t)$ is derived as

$$\begin{aligned} \mu &= \mathbb{E}[\Upsilon_2(t)] \\ &= \frac{q(t)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \mathbb{E}[w_{m,n}(t - T_D)] + \frac{q(t - T_D)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} \mathbb{E}[w_{m,n}(t)] = 0. \end{aligned} \quad (\text{E.2})$$

The variance of $\Upsilon_2(t)$ is

$$\begin{aligned}
 & Var[\Upsilon_2(t)] \\
 &= \mathbb{E} \left[\left(\frac{q(t)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t - T_D) + \frac{q(t - T_D)}{N_s N_f} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} c_{n-\tilde{u}} w_{m,n}(t) \right)^2 \right] \\
 &= \frac{q^2(t)}{N_s^2 N_f^2} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \mathbb{E}[w_{m,n}^2(t - T_D)] \\
 &+ \frac{q^2(t)}{N_s^2 N_f^2} \sum_{m_1=0}^{N_s-1} \sum_{n_1=0}^{N_f-1} \sum_{\substack{m_2=0, \\ m_2 \neq m_1}}^{N_s-1} \sum_{\substack{n_2=0, \\ n_2 \neq n_1}}^{N_f-1} \mathbb{E}[w_{m_1,n_1}(t - T_D) w_{m_2,n_2}(t - T_D)] \\
 &+ \frac{q^2(t - T_D)}{N_s^2 N_f^2} \sum_{m=0}^{N_s-1} \sum_{n=0}^{N_f-1} \mathbb{E}[w_{m,n}^2(t)] \\
 &+ \frac{q^2(t - T_D)}{N_s^2 N_f^2} \sum_{m_1=0}^{N_s-1} \sum_{n_1=0}^{N_f-1} \sum_{\substack{m_2=0, \\ m_2 \neq m_1}}^{N_s-1} \sum_{\substack{n_2=0, \\ n_2 \neq n_1}}^{N_f-1} \mathbb{E}[w_{m_1,n_1}(t) w_{m_2,n_2}(t)] \\
 &+ \frac{q(t)q(t - T_D)}{N_s^2 N_f^2} \sum_{m_1=0}^{N_s-1} \sum_{n_1=0}^{N_f-1} \sum_{m_2=0}^{N_s-1} \sum_{n_2=0}^{N_f-1} \mathbb{E}[w_{m_1,n_1}(t) w_{m_2,n_2}(t - T_D)] \quad .
 \end{aligned} \tag{E.3}$$

Note that $w_{m_1,n_1}(t - T_D)$ and $w_{m_2,n_2}(t - T_D)$ are uncorrelated when conditions $m_1 \neq m_2$ and $n_1 \neq n_2$ are met simultaneously. Under the same conditions, $w_{m_1,n_1}(t)$ and $w_{m_2,n_2}(t)$ are also uncorrelated with each other. Therefore, we will have $\mathbb{E}[w_{m_1,n_1}(t - T_D) w_{m_2,n_2}(t - T_D)] = \mathbb{E}[w_{m_1,n_1}(t) w_{m_2,n_2}(t)] = 0$. Further more, if $T_D \gg 1/W$, $w_{m_1,n_1}(t)$ is approximately uncorrelated with $w_{m_2,n_2}(t - T_D)$. With $\mathbb{E}[w_{m,n}^2(t - T_D)] = \mathbb{E}[w_{m,n}^2(t)] = \sigma_w^2$, (E.3) can be simplified as

$$Var[\Upsilon_2(t)] = \frac{(q^2(t) + q^2(t - T_D)) \sigma_w^2}{N_s N_f} . \tag{E.4}$$

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