Spatio-Temporal Processing for Indoor UWB Array Propagation Channels

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In Loving Memory of My Mum
Summary

This thesis mainly focuses on the study of the spatio-temporal processing for indoor UWB array systems. The main objective of this work was to investigate the spatial and temporal characteristics of UWB signals captured by planar array and also linear array systems. These characteristics include space-time array processing and application based spatio-temporal results on UWB receiver systems.

Firstly, simulation of UWB array systems based on multipath clusters and distortion effects is investigated. This modified simulation contains spatio-temporal characteristics of UWB arrays such as discrete channel impulse responses based on classifications of multipath clusters, propagation environments and physics-based pulse distortion effects. Frequency dependent characteristics are also analysed in the simulation. Thus, discrete impulse responses and distorted pulse signals are simulated in each frequency sub-band corresponding to particular channel parameters of directions of arrival, times of arrival and distortion effects. Furthermore, quantifications of distortion impact on UWB system are evaluated; UWB simulated data is measured corresponding to propagation scenarios in order to determine system performances.

Secondly, analysis of the spatial correlation based array structure is presented. This novel spatial correlation technique is processed between surrounding adjacent antenna positions for each multipath component time bin. Therefore, variations of spatial correlations with excess delays can be temporally analysed. The contributions of this technique and preliminary results are also given as the application on estimation of times of arrival of multipath components. In addition, based on the spatial correlation analysis, the applications of angles of arrival estimation are also described in this thesis. To simplify this technique, the linear array configuration is exploited cooperatively with the relative phase difference method. UWB linear array measurements are also carried out in an anechoic chamber with the controlled conditions to evaluate estimated results. Additionally, the research also presents a complex spatial correlation analysis, which can be used as the identification of distortion effects at each individual time bin.

Finally, UWB RAKE receiver systems are investigated in order to highlight the application of UWB space-time array processing, which is proposed as the adaptive multipath searching unit. The multipath searching and distortion selection processors are constructed to support RAKE selection and weight estimator functions. The implementation of this proposed structure associated with the finger selection strategies and the time-reversal mirror technique can improve UWB array link performances in dense channels with less complexity.
Key words: UWB propagation channel, UWB array processing, UWB spatio-temporal characteristics, spatial correlation
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<tr>
<td>AOA</td>
<td>Angle-of-Arrival</td>
</tr>
<tr>
<td>AOD</td>
<td>Angle-of-Departure</td>
</tr>
<tr>
<td>A-RAKE</td>
<td>All RAKE</td>
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<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
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<tr>
<td>BPF</td>
<td>Band Pass Filter</td>
</tr>
<tr>
<td>CCDF</td>
<td>Complementary of Cumulative Distribution Function</td>
</tr>
<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
</tr>
<tr>
<td>CEPt</td>
<td>Conference of Postal and Telecommunications</td>
</tr>
<tr>
<td>CIR</td>
<td>Channel Impulse Response</td>
</tr>
<tr>
<td>CRLB</td>
<td>Cramer-Rao Lower Bound</td>
</tr>
<tr>
<td>CTF</td>
<td>Channel Transfer Function</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DOA</td>
<td>Direction-of-Arrival</td>
</tr>
<tr>
<td>DSO</td>
<td>Digital Sampling Oscilloscope</td>
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<tr>
<td>DS-SS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>DS-UWB</td>
<td>Direct-Sequence UWB</td>
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<tr>
<td>EGC</td>
<td>Equal Gain Combining</td>
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<tr>
<td>EIRP</td>
<td>Effective Isotropic Radiated Power</td>
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<tr>
<td>EM</td>
<td>Expectation and Maximisation</td>
</tr>
<tr>
<td>ESPRIT</td>
<td>Estimation of Signal Parameters via Rotational Invariance Techniques</td>
</tr>
<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
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<tr>
<td>FD</td>
<td>Frequency Domain</td>
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<tr>
<td>GO</td>
<td>Geometric Optic</td>
</tr>
<tr>
<td>GPIB</td>
<td>General Purpose Interface Bus</td>
</tr>
<tr>
<td>GTD</td>
<td>Geometric Theory of Diffraction</td>
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<tr>
<td>ICA</td>
<td>Independent Component Analysis</td>
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<tr>
<td>IFT</td>
<td>Inverse Fourier Transform</td>
</tr>
<tr>
<td>IR</td>
<td>Impulse Radio</td>
</tr>
<tr>
<td>IR-UWB</td>
<td>Impulse Radio Ultra Wideband</td>
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<tr>
<td>ISI</td>
<td>Inter Symbol Interference</td>
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<tr>
<td>JADE</td>
<td>Joint Angle of Arrival and Delay of Arrival Estimation</td>
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<tr>
<td>LNA</td>
<td>Low Noise Amplifier</td>
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<tr>
<td>Abbreviation</td>
<td>Definition</td>
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<tr>
<td>LOS</td>
<td>Line-of-Sight</td>
</tr>
<tr>
<td>MEA</td>
<td>Method of Equal Areas</td>
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<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
</tr>
<tr>
<td>MISO</td>
<td>Multiple Input Single Output</td>
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<tr>
<td>MPC</td>
<td>Multipath Components</td>
</tr>
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<td>MRC</td>
<td>Maximal Ratio Combining</td>
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<tr>
<td>MUSIC</td>
<td>Multiple Signal Classification</td>
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<tr>
<td>NLOS</td>
<td>Non-Line-of-Sight</td>
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<tr>
<td>PAM</td>
<td>Pulse Amplitude Modulation</td>
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<tr>
<td>PAS</td>
<td>Power Angular Spectrum</td>
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<tr>
<td>PDADS</td>
<td>Power Delay Azimuth Density Spectrum</td>
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<td>PDP</td>
<td>Power Delay Profile</td>
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<td>PDS</td>
<td>Power Delay Spectrum</td>
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<td>PHY</td>
<td>Physical Layer</td>
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<tr>
<td>PN</td>
<td>Pseudonoise</td>
</tr>
<tr>
<td>PPM</td>
<td>Pulse Position Modulation</td>
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<tr>
<td>PPM-TH</td>
<td>Pulse Position Modulation Time-Hopping</td>
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<tr>
<td>P-RAKE</td>
<td>Partial RAKE</td>
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<tr>
<td>RF</td>
<td>Radio Frequency</td>
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<tr>
<td>RMS</td>
<td>Root Mean Square</td>
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<tr>
<td>RMSE</td>
<td>Root Mean-Squared Error</td>
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<td>SAGE</td>
<td>Space Alternating Generalised Expectation-Maximisation</td>
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<td>SD</td>
<td>Selection Diversity</td>
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<tr>
<td>SIC</td>
<td>Successive Interference Cancellation</td>
</tr>
<tr>
<td>SIMO</td>
<td>Single Input Multiple Output</td>
</tr>
<tr>
<td>SINR</td>
<td>Signal-to-Interference-Plus-Noise Ratio</td>
</tr>
<tr>
<td>SISO</td>
<td>Single Input Single Output</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
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<td>S-RAKE</td>
<td>Selective RAKE</td>
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<td>SSA-PDP</td>
<td>Small-Scale Averaged PDP</td>
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<td>S-V</td>
<td>Saleh and Valenzuela</td>
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<tr>
<td>TD</td>
<td>Time Domain</td>
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<tr>
<td>TDL</td>
<td>Tapped Delay Line</td>
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<td>TH</td>
<td>Time Hopping</td>
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<td>TOA</td>
<td>Time-of-Arrival</td>
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<td>TR</td>
<td>Transmitted Reference</td>
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<td>Description</td>
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<tr>
<td>TRM</td>
<td>Time-Reversal Mirror</td>
</tr>
<tr>
<td>ULA</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td>UTD</td>
<td>Uniform Theory of Diffraction</td>
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<tr>
<td>VNA</td>
<td>Vector Network Analyser</td>
</tr>
<tr>
<td>WBAN</td>
<td>Wireless Body Area Network</td>
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<tr>
<td>WPAN</td>
<td>Wireless Personal Area Network</td>
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<td>$\alpha^2$</td>
<td>Shape Factor</td>
</tr>
<tr>
<td>$\alpha_n$</td>
<td>Energy-Normalised Channel Gain Parameter</td>
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<tr>
<td>$\alpha_{c,(f,\tau)}$</td>
<td>Spatial Complex Gain</td>
</tr>
<tr>
<td>$\beta$</td>
<td>RMS Bandwidth</td>
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<tr>
<td>$\gamma_k(\tau)$</td>
<td>Distortion of the $k^{th}$ component in the $i^{th}$ cluster</td>
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List of Symbols

\( \sigma \) Conductivity of Reflecting Surface
\( \sigma^2 \) Variance
\( \sigma_{\theta s} \) Angular Spread
\( \sigma_{css} \) Delay Spread
\( \tau \) Time Delay
\( \tau_e \) Timing Error
\( \tau_{fed} \) Feeding Network Time Delay
\( \tau_{i,k}(s) \) Time of Arrival of the \( k \)th Multipath component in the \( i \)th Cluster
\( \tau_m \) Mean Excess Delay
\( \tau_{max} \) Maximum Propagation Delay
\( \tau_n \) Propagation Delay of the \( n \)th Multipath Component
\( \{ \tau_n' \} \) Estimated Tapped Delay Positions
\( \tau_{rms} \) Root Mean Square (RMS) Delay
\( \theta \) Elevation Angle
\( \theta_d \) Angle of Diffraction
\( \theta_i \) Angle of Incident
\( \theta_r \) Angle of Reflection
\( \theta_t \) Angle of Transmission
\( \sigma_{i,k} \) Individual Tap Weighting Coefficients
\( w \) Combining Weighting Vector for RAKE Receiver
\( w_q \) Weighting Vector for the \( q \)th Spatial Combiner
\( a_{i,l} \) Amplitude of Multipath Component
\( a_n \) Impulse Response Amplitude of the \( n \)th Multipath Component
\( \{ a_n' \} \) Estimated Tapped Delay Gains
\( \frac{1}{\sigma_{0,0}^2} \) Expected Value Power of the First Arriving Multipath Component
\( a_{m}(x) \) Path Transfer Function for each \( m \)th Sensor
\( AF(t) \) Time Domain Array Factor
\( A_s \) Phase Matrix for the \( s \)th Subband
\( B \) Bandwidth
\( B_f \) Fractional Bandwidth
\( B_{\text{planar}}(\tau, \theta, \phi) \) Delay-and-Sum Beamforming for Planar Array Structure
\( B_{\text{ULA}}(\tau, \theta, \phi) \) Delay-and-Sum Beamforming for Uniform Linear Array Structure
\( c \) Light Velocity = 3×10^8 m/s
\( c(t_{n} - t_{ref}) \) Excess Distance Relative to the Reference Array
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
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<td>Pseudorandom Code</td>
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<td>$C_f$</td>
<td>Pseudorandom TH Code for the $k^{th}$ Source</td>
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<td>$C(\tau, \theta, \phi)$</td>
<td>Clean Map</td>
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<td>$d$</td>
<td>Inter-Element Spacing Distance</td>
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<td>$\bar{d}_{(i,j),\text{ref}}$</td>
<td>Gap Distance between the $(i,j)$ Element and the Reference Array</td>
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<td>$d_{\text{CRLB}}$</td>
<td>Cramer-Rao Lower Bound Distance Error</td>
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<td>$df$</td>
<td>Small Frequency Range in the $s^{th}$ Subband</td>
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<td>$d^k$</td>
<td>The $i^{th}$ Binary Data Bit Transmitted by the $k^{th}$ Source</td>
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<td>$d_{n,\text{ref}}$</td>
<td>Gap Distance between each $n^{th}$ Antenna and the Reference Antenna</td>
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<td>$d_{\text{thickness}}$</td>
<td>Slab Thickness</td>
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<tr>
<td>$dx$</td>
<td>Sensor Spacing in $x$-Direction</td>
</tr>
<tr>
<td>$dy$</td>
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<tr>
<td>$D$</td>
<td>Transmitter-to-Receiver Distance</td>
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<td>$D_0$</td>
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<td>Dirty Map</td>
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<td>$\text{erfc}$</td>
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<td>$E[\cdot]$</td>
<td>Expected Value Operation</td>
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<tr>
<td>$E_h$</td>
<td>Received Energy per Bit</td>
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<td>$E_{\text{Rx}}$</td>
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<td>$f_u$</td>
<td>Upper Bandwidth</td>
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<td>$f_l$</td>
<td>Lower Bandwidth</td>
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<td>$f_s$</td>
<td>Centre Frequency of the Considered Frequency Subband</td>
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<td>Total Multipath Gain</td>
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<td>$G(\tau)$</td>
<td>Energy-Normalised Gaussian Pulse</td>
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<tr>
<td>$h(\tau)$</td>
<td>Channel Impulse Response</td>
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<tr>
<td>$h_{\text{TM}}(\tau, \theta, \phi)$</td>
<td>Time Domain Transfer function of Antenna Arrays</td>
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<td>$h(s, \tau, \theta, \phi)$</td>
<td>Combined Time-Varying Channel Impulse Response</td>
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<td>$h_{\text{planar}}(\tau, \theta, \phi)$</td>
<td>Channel Impulse Response for Planar Array Structure</td>
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<td>Channel Impulse Response for Uniform Linear Array Structure</td>
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<tr>
<td>$h_{\text{R}_T, \text{R}_L}$</td>
<td>Transmitter Height, Receiver Height</td>
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<tr>
<td>$H(d, f)$</td>
<td>Distance-Frequency-Dependent Channel Transfer Function</td>
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<tr>
<td>$H(f)$</td>
<td>Frequency Domain Channel Response</td>
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</table>
List of Symbols

\( \mathbf{h}_q \)  
Impulse Response Vector for the \( q \)th Spatial Combiner

\( \mathbf{H}_j \)  
Reference Channel Transfer Function

\( I \)  
Array Element Number in Column Position for Planar Array Structure

\( IC \)  
Independent Component

\( J \)  
Array Element Number in Row Position for Planar Array Structure

\( K \)  
Ray or Contribution Number

\( K(l) \)  
Multipath component Number of the \( l \)th Cluster

\( L \)  
Cluster number

\( l(p,q) \)  
Path Number Selected by the \( q \)th Correlator at the \( p \)th Antenna

\( L(s) \)  
Cluster Number of the \( s \)th Subband Channel at each Time Instant \( t \)

\( L_r \)  
Propagation Path Number for RAKE Receiver

\( m_q(\tau) \)  
Correlator Mask for the \( q \)th Correlator Finger

\( m(\tau) \)  
Receiver Template

\( M \)  
Array Element Number for Uniform Linear Array Structure

\( n(\tau) \)  
Additive White Gaussian Noise

\( N \)  
Incoming Multipath Component Number

\( N_h \)  
Hop Numbers

\( N_o \)  
Noise Power Spectral Density

\( N_{ma} \)  
Frequency Resolution or Frequency Point Number

\( N_s \)  
Pulse Numbers per One Data Bit or Repetition Code Length

\( p(\tau) \)  
Gaussian Pulse

\( p(\phi) \)  
Uniform AOA Distribution

\( P \)  
Array Antenna Number for Array RAKE Receiver

\( P_{ij} \)  
Signal Power captured at \((i,j)\) Position

\( P_{kl} \)  
Average power of the \( k \)th Multipath component in the \( l \)th Cluster

\( PL_0 \)  
Mean Path Loss Reference

\( PL(d,f) \)  
Distance-Frequency-Dependent Path Loss

\( Pr_b \)  
Probability of Bit Error Rate

\( P(\tau) \)  
Power Delay Profile

\( P(\tau,\phi) \)  
Power Delay Azimuth Density Spectrum

\( P(R_{ij}/R_{i-1,j}) \)  
Ray Relative Power

\( P(T(T_{ij}) \)  
Cluster Relative Power

\( Q \)  
Correlator Finger Number

\( r_s(f) \)  
Scattering Loss

\( R(\tau) \)  
Received Signal
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tr>
<td>$R_{RR}$</td>
<td>Cross-Correlation between Beamformer Response Function and Received signals</td>
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<tr>
<td>$R_{n,k}$</td>
<td>MPC Signal Received at the $n^{th}$ Receive Antenna at the $k^{th}$ Time Bin</td>
</tr>
<tr>
<td>$R_{ref,k}$</td>
<td>MPC Signal Received at the Reference Sensor at the $k^{th}$ Time Bin</td>
</tr>
<tr>
<td>$R_{sh}()$</td>
<td>Total Reflection Coefficient</td>
</tr>
<tr>
<td>$R_{xx}$</td>
<td>Auto-Correlation Function</td>
</tr>
<tr>
<td>$\bar{R}_{ij,k}$</td>
<td>Average Correlation Coefficients</td>
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<td>$r_{q}[n]$</td>
<td>Output Vector of the $q$th Spatial Combiner</td>
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<tr>
<td>$R_{i,j,k}$</td>
<td>Spatial Correlation Matrix for Planar Array Structure</td>
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<tr>
<td>$R_{n,ref,k}$</td>
<td>Spatial Correlation Matrix for Uniform Linear Array Structure</td>
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<tr>
<td>$s^k(t,l)$</td>
<td>The $k^{th}$ User's Signal Conveying the $l^{th}$ Data Bit</td>
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<td>$S$</td>
<td>Frequency Subband Number</td>
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<td>$T_b$</td>
<td>Bit Duration</td>
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<tr>
<td>$T_c$</td>
<td>TH Chip Period</td>
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<tr>
<td>$T_i(\theta,\phi)$</td>
<td>Steering Delay</td>
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<tr>
<td>$T_i(s)$</td>
<td>Time of Arrival of the $i^{th}$ Cluster</td>
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<tr>
<td>$T_f$</td>
<td>Average Pulse Repetition Period or Frame Duration</td>
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<tr>
<td>$T_{sh}(\tau)$</td>
<td>Total Transmission Coefficient</td>
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<tr>
<td>$\overline{TOA}$</td>
<td>Average TOA</td>
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<tr>
<td>$U[:::]$</td>
<td>Uniform Distribution</td>
</tr>
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<td>$v$</td>
<td>Array Response Vector</td>
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<tr>
<td>$V$</td>
<td>Array Steering Matrix</td>
</tr>
<tr>
<td>$V(\phi)$</td>
<td>Array Manifold Vector</td>
</tr>
<tr>
<td>$w_i(\tau)$</td>
<td>Total Interference and Noise for $(i,j)$ position</td>
</tr>
<tr>
<td>$W(\tau)$</td>
<td>Transmitted Signal</td>
</tr>
<tr>
<td>$W'(\tau)$</td>
<td>Transmitted Information Bit Signal</td>
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<tr>
<td>$w_a$</td>
<td>Additive White Gaussian Noise Vector</td>
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<tr>
<td>$x_i(\tau)$</td>
<td>Received Continuous-Time Signal</td>
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<td>$x(\tau)$</td>
<td>Array Signal Vector for Tap Delay Line</td>
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<tr>
<td>$X$</td>
<td>Tap Delay Line Array Signal Matrix</td>
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<td>$y'(\tau)$</td>
<td>Coded Pulse Signal</td>
</tr>
</tbody>
</table>
Chapter 1

1 Introduction

1.1 Overview

The availability of higher frequency bands has made wireless communications an attractive and advantageous function for broadband digital communications. In wireless network systems, all four dimensions of radio transmission: time, frequency, code and space, can be analysed to distinguish radio links and to optimise frequency bandwidth utilisation. The synthesis of an antenna directivity pattern and diversity of antenna arrays can be interpreted as spatial processing. Operations such as demodulation, filtering or Fourier analysis of the antenna output signal are temporal processing techniques. This processing is required whenever there is a dependent function between the spatial and temporal variables. Space-time processing techniques can typically be applied in areas such as: interference suppression for broadband radar, wideband interference rejection in receive arrays, cancellation of main beam interference due to multipath, space-time coding for an array for simplification of beamforming, simultaneous frequency and direction-of-arrival (DOA) estimation in a multiple source environment, and so on [Kle02].

The use of Impulse Radio Ultra Wideband systems (IR-UWB) is increasing, especially for short-range communications such as wireless personal area networks (WPANs) [Mol03], [Par07], wireless body area networks (WBANs) [For06], [GAN08] and sensor networks [Gez05], [Han06c], and [Pat05]. This technology offers many advantages over conventional narrowband and wideband systems. In addition, UWB radio systems are severely limited in terms of transmitted power, but due to the wide bandwidth they suffer less from multipath fading effects that are a great disadvantage in narrowband radio [Mal08c]. Therefore, knowledge about the characteristics of the radio channel using these transmission links is required. Consequently, numerous channel measurements have been carried out covering various propagation scenarios and, in parallel, channel models have been proposed to facilitate simulations. UWB signals are fundamentally different from narrowband signals. The UWB systems spread the energy from several hundred MHz to a few GHz. The channel is extremely frequency selective and the received signal is composed of a significant number of multipath components that have different delays in the order of nanoseconds [Win00], [For02a]. These are caused by significant variation
of electromagnetic propagation parameters over the wide frequency range. These differences should be taken into account for designing UWB receiving architectures [Kou08].

In order to provide a wide variety of services through reliable high-data-rate wireless channels and to improve the wireless transmission quality, techniques that can improve spectral efficiency and combat channel impairments have been developed. Thus, the investigation of antenna array systems and smart antennas becomes popular. They are mainly focused on the channel exploitation for channel capacity and, also, on space diversity to achieve maximum diversity and coding gain [Kei06], [Mal07], [Ali07]. These systems can offer a linear capacity increase and a diversity gain that can be used to solve the significant problems of fading and multi-user interference. Smart antennas employing Multiple Input Multiple Output (MIMO) systems have also been investigated. The initial concept of separating signal transmissions in a temporal and spatial domain can be found in systems which utilise a single transmitter and multiple antennas at the receiving side—Single Input Multiple Output (SIMO) [Ger05]. In general, systems with multiple antennas require channel models that characterise both temporal and spatial characteristics of the channel. This can lead to the deployment of new advanced transmission technologies and signal processing principles; for instance, adaptive beamforming and RAKE diversity techniques.

1.2 Motivation

Many extensive studies of UWB array propagation channels have been carried out in order to gain more profound knowledge, and numerous channel models are proposed for propagation channels via multi-antennas in both real and virtual arrays as conducted in [Win02], [Han04], [Cho05b], [Muq05]. Thus, characteristics of the temporal and the spatial channels are provided simultaneously. When considering the spatial aspect, UWB spatial correlation should be characterised to determine the effectiveness of diversity combining schemes or beam-forming techniques. Conventional narrowband work can be effectively operated when spatial correlation between received array signals represents a low correlation value, whereas beam-forming schemes correspond to a high degree of correlation. There are several factors that relate to the degree of spatial correlation, i.e. distance between the transmitter-receiver ends [Kyr03], gap distance between array elements [Ozd04], [Raj08], array orientation [Xin04a], polarisation [Liu05], and angular energy distribution [Sal94], [Tsa02]. Furthermore, since the geometry of obstructions and building architecture along propagation paths significantly affect UWB incoming contributions, spatial features in the prospect of azimuth and elevation domains should also be focused on.
These parameters can boost the spatial correlation characteristics and become the significant research issue.

In addition, since each multipath component corresponds to its own impulse response, time domain characteristics are also taken into account. Various temporal domain features are investigated such as temporal dispersion and distorted pulse waveforms regarding propagation path characteristics. There are other reports recently published regarding the issue of temporal domain UWB array propagation channels [Kai07], [Kei06]. The first paper presented the signal processing framework for MIMO UWB channels. The time domain impulse responses matching with particular attenuation delays, angles-of-departure (AODs) and angles-of-arrival (AOAs) were simulated. In the latter paper, Kei et. al. reported cross correlation between each array element and the reference array computing at each time bin. High degrees of the correlation value at the first bin were implied as the presence of the first incoming multipath arrival. Moreover, results were no longer correlated for the latter time bins. Recently, there has been another publication reported about spatial correlation studies in UWB channels conducted by frequency domain measurements [Mal08a]. This research is originated from the enthusiasm for investigation of UWB spatial correlations which is rarely conducted and not completely known at present. Similarly to the contribution of this thesis, the investigation results involved both magnitude and complex spatial correlation analysis and also correlation characteristics depending on array orientations.

Despite no existence of spatio-temporal works regarding the above issues at the beginning of the research, the appearance of these subsequent publications, which were simultaneously launched throughout the research course, can support the significance of space-time array UWB propagation channels. In particular, spatial correlation research considering the time bin cross-correlations becomes interesting and it is the important indicator for the classical RAKE receivers in the multi-antenna condition [Kei06]. However, the approach proposed by this thesis which gains contributions one step further than others is the investigation of distortion effects and propagation characteristics surrounding adjacent antennas. Implementing the proposed spatial correlation computation technique can define information of distorted UWB signals. Moreover, the applications of pointing the transmit antenna to different directions and moving receiving arrays in 0°-360° orientation can gain deep insight and quantitative measurement results to be analysed.
1.3 Research Objectives and Study Approach

The objective of this research is to investigate spatial and temporal characteristics of indoor UWB antenna array systems in all aspects, i.e. frequency dependence, classification of multipath clusters and distortion effects. SIMO propagation channel in indoor environments is the major reference issue to be investigated. Additionally, in order to evaluate the practical usefulness of the proposed spatial correlation technique for UWB array processing, further analysis and measurement campaigns were also carried out.

The serial path of objectives corresponding to study approach is described as follows:

1) To gain the comprehension of the scope and directions of this research, the UWB characteristics, UWB propagation channels and array processing systems were reviewed in detail.

2) The next objective is to gain an understanding of frequency dependent distorted channels. Research started from analysing the UWB channel impulse response (CIR) by modifying simulation of impulse response signals over frequency sub-bands regarding classified multipath clusters and distortion mechanisms. Data simulated from this task was used for link performance computation.

3) To investigate the spatial correlation between each array antenna, there is the initiation of a new computation technique for considering the spatial correlation coefficients at each time bin, based on data captured by UWB antenna arrays. The application of this technique could lead to UWB multipath cluster identification for the time of multipath arrivals.

4) To determine AOAs more accurately, complex correlation analysis and relative phase difference techniques were analysed. Linear array measurements were also conducted to gain sufficient knowledge of incoming multipaths in order to evaluate the proposed computation technique.

5) Finally, the results from UWB space-time array processing were taken into account to be utilised in RAKE systems. Time delay tracking for the different multipath contributions, which is generally based on correlation measurements and adaptive weighting factor algorithms were also studied in this work.
1.4 Novel Work Undertaken and Contributions

Figure 1-1 presents the scope of UWB array SIMO propagation channels carried out in this thesis. Multipath arrivals to array elements are taken into account under the assumption that each multipath signal arriving at each array at different delay time. Apart from measuring indoor wireless propagation channels which is normally carried out in other studies, UWB propagation channel simulation, analysis of frequency-dependent distortion effects, and space-time array characterisation are investigated. These contributions are studied regarding classifications of multipath clusters and directions of arriving multipaths.

![SIMO indoor propagation channels](image)

Array orientations are conducted in the experiments in order to observe incoming signals being scattered from obstructions in various directions. Furthermore, in this research spatial correlations are analysed at each high resolution time bin which means correlation characteristics are taken into account along with the pulse distortion effects. Significant results of complex correlation are obtained when processing both magnitude and complex measurement data differently from the conventional methods as conducted in [Win02], [Han04], [Cho05b], [Muq05] which are considered only the magnitude term to analyse UWB multipath parameters.

The main achievements in this research are:

1) Modified simulation structure of spatio-temporal UWB discrete CIRs based on classifications of multipath cluster and pulse distortion effects over each frequency sub-band. Frequency dependent characteristics of pulse signals are also generated.
2) Proposal of spatial correlation analysis based on array processing between each array element and its surrounding elements at each particular time bin. Its application to time-of-arrival (TOA) estimation of multipath clusters is also presented.

3) Analysis of the complex spatial correlation combining with the relative phase difference technique in a linear array aspect is proposed in order to determine AOA estimation and distortion effect at each individual time bin.

4) Investigation of the application of spatio-temporal processing unit on UWB receiver systems. Proposal of adaptive multipath searching unit in UWB array RAKE receiver systems is also highlighted.

The results of these research contributions have been published and submitted in journals and international conferences in the following papers.

**Journal Papers:**


**Conference Papers:**


1.5 Thesis Structure

The thesis structure can be described by the layout diagram of the thesis as depicted in Figure 1-2, which guides the concise methodologies used in each chapter. The remainder of the thesis is described as follows:

Chapter 2 describes a review of UWB propagation channels and UWB propagation models to be used in the simulation and analytical parts. Moreover, UWB array systems and space-time array processing are also reviewed.

Chapter 3 provides the modified UWB simulation over frequency sub-bands regarding multipath cluster classification and distortion effects. Thus both frequency-dependent-UWB CIRs and distorted waveforms are simulated. Distortion quantification is also analysed.

Chapter 4 proposes the spatial correlation technique computed on a UWB planar array database. Its application of TOA estimation is evaluated by other standard method and generic multipath UWB model.

Chapter 5 focuses on AOA estimation at each individual time using the complex spatial correlation analysis and the relative phase different technique. The linear array UWB measurement operated under control situations is also provided to simply evaluate this computation technique.

Chapter 6 presents analysis of UWB link performances in various scenario cases regarding to spatio-temporal applications on array RAKE receiver. The solution for distortion channels and multipath cluster searching is also derived herein.

Chapter 7, finally, draws conclusions on all the works, and rough ideas and scopes of further investigations are also provided.
Figure 1-2: Layout diagram of the thesis structure
Chapter 2

2 UWB Propagation Channel and Related Works

In recent years, many extensive studies have been carried out in order to gain more profound knowledge of UWB propagation channels. Its characteristic gains a great deal of interest for future short-range wireless communications due to ultra high data rate and low power transmission in an extremely large transmission bandwidth [Bat03]. UWB is also defined as the carrierless short pulse signal in the radar community, which is mainly exploited in military applications [Tay95]. More recently, UWB technology has been utilised for spread spectrum communications employing its fine delay resolution property [Win98a], [Win00a], [Ram01]. Furthermore, these significant characteristics can also lead UWB technology become the candidate for communication systems in dense multipath environments [Win98b].

2.1 Definition of UWB Radio Signal

2.1.1 Operating Frequency

The United States Federal Communications Commission (FCC), [FCC02] has defined a bandwidth measure for UWB signals as the fractional bandwidth, $B_f$, which is defined as the ratio of the energy bandwidth and the centre frequency.

\[
B_f = \frac{(f_H - f_L)}{\frac{f_H + f_L}{2}}
\]  (2.1)

where $f_H$ and $f_L$ are the upper and lower -10 dB bandwidth, respectively. The FCC defined UWB signals as those which have $B_f > 0.20$ or a bandwidth greater than 500 MHz. Table 2-1 presents the radiation limits for indoor and outdoor data communication applications schemed by the FCC.
UWB Propagation Channel and Related Works

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Indoor EIRP (dBm/MHz)</th>
<th>Outdoor EIRP (dBm/MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>960-1610</td>
<td>-75.3</td>
<td>-75.3</td>
</tr>
<tr>
<td>1610-1990</td>
<td>-53.3</td>
<td>-63.3</td>
</tr>
<tr>
<td>1990-3100</td>
<td>-51.3</td>
<td>-61.3</td>
</tr>
<tr>
<td>3100-10600</td>
<td>-41.3</td>
<td>-41.3</td>
</tr>
<tr>
<td>Above 10600</td>
<td>-51.3</td>
<td>-61.3</td>
</tr>
</tbody>
</table>

Table 2-1: FCC radiation limits for indoor and outdoor communication applications

<table>
<thead>
<tr>
<th>Frequency range (GHz)</th>
<th>Maximum mean EIRP density (dBm/MHz)</th>
<th>Maximum peak EIRP density (dBm/50MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Below 1.6</td>
<td>-90.0</td>
<td>-50.0</td>
</tr>
<tr>
<td>1.6 to 3.4</td>
<td>-85.0</td>
<td>-45.0</td>
</tr>
<tr>
<td>3.4 to 3.8</td>
<td>-85.0</td>
<td>-45.0</td>
</tr>
<tr>
<td>3.8 to 4.2</td>
<td>-70.0</td>
<td>-30.0</td>
</tr>
<tr>
<td>4.2 to 4.8</td>
<td>-41.3 (until December 31st, 2010)</td>
<td>0.0 (until December 31st, 2010)</td>
</tr>
<tr>
<td>4.8 to 6.0</td>
<td>-70.0 (beyond December 31st, 2010)</td>
<td>-30.0 (beyond December 31st, 2010)</td>
</tr>
<tr>
<td>6.0 to 8.5</td>
<td>-41.3</td>
<td>0.0</td>
</tr>
<tr>
<td>8.5 to 10.6</td>
<td>-65.0</td>
<td>-25.0</td>
</tr>
<tr>
<td>Above 10.6</td>
<td>-85.0</td>
<td>-45.0</td>
</tr>
</tbody>
</table>

Table 2-2: CEPT maximum EIRP densities in the absence of appropriate mitigation techniques

Due to using very wide spectrum range, UWB systems are not intended to operate under any specific allocation. However, the European approach is more cautious than FCC giving cause for concern that a new technology should cause little or no harm to existing systems. In order to avoid harmful interference to some of the existing radio communication systems, the European Conference of Postal and Telecommunications (CEPT) working group considered the appropriate frequency bands for UWB and indicated that radiation limits of UWB devices for communication applications would need to be lower than the FCC limits [CEP03]. In 2007, the European communities published the commission Decision on allowing the radio spectrum utilisation for equipment using UWB technology in communities as described in Table 2-2 [EC07]. The effective isotropic radiated power (EIRP) stated by this decision can be operated in both indoor...
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Figure 2-1: Comparison of FCC and CEPT UWB radiation masks

and outdoor propagation. This Decision is operated indoors in assuming that the antenna will stop transmission within ten seconds unless it receives an acknowledgement from the receiver that its transmission is being received. Additionally, the CEPT mask is also operated outdoors provided that these frequency masks should not be transmitted at a fixed outdoor location or connected to a fixed outdoor antenna or in vehicles. Figure 2-1 demonstrates the UWB radiation masks defined by FCC and CEPT with the maximum mean values. The maximum average power spectral density of all cases is limited by the radiation threshold proposed by the FCC Part 15 regulations [FCC02].

Furthermore, the IEEE also established two study groups to define new physical layer concepts utilising UWB technology: the IEEE 802.15.3a Study Group (already disbanded) for short-range and high data rate applications and the IEEE 802.15.4 for low data rate applications. The purpose of the IEEE 802.15.3a was to provide a higher speed physical layer (PHY) candidate, with a minimum data rate of 110 Mbps at 10 m, for the IEEE 802.15.3 medium access standard. The applications of this standard involve cable replacement in a short distance such as image, multimedia links and WPAN [IEEE 802.15.3-2003]. The model analysis of the UWB radio channel is also reported in this standard [IEEE 802.15.SG3a]. On the other hand, another established IEEE standard for low data rate communications, the IEEE 802.15.4, is intended to operate with very low power and very low complexity systems in unlicensed and international frequency bands such as applications in sensors, interactive toys and remote controls [IEEE 802.15.4-2003].
2.1.2 Impulse Radio Schemes

UWB impulse radio is a spread spectrum technology which is different from other conventional spread spectrum techniques whereby instead of transmitting data information on modulated continuous carrier signals, UWB transmits information through a short pulse or *chirped* signal. Thus, UWB can gain many significant advantages such as simple structure of the transmitter, low transmission power, extremely large bandwidth which makes signal detection by unassigned users become difficult, and high bit rate of more than a few hundreds of Mbps. In the radar field, UWB radio is based on the waveform radiation formed by a sequence of very short pulses or pulsed waveforms (monocycles). This transmission technique is called as *Impulse Radio* (IR) with pulse duration typically about a few hundred picoseconds. Each pulse contains large spectrum which must meet the spectral mask requirement. IR-based UWB is similar to spread-spectrum systems where the major difference is that the spread-spectrum *chip* is replaced by a discontinuous UWB *pulse*. This type of transmission does not employ the additional carrier modulation since the pulse can propagate well in the propagation channels using the baseband signalling technique. Information to be transferred is presented in a digital form (0 or 1) by a binary sequence. Each transmitted bit is spread over one or more pulses in a code-repetition pattern. This can provide resistance to noise and interference [Ben06].

To occupy 500 MHz bandwidth, UWB signals can be produced by transmitting a very high data rate of independent pulses. The pulses might satisfy the Nyquist criterion at an operating pulse rate $1/T$ with a minimum bandwidth of $B=1/(2T)$. Generally, the most adopted pulse shape that is currently used in UWB transmission systems is modelled as the second derivative of the Gaussian function, Gaussian monocycle or doublet [Win00a]. With regard to a Gaussian pulse, $p(\tau) = \pm \exp(-\tau^2/(2\sigma^2)) / \sqrt{2\pi}\sigma = \pm \sqrt{2}/\alpha \cdot \exp(-2\pi^2/\alpha^2)$, the second derivative of a Gaussian function can be expressed by

$$g(\tau) = \frac{d^2 p(\tau)}{d\tau^2} \propto \left(1 - 4\pi \frac{\tau^2}{\alpha^2}\right) \exp\left(-\frac{2\pi^2}{\alpha^2}\right)$$

(2-1)

where $\alpha^2=4\pi \sigma^2$ is the shape factor and $\sigma^2$ is the variance. Ideally, when feeding the UWB transmitting antenna with a current pulse shaped as the first derivative of a Gaussian waveform, a second derivative Gaussian pulse can be detected at the output of the transmitting antenna. This output pulse is referred to as the pulse at the receiver additionally [Ben04]. The waveforms of the Gaussian pulse (the first picture) and its first 15 derivatives are plotted in Figure 2-2.
There are various common and traditional ways of transmitting UWB signals. These techniques rely on the simple modulation scheme where information data symbols modulate pulses, such as Pulse Position Modulation (PPM) and Pulse Amplitude Modulation (PAM). In order to shape spectrum of generated-signals, the data symbols are encoded using pseudorandom or pseudonoise (PN) codes. The most common IR based UWB approach is relied on PPM time-hopping (PPM-TH), which encoded data symbols introduce a time dither or randomisation on generated pulses. Typical PPM-TH-UWB signals can be derived by [Hu04]

$$s^k(\tau,i) = \sqrt{\frac{E_b}{N_s}} \sum_{j=-N_s}^{N_s-1} g(\tau - jT_s - c^k T_c - d^k \varepsilon)$$  \hspace{1cm} (2-2)$$

$s^k(\tau,i)$ is the $k^{th}$ user's signal conveying the $i^{th}$ data bit. $g(\tau)$ is the energy-normalised Gaussian pulse. $E_b$ is the bit energy; $N_s$ is the pulse numbers for transmitting one data bit or the length of the repetition code. $T_s$ is the average pulse repetition period or the time duration of frame, thus the bit duration $T_b\approx N_s T_s$. $T_c$ is the TH chip period which satisfies $N_h T_c \leq T_s$ where $N_h$ is the number of hops. $\{C^k\}$ represents the pseudorandom TH code for the $k^{th}$ source. $d^k$ presents the $i^{th}$ binary data bit transmitted by the $k^{th}$ source which the data bits are $\{0,1\}$ or $\{1, -1\}$, and $\varepsilon$ is the PPM time shift.
Furthermore, for radiating very short pulses, direct-sequence UWB (DS-UWB), an amplitude modulation scheme of basic IR pulses by encoded data symbols seems particularly attractive [For02b]. In particular, TH-UWB and DS-UWB signals are capable of generating UWB signals with spectrum expansion by using very short pulses. An example of PPM-TH-UWB or 2PPM-TH-UWB (binary PPM-TH-UWB) and PAM-DS-UWB are depicted in Figure 2-3. PPM-TH-UWB and PAM-DS-UWB signals are generated by TH-code = \([2 0 0 0 0]\) and DS-code = \([1 -1 1 1 -1 1 1 -1 1 -1 1]\) respectively.

### 2.2 Characteristics of UWB Propagation Channel

UWB signal propagation in indoor-outdoor environments is the most important issue with significant impacts on the success of UWB technology. If the channel is well characterised, the disturbance effect can be reduced by suitable designs for transmission and reception. The typical UWB propagation channel does not mainly depend on geometry of the environment. Rough information of surroundings is supposed to be sufficient for its characterisation [Gha04]. The main difference between UWB propagation channels and conventional narrowband channels is based on the frequency dependent transfer functions. Frequency dependence in narrowband channels typically varies within a few MHz bandwidths due to different multipath components (MPCs). However, UWB frequency dependence presents variations of averaged transfer functions caused by different attenuations relative to the whole system bandwidth.
A categorisation of UWB propagation channels can be examined regarding the arrivals of MPCs. The channel that allocates time intervals between each incoming MPC larger than the inverse of the channel bandwidth can be identified as the sparse channel. It can be implied that the larger the bandwidth, the more likely the channel is sparse. Thus, not all resolvable time bins carry significant amounts of energy. On the other hand, another type of UWB channel is called the dense channel; the inter-arrival time of multipath components (MPCs) is less than the resolvable bin width. Dense channels commonly appear in the severe scattering environments such as a propagation path with a large number of reflecting and diffracting obstructions. It can occur even in the extremely large bandwidth. This has an important consequence for RAKE receivers. In particular, the multipath arrival statistics of sparse channel can be observed much more easily than in narrowband channels [Mol05b].

2.2.1 Statistical Channel Models

To evaluate the coverage and the system capacity, analysis and simulation relying on link budget calculations are required. Since link budget analyses are based on the accurate path loss model of UWB channels, UWB propagation losses become more critical due to interferences from other coexistent sharing radio frequency (RF) spectrum systems. Furthermore, the low transmitted power of UWB signals can lead to the significant degradation of communication links. Statistical channel models that predict signal strength are considered in two aspects. The first one is the large scale model which signal power over large distances (several hundreds or thousands meters in outdoors, or tens of meters in indoors) between the transmitter and the receiver is characterised. Secondly, the small scale model or fast fading considers signal power over short distances up to 15-30m in outdoors or up to a few meters in indoors. Signals have deep fading and vary rapidly. The large scale model predicts the average power of small scale variations.

2.2.1.1 Large Scale Channel Characterisation

Large scale fading or shadowing is the variation of local attenuations around the deterministic mean path loss. It is related to diffraction and reflection effects on MPCs anticipating frequency dependence along propagation paths. Generally, the large scale fading from house to house can be modelled as a lognormal distribution, which is consistent with the narrowband shadowing, with shadow variances, $\sigma^2$, of 1-2 dB for line-of-sight (LOS) and 2-6 dB for non-line-of-sight (NLOS). However, there is another report of normal distribution for shadowing suggested by [Gha02]. The total attenuation due to shadowing and path loss can be written by (2-3) with $PL_0$ and $n$ notify the mean path loss reference and the path loss exponent respectively. $\sigma$ is the standard deviation of shadow fading.
UWB Propagation Channel and Related Works

\[ [PL_0 + 10\mu_n \log_{10}(d)] + [10n_1\sigma_n \log_{10}(d) + n_2\mu_\sigma + n_3\sigma_\sigma] \]  \hspace{1cm} (2-3)

where \( n_1, n_2, \) and \( n_3 \) are zero-mean, unit-variance Gaussian variables with the range of \( n_i \) between \([-0.75, 0.75]\) and \( n_2, n_3 \) between \([-2, 2]\) to avoid nonphysical values of the attenuation [Ben06]. The mean values of \( \mu_n, \sigma_n, \mu_\sigma, \) and \( \sigma_\sigma \) respectively, and their standard deviation values, \( \sigma_n \) and \( \sigma_\sigma \), can be found in Table 1 in [Gha02].

Alternatively, due to the significant properties of UWB channels, the UWB path loss can simply be derived depending on both distance and frequency; \( PL(D, f) \propto D^n \cdot f^m \) with \( n \neq m \). \( PL \) is the mean path loss where \( n \) is the mean path loss exponent representing that how fast path loss increases with distance \( D \). This term is obtained by performing a linear regression of path loss dB values versus the logarithmic distances. Similarly for the frequency dependent term, the coefficient \( m \) varies between positive and negative values corresponding to propagation environments. This value varies between 0.8 and 1.4 including antenna effects [Kun02] and -1.4 in industrial environments and +1.5 in residential environments. The path loss dependence on distance and frequency can be expressed by the product between the two independent functions of distance and frequency:

\[ \overline{PL}(D, f) = \overline{PL}_D(D) \cdot \overline{PL}_f(f) \]  \hspace{1cm} (2-4)

The distance dependence path loss can be modelled by (2-5) which is similar to narrowband channel model. This equation describes the summation of the path loss in dB from the transmitter to the reference distance \( D_0 \) and the additional path loss. \( n \) can be extracted by UWB experimental data; ideally, \( n=2 \) for free space propagation.

\[ \overline{PL}(D)_{dB} = PL(D_0)_{dB} + 10n \log_{10}\left( \frac{D}{D_0} \right) \]  \hspace{1cm} (2-5)

Frequency dependence in UWB propagation channels is significant because antenna responses cannot be isolated from the channel behaviour. The total frequency-averaged path loss can be obtained by integrating the channel frequency response over the frequency band of interest as given in (2-6) where \( E[\cdot] \) is the expected value operation. \( H(D, f) \) is the channel transfer function, and \( \Delta f \) is small enough that diffraction coefficients, reflection coefficients, dielectric
constants, etc. can remain constant over that bandwidth. The frequency path loss is usually given as \(PL(f) \propto f^{-\alpha}\) [Qiu99].

\[
\overline{PL}(D, f) = E \left[ \int_{f-\Delta f/2}^{f+\Delta f/2} |H(D, f)|^2 df \right]
\]  

### 2.2.1.2 Small Scale Channel Characterisation

In small scale fading, the relative amplitude over small scale areas is taken into consideration. Amplitude fading statistics of conventional narrowband channels are described by Rayleigh and Rice distributions which model the severe multipath conditions without LOS. In UWB propagation, since the extremely wide bandwidth is occupied, leading to a high temporal resolution capability, a single path incoming at a certain delay bin can be resolved. As a result, amplitude distribution in each delay bin becomes different from the Rayleigh fading distribution. Small scale amplitude distribution can be empirically determined by calculating bin data at specific excess delays. Increase in deep fading can be found with increasing of time delays. These UWB small-scale statistics in various environments are matched to typical theoretical distributions such as Nakagami distribution, Rice distribution which describes the envelope of summation of one dominant component and other smaller components, Lognormal distribution with its advantage that small scale and large scale fading remains the same statistic form, Weibull distribution [Alv03], and Rayleigh distribution which is measured in an industrial environment even for a 7.5 GHz operated frequency bandwidth, very small resolvable binwidth [Mol05b], [Ars06]. Moreover, POCA and NAZU distributions are also suggested for UWB applications [Zha02]. The amplitude distribution is modelled by the superposition of a small number of equal-strength MPCs (POCA) including the possible additional strong specular component (NAZU).

Considering UWB multipath contribution characteristics, the IEEE channel modelling subcommittee finally converged on the model based on the cluster approach proposed by Turin [Tur72] and further formalised by Saleh and Valenzuela [Sal87] (S-V model). The S-V model is based on the observation that multipath contributions generated by the same pulse usually arrive at the receiver and are grouped into clusters. The TOA of clusters is modelled as a Poisson arrival process with rate \(A\) as given in (2-7) where \(T_i\) and \(T_{i-1}\) are the TOAs of the \(i^{th}\) and the \((i-1)^{th}\) clusters respectively.

\[
p(T_i \mid T_{i-1}) = A \exp[-A(T_i - T_{i-1})]
\]  

(2-7)
Within each cluster, subsequent multipath contributions also arrive according to a Poisson process with rate $\lambda$:

$$p(\tau_{k,l} | \tau_{(k-1),l}) = \lambda \exp[-\lambda(\tau_{k,l} - \tau_{(k-1),l})] \tag{2-8}$$

where $\tau_{k,l}$ and $\tau_{(k-1),l}$ are the TOAs of the $k^{th}$ and $(k-1)^{th}$ contributions respectively within the cluster $l$. TOAs of the first contribution within each cluster $\tau_{1,l}$, for $l=1,\ldots,L$, are set to zero. In this model, the gain of the $k^{th}$ ray in the $l^{th}$ cluster is a complex random variable with amplitude $a_{k,l}$ and phase $\theta_{k,l}$. The first values, $a_{k,1}$, are assumed to be statistically independent and Rayleigh distributed positive random variables, whereas the latter values, $\theta_{k,l}$, are assumed to be statistically independent uniform random variables over $[0, 2\pi)$.

$$p(a_{k,l}) = \frac{2a_{k,l}}{E[|a_{k,l}|^2]} \exp \left( \frac{-a_{k,l}^2}{E[|a_{k,l}|^2]} \right) \tag{2-9}$$

$$p(\theta_{k,l}) = \frac{1}{2\pi} \quad \text{with} \quad 0 \leq \theta_{k,l} < 2\pi$$

$$\overline{a_{k,l}^2} = a_{0,0}^2 \cdot \exp(-T_1/\Gamma) \cdot \exp(-\tau_{k,l}/\gamma) \tag{2-10}$$

$\overline{a_{0,0}^2}$ is the expected value of the power of the first arriving MPC. $\Gamma$ and $\gamma$ are the exponential decay factors of clusters and rays respectively. With respect to (2-10), the average power delay profile (PDP) is characterised by an exponential decay of the cluster amplitude, and a different exponential decay for the ray, or arrival, amplitude within each cluster, as illustrated in Figure 2-4 [Spe00].

![Figure 2-4: Typical PDP for S-V channel model](image)

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2.2.2 UWB Propagation Models

2.2.2.1 Impulse Response Model

Random and complicated indoor radio propagation channels can be characterised using the impulse response approach as,

\[ h(\tau) = \sum_{n=1}^{N} a_n \delta(\tau - \tau_n) e^{i\phi_n} \]  

(2-11)

where \( N \) is the total number of resolved time bins, \( a_n, \tau_n, \) and \( \phi_n \) are random amplitude, propagation delay and carrier phase sequences respectively. \( \delta(.) \) is the Dirac delta function. This model was first suggested by Turin; all parameters characterising the channel are random variable with specific distributions [Tur56].

The parameter that is usually derived from the impulse response model in (2-11) is the total multipath gain, \( G \). This parameter measures the total energy collected over the \( N \) received pulses with unitary energy transmitted. It can be defined as

\[ G = \sum_{n=1}^{N} |a_n|^2 \]  

(2-12)

Thus, impulse response can be rewritten as (2-13) where \( a_1, ..., a_N \) are the energy-normalised channel gain parameters, \( \sum_{n=1}^{N} |a_n|^2 = 1 \) [Ben04], [Zha06].

\[ h(\tau) = \sqrt{G} \sum_{n=1}^{N} a_n \delta(\tau - \tau_n) e^{i\phi_n} \]  

(2-13)

Note that \( G \leq 1 \) and is related to the attenuation during propagation. In multipath environments, \( G \) decreases with distance; \( G = G_0/D^\gamma = 10^{A_0/10}/D^\gamma \). \( A_0 \) (dB) is the path loss at the reference distance \( D_{ref}=1 \) m: \( A_0 \) (dB) = 10log\((E_{Tx}/E_{Rx0})\). For LOS, the reference path loss \( A_0=47 \)dB, \( \gamma = 1.7 \), and \( A_0=51 \)dB, \( \gamma = 3.5 \) for NLOS.

Another parameter that can be derived from the impulse response characteristics is time dispersion or delay spread. Normally, this effect is characterised by the first central moment and the square root of the second central moment of the PDPs. Thus, those values are represented by...
the mean excess delay, $\tau_m$, and the root mean square (rms) delay, $\tau_{rms}$, as defined by (2-14) and (2-15) respectively [Cic05], [Ben04].

$$\tau_m = \frac{\sum_{n=1}^{N} a_n^2 \tau_n}{\sum_{n=1}^{N} a_n^2}$$  \hspace{1cm} (2-14)

$$\tau_{rms} = \sqrt{\bar{\tau}^2 - \tau_m^2} \quad \text{where} \quad \bar{\tau}^2 = \frac{\sum_{n=1}^{N} a_n^2 \tau_n^2}{\sum_{n=1}^{N} a_n^2}$$  \hspace{1cm} (2-15)

### 2.2.2.2 Discrete Time Impulse Response

Hashemi [Has93] suggested a convenient discrete impulse response model for characterising channels affected by multipaths. In this model, the time axis is divided into small time intervals or bins, which are assumed to contain either one MPC, or no MPC. According to (2-11), the discrete time channel model can be rewritten as

$$h(\tau)_{\text{discrete}} = \sum_{n=1}^{N} a_n \delta(\tau - n\Delta \tau) e^{j\phi_n}$$  \hspace{1cm} (2-16)

$\Delta \tau$ is the time duration of each bin or the minimum resolved time bin.

In particular a log-normal distribution is suggested for characterising the multipath gain amplitudes, and additional log-normal variable is introduced for presenting the fluctuations of the total multipath gain. The channel coefficients are assumed to be real rather than complex variables, that is phase term $\phi_k$ will be assumed as $\pm \pi$ with equal probability for representing pulse inversion due to reflection from dielectric surfaces. Consequently, the CIR of the IEEE model [Mol05a] can be commonly expressed as follows:

$$h(\tau)_{\text{discrete}} = \sum_{l=0}^{L} \sum_{k=0}^{K_l} a_{k,l} \delta(\tau - T_l - \tau_{k,l})$$  \hspace{1cm} (2-17)
where $a_{k,l}$ is the multipath gain coefficient of the $k^{th}$ MPC in the $l^{th}$ cluster. $L$ and $K_l$ represent the number of clusters and the number of MPCs within that $l^{th}$ cluster respectively. $T_i$ is the delay of the $l^{th}$ cluster or the TOA of the first arriving MPC within that $l^{th}$ cluster. $\tau_{k,l,i}$ denotes the $k^{th}$ path arrival delay with respect to the first arriving MPC in the $l^{th}$ cluster. $\delta(\cdot)$ is the Dirac delta function.

Note that, due to the frequency selectivity of scatterers, each MPC will experience distortion, which can be taken into account as follows [Cho04]:

$$h(\tau)_{\text{discrete}} = \sum_{l=0}^{L} \sum_{k=0}^{K_l} a_{k,l} \chi_{k,l}(\tau) * \delta(\tau - T_i - \tau_{k,l})$$

$\chi_{k,l}(\tau)$ denotes the distortion of the $k^{th}$ component in the $l^{th}$ cluster due to the frequency selectivity of scatterers. For instance, reflection from and transmission through dielectric or conductive objectives are other important effects show frequency dependence. The dielectric properties of most of materials show significant variations over the frequency ranges of interest. Reflection by a rough surface shows strong dependence on the considered frequency. Furthermore, diffraction at the edge of a screen or wedge also shows strong frequency dependence: the diffraction loss increases with frequency.

When considering a deterministic representation of a single impulse response, the pulse distortion does not lead to a fundamental change of the description method. As long as the system is band limited, any deterministic impulse response can be represented by a tapped delay line model [Mo05], where the tap spacing is at least as dense as required by the Nyquist criterion. On the other hand, the number of taps required to represent the impulse response can increase due to the pulse distortion.

### 2.2.3 UWB Channel Measurements

There are two channel sounding methods to perform measurement of UWB propagation channels. Firstly, the channel can be measured in the time domain (TD) based on pulse signal transmission or direct sequence spread spectrum signalling. The second method is frequency domain (FD) channel sounding using a frequency sweeping technique. However, this method is not suitable for the non-stationary channels differently from the TD method that the movement of obstructions between the transmitter and receiver ends can be solved. Furthermore, there are several aspects which describe advantages and disadvantages of using these channel measurements as discussed below. To observe the small scale channel characteristics, the real or synthetic array antennas are used in measurement systems. In each receiver location, channel
response measurements can be conducted by using a small-scale square grid or the X-Y scanner to synthesise virtual arrays [Win02], [Han04], [Han05], and [Cho05b]. During each phase of the multipath profile measurement, both the transmitter and receiver are located in a stationary position.

2.2.3.1 Time Domain Measurement Techniques

Firstly, the time domain measurement can be commonly performed by probing the channels periodically with a very short Gaussian-like pulse in nanoseconds. Using the simple pulse shape can make the receiver perform the deconvolution easily in the post-processing stage. The receiver records the received responses using a digital sampling oscilloscope (DSO). The bandwidth of the channel sounder depends on the pulse shape and the pulse duration of the single pulse, inversely proportional to the bandwidth of the transmission. This can determine the multipath resolution, or in other words, the minimum discernible path between individual multipath contributions. Hence, by changing the transmitted pulse width, the spectral allocation can also be changed. Furthermore, this impulse-based-measurement system requires an additional probe antenna in proximity to the transmit antenna. This antenna is used for triggering purposes, so that all received multipath profiles have the same reference delays; time delay measurements of incoming signals propagating through different paths can be determined. The trigger signal is transmitted to the DSO by a long fixed length coaxial cable. Since this cable is used for transferring triggering pulses only, a high quality expensive cable is not necessary. According to Figure 2-5, the TD based impulse channel sounding consists of the periodic pulse generator, transmitting and receiving antennas, a trigger probing antenna, a wideband low noise amplifier (LNA), and a DSO [Win97], [Cas01], [Qia03], and [Muq03].

In addition, another TD sounding system relies on direct sequence spread spectrum technique (DS-SS) and a correlation receiver [Opp04]. In theory, periodic pulse signals are generated using the maximum length code (m-sequence), and its autocorrelation is computed at the receiver site. The optimal pseudorandom code is the popular m-sequence which is used to spread the transmitted signal energy over the wide frequency band. The bandwidth of the transmitted signals is twice the bit rate of the m-sequence, or the chip rate. Therefore, the delay resolution is inversely proportional to the chip rate. Figure 2-6 presents the diagram of a UWB radio channel sounding system using DS-SS technique where \(c(t)\), \(W(t)\), \(h(t)\) and \(R(t)\) are the pseudorandom code, the transmitted signal, the CIR and the received signal respectively. The correlator is sampled equally to the chip rate. Each sample value exhibits the strength of the corresponding propagation channel at the certain delay. Nevertheless, the technique of DS-SS sounding in UWB systems is limited by the utilisation of the chip rate corresponding to measured channel bandwidths, i.e. the minimum bandwidth of 500MHz can be achieved by transmitting
pseudorandom codes with the chip rate of 250MHz, or if the bandwidth is larger than 1GHz, the chip rate is required to exceed 500MHz. These high chip rate signals are complicated and cost many resources to be generated.

### 2.2.3.2 Frequency Domain Measurement Techniques

Due to the drawback of TD channel measurement, which is inevitably affected by operating hardware such as receiver noise, jitter, and synchronisation, the frequency domain measurement is considered as a good candidate to measure UWB channel characteristics in short-range wireless scenarios. The vector network analyser (VNA) is used as the frequency domain channel sounder. Furthermore, FD measurements can gain more advantages of large dynamic range, i.e. more than 130 dB for Rohde & Schwarz VNA, and obtained phase information. S-parameter, $S_{21}$, which essentially corresponds to the channel transfer function (CTF) is measured and stored. In order to get reliable channel models, the sweeping time of VNA should not exceed the channel coherence time, otherwise, the channel may change during the sweep.
Figure 2-7 shows the configuration of FD measurement apparatus for acquiring the complex frequency responses of the UWB channel via the $S_{21}$ parameter. The example of synthesised antenna arrays at the transmitter and the receiver is also shown. The measurement system consists of a VNA, a wideband power amplifier, and a LNA. The virtual array antennas are controlled by the X-Y positioner via the general purpose interface bus (GPIB) interface. During the measurements, the VNA was set to transmit continuous wave tones uniformly distributed over the interest bandwidths. The frequency resolution gives the maximum excess delay and also maximum distance range to be able to measure. The upper bound for the detectable delay, $\tau_{\text{max}}$, can be defined by (2-19) where $N_{\text{pnt}}$ is the frequency resolution or numbers of frequency points used per sweep, i.e. 801, 1601 points, and $B$ is the frequency span to be swept [Opp04].

$$\tau_{\text{max}} = \left( N_{\text{pnt}} - 1 \right) / B$$  \hspace{1cm} (2-19)

The effects of the measurement system are removed in order to ensure that only the propagation channel transfer function will be used for further analysis. The hardware effects should be removed from the channel measurements by calibrating the measurement equipment in an anechoic chamber with respect to a reference distance, in general, 1 m. The same cables, adapters, and other components are still used during the calibration process except amplifiers which are isolated in the reverse direction when being calibrated. Thus they will be measured separately. All measurement signals are calibrated with these measurements calibrated for data...
UWB Propagation Channel and Related Works

In addition, cabling effects can be also considered negligible due to large ground planes on the antenna [Kot01a]. Since the antenna is driven for measuring signals at each grid position, the bending cable attached to the receiver terminal can induce strong cross polarise fields [Kot01b]. To eliminate the disturbance of conducting cables, the antenna element employed for the receiver should be designed as small volume as possible, i.e., monopole-like antenna, for the required bandwidth and should be attached on the large ground plane. As a consequence, cabling effect can be assumed negligible due to very small wavelength comparing to the receiver size [Bro07].

For the post-processing, since the measurement is operated with no movement being allowed during capturing signals, the channel is assumed to be static. At each grid point, all snapshots of CTFs are averaged over the time domain, and then, these time-averaged CTFs are processed by frequency domain windowing to reduce the leakage, but this can cause the degradation of time-domain resolution. Generally, the Hamming window is selected due to its very high side lobe suppression (about 41 dB) [Opp99]. Next, the windowed time-averaged CTFs are transformed into the CIRs by the inverse Fourier transform (IFT). The IFT can be usually taken directly from the measured raw data if the measurement system is the classical carrier-based system. This technique is called as the complex baseband technique. Its spectrum can be obtained by a down-conversion stage with a simple mixer device as illustrated in Figure 2-8(a). In contrast, the real passband IFT processing technique, which is described in Figure 2-8(b) and (c), is more suitable for carrierless pulse-based system. Moreover, this real passband technique can gain the better temporal resolution than the complex baseband technique due to the zero-padding operation. This technique can support the improvement in accuracy of separating UWB MPCs.

There are two real passband techniques commonly used for converting signal to the time domain, which both can gain approximately the same results. The first one is the Hermitian passband which involves first zero-padding of the CTF from the lowest frequency down to the direct current (DC), taking the complex conjugate of the signals, and then flipping the conjugate positive spectrum to the negative frequencies. Thus the complete symmetric spectrum can be obtained as illustrated in Figure 2-8(b). The doubled-sided spectrum corresponds to a real signal, furthermore, the time resolution is as much as twice that achieved by the complex baseband technique. Another technique is the conjugate approach which is less complex than the previous one. This technique can be operated by reflecting the conjugate zero-padded passband signals just only the left side of the spectrum. As a consequence, the signals are converted by IFT with the same window size as the Hermitian technique. This approach is described in Figure 2-8 (c) [Opp04], [Cho05b].

Consequently, all captured CIRs in TD measurements or converted CIRs in FD measurements, \( h(\tau) \), are used to determine PDPs by \( P(\tau) = |h(\tau)|^2 \). In order to determine the small
scale characteristics, a sufficient number of measurement points should be taken into account in a location where large scale parameters are constant. Thus all PDPs are averaged over small-scale grid points to give a small-scale averaged PDP (SSA-PDP) particularly for each location.

Figure 2-8: IFT signal analysis techniques (a) Complex baseband (b) Hermitian passband (c) Conjugate approach

2.2.4 High-Resolution Algorithms for MPC Determination

Since UWB signals are capable of providing fine time resolution, in order to determine multipath parameters accurately, exploiting the high-resolution algorithm used for extracting parameters is another open issue to be investigated. Results from high-resolution algorithms should gain better accuracy than the ones analysed from a common Fourier-based computation; i.e. delay resolution extracted from an algorithm is more accurate than the result calculated from the inverse bandwidth. There are varieties of high-resolution algorithms being used in UWB systems described as follows.

2.2.4.1 Sensor CLEAN Algorithm

The CLEAN algorithm is originally used in astronomy to enhance radio astronomical maps of sky [Hög74]. This technique is based on a serial-interference cancellation algorithm [Ami02],
[Mol05] by subtracting the similarity between measurement data or the dirty map and the a priori information or the template, and reconstructing the estimated data or the clean map based on those detected similarities. This technique has been widely used in both narrowband and UWB communications. Cramer [Cra02] proposed the modification of CLEAN algorithm for extracting MPCs based on array UWB measurements with the requirement of a minimum priori information, in which beamformer’s responses are constructed as a function of beam directions, elevation and azimuth AOAs ($\theta , \phi$), and arrival times ($\tau$) by delay-and-sum beamforming. Thus these digitised pulse response functions are known as the template, $B(\theta , \phi , \tau)$. Firstly, the dirty map is generated by correlating between the received signals $R(\tau)$ with the template $B(\theta , \phi , \tau)$. Then the highest peak of correlation is selected and identified its amplitude and delay. The contribution regarding to that identified correlation peak is subtracted from the correlated set (dirty map). Next, the identified amplitude and delay are used to generate the clean data and stored in the clean map. This process is repeated until the energy of the highest correlation peak falls below the threshold. Further details of the sensor CLEAN algorithm are explained below.

1) Generating sets of beamforming responses corresponding to beam directions and delay times, $B(\theta , \phi , \tau)$.

2) Initialise the clean map, $C(\theta , \phi , \tau) = 0$, and the dirty map, $D(\theta , \phi , \tau) = R_{\text{BB}}$ which is the cross-correlation between the beamformer response function and the received signals.

3) Shift each signal in the correlation matrix in the dirty map corresponding to delay, $\tau$, associated by AOAs, $\theta$ and $\phi$.

4) Compute the auto-correlation of the correlation matrix, $R_d$, and determine the joint maximum correlation peak. The amplitude, TOAs and AOAs associated with the correlation peak are identified and stored as $a_{\text{est}}, \tau_{\text{est}}$ and $\theta_{\text{est}}$ respectively.

5) If the value of amplitude $a_{\text{est}}$ is higher than the threshold level, go to the next step. Otherwise, stop the process.

6) Clean the dirty map by subtracting the peak from the correlation matrix: $D(\theta , \phi , \tau)_{\text{new}} = D(\theta , \phi , \tau) - a_{\text{est}} R_{\text{BB}}(\tau - \tau_{\text{est}})$.

7) Update the clean map: $C(\theta , \phi , \tau)_{\text{new}} = C(\theta , \phi , \tau) + a_{\text{est}} \delta (\theta_{\text{est}}, \phi_{\text{est}}, \tau - \tau_{\text{est}})$, and then repeat the process from step 4 until the iteration matches the criteria. Then stop the process.

8) Finally, evaluate the estimated impulse response from the clean map: $h(\theta , \phi , \tau) = C(\theta , \phi , \tau)$. 

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2.2.4.2 SAGE Algorithm

There are several works reporting about employing the Space Alternating Generalised Expectation-maximisation algorithm (SAGE) to investigate the UWB channel parameters, in AOs, TOAs, and spectrum of each array path [Han03], [Han04a], [Han04b], [Saa07]. This algorithm analyses the spatially sampled transfer functions in which the frequency dependent complex amplitude is an estimation of interest. Both sensor CLEAN and SAGE algorithms adopt the successive interference cancellation (SIC) for removing the estimated components from the received sequence after determining wave parameters [Ami02], [Mol05]. Thus, contributions from estimated MFCs are subtracted from the considered signal. However, the erroneous difference between wrong subtracted contributions and received MPC sequences can lead to the ghost signal. The SAGE algorithm is based on the frequency sub-band processing to estimate a frequency dependent spectrum for each path with the spherical wave incident model. According to a UWB propagation path based ray model, DOAs and TOAs contain a frequency dependent complex gain with the transfer function of one single ray which can be described by the relation

\[ y_n = x_n(f, \tau_n) r_n(f) D(f, \theta_n, \phi_n) \]  

where \( x_n \) is the complex gain of spatial propagation; \( r_n(f) \) is the scattering loss, and \( D(f, \theta_n, \phi_n) \) is the radiation pattern of the received antenna for a single polarisation. As a consequence, the measured transfer function of the frequency sub-band \( s \) observed at the array position can be shown by

\[ y_{k_1,k_2,k_3} = \sum_{n=1}^{N} \left[ \alpha_n(f_s, \tau_n) r_n(f_s) D(f_s, \theta_n, \phi_n) \prod_{i=1}^{3} \exp(jk_i \beta_n^{(i)}) \right] + w_{k_1,k_2,k_3} \tag{2-20} \]

\( k_i \) shows the index number of the interested domains, where \( 0 \leq k_i < K_i - 1 \) with \( 1 \leq i \leq 3 \), i.e. \( k_1 \) and \( k_2 \) represent position index of array antennas. Since the SAGE algorithm is based on frequency dependent characterisation, the whole bandwidth is divided into several frequency sub-bands. Accordingly, \( k_3 \) represents the index of frequency sweeping within the \( s^{th} \) sub-band where \( f_s \) is the centre frequency of the considered frequency sub-band. \( n \) is the number of incoming multipaths \((n=1...N)\), and \( w_{k_1,k_2,k_3} \) is the zero-mean-additive white Gaussian noise. According to the configuration of impinging multipaths on the array antenna which will be explained in the following, \( \beta_n^{(i)} \) can be denoted by

\[ \beta_n^{(1)} = \frac{2 \pi f_s}{c} dx \sin \theta_n \cos \phi_n \]
\[ \beta_n^{(2)} = \frac{2 \pi f_s}{c} dy \sin \theta_n \sin \phi_n \]
\[ \beta_n^{(3)} = 2 \pi df \tau_n \]

and \( dx \) and \( dy \) are gap distances between elements in \( x \) and \( y \)-directions. \( df \) is small frequency range in the \( s^{th} \) sub-band.

To simplify the notation, measured transfer function can be vectorised as \( y_s = [y_{1,1,s}, y_{1,2,s}, y_{1,3,s}, y_{2,1,s}, y_{2,2,s}, y_{2,3,s}, y_{3,1,s}, y_{3,2,s}, y_{3,3,s}]^T = A_s H_s + w_s \). \( A_s \) is the multi-dimensional phase
matrix and can be expressed by the Kronecker product, $\otimes$, of each row of the matrix so they are $A_s = A_s(\beta_1) \otimes A_s(\beta_2) \otimes A_s(\beta_3)$ where $A_s(\beta_s) = [a_s \beta_s^0_{1,s} \quad a_s \beta_s^0_{2,s} \quad \cdots \quad a_s \beta_s^0_{N_s}]^T$ where this expands again to $a_s \beta_s^0_{n,s} = [1 \quad \exp(j \beta_s^0_{n,s}) \quad \exp(j 2 \beta_s^0_{n,s}) \quad \cdots \quad \exp(j (K-1) \beta_s^0_{n,s})]^T$. $\mathbf{H}_s$ represents a reference transfer function measured at the array centre; this vector contains characteristics of channel propagation and antennas at each incoming wave, $\mathbf{H}_s = [H_{1,s} \quad H_{2,s} \quad \cdots \quad H_{N_s}]$, where $H_{n,s} = a_n(\tau_n, \phi_n) R_n(f) D(f, \theta_n, \phi_n)$. $\mathbf{w}_s$ is the vector of AWGN.

Maximum likelihood estimation (MLE) is taken into account in the SAGE algorithm to estimate the contribution that maximises the likelihood function. The measured UWB data is assumed to follow a certain Gaussian distribution due to noise perturbation. Accordingly, with the assumption that the noise vector, $\mathbf{w}_s$, is independently and identically distributed, the probability of the generating measured data vector $\mathbf{y}_s$ from the signal component vector $\beta_s$ can be given by [Han03], [Han04a].

$$p(\mathbf{y}_s | \beta_s) = \prod_{k_1=1}^{K_1} \prod_{k_2=1}^{K_2} \prod_{k_3=1}^{K_3} \left[ \frac{1}{\pi \sigma^2} \exp \left( - \frac{\|y_{k_1,k_2,k_3,s} - \beta_{k_1,k_2,k_3,s}\|^2}{\sigma^2} \right) \right] \quad (2-21)$$

Hence $K = K_1 K_2 K_3$ denotes the total number of measured data, and $S$ is the total number of frequency sub-bands. When taking the logarithm of the above equation, the log-likelihood function can be simplified in (2-22). Therefore, the maximum likelihood estimation of parameters which maximises the likelihood function in (2-21) is given by (2-23) [Han06].

$$\log[p(\mathbf{y}_s | \beta_s)] = -K \log \pi \sigma^2 - \frac{\|y_s - \mathbf{A}_s \mathbf{H}_s\|^2}{\sigma^2} \quad (2-22)$$

$$\beta' = \arg \max_\beta \log p(\mathbf{y}_s | \beta) = \frac{1}{\sigma^2} \arg \min_{\theta_n, \phi_n, \mathbf{H}_s} \|y_s - \mathbf{A}_s \mathbf{H}_s\|^2 \quad (2-23)$$

Nonetheless, since estimating $\beta'$ in (2-23) requires large simultaneous search dimensions, the expectation and maximisation (EM) algorithm [Dem77] is employed to reduce this complexity. The EM algorithm estimates the transfer function of complete data $\mathbf{x}_{n,s}$ from the incomplete data $\mathbf{y}_s$ as given by $\mathbf{x}_{n,s} = a_{n,s} H_{n,s} + \chi_{n,s}(\mathbf{y}_s - \mathbf{A}_s \mathbf{H}_s)$. Complete data means the data of each
incident path which is not directly detected in the measurement, and the incomplete data corresponds to the measured data. \( y_n-A_nH_n \) associates with the \( w_n \) term, thus the term \( x_{n,d} \) represents as a part of the measured noise component. \( x_{n,d} \) is a positive number which has a constraint \( \sum_{n=1}^{N} x_{n,d} = 1 \). As a consequence, the modified log-likelihood function for complete data \( x_n \) is analogous to (2-23).

\[
\beta_n' = \arg \max_{\beta} \log p(x_{n,s} | \beta) = \arg \min_{\beta_n} \| x_{n,s} - A_nH_n \|^2 \tag{2-24}
\]

where \( \beta_n \) is a set of model parameters for the \( n^{th} \) multipath: \( \beta_n = [\theta_n, \phi_n, \tau_n, H_n] \). The simultaneous search dimension becomes reduced due to dividing a whole data into each path. Only the parameters of multipaths containing the largest power of the complete data are estimated. If only 3-dimensional simultaneous search is taken into account, the procedure can be minimised as

\[
(\theta_n', \phi_n', \tau_n') = \arg \max_{(\theta, \phi, \tau)} | z(\theta, \phi, \tau, x_{n,s}) | \tag{2-25}
\]

\( z(\theta_n', \phi_n', \tau_n') \) can be denoted by \( z(\theta_n', \phi_n', \tau_n', x_{n,s}) = E[A_nH_n x_{n,d}] \), and the complex transfer function of the individual wave can be obtained from the estimated parameters by \( H_n = z(\theta_n', \phi_n', \tau_n', x_{n,s}) / K \) with the path gain \( P_{n,d} = |H_n|^2 \).

The SAGE algorithm considers subspaces of parameters of interest. To estimate multipath parameters, the whole search space of the EM algorithm is divided and processed sequentially in the following order where the estimated parameters are updated and used for the next process. The iteration of the algorithm is processed until the likelihood achieves the certain maximum value or some converging fixed values.

\[
\theta_n' = \arg \max_{\theta} | z(\theta, \phi_n, \tau_n', x_{n,s}) | \\
\phi_n' = \arg \max_{\phi} | z(\theta_n', \phi, \tau_n', x_{n,s}) | \\
\tau_n' = \arg \max_{\tau} | z(\theta_n', \phi_n', \tau, x_{n,s}) | \tag{2-26}
\]
2.2.4.3 MUSIC Algorithm

The high resolution multiple signal classification (MUSIC) algorithm is used in estimating both the accurate signal arrival times and angles associated with spectral-based methods [Wan04]. MUSIC employs the specific structure and properties of the array correlation matrix. The eigen-decomposition of this matrix can be derived from [Sch86] based on the frequency subspace technique. According to time domain CIR in (2-11), the frequency domain channel response and received signal can be given by (2-27) and (2-28) respectively.

\[
H(f) = \sum_{n=1}^{N} a_n e^{-j2\pi f_n} \tag{2-27}
\]

\[
x_n = H(f_n) + w_n = \sum_{n=1}^{N} a_n e^{-j2\pi (f + n\Delta f) \tau_n} + w_n \tag{2-28}
\]

where \(w_n\) is the additive white Gaussian noise (AWGN) with mean zero and variance. The signal model can be rewritten into the vector form as

\[
x = H + w = V \alpha + w \tag{2-29}
\]

where \(x = [x(1) \ x(2) \ ... \ x(N)]^T\), \(H = [H(f_1) \ H(f_2) \ ... \ H(f_N)]^T\), \(w = [w(1) \ w(2) \ ... \ w(N)]^T\), \(V = [v(\tau_1) \ v(\tau_2) \ ... \ v(\tau_N)]\) is the \(N \times M\) source direction matrix, which contains individual vectors \(v(\tau_m) = [\exp(-j2\pi \Delta f \tau_m) \ \exp(-j2\pi 2\Delta f \tau_m) \ ... \ \exp(-j2\pi N\Delta f \tau_m)]^T\), \(\alpha = [\alpha'_1 \ \alpha'_{2} \ ... \ \alpha'_{M}]^T\) is the \(M \times 1\) vector of the source waveform, \(\alpha'_m = a'_m \exp(-j2\pi \Delta f \tau_m)\). Consequently, autocorrelation matrix of the received signal modelled in (2-29) is:

\[
R_{xx} = E[xx^H] = VAV^H + \sigma_w^2 I \tag{2-30}
\]

\(A=E[\alpha \alpha^H]\) is the diagonal matrix containing the corresponding eigen values. It is symmetric and positive definite with rank \(N\). \(V\) contains the signal subspace eigenvectors of \(R\), and the superscript \(H\) represents conjugate transpose operation of a matrix. \(\sigma_w^2 I\) is the autocorrelation matrix of noise. Let the number of samples \(M\) be larger than the number of multipaths \(N\), hence
the rank of $\mathbf{VAV}^H$ becomes $N$. When using the eigenvectors to form the signal and noise subspaces, correlation matrix in (2-30) can be rewritten by

$$
\mathbf{R}_{xx} e_i = \mathbf{VAV}^H e_i + \sigma_w^2 e_i = (\mu_i + \sigma_w^2) e_i = \lambda_i e_i
$$

(2-31)

e_i is the eigenvector associated with eigenvalue $\lambda_i$, and $\lambda_1 \geq \lambda_2 \geq \ldots \lambda_M = \sigma_w^2$. Let the signal subspace can be given by $\mathbf{U}_s=[e_1, e_2, \ldots, e_N]$, and the noise subspace can be exhibited by $\mathbf{U}_n=[e_{N+1}, e_{N+2}, \ldots, e_M]$. When the noise subspace and the column vector of the signal subspace is orthogonal to each other, the eigenvalue is equal to autocorrelation of noise, $\lambda = \sigma_w^2$. Hence (2-31) can be given by $(\mathbf{R}_{xx} - \sigma_w^2) e_i = \mathbf{VAV}^H e_i = 0$, where $i=N+1, \ldots, M$. As a consequence, this property enables to find the multipath parameters of time delays from the locations of the highest peaks of the frequency domain MUSIC pseudospectrum [Sch86], [Böl06].

$$
\hat{f}_{\text{MUSIC}}(\tau) = \frac{1}{\mathbf{v}^H(\tau) \mathbf{U}_w \mathbf{U}_w^H \mathbf{v}(\tau)}
$$

(2-32)

When a linear array structure is taken into account, to determine the AOAs, the matrix $\mathbf{V}$ can be defined as $\mathbf{V} = [\mathbf{v}(\theta_1) \mathbf{v}(\theta_2) \ldots \mathbf{v}(\theta_N)]$ with the array steering vector $\mathbf{v}(\theta_i) = [\exp(-j2\pi f \sin \theta_i) \exp(-j2\pi f(2) \sin \theta_i) \ldots \exp(-j2\pi f(M) \sin \theta_i)]^T$. Thus, AOA parameters can be estimated at the highest peak locations of MUSIC spectrum function.

$$
\hat{f}_{\text{MUSIC}}(\theta) = \frac{1}{\mathbf{v}^H(\theta) \mathbf{U}_w \mathbf{U}_w^H \mathbf{v}(\theta)}
$$

(2-33)

According to these high resolution algorithms for UWB parameter estimation, to achieve the accuracy of parameter estimation, these techniques are operated with some specific constraints. The Sensor-CLEAN algorithm serially proceeds and identifies incident waveforms, hence time domain sequential data is required as the input to be analysed. Although there is a close resemblance between reconstructed waveforms and received signals, some incorrect estimation still exists due to incomplete removal of ghost signals. Furthermore, its performance can be, in general, evaluated by the mean square error, the correlation coefficient, the energy-capture ratio, and the relative error [Liu07]. The SAGE algorithm also updates the estimated
parameters in a serial manner leading to more practical and faster convergence than its originated expectation-maximisation technique, which proceeds signal parameters in parallel [Xu08]. It replaces the high-dimensional optimisation process by performing separation of several one-dimensional maximisation procedures sequentially. This algorithm can reduce the computational cost and is an advantage when comparing to the MUSIC algorithm, which is complicated for system implementation and deteriorates rapidly when there is significant correlation among signals received from different AOAs [Pic03].

In addition, in SAGE, the detection and cancellation of paths are conducted in an accurate manner notwithstanding being deteriorated by the error propagation, which is the main drawback of the SIC-type path detection [Liu07]. This is because the problem can be alleviated by the fine time resolution offered by UWB signals. On the other hand, one of the disadvantages of the SAGE technique is that it is considerably unstable when estimating incoming signals with low signal-to-noise ratio (SNR) and small number of snapshots [Pei01]. SAGE is sensitive to an initial guess that is convergence to the required global maximum can be obtained when the initial estimate is within the significant range of the global maximum. Nevertheless, the process tends to converge to the suboptimum solution, or the local maxima, if careless initialisation is taken [Xu08].

2.3 UWB Array Systems

Because of the limitation of strict power regulations, the usage of highly efficient UWB antenna arrays is considered as the solution to be implemented in order to challenge power restriction and allow wireless systems to occupy UWB frequency bandwidth. Furthermore, UWB array features such as beamforming, spatial diversity and direction estimation are also exploited in present research activities.

2.3.1 Modelling UWB Antenna Arrays

In narrowband applications, electrical properties of antenna arrays, i.e. input impedance, efficiency, gain, effective area, radiation pattern, and polarisation properties are investigated at the centre frequency of radio systems [Rom03]. However, for UWB bandwidth, these properties become highly frequency dependent, though evaluating these features in frequency functions lacks information for characterising transient radiation behaviours. In fact, only frequency domain theory is not sufficient to characterise the influence of UWB antenna arrays. Thus, the transient radiation behaviours of UWB antenna arrays in time domain in terms of array transient response, dispersion, and ringing are commonly analysed.
It is assumed that a single source signal travels into identical array elements which are identically oriented. Hence, the time domain transfer function of the antenna arrays, \( h_{m}(\tau) \), can be obtained from a convolution of the single element transient response or impulse response, \( h(\tau) \), and the time domain array factor, \( AF(\tau) \): \( h_{m}(\tau) = h(\tau) \ast AF(\tau) \). Thus realistic results for dispersive behaviour of the whole array can be obtained. If the inter-element space between arrays is large enough (more than \( \lambda/2 \)), there is not any effect of mutual coupling due to reflections and re-radiations. Therefore, dispersion and distortion can be neglected. Furthermore, the degree of mutual coupling effect also depends on type of each array; mutual coupling effects can be observed in an omnidirectional array antenna much more than a directional array antenna [Kot05]. Assuming mutual coupling and antenna mismatch being neglected throughout UWB bandwidth, the array factor can be defined by summation of the path transfer functions, \( A_{m}(\tau) \) for each \( m^{th} \) sensor. This weighting function contains the transfer function of the active or passive transmission in array systems; for passive feeding networks \( \sum_{m} |A_{m}(\tau)|^2 \leq 1 \). Denoting that there are \( M \) array elements of interest, hence

\[
AF(\tau) = \sum_{m=1}^{M} A_{m}(\tau) \ast \delta(\tau - \tau_{m}) \tag{2-34}
\]

The time delay \( \tau_{m} \) is the time difference between TOA at the \( m^{th} \) element relatively to the one at the reference point which is located at the centre of the array structure. If considering an ideal feeding network, multiple reflections due to impedance mismatch are neglected, \( A_{m}(\tau) \) can be given by \( A_{m}(\tau - \tau_{\text{feed},m}) \) with a uniform amplitude distribution of \( a_{m} = 1/\sqrt{M} \) where \( \tau_{\text{feed},m} \) is the time delay due to feeding networks [Mic08]. This term is used to decrease the ringing signal travelling from boresight directions. As a consequence, the time domain array factor can be simplified by [Ben06]

\[
AF(\tau) = \sum_{m=1}^{M} a_{m} \delta(\tau - \tau_{m} - \tau_{\text{feed},m}) \tag{2-35}
\]

2.3.2 UWB Array Characteristics

UWB multipath signals propagate spatially through various obstructions along propagation paths, and their impinging contributions are sensed by array sensors at the receiver. Typically, UWB multipaths propagate towards an array receiver in a clustered pattern. The
transmitted signals from some main directions incident to received arrays. In some cluster beamforming where the main lobe width fits the incoming-cluster width, enough energies can be detected, whereas undesired clusters are suppressed with the array gain. As a consequence, the signal-to-interference-plus-noise ratio (SINR) can be increased with shortened channel delay spread. Higher data rates can also be achieved [Cra99], [Spe00]. The feature of the array processing is concerned with the extraction of information of channel parameters such as TOAs and AOAs apart from amplitudes of received signals. In this section, all aspects of characteristics of UWB arrays, array signal processing, and parameter estimation are described. Indoor UWB communication systems are characterised by dense multipath environments. The large number of multipath arrivals seems to limit the usefulness of conventional beamforming algorithms. Other multipaths beyond the main lobe direction can be considered as interfering. Corresponding to fine time resolution of UWB impulse signals which is shorter than the average travel time across the array, several significant UWB beamforming characteristics are briefly summarised below [Ars06], [Rie06], [Mal06];

- Similar to conventional narrowband and broadband beamforming, the width of UWB main lobe depends on the ratio of centre wavelength over the array size. Furthermore, the narrow beamwidth of the UWB main lobe can be simply constructed by increasing the spacing gap between each element: the longer the distance between elements, the narrower the beam pattern. However, this is contrary to the narrowband and broadband beamforming, that more antennas are required to sharpen the main lobe without undesired ambiguities caused by grating lobes. The width of the beam pattern decreases with the increasing of the number of array elements [Hus02].

- For conventional bandwidth, increasing the space between elements more than half of the centre frequency wavelength can increase the directivity but induce the grating lobe effect with maximum amplitude equal to the main lobe magnitude. However in UWB arrays, since the grating lobes occur at various angles for different frequencies unlike the main lobe which is located at the same angle for any frequency. Thus there is rarely existence of grating lobes for UWB signals due to averaging out over UWB frequency. It can be implied that there is beneficial suppression of grating lobes in UWB beamforming. As a consequence, the inter-element spacing between array elements is not longer limited by half a wavelength; when employing only a few arrays, the gap distance between elements can be increased to achieve high resolution without any effect from grating lobes.

- Moreover, UWB sidelobe levels are fixed for a wide range of bearings. Since the travel time across the array might become larger than UWB pulse width when propagating outside the main lobe direction, an overlap at the beamformer output does not occur. Hussain reported that there is not any significant difference of UWB sidelobe amplitude level when increasing the inter-
element distance, \( d \), in an order of half wavelength. Additionally, power and energy pattern of UWB can be recognised as the sidelobe-free directivity pattern \([\text{Hus02}]\). Accordingly, some research commented about non-necessity of occupying the unequal amplitude weighting for individual antennas to increase main-to-sidelobe ratios as commonly operated in narrowband and broadband cases \([\text{Hus02}], [\text{Rie06}]\). Cramer also stated that the incoming signals do not interfere in the sidelobe region, thus the best peak-to-sidelobe ratio can be achieved by using only the uniform weighting \( a_m = a_n \) for all elements \( m \) and \( n \).[\text{Cra99}]. Nonetheless, the adjustable amplitude weighting can still be used to achieve electronic beam steering for the purpose of optimising signal reception in the specific look directions of angles of arrival \([\text{Cra99}], [\text{Tan03}], [\text{Hon04}], [\text{Thi05}], [\text{Has08}], \) and \([\text{Man08}]\).

- A double-dB gain can be obtained for UWB particularly relevant to directions of incoming signals. For example, when UWB multipath wavefronts are arriving from the broadside of two antennas with large spacing, incident signals are added constructively by the beamformer; leading to double magnitude output signals, or four times the peak power. On the other hand, arriving multipaths do not overlap at the beamformer output; the signal power equals the single signal only. This latter case can be observed when UWB signals travelling along two arrays from the endfire direction.

- Nonetheless, the drawback of UWB properties can be addressed for interference rejection. Unlike side lobe level suppression, UWB interference rejection requires filters subject to delays of arriving interference signals in each antenna branch. Undesired signals from selected directions are nulled over wide frequency range. However, interference rejection becomes more complicated to become feasible in dense UWB multipath environments; a large number of filters on behalf of dense multipath delays are required. Even though the main lobe perfectly steers towards the DOA of desired signals, all other numerous paths coming from other directions are considered as interfering. Accordingly, this leads to limitation of conventional beamforming algorithms.

Basically, UWB array signal model can be simply examined by considering a uniform linear array (ULA) antenna with equal inter-element spacing, \( d \), between all array elements. Since all elements are equally spaced, a difference in propagation paths between any two successive arrays can be given by \( d \sin \phi \). Figure 2-9 depicts the impinging of the incoming signal on ULA. The array response vector of the \( M \)-element-ULA can be written as a vector containing the centre frequency and individual time delays to the \( m \)\textsuperscript{th} element with respect to the first element:
\[
v(\phi) = \begin{bmatrix}
1 & \exp(-j2\pi c \tau_1(\phi)) & \exp(-j2\pi c \tau_2(\phi)) & \cdots & \exp(-j2\pi c \tau_{M-1}(\phi))
\end{bmatrix}^T
\]

or with respect to the centre element \((M+1)/2\):

\[
v(\phi) = \begin{bmatrix}
\exp(-j2\pi(M+1)/(d \sin \phi / \lambda)) & \exp(-j2\pi(M+1)/2-1)/(d \sin \phi / \lambda)) & \cdots & 1 & \cdots & \exp(-j2\pi(M+1)/2-(M-1))/(d \sin \phi / \lambda))
\end{bmatrix}^T
\]

(2-36b)

Figure 2-9: Plane wave incidence for linear antenna arrays

Figure 2-10: Plane wave incidence for planar array

In addition to simple geometry of ULA, in order to resolve both elevation angle and azimuth angle in one hemisphere for realistic configuration of incoming signals, another
dimension is added into the array configuration. This leads to the \( I \times J \) planar structure as shown in Figure 2-10. \( M \) is the number of sensors per row of the array, and \( N \) is the number of sensors per column of the array. In this case, the incremental delay in the \( x \)-direction is given by simple trigonometry \( (dx \sin \theta \cos \phi)/c \), and for the \( y \)-direction can be given by \( (dy \sin \delta \sin \phi)/c \) where \( dx \) and \( dy \) are the sensor spacing in the \( x \) and \( y \) directions respectively with an elevation angle of \( \theta \) and an azimuth angle of \( \phi \). When considering the elevation angle of incidence, \( \theta_0 \), and the azimuth angle of incidence represented by \( \phi_0 \) corresponding to the reference centre element of \( ((I+1)/2, (J+1)/2) \) for the \( I \times J \) planar arrays, the array factor at a single frequency can be derived by (2-37) \[ \text{[Con05], [Ali07]. While } \theta \text{ represents the beamformer elevation look direction, and } \phi \text{ is denoted as the beamformer azimuth look direction. For the general case of a planar array, the array factor at a single frequency is given as} \]

\[
AF(\theta, \phi) = \frac{\sin \left( \frac{I+1}{2} \psi_x \right) \sin \left( \frac{J+1}{2} \psi_y \right)}{\sin \left( \frac{1}{2} \psi_x \right) \sin \left( \frac{1}{2} \psi_y \right)}
\]

(2-37)

where

\[
\psi_x = \frac{2\pi}{\lambda} dx (\sin \theta \cos \phi - \sin \theta_0 \cos \phi_0)
\]

(2-38)

\[
\psi_y = \frac{2\pi}{\lambda} dy (\sin \theta \sin \phi - \sin \theta_0 \sin \phi_0)
\]

2.3.3 UWB Delay-and-Sum Beamforming

For a TD description of UWB ULA and planar array systems, the impulse response with uniform spacing in the \( x \) and \( y \)-directions is derived from the delay-and-sum operations as given in (2-39)-(2-40) respectively. These TD approaches are used to characterise UWB beamforming as mainly described in this section.

\[
h_{\text{ULA}}(\tau, \theta, \phi) = \sum_{m=0}^{M-1} \delta(\tau - m\zeta)
\]

(2-39)

\[
h_{\text{planar}}(\tau, \theta, \phi) = \sum_{j=0}^{I-1} \sum_{i=0}^{J-1} \delta(\tau - i\alpha - j\beta)
\]

(2-40)

where \( \zeta = \frac{d}{c} (\sin \theta \cos \phi - \sin \theta_0 \cos \phi_0), \alpha = \frac{dx}{c} (\sin \theta \cos \phi - \sin \theta_0 \cos \phi_0) \) and \( \beta = \frac{dy}{c} (\sin \theta \sin \phi - \sin \theta_0 \sin \phi_0) \).
Both the frequency domain and the time domain representations make the assumption that UWB multipath contributions are detected from all sensors, in other words no sensor is shadowed. Array signal processing or beamforming relates to the signal manipulation induced on array elements of receiving system. In a conventional narrowband beamformer, signals corresponding to each element are multiplied by complex weights to form the array output. Performance of the narrowband beamformer starts to deteriorate as the signal bandwidth increases. This is because of the changing of the phase and the desired angle provided for each element in different frequency components. According to a delay-and-sum beamformer for ULA in Figure 2-11, the UWB beamformer can be generated by summing of received signals from the various sensors corresponding to their individual relative delays and amplitude weighting. Thus a general equation of the delay-and-sum beamformer output as the most applicable to UWB array processing can be considered as

$$B_{\text{ULA}}(\tau, \theta, \phi) = \sum_{m=1}^{M} \alpha_m r(\tau + T_m(\theta_\theta, \phi_\theta) - \tau_m(\theta, \phi)) \quad (2-41)$$

Figure 2-11: A delay-and-sum beamformer with the steering delay for ULA sensors

where the delay at the mth sensor, \(\tau_m(\theta, \phi)\), is explicitly a function of the beamformer elevation and azimuth look direction at the mth sensor. \(T_m(\theta_\theta, \phi_\phi)\) is the steering delay used to steer each individual mth array into a reference broadside direction \((\phi_\phi, \theta_\theta)\). \(\alpha_m\) refers to the weighting value for received signal at the mth sensor. \(R(\tau)\) is a received impulse signal measured at the center position of arrays. The above equation exhibits the beamforming of ULA, whereas the delay-and-sum beamforming for planar arrays as depicted in Figure 2-12 is represented by \((2-42)\).
All received signals at \((i,j)\) sensors are delayed and weighted individually by \(\tau_{ij}(\theta, \phi)\) and \(a_{ij}\) respectively with steering delay of \(T_{ij}(\theta_0, \phi_0)\) before summing all these signals to generate the array outputs. Alternatively, in order to achieve a time-independent beam pattern, \(B(\theta, \phi)\), the total energy of the beamformer output is taken into consideration instead as described in (2-43). This can reduce the above UWB beam pattern to the conventional narrowband beam pattern if \(R(\tau)\) is one of the narrowband type [Ben06]. Alternatively, \(B(\theta, \phi)\) can also be given by \(\max, \|B(\tau, \theta, \phi)\|\) [Mur97] as shown in (2-43).

\[
B(\theta, \phi) = \left[ \int_{-\infty}^{\infty} |B(\tau, \theta, \phi)|^2 d\tau \right]^{1/2} \tag{2-43}
\]
2.3.4 **Tapped-Delay Line Structure**

For broadband signal processing, a tapped delay line (TDL) is commonly implemented on each branch of the array. In the TDL system, each array element has a frequency-dependent phase response. Therefore, when travelling along the same given propagation distance, systems can be compensated for the fact that lower frequency signal components have less phase shift, whereas higher frequency signal components have a larger phase shift. This configuration operates as an equaliser which can make the array responses remain the same over extremely large frequency bandwidths. Furthermore, gain weighting and phase adjustment over the frequency band can be seen in the TDL system as presented in Figure 2-13 [God04]. This TDL configuration is described for UWB planar array reception. It can be seen from Figure 2-13 that the steering delays $T_{i,j}(\theta_0, \phi_0)$ placed immediately after each array are used to steer the array in a given look direction $(\theta_0, \phi_0)$ with gain weighting $\alpha_{i,k}$ at each $k^{th}$ tap delay. $	au_{i,j}(\theta, \phi)$ denotes the time taken by the plane wave arriving from direction $(\theta, \phi)$. This delay term is measured at the $(i,j)$ element relatively to the reference point or the centre element. Hence, the actual delay difference where all desired signal components at each sensor output are aligned exactly in phase can be determined by (2-44). Consequently, the term of time delay $T_{i,j}(\theta_0, \phi_0)$ added to each sensor is able to completely compensate for the actual delay difference [Zou04].

\[
T_0 = T_{i,j}(\theta_0, \phi_0) - \tau_{i,j}(\theta, \phi) \tag{2-44}
\]

![Figure 2-13: UWB beamformer processor with tapped delay line structure](image)
As a result, the continuous-time signal received by the \((i,j)\) sensor is designated by \(x_{ij}(\tau)\) which can be written as

\[
x_{ij}(\tau) = R(\tau - (T_{ij}(\theta_0, \phi_0) - \tau_{ij}(\theta, \phi))) + \omega_{ij}(\tau)
\]

\[
= R(\tau - T_0) + \omega_{ij}(\tau)
\]

(2-45)

where \(R(\tau)\) is the desired signal induced on the reference element at the centre of the coordinate system in the look direction \((\theta_0, \phi_0)\). Thus the signal induced on the \((i,j)\) element can be given by \(R(\tau - T_0)\), and \(\omega_{ij}(\tau)\) is denoted as the total interference and noise observed at the \((i,j)\) sensor.

In Figure 2-13, a pre-steered linearly constrained adaptive beamformer can steer the array response to the direction of interest by employing the steering delays \(T_{ij}(\theta, \phi)\) which are located at the front end and can be either digital or analogue time delays. In general, these pre-steering delays are analysed based on the knowledge of the DOA of the desired signal and the array steering vectors. The vital function of this pre-steered adaptive beamforming is to completely compensate for the actual delay difference. However, there might be some errors; the real delay difference, \(T_{ij}(\theta, \phi) - \tau_{ij}(\theta, \phi)\) is impossible to be exactly compensated for due to practical imperfections such as steering vector errors. This problem can cause desired signals to be regarded as interference [Zou04].

In ULA when only one dimensional incident plane wave \((\theta=90^\circ)\) is considered, signals \(x_m(\tau)\) depend on the azimuth AOA, \(\phi\), and the inter-element spacing, \(d\). When considering the first array sensor as the reference, a time delay, \(\tau_m(\theta=90^\circ, \phi)\), existing between the received signals at the \(m^{th}\) element referring to the first element can be obtained by (2-46a). In addition, if the reference element in ULA is considered at the centre of linear arrays, thus \(\tau_m(\theta=90^\circ, \phi)\) can be written as shown in (2-46b).

\[
\tau_m(\theta = 90^\circ, \phi) = (m - 1) \frac{d}{c} \sin \phi : \text{the reference at the first array} \quad (2-46a)
\]

\[
\tau_m(\theta = 90^\circ, \phi) = \left(\frac{M+1}{2} - m\right) \frac{d}{c} \sin \phi : \text{the reference at the centre array} \quad (2-46b)
\]

It is assumed that the incoming signal is spatially concentrated around the angle \(\phi\) [Gha04]. In addition, if considering UWB signals captured by planar arrays with the reference element at the middle position, \(((J+1)/2, (J+1)/2)\), regarding (2-40), the time delay at any \((i,j)\) antenna element...
can be given in (2-47) where $\theta$ and $\phi$ is an elevation angle and an azimuth angle respectively, for an incidence of a plane wave with $\theta_0$ and $\phi_0$.

\[
\tau_{i,j}(\theta, \phi) = \left(\frac{l+1}{2} - j\right) \frac{dx}{c} \left(\sin \theta \cos \phi - \sin \theta_0 \cos \phi_0\right) + \\
\left(\frac{l+1}{2} - j\right) \frac{dy}{c} \left(\sin \theta \sin \phi - \sin \theta_0 \sin \phi_0\right)
\] (2-47)

Considering a planar array receiver, each array element in Figure 2-13 is connected to a delay line section of $(K-1)$ delays, with an inter-tap delay spacing of $T$ seconds. $w_{i,j,k}$, denoting the weight on the $k^{th}$ tap of the $(i,j)$ antenna where $1 \leq i \leq I$, $1 \leq j \leq J$ and $1 \leq k \leq K$, is multiplied by the delayed input signal of each element. These adaptive weights are iteratively adjusted to minimise output noise-and-interference power. The $k^{th}$ tap output corresponds to the output signal after $(k-1)$ delays. It means that the first tap output corresponds to the output of pre-steering delays, $T_0(\theta, \phi)$; the second tap output corresponds to the output after one delay and the $K^{th}$ tap output corresponds to the output after $(K-1)$ delays.

Finally, the signals $y(\tau)$ at the output of the beamformer, which are determined by summation of all intermediate signals, can be written in (2-48).

\[
y(\tau) = \sum_{i,j=1}^{I,J} \sum_{k=1}^{K} x_{i,j}(\tau - (k-1)T)w_{i,j,k} 
\] (2-48)

where the weighting vector, $\mathbf{w}$, can be defined by

\[
\mathbf{w}^T = [\mathbf{w}_1^T \mathbf{w}_2^T \mathbf{w}_3^T \ldots \mathbf{w}_K^T]_{I\times K}
\] (2-49)

Giving that $IJ$ is the total number of $IJ$-planar arrays, hence $IJ \times K$ are the weights of the whole structure, with $w_{i,j}$ denoting the column of $IJ$ weights on the $k^{th}$ tap. Array signals after presteering delays can be defined as an $IJ$-dimensional vector $x(\tau)$; therefore $IJK$-dimensional vector $x(\tau)$ is represented as total array signals across the TDL structure as given in (2-50) and (2-51) respectively.

\[
x(\tau) = [x_{1,1}(\tau), x_{1,2}(\tau), \ldots, x_{1,J}(\tau), x_{2,1}(\tau), \ldots, x_{I,J}(\tau)]^T_{1\times IJ}
\] (2-50)
\[ X^T = [x^T(r) \ x^T(r-T) \ x^T(r-2T) \ \ldots \ x^T(r-(K-1)T)]_{U \times K} \]  

(2-51)

As a consequence, following (2-48)-(2-49), the output \( y(r) \) of beamformer processing can be derived in the vector notation by

\[ y(r) = \omega^T X(r) \]  

(2-52)

Thus, the mean output power \( E[y^2(r)] \) can be determined by

\[ E[y^2(r)] = \omega^T R \omega \]  

(2-53)

where \( R = E[XX^T] \) is the array auto-correlation matrix with its elements representing the correlation between various tap outputs. Moreover, the correlation between the outputs at specific \( k^\text{th} \) tap delay on the \((i,j)\) element and at \( k^\text{th} \) tap delay on the \((i',j')\) element can be expressed by the following equation, where \( R_{x,x}(\tau_d) \) is the correlation function of \( x(r) \), that is \( R_{x,x}(\tau_d) = E[x(r)x^*(r+\tau_d)] \).

\[ R_{x,i,j,\tau_d}(\tau_d) = E[x_{i,j}(r-(k-1)T)x_{i',j'}^*(r-(k'-1)T) \]

\[ = \hat{R}_{x,i,j,\tau_d}(k-k' \tau_d + T_{i,j}(\theta_0, \phi_0) - T_{i',j'}(\theta_0, \phi_0) - \tau_{i,j}(\theta, \phi) + \tau_{i',j'}(\theta, \phi)) \]

\[ \tau_d = (k-k')T + T_{i,j}(\theta_0, \phi_0) - T_{i',j'}(\theta_0, \phi_0) - \tau_{i,j}(\theta, \phi) + \tau_{i',j'}(\theta, \phi) \]  

(2-54)

### 2.3.5 Spatial Diversity

Spatial diversity can gain various advantages for radio systems corresponding to its behaviour that this technique takes advantage of signal variability from one element to another. Moreover, more energy can be detected when using several antennas at the receiver. As a result, the robustness of communication links, SNR and SINR can be improved. Furthermore, the range and link capacity also gain positive effects from the space diversity technique. To predict the performance of a diversity-based multisensor receiver, spatial correlation functions are taken into consideration. In following equation, normalised correlation between antenna elements is defined which describes the relation between the output signal of one antenna to its value of another and and presented as the spatial diversity descriptors below.
y_m(τ) and y_n(τ) are output signals for the m\textsuperscript{th} and n\textsuperscript{th} arrays respectively. Both terms are the correlation results between the received signal, R(τ), and a reference template signal, W(τ). That is
\[ y(τ) = R(τ)W(τ - τ_d)dτ. \]
This equation is the general definition to indicate the signal output issued from one element comparing to the signal obtained from the given sensor. Generally, there are various statistical computation techniques of spatial correlation between array sensors regarding the method the receiver is operated [Ben06].

- **Synchronised reference descriptor, \( \hat{ρ}_{SR} \):** Assuming that optimum synchronisation can be achieved when considering the reference array, this leads to the maximum correlation value. Thus, to estimate correlation between any n\textsuperscript{th} element and the reference element, spatial correlation is considered at the time of the maximum output signal relative to the reference sensor signal as presented by (2-56). Denote \( τ_{\text{max,ref}} = \text{argmax}_τ (y(τ)) \).

\[
\hat{ρ}_{SR} = \frac{E[y_{\text{ref}}(τ_{\text{max,ref}}) y_n^*(τ_{\text{max,ref}})]}{E[|y_{\text{ref}}|^2]} \tag{2-56}
\]

- **Synchronised sensor descriptor, \( \hat{ρ}_{SS} \):** This correlation case emphasises estimation at the maximum output signal from n\textsuperscript{th} array instead as given in (2-57) where \( τ_{\text{max,n}} = \text{argmax}_τ (y_n(τ)) \).

\[
\hat{ρ}_{SS} = \frac{E[y_{\text{ref}}(τ_{\text{max,ref}}) y_n^*(τ_{\text{max,n}})]}{E[|y_{\text{ref}}|^2]} \tag{2-57}
\]

Next, the following spatial correlation functions are described in case of operating in array RAKE systems (the part of array RAKE systems will be explained in Chapter 6).

- **RAKE decorrelation descriptor, \( \hat{ρ}_{RD} \):** In this case, it is assumed that perfect RAKE receiver can be obtained at the reference sensor. Therefore, the received signal is determined by operating convolution between the received signal from n\textsuperscript{th} array, \( R_n(τ) \), and the reference CIR, \( h_{\text{ref}}(τ) \): \( R_n(τ) = h_{\text{ref}}(τ) * R_n(τ) \). Thus, the output signal \( y_n^*(τ) = R_n(τ)W(τ - τ_d)dτ \) and \( τ_{\text{max,ref}} = \text{argmax}_τ (y_n(τ)) \). Hence,
RAKE combining descriptor, $\hat{\rho}_{\text{RC}}$: Contrary to the previous case, RAKE combining on each individual array is assumed to be perfect. Output signals are calculated relative to their individual signals. Therefore, $R'_n(\tau) = h_n(\tau) * R_n(\tau)$ and $y'_n(\tau) = |R'_n(\tau) W(\tau, \tau_0) d\tau$. Thus, the spatial correlation function for RAKE combining can be expressed by (2-59) where $\tau_{\text{max},n} = \arg \max_{\tau} (y'_n(\tau))$.

$$\hat{\rho}_{\text{RC}} = \frac{E[y'_n(\tau_{\text{max},n}) y'_n^*(\tau_{\text{max},n})]}{E[|y'_n|^2]}$$  \hspace{1cm} (2-59)

### 2.4 Conclusion

Definition of UWB impulse radio signal, characteristics of UWB propagation channels and UWB propagation models were derived to describe physical characteristics of UWB propagation. The configuration of PPM-TH-UWB signals and discrete time CIR based Saleh and Valenzuela propagation model were mainly used in the research. Results obtained from the proposed issues were compared with these standard UWB propagation characteristics.

Furthermore, UWB array systems including of the UWB array antennas, characteristics of array channels, tapped-delay line model, UWB beamforming processor, and spatial diversity were explained in both $M$-ULA and $MxN$-planar array systems. Array spatio-temporal computational process implemented in this research was based on these methodological series.
Chapter 3

3 Modified Spatio-Temporal Simulation and Distortion Channels

In Chapter 2, the basic characteristics and the literature review of UWB communication systems have been described in general. Further details of its propagation channels including performance degrading effects due to propagation environments will be presented in this chapter. To study realistic UWB propagation channels, high-resolution measurement data with array systems become very significant to enable a full visualisation of the physical ray paths and propagation directions. However, operation of UWB measurement systems in realistic propagation environments can cost time and resources to perform precise investigation. Unfortunately, measured UWB signals might include receiver noise, jitter, and synchronisation which are the main degrading factors for the systems. Thus another way that can be simply implemented to achieve UWB radio channel characterisation is channel simulation to perform all realistic possible paths without any noise or unwanted data. All related parameters such as AODs, AOAs, TOAs and obstruction positions etc. can be specifically generated in various sampled situations.

Consequently, in this chapter the UWB simulation modelling in both time and frequency domains based on array systems will be proposed. The classification of clusters along the propagation paths and physics-based distortion mechanisms are also generalised to be included into the simulation algorithm. Furthermore, simulated channel results with distortion effects, frequency dependent characteristics and distortion impacts on system performances will be illustrated herein.

3.1 Introduction

There is extensive work in channel investigations discussing UWB multipath clusters and their dependency on measurement bandwidth and the considered environment. In general, the UWB multipath cluster model is described by the classical S-V model [Sal87] and by some
modified Spatio-Temporal Simulation and Distortion Channels

modifications as proposed by Chong et al. [Cho05a] and Spencer et al. [Spe00] where, in the latter, the combined spatio-temporal statistical model for indoor multipath propagation was presented herein. In order to investigate UWB channel modelling, its propagation paths are characterised as the stationary condition where the channel characteristics remain constant within finite space, time, and frequency intervals. However, most of the current available UWB channel model simulations imply only channel impulse responses without any details of obstruction clusters and distortion effects to support realistic UWB propagation channels. Furthermore, some research proposed a generic statistical channel model for a wideband environment [Che06]. This work included the frequency dependent characteristics such as angular/delay spreads. The departure and arrival of multipath rays were grouped into a cluster in which a set of AOAs, AODs, and TOAs were generated for each frequency interval. However, due to the lack of information of obstructions throughout propagation paths, physics-based distortion characteristics due to multipath effects were not considered in this simulation.

In order to investigate both the realistic UWB spatio-temporal channel statistics and the geometrically-based pulse distortion model, the simulation proposed in this chapter is a combination of simulating characteristics of UWB multipath clusters and physics-based pulse distortions captured by planar arrays. In addition to time domain characteristics, this simulation is extended to include the frequency-dependent electromagnetic properties of obstruction materials. This deterministic model simulation can provide the capability of generating accurate site specific UWB CIRs and their realistic pulse waveforms. Alternatively, this proposed simulation also gain a viable and affordable way to investigate the UWB CIRs with the advantage of a short simulation time due to generating only specific dominant multipath signals. This can reduce simulation time when comparing with the ray tracing technique in which all propagation paths are computed [Zha04b]. Finally, UWB receiving system performance is also presented to quantify distortion effects. All cases of multipath channels are taken into account over all frequency sub-bands.

3.2 UWB Multipath Clusters and Channel Impulse Response

3.2.1 Classification of UWB multipath clusters

According to multipath cluster investigations derived from many indoor UWB measurements [Sal87], [Cho05a], [Spe00], [Han06b], and [Cra02], each multipath cluster could be identified by a group of MPCs, which is scattered or reflected from obstructions, with similar AOAs, AODs, and TOAs. The received multipath clusters from dominant propagation paths from a transmitter to a receiver are expected to come from three types of radio propagation paths. The
first group corresponds to scattering nearby the transmitter site. Similarly, another group of clusters can be observed at the receiver site due to the scattering objects in the neighbouring area of the receiver. Finally, LOS components between the transmitter and the receiver are considered. Consequently, these different propagation clusters can be classified into three classes as Class-I, Class-II and Class-III type of clusters respectively [Che06] as illustrated in Figure 3-1 where the diagrams of four channel propagation cases corresponding to cluster classifications are depicted herein.

Figure 3-1: Configuration of 4 cases of indoor UWB channel propagation
When defining significant clusters specifically related to propagation scenarios, in the LOS scenario of Case-A describing a small furnished office room, the channel is dominated by Class-III clusters including of LOS components, single reflection and multiple reflections from a wall or a floor as depicted in Figure 3-1(a). Next, the NLOS scenario Case-B is considered where a transmitter and a receiver are in the same office room with a light wall or a cloth partition separation between both ends; the channel would still be dominated by Class-III cluster. Besides single reflection and multiple reflections, refraction by a partition can be observed as shown by Figure 3-1(b). However, when propagation paths in a larger furnished office room -NLOS scenario Case-C (Figure 3-1(c))- are considered, all three classes of clusters are presented. Finally, according to Figure 3-1(d), if a transmitter and a receiver are located in a different furnished room or separated by a thick wall, the channel would be dominated by Class-I and Class-II clusters corresponding to the extreme NLOS condition Case-D. Only reflection from a wall and multiple reflections significantly affect the propagation channels. Physics-based pulse distortion mechanisms of all obstructions will be defined in Section 3.3.

3.2.2 Channel Impulse Response Modelling

Unlike typical IEEE UWB channel models [Mol04], this simulation relies on the dynamic UWB channel model which can be utilised by time-variant channels. Extending the work of the previous simulation model [Che06], inter-cluster and intra-cluster characteristics are also examined. Several statistics such as AOAs, AODs, TOAs, Power Angular Spectrum (PAS), Power Delay Spectrum (PDS), etc., have to be known first in order to parameterise the models.
These parameters can be obtained from various channel measurements [Sal87], [ChoOSa], [Spe00], [Han06b], and [Cra02]. Generally, impulse response can be defined by

\[ h(s, \tau, \theta, \phi) = \sum_{l=1}^{L(s)} \sum_{k=1}^{K(l)} \sqrt{P_{k,l}(s)} \cdot \delta(\tau - T_i(s) - \tau_{k,i}(s), \theta - \Theta_i(s) - \theta_{k,i}(s), \phi - \Phi_i(s) - \phi_{k,i}(s)) \]  

(3-1)

where \( s \) is the index of the UWB sub-band channel divided from the entire frequency spectrum. \( L(s) \) is the number of clusters for the \( s^{th} \) sub-band channel at each time instant \( \tau \) and \( K(l) \) is a total number of multipath components (MPCs) in the \( l^{th} \) cluster. \( T_i(s) \) is the TOA of the \( l^{th} \) cluster, and \( \tau_{k,i}(s) \) is the delay of the \( k^{th} \) MPC in the \( l^{th} \) cluster. \( \Theta_i(s) \) and \( \Phi_i(s) \) are the mean of all elevation and azimuth angles of arriving contributions in the \( l^{th} \) cluster, and \( \theta_{k,i}(s) \) and \( \phi_{k,i}(s) \) are the elevation and azimuth AOAs of the \( k^{th} \) MPC in the \( l^{th} \) cluster respectively. \( P_{k,i} \) is the average power of the \( k^{th} \) MPC in the \( l^{th} \) cluster including the PDS and PAS of inter-cluster statistics as probability density functions, \( p_{\text{inter}}(T_{i}, \Theta, \Phi) = p(T_{i}) p(\Theta) p(\Phi) \), and intra-cluster statistics, \( p_{\text{intra}}(\tau_{k,i}, \theta_{k,i}, \phi_{k,i} | \tau_{k,1}, \theta_{k,1}, \phi_{k,1}) = p(\tau_{k,1}) p(\theta_{k,1}) p(\phi_{k,1}) \). For the inter-cluster, the conditional distributions of \( T_i, \Theta_i \) and \( \Phi_i \) depend on the given TOA of the previous multipath cluster, \( T_{k-1} \), elevation angle and azimuth angle of the first cluster respectively.

However, for the contribution statistic, it is assumed that the relation between cluster and ray statistics and, that, the relation between TOA and AOA distributions are independent. Only the dependence of TOA of MPC on TOA of the previous MPC in the same cluster is defined, \( p(\tau_{k,i} | \tau_{k-1,i}) \). These TOA and AOA distributions are described by two Poisson processes with exponential decaying and a zero-mean Laplacian distribution with standard deviation \( \sigma \) respectively as shown in [Spe00], [Cho03a].

Consequently, the combined time-varying CIR for each scenario can be characterised as follows. Firstly, in LOS Case-A and NLOS Case-B, Class-III cluster-CIRs are taken into account as shown in (3-2) and (3-3) respectively.

\[ h_{\text{case-A}}(s, \tau, \theta, \phi) = h_{\text{LOS}}(s, \tau, \theta, \phi) + h_{\text{III}}(s, \tau, \theta, \phi) \]  

(3-2)

\[ h_{\text{case-B}}(s, \tau, \theta, \phi) = h_{\text{LOS}}(s, \tau, \theta, \phi) + h_{\text{III}}(s, \tau, \theta, \phi) \]  

(3-3)
The CIR term for LOS components which is considered in (3-2) and (3-3) is defined by
\[ h_{\text{LOS}}(s, \tau, \theta, \phi) = \sqrt{P_{\text{LOS}}(s)} \cdot \delta(\tau - d(s)/c, \theta - \theta_{\text{LOS}}, \phi - \phi_{\text{LOS}}) \]
where \( P_{\text{LOS}} \) is the power of LOS component, \( d \) is the distance between both ends, \( c = 3 \times 10^8 \) m/s, \( \theta_{\text{LOS}} \) and \( \phi_{\text{LOS}} \) are the elevation and azimuth AOAs of the direct path. Next, for the CIRs of the Case-C and Case-D NLOS scenarios,

\[ h_{\text{case-C}}(s, \tau, \theta, \phi) = h^I(s, \tau, \theta_{\text{TX}}, \phi_{\text{TX}}) \]

\[ h_{\text{case-D}}(s, \tau, \theta, \phi) = h^I(s, \tau, \theta_{\text{TX}}, \phi_{\text{TX}}) \]

\( g'(s, \tau, \theta_{\text{TX}}, \phi_{\text{TX}}) \) is the single-directional-channel response of Class-I clusters at the transmitter site.

\[ g'(s, \tau, \theta_{\text{TX}}, \phi_{\text{TX}}) = \sum_{l=1}^{L(s)} \sum_{k=1}^{K(l)} \left[ P_{l,k}(s) \cdot \delta(\theta_{\text{TX}} - \theta_{\text{TX},l,k}(s)) \cdot \delta(\phi_{\text{TX}} - \phi_{\text{TX},k,l}(s)) \right] \]

\( L(s) \) is the number of Class-I clusters for the \( s \)-th sub-band channel and \( K(l) \) is the number of AODs in the \( l \)-th Class-I cluster. \( \theta_{\text{TX},l,k}(s) \) and \( \phi_{\text{TX},k,l}(s) \) are the elevation and azimuth AODs of the \( k \)-th MPC in the \( l \)-th cluster respectively, and finally, \( P_{l,k} \) is the average power of the \( k \)-th MPC in the \( l \)-th cluster which is determined by PAS. In order to generalise the TOA, AOA, and AOD profiles to be used for the model simulation, the method of equal areas (MEA) [Che06], [Pat04] is applied. Detail of derivation of MEA is illustrated in Appendix A.

### 3.3 Pulse Distortion Modelling

In addition to CIRs to be employed, distortion characteristics are also considered for constructing the simulation model. According to UWB channel properties for multipath environments, more than one multipath ray can be observed within a short time bin. These overlapped ray arrivals essentially originate from the same propagation routes and can inevitably cause pulse shape distortions. As a result, this section describes an UWB time-domain distortion model, based on geometric optic (GO) technique to describe direct and reflected rays, and based on geometric/uniform theory of diffraction (GTD/UTD) technique to describe the diffracted rays. The simplified UWB received signal is \( R(\tau) = W(\tau) * h(\tau) \) where \( W(\tau) \) is the transmitted pulse. Typically, the total response, \( h(\tau) \), from a complex multipath channel can be modelled by the
summation of all impulse responses of local scattering with the closed form expression of specific geometric configurations as presented in (3-7) [Qiu06b].

![Diagram of distortion mechanism configuration](image)

Figure 3-2: Diagram of distortion mechanism configuration

\[ h(\tau) = h(\tau, \theta, \phi) = \sum_{n=1}^{N_{GO}} a_n(\theta, \phi) \cdot \delta(\tau - \tau_n) + \sum_{n=1}^{N_d} h_n(\tau, \theta, \phi) * \delta(\tau - \tau_n) \]  

(3-7)

where \( N_{GO} \) and \( N_d \) are the geometric-optic-ray and diffracted-ray-numbers respectively. In general, three significant propagation mechanisms which cause physics-based distortion in UWB propagation channels are considered: GO rays (LOS connection), diffracted rays due to a half plane, and finally, diffracted rays by a dielectric slab [Qiu04]. Illustration of UWB pulse distortion mechanisms is also illustrated in Figure 3-2 where the omnidirectional antenna is considered as the transmitter, and 3x3 planar array antenna is the receiver.

### 3.3.1 Geometric Optic Rays and Multiple Reflections

In practice, the distortion due to GO rays can be simply presented by the two-ray model, a direct signal and a single reflection signal, as can be defined by [Qiu04], [Qiu06b].
where \( R_1(\tau) = K\{\pm \bar{\alpha}(\tau)+2k(1-k^2)\exp[-(1+K)a\tau]\}U(\tau), a \leq 1 \). \( K=(1-k)/(1+k) \) and \( k=\sqrt{\varepsilon_r \cos^2 \theta_r / \varepsilon_r \sin^2 \theta_r} \) for vertical or horizontal polarisation respectively with \( \theta_r=\arctan((h_r + h_\alpha)/d) \). \( \tau=(r_{\text{direct}}+r_{\text{direct}})/c, r_{\text{direct}}=\sqrt{(h_{tx} - h_{rx})^2 + d^2} \) and \( r_{\text{reflect}}=\sqrt{(h_{tx} + h_{rx})^2 + d^2} \)

\[ R_M(\tau) = K^M \delta(\tau) + MK^{M-1} \cdot R_{01} + \sum_{m=2}^{M} \frac{M!}{(M-m)!m!} K^{M-m} F_{0m}(k, \tau) \] (3-9)

\( F_{0m}(k, \tau) \) can be derived as described in [Qiu04] where \( F_{0m}(k, \tau)=4k(1-k^2)^m \cdot \exp(-a \tau) \cdot \sum_{k=0}^{\infty} f_{km} \cdot m \geq 1 \) with \( f_{km} = 1/2K(k!) \cdot (x/k^2)^k (-1)^k \cdot ((x+k)\exp(-x) - \sum_{l=0}^{m+k-1} [(-1)^l x^l / l!]) \)

\[ +(-1)^{m+k+1} x^m k / (m+k-1)! \cdot f_{km} = 1/2K \cdot (x\exp(-x) - \sum_{l=0}^{m-1} [(-1)^l x^l / l!]) + (-1)^{m-1} x^m (m-1)! \].

### 3.3.2 Diffraction by a Perfectly Conducting Half-Plane

This diffraction phenomenon is commonly produced by edges of tables, counters, other furniture etc. as shown in Figure 3-2. For any incident angle \( \theta_i' \) at a half plane, the impulse response can be expressed by [Qiu04]

\[ h_{\text{half-plane-diff}}(\tau) = \sqrt{2r_{\text{direct}}/c} \frac{1}{2\pi} \left[ \frac{\cos((\theta'_r - \theta'_i)/2)}{\tau + r_{\text{direct}}/c \cos(\theta'_r - \theta'_i)} \pm \frac{\cos((\theta'_r + \theta'_i)/2)}{\tau + r_{\text{direct}}/c \cos(\theta'_r + \theta'_i)} \right] \] (3-10)

\[ \frac{1}{\sqrt{\tau - r_{\text{direct}}/c}} U(\tau - r_{\text{direct}}/c) \]
3.3.3 Propagation Through d-thickness-Slab Obstructions

This propagation mechanism can be generally observed in indoor UWB propagation such as propagating through walls, doors, partitions, etc. Transmitted pulses can be transmitted, reflected or even diffracted by a thin slab in different regions. Moreover, multiple reflections and transmissions are generated when propagating through thick walls or obstacles. The closed form time-domain expressions of the total transmission coefficient, \( T_{\text{slab}}(\tau) \) and the total reflection coefficient, \( R_{\text{slab}}(\tau) \), for propagation through any \( d \)-thickness slabs can be derived as follows [Qiu04],

\[
T_{\text{slab}}(\tau) = \delta(\tau - \tau_{2,1}) + R_2(\tau) \cdot [\delta(\tau - \tau_{1,1}) - \delta(\tau - \tau_{2,1})] + \sum_{n=2}^{\infty} R_{1,2n}(\tau) * \delta(\tau - \tau_{1,n}) - \sum_{n=1}^{\infty} R_{1,2n}(\tau) * \delta(\tau - \tau_{2,n})
\]

where \( \tau_{1,n} = 2\left(-\sqrt{\tau \sin \theta_i \sin \theta_d} + \tau_{1,1}\right) \) and \( \tau_{2,n} = [-(2n-1)\sqrt{\tau_i} + (2n-2)\sin \theta_i \sin \theta_d + \cos(\theta_i - \theta)]nllc \) with \( l = d_{\text{thickness}}/\cos \theta \) and \( d_{\text{thickness}} \) is thickness of a slab.

\[
R_{\text{slab}}(\tau) = -R_1(\tau) \cdot [\delta(\tau) + \delta(\tau - \tau_1)] - \sum_{n=2}^{\infty} R_{1,2n-1}(\tau) * \delta(\tau - \tau_n) + \sum_{n=1}^{\infty} R_{1,2n+1}(\tau) * \delta(\tau - \tau_n)
\]

where \( R_{1,1} = K_1 \delta(\tau) + MK_1^{-1} R_0(\tau) + \sum_{m=2}^{M} M!((M-m)!m!)K_1^{-m} F_m(k_1, \tau); M \geq 1 \), and the \( n^{th} \) delay \( \tau_n = 2\left(1 + \sin \theta_i \sin \theta_d \right)nllc \). Additionally, if UWB pulses propagate through a thin thickness slab, the total diffraction coefficient, \( D_{\text{slab}}(\tau) \), will be considered as shown in (3-13).

\[
D_{\text{slab}}(\tau) = 1 - T_{\text{slab}}(\tau) * D_{\text{half-plane}}(\tau, \theta_d - \theta_i) + R_{\text{slab}}(\tau) * D_{\text{half-plane}}(\tau, \theta_d + \theta_i)
\]
where $D_{\text{half-plane}}(\tau, \theta, \phi) = -L\cos((\theta_{d\tau} + \phi)/2)/(2\pi\sqrt{2\tau}(\tau+L/2c)\cos^2((\theta_{d\tau} + \phi)/2))$. These pulse distortion characteristics will be combined into the UWB CIR model as mentioned in the previous section in order to include the distortion effects into the model simulation.

### 3.4 Modified Spatio-Temporal Simulation and Frequency-Dependent Results

Consequently, in order to construct realistic UWB propagation channels, classified MPC clusters and pulse distortion effects are simulated together. Typically, the total response, $h(\tau)$, from a complex multipath channel can be modelled by the summation of all impulse responses of local scattering with the closed form expression of specific geometric configurations. Unlike typical UWB channel models [Mol04], this simulation relies on combining pulse distortion characteristics into the UWB CIR model. In addition, this modified simulation also includes frequency-dependent effects as each CIR model is considered at each frequency sub-band, $s$. For instance, in the LOS scenario of Case-A describing a small furnished office room, the channel would be dominated by Class-III clusters and distorted by multiple reflections and half-plane diffraction as defined by

$$h_{\text{case-A}}(s, \tau, \theta, \phi) = h_{\text{LOS}}^{\text{III}}(s, \tau, \theta, \phi) + h_{\text{GO}}^{\text{III}}(s, \tau, \theta, \phi) + h_{\text{half-plane}}^{\text{III}}(s, \tau, \theta, \phi) \quad (3-14)$$

Next, the NLOS scenario Case-B is considered where a transmitter and a receiver are in the same office room with a-light-wall or a-cloth-partition separation between both ends; the channel would be still dominated by Class-III clusters and distorted by the similar effects as Case-A. However, thin slab diffraction is added into the distortion model, $h_{\text{thin-slab}}^{\text{III}}(s, \tau, \theta, \phi)$ to construct $h_{\text{case-B}}$.

$$h_{\text{case-B}}(s, \tau, \theta, \phi) = h_{\text{LOS}}^{\text{III}}(s, \tau, \theta, \phi) + h_{\text{GO}}^{\text{III}}(s, \tau, \theta, \phi) +$$

$$h_{\text{half-plane}}^{\text{III}}(s, \tau, \theta, \phi) + h_{\text{thin-slab}}^{\text{III}}(s, \tau, \theta, \phi) \quad (3-15)$$

When propagation paths in a larger furnished office room, NLOS Case-C, are considered, all three classes of clusters are presented. All distortion effects might possibly appear in Class-III
similarly to LOS-A, but distortions due to thin slab diffraction and thick slab reflection are modelled for Class-II clusters. CIRs of Case-C can be described by (3-16) where $g(s,\theta_x,\sigma_x)$ is the single-directional CIR of Class-I clusters at the transmitter site. $\theta_x$ and $\phi_x$ are the elevation and azimuth AOD of the MPCs. Finally, if a transmitter and a receiver are located in a different furnished room or separated by a thick wall, the channel would be dominated by Class-I and Class-II clusters corresponding to the extreme NLOS condition (Case-D). Only thick slab reflection is considered. Thus only the last two terms in (3-16) are taken into account for $h_{\text{case-D}}$.

$$h_{\text{case-C}}(s,\tau,\theta,\phi) = h_{\text{GO}}^{\text{III}}(s,\tau,\theta,\phi) + h_{\text{half-plane}}^{\text{III}}(s,\tau,\theta,\phi) +$$

$$h_{\text{thin-slab}}^{\text{II}}(s,\tau,\theta,\phi) + h_{\text{thick-slab}}^{\text{II}}(s,\tau,\theta,\phi) + g^I(s,\theta_x,\phi_x)$$

(3-16)

$$h_{\text{case-D}}(s,\tau,\theta,\phi) = h_{\text{thick-slab}}^{\text{II}}(s,\tau,\theta,\phi) + g^I(s,\theta_x,\phi_x)$$

(3-17)

### 3.4.1 Methodology of Modified Spatio-Temporal Simulation

This section proposes an UWB pulse simulation method based on planar array systems, a different approach than the one proposed in [Che06]. Furthermore, this modified simulation also incorporates the physics-based pulse distortion model. Using priori modelled parameters [Sal87], [Cho05a], [Spe00], [Han06b], and [Cra02], which are presented in Table 3-1, the structure of this simulation can be described as follows with a uniform distribution, $U[\cdot;\cdot]$, denoted in some parameters.

1. The entire frequency (2-11 GHz) will be divided into 10 sub-bands with each sub-band occupying 1 GHz bandwidth. CIRs and distorted pulses are generated representing the arrival rays with the UWB transmitted pulse consists of the second derivative of a Gaussian function defined as $W(\tau) = \tau \exp(-\pi(\tau/\tau_d)^2)$). All following processes are applied to each sub-band. Although using 500 MHz bandwidth can assure the constant value of the dielectric constant within a sub-band, thus no effect of frequency dependence appears within each sub-band, and time bin resolution is considerably long, 2ns, to observe each individual distorted pulse. On the other hand, it is neither compatible for using larger bandwidth, i.e. 2 GHz, since the dielectric constant cannot be kept remain constant leading to the high probability of frequency dispersion even in the short pulse duration (0.5ns).
2. The propagation scenario of interest for instances Case-A, Case-B, Case-C or Case-D is specified, and the cluster numbers of each associated cluster class corresponding to the environment conditions are generated.

3. Frequency-dependent angular spread \( \sigma_\phi \) and delay spread \( \sigma_\tau \) for each sub-band are defined, where \( \sigma_\phi = \beta_\phi \exp(\gamma_\phi f) \) and \( \sigma_\tau = \beta_\tau \exp(\gamma_\tau f) \), \( f \) is the centre frequency of each sub-band. \( \beta_\phi \) and \( \beta_\tau \), \( \gamma_\phi \) and \( \gamma_\tau \) are amplitude parameters and exponential decay factors of angular and delay spread functions respectively.

4. The discrete AODs, AOAs and TOAs are generated using the MEA (Appendix A) with \( \phi_x = \phi_0 - \frac{1}{\sqrt{2}} \cdot \sigma_\phi \ln(-2X/X) \) where \( \phi_0 = 2\phi_0 - \phi_{x+1}, x=1,2,..., X/2 \) and \( X=30 \). These formulas can be used to determine both azimuth and elevation angles with \( \phi_0 = U[0,360] \) used for a uniform distribution of azimuth angles and \( \theta_0 = U[-90,90] \) used for elevation angles. \( \gamma = \gamma_\tau + (-\sigma_\tau \ln(1-y/Y)), \gamma=1,2,..., Y-1 \) with \( Y=20 \).

Moreover, when Cluster-I and Cluster-II are considered, it is assumed that the AODs represent the incident angles of the transmitted signal when propagating through the obstructions. Additionally, azimuth AOAs represent arriving angles arising from diffraction or reflection contributions from horizontal directions, and elevation AOAs represent diffracting or reflecting angles coming from vertical directions. In particular, due to the large distance between the transmitter and the receiver, AOAs generated for Class-II are coupled into AODs generated for Class-I corresponding to the possible directions of signal arriving. As a result, all AOAs of Class-II clusters are mapped with some AODs from the transmitter dominated by Class-I.

<table>
<thead>
<tr>
<th>Scenario Cases</th>
<th>Dominant Cluster</th>
<th>Simulating Channel Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>LOS-A (3-18 m)</td>
<td>Class-III / GO, M-reflection, Half-plane diffraction</td>
<td>U[1,5] U[1,3] U[0,5] U[0,2] 0 0 U[10,60] 4.3 7.1</td>
</tr>
<tr>
<td>NLOS-C (9-30 m)</td>
<td>Class-I/II / Thin slab diffraction, Thick slab reflection</td>
<td>U[5,10] U[5,10]</td>
</tr>
</tbody>
</table>

Table 3-1: Typical values of model parameters for CIR generalisation
3.4.2 Simulation Results

Using this methodology, UWB multipath CIRs and their distorted pulses can be generalised based on specific environments. Modelling parameters and types of distortion mechanisms in Table I are selected corresponding to propagation path cases and related classes of clusters. The height of the transmitter and the receiver is 1.325m. The time resolution of generated signals in this simulation is 83.32ps with 6144 data points. Accordingly, the maximum time delay allocating MPCs is approximately 500ns. However, very low level of some generated impulse signals obtained from the simulation are not considered although they appear earlier than 500ns. Since a whole 10-frequency-sub-band received signals are simulated with 3x3 planar arrays, this can lead to complex and congested results (90 values for one MPC). Hence only simulated results from one frequency sub-band observed by the reference (3,3) antenna are shown.

![Figure 3-3: Simulated results for LOS Case-A scenario](image)

According to the simulation process mentioned in the previous section, examples of general characteristics of CIRs results are obtained as depicted in Figure 3-3 where the UWB channel for LOS Case-A simulated at sub-band-frequency 6GHz (6-6.999GHz) are shown. Figure
3-3 (a) depicts the power density of clusters and MPCs at particular azimuth AOAs and TOAs by power delay azimuth density spectrum (PDADS, $P(\tau, \phi)$). $P(\tau, \phi)$ can be given by $P(\tau, \phi) = E[|h(\tau, \phi)|^2]$ where $E[\cdot]$ and $|\cdot|$ denote the expectation and absolute value respectively [Cho03a]. Figure (b) and (c) presents discrete amplitude and power signals of CIRs with temporal domain respectively. Then, received LOS pulse and distorted pulses due to multiple reflections are shown in Figure (d).

Figure 3-4: Examples of pulse distortions due to reflection and diffraction mechanisms

Figure 3-4 depicts the examples of generalised 5 UWB MPCs arriving at the receiver at 30ns, 34ns, 55ns, 123ns and 175ns respectively. The first arrival path represents a LOS received signal with AOA=81°. The second component represents the 3-multiple reflection signal between two wallboards located both sides along the propagation path to the receivers. It is assumed that reflected angles between both wallboards are identical with 87°. The third MPC is specified as the diffraction from an aluminium edge of a bookcase with $r_{direct} = 3.2m$ and $\theta_d = 87^\circ$. When considering propagation through a thin slab obstruction, a wooden door ($d_{thickness}=4.44cm$) with $\theta_d = 76^\circ$, the sample of the distorted pulse is illustrated as the fourth MPC. Finally, the distortion effect due to propagation through a thick slab such as a concrete wall ($d_{thickness}=19.45cm$) can be shown in the last MPC.
Modified Spatio-Temporal Simulation and Distortion Channels

Figure 3-5: Comparison of simulated 3D CIR between LOS Case-A and NLOS Case-B

Figure 3-6: Comparison of simulated pulse distortions between LOS Case-A and NLOS Case-B
Figure 3-5 and Figure 3-6 present the comparison of simulated results of LOS Case-A and NLOS Case-B operating at 6 GHz. CIRs are plotted in three dimensional views of azimuth AOAs and TOAs. Received pulse signals for LOS Case-A, are simulated as a direct path signal with a reflection from a floor and also multiple reflections from surrounding obstructions. The aluminium plate is considered as the plate from which signals diffract to the receiver. Furthermore, simulated UWB pulse signal propagating in an office room (NLOS Case-B) where some areas are separated by a thin wall board \((d_{\text{thickness}}=1.17\text{cm})\) and a cloth partition \((d_{\text{thickness}}=5.93\text{cm})\) is also presented in these figures. Signal arrivals at 65ns, 73ns and 78ns present propagations through a cloth partition and a wallboard respectively. Comparing between these two scenarios, number of MPCs in Case-A are less than Case-B but with stronger strength of received signals.

In addition, UWB propagation in a larger office room where a thick wall is located along propagation paths, NLOS Case-C, is simulated and its results are shown as follows. This scenario case consists of all three classes of clusters Class-I, Class-II and Class-III. Consequently, angle coupling between cluster Class-I and Class-II is taken into account. According to (3-3) and (3-6), the number of both elevation, \(\theta_{\text{tx,kl}}(s)\), and azimuth, \(\phi_{\text{tx,kl}}(s)\), AODs for Class-I, are mapped into each AOA for class-II. The mapping function for azimuth angles, \(M_\phi\), with the average number of azimuth AODs can be formulated by:

\[
M_\phi : \{\phi_{T,1}, \phi_{T,2}, \phi_{T,3}, \ldots, \phi_{T,C}\} \rightarrow \phi_{R,k,l}^I
\]

where \(\{\phi_{T,1}, \phi_{T,2}, \phi_{T,3}, \ldots, \phi_{T,C}\} \subseteq \{\phi_{T,1,1}, \phi_{T,1,2}, \ldots, \phi_{T,k,l}, \ldots, \phi_{T,K,L}\}\) (3-18)

\(C\) is the number of AODs selected to be matched. It can be described that each AOA \(\phi_{R,k,l}^I\) is coupled with \(C\)-AODs \((\phi_{T,1}, \phi_{T,2}, \ldots, \phi_{T,C})\) chosen from the set of all AODs, \(\{\phi_{T,1,1}, \phi_{T,1,2}, \ldots, \phi_{T,K,L}\}\) as illustrated in Figure 3-7 and Figure 3-8. Subsequently, this is also worth mentioning that each AOD is assumed to be mapped bijectively to the corresponding AOAs. This is also similarly formulated for elevation angle mapping, \(M_\theta\) [Che06].

Comparing between simulated MPCs scattering from three clusters, according to the characteristics of clusters, Class-III clusters are dominated by the direct signals, single reflection and multi-reflection between LOS connections. Therefore, its signal strength, indicated by the red line, is higher than pulse signals dominated by Class-I and Class-II clusters as indicated by the green line. Furthermore, pulse shapes of MPCs in Class-III clusters are less distorted than MPCs in Class-I and Class-II clusters as illustrated in Figure 3-9. Multipath clusters arriving at 182.7ns
Modified Spatio-Temporal Simulation and Distortion Channels

Figure 3-7: A set of all azimuth AODs of Class-I to be matched with AOAs of Class-II

Figure 3-8: Correspondent AOAs of Class-II being matched with AODs of Class-I
are dominated by diffraction from a wooden door with $d_{\text{diff}}=4.44\text{cm}$, and at 203.7ns and 223.2ns are distorted by reflection from a concrete wall ($d_{\text{wall}}=19.45\text{cm}$).
Figure 3-10 presents configuration of simulated pulses for distorted UWB propagation channel of NLOS Case-D which is considerably affected by thin-slab diffraction and thick-slab reflection. As a result, the least amplitudes of simulated pulse signals are obtained comparing to other scenario cases. Furthermore, TOAs of clusters and MPCs can be observed at later time delays (135ns) with severe distortions. These are caused by propagating from the longest distances or between rooms with various obstructions along the propagation paths. Further investigation of distorted UWB channels corresponding to frequency dependent characteristics is presented in the next section.

### 3.4.3 Frequency-Dependent Results

Referring to dielectric constants of furniture materials in propagation paths [Saf02], [Muq05], properties of dielectric constants with frequency variation are also taken into account, thus simulation at each sub-band examines pulse distortions from different dielectric constants. This is different from previous results in which only the results simulated over 6GHz sub-band, 6-6.999GHz, are illustrated. The number of clusters and MPCs arriving within each cluster corresponds to cluster arrival rate, $A$, and ray arrival rate, $\lambda$, and also corresponds to cluster decay factor, $\tau$, and ray decay factor, $\gamma$, as described in Table 3-1. Accordingly, it can be supposed that all simulated MPCs at each frequency sub-band rely on UWB multipath channel characteristics.

The effects of frequency dependent dielectric constants of materials for individual UWB propagation paths are characterised and their results are shown in Figure 3-11. Total energies can be computed by summing up all power of simulated MPCs at each frequency sub-band. Power in each MPC is normalised over total power for each obstruction material within each frequency sub-band. It can be seen that, there are frequency-selective characteristics for propagation through each material as total energy variations fluctuate in the overall sub-band. Frequency dependent distortions in LOS Case-A and NLOS Case-B are rarely observed, except the energy degradation at 5 GHz and at 4 GHz respectively. In NLOS Case-C where includes dominant distortion effects from all cluster classes, Frequency dependent variations can be seen. Total energy fluctuates during all frequency sub-bands depending on each material obstruction. However, since dielectric constant value of glass is constant throughout all frequency sub-bands, small fluctuation of total energy of received signal scattering from glass can be observed. This also appears in other scenario cases. Furthermore, small frequency dependent variation of a brick wall is presented in the result in LOS Case-A, NLOS Case-B and C since small numbers of MPCs reflection from a brick wall are selected in the simulation process due to its scenario condition that should be taken into account in the extreme NLOS condition.
Modified Spatio-Temporal Simulation and Distortion Channels

Figure 3-11: Comparison of total energy variations of received signals simulated for all frequency sub-bands

Figure 3-12: Frequency-dependent energy distribution of all scenario cases classified by types of distortion mechanism
Dramatic variation of total energies can be seen in NLOS Case-D propagation channel. This is caused by various obstructions along the long distance of propagation paths which can distort and attenuate the simulated signal strength. Moreover, weak energy signals can be obtained during 3-6 GHz affected by thin slab diffraction from cloth partition and styrofoam wall. Thus, incorporating with multiple reflections occur before or after thin slab diffractions, received signal strength is extremely degraded leading to low values and fluctuation of total energy in this case.

Furthermore when considering UWB energy distribution of all UWB channels distorted by various obstructions over all frequency sub-bands, energy variation classified by distortion mechanisms can be determined as illustrated in Figure 3-12. Summation of total energies of all MPCs from all clusters distorted by particular obstructions is calculated. Total energies for MPCs of LOS and GO ray mechanism are less than others. This is caused by the least total numbers of all LOS contributions to be calculated although its received signal energies are maximum when considering individual path cases. In contrast, most numbers of MPCs are distorted by other mechanisms, such as multiple reflections, half-plane diffractions, thin slab diffractions and thick slab reflections, thus due to these large numbers of distorted MPCs, the highest normalised total energies in received signals can be observed in these distortion mechanisms. Due to the large numbers of clusters and MPCs as indicated in Table 3-1, energy distributions of these distortion effects are approximately in the same high level, especially for thin and thick slabs distortions. When considering normalised total energy distribution specifically in each distortion mechanism, high level of energy distribution can be seen during 3-6GHz for LOS/GO ray and multiple reflections from a wallboard mechanisms. However, there is considerable degradation of energy distribution for diffraction from aluminum plate at 8GHz and for thin slab diffraction from a cloth partition and thick slab reflection from a concrete block at 7 GHz.

3.5 Quantification of Distortion Effects on System Performance

UWB pulse distortions are inherently characterised by the extremely large bandwidth. Since quantifying the impacts of pulse distortion on UWB system performance appears to be novel, recently there are several works reported about quantification of UWB distortion effects in addition to the physics-based pulse distortion issues which have been addressed in previous research [Qiu04], [Wan05], [Muq05], and [Qiu06b]. Inter symbol interference (ISI) and probability of bit error rate (BER) due to distortion effects in various complex propagation conditions are commonly investigated as researched by [Win02], [Zha06], and [Mak08b]. Furthermore, Zhou [Che07] also reported loss of SNR as high as 4dB in template mismatches due to distortion effects. Moreover, to evaluate the timing error due to distortion, timing errors of
received distorted signals were compared with the benchmark value. The error range much larger than the Cramer-Rao lower bound (CRLB) was found in this study. These lead to errors that can limit the accuracy of TOAs of received multipath signals. The very high temporal resolution of UWB pulse signals makes UWB signals become the ideal candidates for combined communications and positioning. If the TOAs of incoming multipath signals are known with little uncertainty, the computation of propagating distances from the source to the receivers is still possible with a few errors in estimation. Yuan et al. reported the deterioration of UWB positioning performance due to distortions even though the modified phase-only correlator method for high-resolution multi-target ranging was employed [Yua08].

Consequently, to quantify pulse distortion impacts on UWB system performance, BER probability and ranging error are analysed in this section subject to specific scenarios, obstruction materials and individual MPCs. Thus characteristics of signal dispersion due to various multipath distortion conditions are examined. Binary PPM-TH-UWB is generated as the transmitted signal conveying 100,000 bits through 100,000 pulses, thus code repetition coding is not applied. The average pulse repetition period is 60ns guaranteeing the absence of ISI in LOS. These generated signals, \( W(r) \), are simulated and transmitted over the synthesised multipath channels as stated in Section 3.4; therefore CIRs, \( h(r) \), including cluster types and distortion effects are taken into account.

### 3.5.1 Probability of Bit Error Rate

An ideal RAKE receiver that processes all multipath contributions is simulated for the receiving system. Consequently, probability of bit error rate, \( P_{b} \), of binary PPM-TH received signals, \( R(r) = W(r) \ast h(r) \), can be identified by

\[
P_{b} = \frac{1}{2} \text{erfc} \left( \sqrt{\frac{E_{\text{rx}}}{2N_0}} \right) = \frac{1}{2} \text{erfc} \left( \sqrt{\frac{\text{SNR}}{2}} \right)
\]  

(3-18)

where error function \( \text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_{x}^{\infty} \exp(-t^2) dt \) and \( E_{\text{rx}} \) represents the received energy per pulse [Ben04]. Ideal all RAKE receiver is considered with maximal ratio combining evaluating the weighting factors to be used in a RAKE receiver.

Error probability results in distorted UWB multipath channels for LOS Case-A, NLOS Case-B, NLOS Case-C and NLOS Case-D are illustrated in Figure 3-13, Figure 3-14, Figure 3-15 and Figure 3-16 respectively where system performances simulated at all frequency sub-bands and compared with the link performance of the AWGN channel. Obstruction materials are chosen...
corresponding to propagation scenarios and classes of clusters. For example, multiple reflections are generated by reflections from a wallboard, structure wood, plywood and glass, and aluminium plate is considered for perfectly-conducting half-plane diffraction (LOS Case-A and NLOS Case-B), a wooden door and a cloth partition are considered as thin-slab obstructions (NLOS Case-B, Case-C and Case-D). Styrofoam, bricks and concrete blocks are taken into account as wall materials between rooms (NLOS Case-D). Comparing performances simulated all sub-bands for LOS Case-A, the best performance could be obtained for UWB multipath channels simulated over 4GHz sub-band (4-4.999GHz); however the best performance for Case-B, Case-C and Case-D can be observed at simulating over 5GHz (5-5.999GHz), 4GHz (4-4.999GHz for Case-C-III), 6GHz (6-6.999GHz for Case-C-II) and 10GHz (10-10.999GHz) sub-bands respectively. Since the number of clusters and MFCs of each scenario are simulated randomly in each sub-band, high impact of distorted UWB received signals on system performances can be observed at different sub-bands. However, when considering performances on the general basis, results can be confirmed that distortion effects can degrade system performances regarding to characteristics of propagation paths with the highest performance in Case-A following by Case-B, Case-C and Case-D respectively.

Furthermore, comparing $P_{b}$ between each scenario case can quantify distortion effects corresponding to cluster classifications and propagation scenarios as depicted in Figure 3-17. The best performances of each case allocated in different frequency sub-bands are compared. It is observed that at BER=0.001, there are approximately 0.2-dB loss in SNR of MFCs originated from cluster Class-III between LOS Case-A and NLOS Case-B and between NLOS Case-B and Case-C-III. Moreover, less than 0.1-dB loss can be obtained between SNR of Class-II clusters in Case-C-II and Case-D-II. Because all channel parameters generated by Class-I clusters such as AODs, TOAs and CIRs are coupled with Class-II clusters as scatterer sources for Class-II, Class-I clusters are not taken into account for simulating system performance.
Figure 3-13: Comparison of error probability simulated for LOS Case-A over all frequency sub-bands

Figure 3-14: Comparison of error probability simulated for NLOS Case-B over all frequency sub-bands
Figure 3-15: Comparison of error probability simulated for NLOS Case-C over all frequency sub-bands.

Figure 3-16: Comparison of error probability simulated for NLOS Case-D over all frequency sub-bands.
3.5.2 Timing Error and Ranging Error

In addition to SNR loss and BER due to the pulse shape mismatch, pulse distortion can also degrade synchronisation and positioning by time shifting of TOAs as well as an amplitude error in the correlation peak. Accurate synchronisation is significant for UWB communication systems due to its extremely short pulse width. The pulse shape mismatch due to distortion can lead to the timing error between correlation peaks of the template and received signals and result in the serious degradation of BER. $R(t)$ and $R'(t)$ are denoted as the received pulse waveform corresponding to undistorted channel and distorted channel respectively. $R(t)$ is considered as the reference signal or the local template which is used to be correlated with other received signals as normally operated in a correlation-based receiver. When estimating timing errors of distorted signals, cross-correlation between distorted signals, $R'(t)$, and the template signal, $R(t)$, is estimated and compared with the correlation results of the auto-correlation of $R(t)$ [Zho07]. To facilitate the comparison of results, these two different signals are normalised to have the same energy before correlating. Definition of timing error estimation is expressed by $\tau_e = \tau_{RR^*} - \tau_{RR}$ which both terms of time located the highest peaks of correlation function can be defined by (3-19) where the cross-correlation term of $R_{rr}(t) = E[R(t)R^*(t+\tau)]$, and the auto-correlation term $R_{aa}(\tau) = E[R(t)R^*(t+\tau)]$. 

Figure 3.17: Comparison between the highest system performances of each scenario regarding to cluster classification.
Modified Spatio-Temporal Simulation and Distortion Channels

\[
\tau_{RR'} = \arg \max_\tau |R_{RR'}(\tau)|
\]

\[
\tau_{RR} = \arg \max_\tau |R_{RR}(\tau)|
\]  

(3-19)

Figure 3-18 shows the estimation of timing errors by time difference of correlation peaks between correlated signals. Figure 3-18 (a) presents the ideal case of free space LOS scenario where autocorrelation results of the received pulse without distortion \(R(\tau)\) shown by the solid line and autocorrelation results of the transmitted pulse \(W(\tau)\), the dashed line, present the same peaks of the waveforms. Both peaks of these symmetric waveforms correctly indicate the timing position. Since, the impact of pulse distortion is examined in this study, \(R(\tau)\) is considered as the local template to be correlated with the received distorted signals \(R'(\tau)\) as shown in Figure 3-18 (b)-(f). The solid line presents the correlation result between different templates, \(R(\tau)\) and \(R'(\tau)\). The dashed curve is the result of correlation with the same template itself or autocorrelation of undistorted received signal, \(R(\tau)\). The symmetric autocorrelation waveform can be observed.

Figure 3-18: Correlation of distorted signals with the template signal
contrary to the different template correlation whose its asymmetric waveform gives an error in the timing position.

The second MFCs scattered from obstruction clusters class-III in scenarios Case-A, Case-B and Case-C are examined as illustrated in Figure 3-18 (b)-(d) respectively. Two supplot pictures at the bottom of each cross correlation results describe two signals being used for correlating. The upper one is the template received signal without any distortion effect, \( R(t) \), and the lower one shows the received distorted pulse signal, \( R'(t) \). Since a few differences in timing errors is rarely observed in the first MFCs corresponding to LOS components or the shortest distance propagation paths, to present noticeable timing error characteristics caused by distortion, the later components of multipath receiving pulses are selected as examples to be shown. The fifth and the fourth MFCs originated from clusters class-II in scenarios Case-C and Case-D are also presented in figure (e) and (f) respectively. Distinguishing the difference between these two peaks can define timing errors as also shown in Table 3-2. Timing errors and ranging errors (multiplying timing error by light velocity \( 3 \times 10^8 \) m/s) due to pulse distortion regarding to classification of scenario cases, cluster types and obstruction materials are summarised herein.

The accuracy of the TOA estimation or timing errors caused by pulse distortion is expressed by the minimum variance of the TOA estimation error, \( \sigma_\tau^2 \), in terms of CRLB. To compare the estimated errors with a benchmark, the limitation of the lower bound performance can be defined by CRLB. This value is related to signal bandwidths and SNR at the receiver. For a single path AWGN channel, the best achievable accuracy of a position estimate, which is derived from TOA estimation, satisfies the following inequality [Gez05], [Urk83]:

\[
\sqrt{\sigma_\tau^2(d)} \geq \frac{c}{2\sqrt{2\pi} \beta \sqrt{\text{SNR}}}
\]  

(3-20)

Thus, the minimum estimated position error less than \( d_{\text{CRLB}} \) m can be obtained when considering SNR=0. \( c = 3 \times 10^8 \) m/s and \( \beta \) is the effective or root mean square signal bandwidth given by function of the Fourier transform of the transmitted signal, \( P(f) \). In this study, frequency domain of PPM-TH-UWB transmitted signals, is considered to compute \( \beta \) over 2-11GHz.

\[
\beta \approx \left[ \int_{-\infty}^{\infty} |f|^2 \left| P(f) \right|^2 df / \int_{-\infty}^{\infty} \left| P(f) \right|^2 df \right]^{1/2}
\]  

(3-21)
In order to present the pulse distortion effects on UWB propagation channels, comparison between lower bound average distance estimation errors computed by CRLB and estimated ranging errors are described. Since $\beta = 3.4\,\text{GHz}$, is calculated from the same transmitted signals over 2-11GHz, all scenario cases take into account the same value of $\beta$ leading to the lower bound distance error of $d_{CRLB}=0.99\,\text{mm}$ in all scenario cases. When multiplying timing errors by the light velocity, $c=3\times10^8\,\text{m/s}$, ranging errors can be defined. Comparing ranging error distances and the lower bound distance error, $d_{CRLB}=0.99\,\text{mm}$, results significantly show that apart from SNR, distortion effect can cause performance degradation in ranging errors much bigger than $d_{CRLB}$. These examples of received pulse distortion are analysed specifically at the highest energy allocation frequency sub-bands, as stated in Figure 3-17, with various types of obstruction material.

<table>
<thead>
<tr>
<th>Distortion Channels</th>
<th>Obstruction Material</th>
<th>RMS Bandwidth $\beta$ (GHz)</th>
<th>$d_{CRLB}$ (mm) at SNR=0</th>
<th>Centre Frequency Sub-band, $f_s$ (GHz)</th>
<th>Order of MFCs</th>
<th>Timing Error (ns)</th>
<th>Ranging Error (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case A-III</td>
<td>Wallboard</td>
<td>3.4</td>
<td>0.99</td>
<td>4</td>
<td>1</td>
<td>0.083</td>
<td>24.9</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>2</td>
<td>0.65</td>
<td>195</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>1</td>
<td>0.083</td>
<td>24.9</td>
</tr>
<tr>
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<td></td>
<td></td>
<td></td>
<td>2</td>
<td>0.33</td>
<td>99</td>
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<td></td>
<td>4</td>
<td>0.70</td>
<td>210</td>
</tr>
<tr>
<td>Case B-III</td>
<td>Wallboard</td>
<td>3.4</td>
<td>0.99</td>
<td>5</td>
<td>1</td>
<td>0.083</td>
<td>24.9</td>
</tr>
<tr>
<td></td>
<td></td>
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<td></td>
<td></td>
<td>2</td>
<td>0.33</td>
<td>99</td>
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<td></td>
<td></td>
<td></td>
<td>3</td>
<td>0.96</td>
<td>288</td>
</tr>
<tr>
<td>Case C-III</td>
<td>Wooden door</td>
<td>3.4</td>
<td>0.99</td>
<td>4</td>
<td>1</td>
<td>0.083</td>
<td>24.9</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>2</td>
<td>0.33</td>
<td>99</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>3</td>
<td>0.96</td>
<td>288</td>
</tr>
<tr>
<td>Case C-II</td>
<td>Partition</td>
<td>3.4</td>
<td>0.99</td>
<td>6</td>
<td>4</td>
<td>0.105</td>
<td>315</td>
</tr>
<tr>
<td></td>
<td>Concrete Block</td>
<td></td>
<td></td>
<td></td>
<td>5</td>
<td>1.15</td>
<td>345</td>
</tr>
<tr>
<td></td>
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<td>0.96</td>
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<td></td>
<td></td>
<td>3</td>
<td>1.15</td>
<td>345</td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td>4</td>
<td>1.57</td>
<td>471</td>
</tr>
</tbody>
</table>

Table 3-2: Pulse distortion impacts on timing error
Furthermore, comparison of pulse distortion effects on each MPC at each propagation channel is also presented in Table 3-2. Generally, timing and ranging errors of the first three MPCs in propagation channels obstructed by cluster Class-III are roughly in the same range and less than errors caused by cluster Class-II. More errors are remarkably estimated in later MPCs of propagation Case C-II. According to the correlation of dense multipath channels of Case C-II propagation as presented by Figure 3-18 (e), the later fourth MPC scattered by concrete block obstruction, cluster Class-II, gives the high timing error of 1.15ns or the ranging error of 345mm. And the highest timing error of 1.57ns timing error or 471mm ranging error can be seen for the fourth MPC of Case D-II due to thick slab reflection from concrete block.

3.6 Conclusion

A modified simulation of UWB multipath channels combined with cluster classification and physics-based pulse distortion mechanisms caused by different mechanisms of single and multiple reflections and diffraction was proposed in this chapter. Spatio-temporal characteristics of multipath clusters which extend the standard IEEE UWB CIR model were specifically generated based on 3x3 planar array systems with regard to scenario types and simulated over 10 frequency sub-bands (2-11 GHz). Thus, frequency dependent characteristics of the propagation channels were also investigated and compared between each scenario both for LOS and NLOS cases. In addition, simulation over 10 frequency-sub-bands could lead to variation of frequency-dependent normalised energy regarding to cluster classification and obstruction materials.

The extremely wide bandwidth of UWB signals can cause pulse distortion due to the frequency selectivity of attenuation in propagation channels. In order to take into account frequency-dependent effects on multipath signals, distorted UWB multipath pulses were generalised at each frequency sub-band regarding to TOAs, AODs, AOAs and dominant obstruction clusters classified for each scenario case. Moreover, results of frequency dependent distortions presented variations of normalised total energies of all simulated pulse signals. Comparing between all scenario cases, NLOS Case-D presented the worst case of total energy frequency-variation as propagating through the longest distance and scattering from various obstructions as resulted in Figure 3-16.

Furthermore, more significant results of the frequency-dependent characteristics and the quantification of distortion effects on UWB multipath channels were also presented. Probability of bit error rate was determined to quantify distortion effects on UWB multipath channels for all frequency sub-bands. Reasonable investigated results were presented as the maximum total energy distribution, and the highest system performance (the lowest probability of bit error rate) could be observed in LOS case. In contrast, propagation between rooms gained the worst results...
of both energy distribution and system performance [Mak08a], [Mak08b]. Finally, due to the distortion effect, this chapter also presented performance degradation in ranging errors. Correlation between distorted pulses and transmitted pulses determined timing and ranging errors, which these ranging errors extremely exceeded the benchmark errors estimated by CRLB. In addition, results also examined that LOS component gained more accuracy of ranging errors than late-arrival NLOS distorted components.
Chapter 4

4 Spatial Correlation based Array Processing for Multipath Cluster Estimation

Modelling the temporal characteristics of the UWB channel has been a major area of interest in UWB. The use of multi-antenna elements provides an opportunity to characterise the temporal and the spatial properties of the propagation channel simultaneously. Furthermore, characteristics of the correlation of UWB signals should be taken into account to determine the effectiveness of diversity combining schemes or beam-forming techniques. Conventional spatial diversity can be effectively operated when the received signals exhibit a low degree of correlation values whereas beam-forming schemes correspond to a high degree of correlation. Since this research relies on both planar (considering azimuth and elevation angles) and linear arrays (considering only elevation angles), the main objective of this chapter is to determine an accurate, flexible and sufficient spatial channel model that incorporates the spatio-temporal features of UWB channels.

4.1 Background

Recently, there has been an interest in utilising antenna array measurements to investigate the radio propagation characteristics of the UWB channel. MPC measurements in particular employ array antenna systems to analyse the MPCs arriving at the receive end from different directions. For emerging UWB multi-sensor radio measurements, real or virtual antenna array systems are typically implemented [Win02], [Cho05b], [Han04b], [Mal08a]. In many cases, due to cost and complexity issues, instead of using real antenna arrays, many measurement campaigns have been conducted through the use of virtual antenna arrays, utilising small-scale grid measurements [Win02], [Cho05b] or synthetic multiple element antennas [Han04b]. These types of measurements can allow the investigation of the effect of multipath delays and directions of arrival in addition to signal power monitoring. Thus, TOAs as temporal and AOAs as spatial analysis of UWB multipath impulses can be investigated.
So far existing work has focused on the characteristics of the temporal and spatial correlation [Cho05b], [Has93], [Cas02], [Pre02], [Kar98]. Both properties of MPCs have been explored in order to determine the characteristics of multipath signals in different scenarios. According to previous studies [Cho05b], [Has93], [Cas02], a corresponding PDP was computed from each UWB CIR which was detected at each antenna array or grid position. UWB PDPs obtained from all spatial points at each location were averaged to give a SSA-PDP. Then, the time axis of these SSA-PDPs was divided into small delay bins which were assumed to contain either one resolvable multipath or no path. The temporal correlation $\rho_t$ was computed by considering the adjacent bins of these SSA-PDPs, $\rho_{t,k,k+1}$ (where $k = 1...K$ are sequence numbers of MPC bins).

It should be mentioned that differential averaged PDP bins are variable due to multiple diffracted and reflected contributions resulting in more randomness and smaller correlation. Hence, the results of the temporal correlation were relatively small and neglected for practical purposes. Further examinations of the spatial correlation with respect to antenna separation distances and AOA were also described in [Pre02]. Moreover, the spatial correlation between adjacent antenna elements at different separations as a function of both the distance and the excess delay have been also reported [Kar98]. Furthermore, [Cho03b], [Cho03a] characterised both temporal and spatial domains of a wideband channel by determining correlation between cluster TOAs and AOAs. However, no information regarding the existence of the amplitude relation of multipaths received at adjacent antenna array positions was provided in these studies.

The significance of this study comes from the fact that it is useful to characterise the spatial correlation of UWB signals in order to examine multipath signals scattering around antenna arrays. In this work, the effect of the AOA is highlighted by the suggested computation technique; this is because impulse signals detected at each spatial point are considered during each time delay bin. The directions of the MPCs arriving at each adjacent point are different and independent of each time delay, hence it can be implied that the AOA factors are also included, and the time variation of any correlation can be clearly observed. Some important findings resulting from the use of the array spatial correlation technique are presented in this work. By examining MPCs arriving at adjacent antenna array positions at each time bin, multipath clusters can be distinguished. In this study, TOAs of multipath clusters could be immediately identified for dense multipath channels when compared to the process used in [Cho05b], [Sal87], [Spe00], [Cas04] where they have been estimated manually due to the small amount of multipath arrivals in sparse channels. The suggested technique is also less complicated to implement when compared to algorithms, such as the Sensor-CLEAN algorithm [Cra02].

Generally, for UWB systems, two high-resolution algorithms, the Sensor-CLEAN and the UWB-SAGE which is an extension of the expectation-maximisation: EM, have been used to extract the multipath components, both in AOAs and TOAs, with a better accuracy than that
obtained from a Fourier-base analysis [Mol05b]. But waveform estimation is not necessary for the SAGE algorithm because the transmitted waveforms are assumed to be known in the receiver side and frequency domain processing is employed [Han03]. In addition, other algorithms such as the MUSIC and the estimation of signal parameters via rotational invariance techniques (ESPRIT) were also taken into consideration [Xin04b]. However, these algorithms deteriorate rapidly when received signals from different AOAs are strongly correlated, and these techniques also increase complexity of system implementation. Consequently, some work extended these techniques to the joint time and space domain as the joint angle of arrival and delay of arrival estimation (JADE) using a known transmitted signal [Pic03]. Nevertheless, this technique is limited by the resolution in the angle-time domain. On the other hand, the Sensor-CLEAN algorithm can determine incident waveforms thus sequential data in time domain is necessary. Consequently, according to time-domain signals obtained from the measurement, the Sensor-CLEAN algorithm is used in this work for the post-processing. Therefore, comparison of results between the proposed technique and the Sensor-CLEAN algorithm will be presented here.

### 4.2 Discrete Array Processing of UWB Impulse Signals

#### 4.2.1 General Discrete Time UWB Channel Model

In wideband systems, impulse responses of multipath components tend to arrive in clusters. A common way to present this process is established by IEEE 802.15.3a [Mol04], which is based on a modified S-V model [Sal87], as follows:

\[
h_{\text{discrete}}(\tau) = \sum_{l=0}^{L} \sum_{k=0}^{K_l} a_{k,l} \delta(\tau - T_l - \tau_{k,l}) e^{j\phi_{k,l}}
\]  

(4-1)

where \(a_{k,l}\) is the multipath gain coefficient of the \(k^{th}\) MPC in the \(l^{th}\) cluster. \(L\) and \(K_l\) represent the number of clusters and the number of MPCs within that \(l^{th}\) cluster respectively. \(T_l\) is the delay of the \(l^{th}\) cluster or the TOA of the first arriving MPC within that \(l^{th}\) cluster. \(\tau_{k,l}\) denotes the \(k^{th}\) path arrival delay with respect to the first arriving MPC in the \(l^{th}\) cluster. \(\delta(\cdot)\) is the Dirac delta function with phases \(\phi_{k,l}\) uniformly distributed.

In this work, the tapped delay line channel model suggested by [Has93] was applied to simplify the UWB discrete impulse response model utilised in the proposed processing technique.
This discrete tapped delay line channel model is based on the resolvable time delay bins of the UWB impulse radio, as given in (4-2). The time axis is divided into small time intervals and delay bins, which are assumed to contain either one multipath component, or no multipath component. $N$ is the total number of resolved time bins, $a_n$ and $\phi_n$ are random amplitude and carrier phase sequences respectively. $\Delta t_n$ is the time duration of each bin or the minimum resolved time bin.

$$h_{\text{discrete}}(\tau) = \sum_{n=1}^{N} a_n \delta(\tau - n\Delta t_n) e^{j\phi_n}$$

(4-2)

### 4.2.2 Spatial Correlation Computation Technique

The analysis performed in this study was based on dense channel measurements originating from time domain UWB measurements [UWB Database], obtained by propagating a pulsed signal between a fixed transmitter position and multiple receiver positions throughout an office building. Each receiver location was measured in a 7x7 square grid at 15 cm spacing under static conditions, assuming that the propagation environment did not change appreciably during the measurement phase. Generally, a transmitted pulse arrives at spatially separated antennas at different times; however, by using a synchronisation trigger, a virtual array antenna can be constructed. The receiving time of each spatial point was controlled by a clock trigger; therefore sequential measurements of all 49 spatial points could be interpreted as simultaneous array measurements [Cra02].

For all UWB signals received at different locations, a different noise floor level of each room was computed separately by averaging the portion of the PDP, which was captured before the first arrival of MPC, over the 49 locations within the room [Cas02]. As a consequence, PDPs which were higher than the cut-off threshold (6dB above the average noise floor) were computed, and a 2ns temporal width was selected as the computational interval in each bin in order to eliminate the unwanted noise spike and to ensure that only the effective multipaths are analysed. This bin width compromises with the high temporal resolution of the transmitted pulse shape (2ns pulse duration) as presented in [Cra02]. Since the receiving signals were detected over a 50ns window with 1024 measured samples, a resolution time of 48.828ps could be obtained; therefore 41 points of discrete UWB impulse signals were computed for one delay bin. In general, there may be more than one path arriving at the same bin; however the content of each bin was referred to as one MPC. Nonetheless, MPC arrival times $> 300$ns were neglected due to the low signal levels expected. The diagram of the measurement campaign is illustrated below.
For the purpose of this study, the measurements conducted by UltraLab [UWB Database] were considered. It is assumed that multipath signals reflected, diffracted, and propagated to receivers from all directions. Therefore, in order to include the effect of AOA of arriving multipaths, calculating spatial correlation coefficients between signals received at each position and at its surrounding array positions (in row, column and diagonal directions) are considered as described in (4-4). Some high degree correlation values might be observed along these directions depending on the direction of MPC arrivals at each time delay. Based on the general equation of the correlation coefficient which was used in [Cho05b], [Has93], [Cas02], [Kar98], spatial correlation coefficients, \( \rho \), of the same PDP bins at adjacent positions were computed as described below, where \( R_{ijk} \) is the spatial correlation matrix and PDP at each bin \( P = |h(\tau)|^2 \).
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

\[ R_{i,j,k} = \rho_{(i,j,k)(i\pm m,j\pm n,k)}(k) \] for \( m, n=\{0,1\} \) and \( m \neq n \neq 0 \) (4-3)

\[
\rho_{(i,j,k)(i\pm m,j\pm n,k)}(k) = \frac{E[P(i,j,k) \cdot P^*(i\pm m,j\pm n,k)] - E[P(i,j,k)] \cdot E[P^*(i\pm m,j\pm n,k)]}{\sqrt{[E[P(i,j,k)]^2 - E[P(i,j,k)]^2] \cdot [E[P^*(i\pm m,j\pm n,k)]^2 - E[P^*(i\pm m,j\pm n,k)]^2]}}
\] (4-4)

where \( P(i,j,k) \) is the PDP- amplitude of \( k^{th} \) multipath component bin at the \((i^th, j^th)\) position. \( i, j \) is the relative coordinate in the 7x7 square grid, hence \( i \) and \( j = 1...7 \) where \( i \) represents the row index and \( j \) represents the column index of the grid. \( k (k=1...K) \) is the \( k^{th} \) bin of \( K \) PDP bins estimated over the entire power delay profile.

Three types of computations can be considered with respect to 7x7 square grid position groups. Firstly, the correlation coefficients of each \( k^{th} \) PDP bin are observed between adjacent row positions, \( \rho_{(i,j,k)(i\pm 1,j,k)}(k) \) where \( m=1, n=0 \). Secondly, correlation coefficients between each \( k^{th} \) PDP bin of signals received between two adjacent column positions, \( \rho_{(i,j,k)(i,j\pm 1)}(k) \) where \( m=0, n=1 \), are computed. Finally, correlation values of each PDP bin with oblique positions can be described by \( \rho_{(i,j,k)(i\pm 1,j\pm 1)}(k) \) where \( m=1, n=1 \).

![Figure 4-2: 49-Small-scale array positions](image-url)
The PDP bin diagram for the 49 receiving array position is illustrated in Figure 4-2. For example, for each \( k \)th PDP bin, 8 values of correlation coefficients of a signal detected at the (2,2)-position can be calculated by using Equation (4-4). Consequently, \( \rho_{2,2x1,2}(k) \), \( \rho_{2,2x3,2}(k) \), \( \rho_{2,2x2,1}(k) \), \( \rho_{2,2x2,3}(k) \), \( \rho_{2,2x1,1}(k) \), \( \rho_{2,2x1,3}(k) \), \( \rho_{2,2i1,3}(k) \), and \( \rho_{2,23,3}(k) \) can be obtained. On the contrary, there are only 5 values of correlation coefficients when considering the correlation between the signal power received at the (6,7)-position and other signal powers received at the 5 adjoining positions.

Accordingly, a total of 156 correlation coefficient values for the overall 49 array positions were analysed for each PDP bin. Computation results are presented in the next section.

4.3 Spatial Correlation Analysis and Results

The spatial correlation characteristics of UWB impulse signals were investigated by considering correlation coefficients of PDPs in each bin at different array positions, and are presented in this section. According to previous studies [Cho05b], [Has93], [Cas02], where temporal correlation coefficients between adjacent SSA-PDP bins were analysed; minor or, even, non-correlated coefficients were obtained. These results are due to multipath signals arriving at antennas due to reflected or diffracted contributions from any clutter in the propagation paths. Hence, unpredictable fluctuation of MPCs could cause signals to become uncorrelated. This assumption can also be supported by Figure 4-3, which illustrates nine auto-correlation functions, \( R_{ij}(\tau) = \text{E}[P_{ij}(t)P_{ij}^*(t+\tau)] \) where \( P_{ij} \) is a signal power captured at the reference point (4,4) and at its neighbouring positions. The components are uncorrelated when excess delays are roughly equivalent to the pulse duration (2ns). Accordingly, computing temporal correlation coefficients over entire excess delays can lead to the relatively low correlation results summarised in earlier studies. Moreover, it can be seen that \( R_{ij}(\tau) \) of all positions are similar with each other within 0.5ns of the excess delay corresponding to the mutual correlation of pulse signals received among adjacent positions. Hence, computing correlation values of signals using this short time bin, only 0.5ns = 15cm / 3x10^6m, might include the mutual impedance effect between neighbouring array positions but the correlation characteristic throughout the whole pulse shape is excluded.
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

In this study, correlation coefficients of each \( k \)th PDP bin were calculated throughout all 49 spatial positions. A temporal axis of MPC time delay was computed with respect to the PDP bin numbers. Results for LOS and NLOS radio propagation are presented where the diagram of measurement locations is illustrated in Figure 4-1. Results present five scenarios of observation: LOS1 (room F1) and LOS2 (room F2) with the distance from the transmitter \( D \approx 9.5 \)m and \( 5.5 \)m respectively. AOAs relative to a line between the transmitter and the receiver (the centre position of the array) are \( \phi \approx 172^\circ \) and \( 203^\circ \) respectively. For NLOS, NLOS1 (room M) with \( D \approx 13.5 \)m, AOA \( \approx 255^\circ \), NLOS2 (room P) with \( D \approx 6 \)m, AOA \( \approx 49^\circ \) and NLOS3 (room B) with \( D \approx 17 \)m, AOA \( \approx 191^\circ \) are additionally characterised. For example in room F1, there were 120 PDP bins in total and the amplitude power in each bin was higher than the threshold level, counted for each array position; therefore, 240ns of excess delays were considered. The first bins of PDPs, which were estimated from all grid positions, were averaged and set to the reference TOA. With regard to propagation delays at each receiver antenna position, the empty column vector, \( D_{\text{tx}} \) was added into the PDP bin matrix at each position, prior to computing the first bins.

In addition, Figure 4-4 presents contour maps of average correlation coefficients, \( \bar{R}_{i,j,k} \) of signals received at each position and its surrounding positions as defined by (4-3) – (4-4). These correlation values were averaged and presented for each of all 49 array positions, (1,1), (1,2), ..., (1,7), (2,1), (2,2), ..., and (7,7), at each delay bin.
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

\[
\bar{R}_{i,j,k} = \frac{\sum R_{i,j,k}}{P}
\]

\[
P = 3 \text{ for any position } (i, j) \text{ where } i, j = \{1, 7\}
\]

\[
P = 5 \text{ for any position } (1, x), (x, 1), (7, x), (x, 7) \text{ where } x = \{2, \ldots, 6\}
\]

\[
P = 8 \text{ for any position } (i, j) \text{ where } i, j = \{2, \ldots, 6\}
\]

According to Figure 4-4 (a)-(e), it is important to note that spatial correlation coefficients calculated for the first delay bins of 34, 22, 24, 50 and 44 ns respectively are in the range of 0.8 ≤ \( \rho \leq 1 \). The first arriving signals in the first bins are assumed to be coming from the direct path (LOS) or from the shortest path for NLOS, thus signal phases do not relatively change due to clutter; therefore the highest values of correlation coefficients can be found. On the other hand, correlation coefficients of subsequent MPCs, which are taken into account from the second bin, consistently decay with increasing excess delays. Traces of high correlations can also be observed at some of the subsequent bins as indicated by arrow symbols.

The analysis suggests that in LOS scenarios, notably, in Room F1, high correlation values exist along column - array positions at MPC time delays of 34ns, 36ns, 52ns, 68ns, 86ns, 102ns, 120ns and 136ns. In Room F2, there also exist high correlation values located at MPC time delays of 24ns, 26ns, 78ns, 112ns, 132ns and 146ns. These may be caused by MPC arrivals due to diffracted or reflected contributions along the propagation paths. However, the relation between spatial correlation analysis and array positions cannot clearly be evaluated in NLOS propagation scenarios due to the different environments and radio propagation through walls. Distances from the transmitter to receiving locations also have a particular effect on spatial correlation coefficients, i.e. the longer the distance the lower the correlation values. For instance, high correlation values could fairly be observed at 24ns, 42ns, 62ns, 103ns, 106ns, 112ns, 122ns, 134ns, 140ns in NLOS1. In addition, in NLOS2 and NLOS3, significant correlation values are rarely seen at 50ns, 52ns, 58ns, 64ns, 66ns and 76ns and at 44ns, 47ns, 85ns, 97ns and 150ns respectively. Nevertheless, comparing the average TOA of MPC between NLOS2 and NLOS3, although the distance between the transmitter and the receiver in NLOS3 (Room B, D=17m.) is longer than NLOS2 (Room M, D=13.5m), the first average TOA of NLOS3 (\( \text{TOA} = 44 \text{ns} \)) could be estimated earlier than NLOS2 (\( \text{TOA} = 50 \text{ns} \)). This corresponds to more propagation paths through several walls between rooms in NLOS2 contrary to NLOS3 where UWB signals propagating through non-obstacle area prior to propagating into Room B.
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Figure 4-4: Average correlation coefficient contour map for all array positions

(a) LOS1 (Room F1, TOA = 34 ns, Distance=9.5m)

(b) LOS2 (Room F2, TOA = 24 ns, Distance=5.5m)
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Figure 4-4 (cont.): Average correlation coefficient contour map for all array positions

(c) NLOS1 (Room P, $\text{TOA} = 24$ ns Distance=6 m)

(d) NLOS2 (Room M, $\text{TOA} = 50$ ns Distance=13.5 m)
4.4 Time-of-Arrival and Multipath Clusters Estimation

High degrees of correlation values obtained from the previous section were calculated at each delay bin, the resolution time of which was assumed to be equal to the pulse width (2 ns) of the transmitted signal [Win02]. It should be noted that these high spatial correlation coefficients express traces of MPC clusters. A cluster can be defined as an accumulation of MPCs arriving with the same TOAs [Cho05b], at some noticeable time bins. Consequently, those impulse signals seemed to be detected simultaneously at adjacent array points. Nevertheless, the predominance of reflections and diffractions due to obstructions within each location might affect some propagation paths leading to low correlation values. To extract TOAs from traces of high spatial correlation values, steps in the modified technique can be summarised as follows:

1) Average the first bins of all PDPs from all sensor positions and set to the reference TOA for each location.

2) Add the empty PDP set of column vector $D_{\text{TL}}$, where $T$ is equal to time bins regarding propagation delays at each receiving location, prior to the original PDP data set of each sensor.
3) In each time bin, calculate spatial correlation coefficients of PDPs between each array position and its surrounding positions as described in (4-4)-(4-6).

4) Determine a histogram of high-valued correlation coefficients. Next, distinguish time delays that indicate a high degree of correlation ($R_{ij,k} \geq 0.5$). Some results of this step are additionally shown in Figure 4-5.

5) Draw the contour map of average correlation coefficients of signals, $\overline{R}_{ij,k}$, between each position and its surrounding positions as described in (4-5). These correlation values are averaged and presented for each of 49 array positions at each delay bin. Results from this step are depicted in Figure 4-4.

6) At some noticeable time bins, high degrees of correlation values express traces of MPC clusters. These correspond to all results either when considering average correlation values over each array point (step 5), or when considering only high correlation values (step 4). Time bins of MPCs, which locate high occurrence numbers of high correlation values, are considered as the arrival time of multipath clusters.

When considering all correlation values of each room, $R_{i,j,k}$, by determining a histogram of high-valued correlation coefficients, time delays which indicate a high degree of correlation ($R_{i,j,k} \geq 0.5$) can be found and are presented in Figure 4-5. These time delays correspond to the results in Figure 4-4, which represents an averaged contour map of correlation values throughout the 49 positions. Variation in the histograms is clearly seen in the LOS cases, but those are rarely seen in NLOS scenarios particularly in NLOS3. Likewise, an examination was carried out in order to identify multipath cluster regions from these correlation coefficient profiles. Time bins of MPCs, which locate high occurrence numbers of high correlation values, were selected as TOAs of multipath clusters. Therefore, arrival clusters of multipath signals could be obtained.
Figure 4-5: Histogram of MPC time delays with ($\rho \geq 0.5$)
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Figure 4-6: Comparison of Cluster estimation with Sensor-CLEAN algorithm
Cramer [Cra02] proposed the modification of the Sensor-CLEAN algorithm to estimate TOA and AOA of UWB multipath components. This technique was applied to the previously mentioned set of UWB propagation measurements. Consequently, a comparison of TOA distributions obtained from this work and Cramer's algorithm was conducted and presented in Figure 4-6. From the results, it can be observed that time delays estimated by both methods are approximately the same. Since both TOAs and clusters were computed by considering PDP bins leading to less numbers of processed data, estimated results do not perfectly fit results produced by the Sensor-CLEAN algorithm. The average errors of estimated time delays are approximately -14.07 ns and -17.86 ns in LOS1 and LOS2 respectively. Average errors for NLOS1, NLOS2 and NLOS3 are 3.79 ns, -3.08 ns and -3.82 ns respectively. However, the larger differences of estimated TOAs between two methods can be clearly seen in LOS than in NLOS, especially at later time bins corresponding to higher errors in LOS cases. These are possibly caused by signal blocking by two supporting pillars nearby the receivers leading to variant spatial correlation values to be computed.

Information of multipath clusters of all rooms obtained from this section will be further analysed regarding UWB multipath cluster characteristics in order to statistically investigate the TOAs and multipath clusters and to evaluate the accuracy of the suggested analysis. Characteristics of TOAs in comparison with other UWB propagation channels are described in the next section.

4.5 Statistical Modelling of Estimated TOAs and Multipath Clusters

In this section, statistical distributions of TOAs and multipath cluster times that were computed by processing the array of PDP bins as described in the previous section are presented. Besides estimation of statistical parameters for all observation locations, a classification of LOS and NLOS cluster parameters was also carried out. In order to ascertain whether these estimated multipath arrival times have similar characteristics to the UWB channel model, all clustering parameters were modelled based on the Saleh-Valenzuela clustering channel model [Sal87].

When considering the UWB channel model and accounting for MPCs in (4-1), the average power of both clusters and rays within the clusters was assumed to decay exponentially. Hence, the average power of a MPC at a given delay \( T_k + \tau_{k,i} \) is defined by [Sal87]:

\[
\bar{a}_{k,i}^2 = \bar{a}_{0,0}^2 \cdot \exp(-T_k / \Gamma) \cdot \exp(-\tau_{k,i} / \gamma)
\]  

(4-6)
where \( q_{0,0}^2 \) is the expected value of the power of the first arriving MPC. \( \Gamma \) and \( \gamma \) are the exponential decay factors of clusters and MPC rays respectively.

In order to estimate \( \Gamma \), the power amplitude of the first cluster arrival in the measured data set of each location was normalised and its time delay was set to zero. All other cluster arrivals of that data set were defined relative to this amplitude and time. Similarly for \( \gamma \), the peak amplitude of the first arrival in each cluster was normalised to an amplitude of one and a zero-time delay, then all later multipath arrivals within that cluster were referred to this normalised amplitude and time. Consequently, all cluster and ray arrivals were superimposed and plotted with a semi-logarithmic scale. Likewise, both parameters \( \Gamma \) and \( \gamma \) could be estimated by the inversion of negative slopes of linear-least-squares curve fitting in the logarithmic plot. Figure 4-7 and Figure 4-8 display the distributions of the inter-cluster and intra-cluster relative power versus time delays respectively where a best-fit line using the method of least squares was calculated for each individual measurement location from which the decay factors \( \Gamma \) and \( \gamma \) could be derived. There are 12 locations to be measured and plotted to determine the inter-cluster and intra-cluster relative power. Each power distribution for each location can gain individual value of cluster and ray decay factors. The exponential trend of cluster and ray relative power distributions can be clearly seen in LOS rooms (RoomF1, RoomF2) and short distance rooms (RoomL \( D = 8m \), RoomP \( D = 6m \), RoomN \( D = 5.5m \)) respectively. Although RoomP and RoomN are located nearer to the transmitter than RoomL, multipath signals arriving into their receivers are in the endfire direction of the transmitter antenna leading to higher values of exponential decay factors. Furthermore, some overflown distribution data of cluster relative power and ray relative power can be observed in extreme NLOS cases (RoomW, RoomH, RoomU, RoomT, RoomM, RoomB, and RoomC) as present in Figure 4-7 and Figure 4-8. Time dispersion of MPCs in these extreme NLOS cases, or dense channels, is measured leading to high values of exponential decay factors. This is different from received MPCs in LOS and sparse channels which become like many single rays and are indicated by low values of exponential decay factors.

Consequently, their overall mean factors from all room locations, \( \Gamma_{mean} \) and \( \gamma_{mean} \) were then derived. The accuracy of the fit has been quantified in terms of the root mean-squared error (RMSE) of linear-least-squares curve fitting in each location with its averaged RMSE values of 2.6575 and 3.502 for \( \Gamma \) and \( \gamma \) respectively. The results were classified into LOS and NLOS cases and compared with other parameters characterised by other researchers as shown in Table 4-1.
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Figure 4-7: Normalised cluster relative power versus cluster relative delay ($\gamma_{\text{mean}} = 29.57$ ns)

Figure 4-8: Normalised ray relative power versus ray relative delay ($\gamma_{\text{mean}} = 55.59$ ns)
Cluster arrival times and ray arrival times, $T_i$ and $\tau_{k,i}$ respectively, can be described by two independent single Poisson Processes [Sal87].

$$p(T_i | T_{i-1}) = \lambda \exp[-\lambda (T_i - T_{i-1})]$$  \hspace{1cm} (4-7)

$$p(\tau_{k,i} | \tau_{(k-1),i}) = \lambda \exp[-\lambda (\tau_{k,i} - \tau_{(k-1),i})]$$  \hspace{1cm} (4-8)

The first Poisson parameter, $\lambda$, representing the mean cluster arrival rate, can be obtained by considering the delay parameters of the conditional probability distribution in (4-7), which has change in time $\Delta T = T_i - T_{i-1}$. The first multipath arrival in each cluster is considered as the beginning of each cluster, regardless of whether it has the maximum peak amplitude. The first arrival time of each following cluster is subtracted from that of its immediate predecessor. Similarly, the second Poisson parameter, $\beta$, which is the mean ray arrival rate, can be estimated by subtracting ray arrival times from their previous ones within each cluster. Therefore, delay parameters, $\Delta \tau = \tau_{k,i} - \tau_{(k-1),i}$, in a set of conditional arrival times given in (4-8) can be produced. $\Delta T$ and $\Delta \tau$ can be referred as the cluster inter-arrival-times and the ray intra-arrival-times respectively. When employing the linear-least-squares fitting of the complementary of cumulative distribution function (CCDF or 1-cdf) of $\Delta T$ and $\Delta \tau$ in the natural logarithmic scale, $\lambda$ and $\beta$ values can be obtained as illustrated in Figure 4-9 and Figure 4-10 respectively. Determination of all parameters ($\gamma, \lambda, \beta$ and $\alpha$) is described in detail in [Cho05b], [Spe00], and according to earlier statistical results [Cho05b], [Spe00], [Cra02], characteristics of all these parameters are assumed to be similar for all clusters.

The result in Figure 4-10 reveals the discrepancy between linear least-squares fitting and the ln(CCDF) distribution of ray intra-arrival times. Furthermore, the difference between the distribution of ray intra-arrival times and its fitting was also illustrated in [Spe00]. It is possible that the single Poisson process is not sufficient to model the ray intra-arrival times. Therefore, the application of two Poisson processes, proposed by [Cho05a], as given in (4-9) was considered instead,

$$p(\tau_{k,i} | \tau_{(k-1),i}) = \beta \lambda_1 \exp[-\lambda_1 (\tau_{k,i} - \tau_{(k-1),i})]$$

$$+ (1 - \beta) \lambda_2 \exp[-\lambda_2 (\tau_{k,i} - \tau_{(k-1),i})]$$  \hspace{1cm} (4-9)
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Figure 4-9: Logarithmic CCDF of the cluster inter-arrival times (1/\lambda = 39.17\,\text{ns})

Figure 4-10: Logarithmic CCDF of the ray intra-arrival times with mixtures of two Poisson processes 
(\beta = 0.0007282, 1/\lambda_1 = 4.0209, 1/\lambda_2 = 0.2856) and the single Poisson process (1/\lambda = 2.12\,\text{ns})
Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

Parameter | UWB (this work) | Cramer | Chong et al. | Saleh | Spencer et al.
<table>
<thead>
<tr>
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<th></th>
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<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>LOS</td>
<td>NLOS</td>
<td>All rooms</td>
<td>LOS</td>
<td>NLOS</td>
</tr>
<tr>
<td>( \Gamma_{\text{mean}} ) (ns)</td>
<td>5.57</td>
<td>34.37</td>
<td>29.57</td>
<td>27.9</td>
<td>22.1</td>
</tr>
<tr>
<td>( \gamma_{\text{mean}} ) (ns)</td>
<td>2.72</td>
<td>70.69</td>
<td>55.59</td>
<td>84.1</td>
<td>14.27</td>
</tr>
<tr>
<td>( 1/\lambda ) (ns)</td>
<td>63.73</td>
<td>36.86</td>
<td>39.17</td>
<td>45.5</td>
<td>8.69</td>
</tr>
<tr>
<td>( \beta )</td>
<td>0.21</td>
<td>2.07</td>
<td>2.12</td>
<td>2.3</td>
<td>0.51</td>
</tr>
<tr>
<td>( 1/\lambda_1 ) (ns)</td>
<td>-</td>
<td>3.34</td>
<td>4.02</td>
<td>-</td>
<td>0.74</td>
</tr>
<tr>
<td>( 1/\lambda_2 ) (ns)</td>
<td>-</td>
<td>0.048</td>
<td>0.29</td>
<td>-</td>
<td>6.68</td>
</tr>
</tbody>
</table>

Chong et al. defined the second term coefficient of mixed Poisson processes in Equation (4-9) as \( \beta \).

Table 4-1: Comparisons of Clustering Channel Model Parameters

As a consequence, it was found that a mixture of two Poisson processes provided the best fitting for distributions of the mean ray arrival times in all rooms and NLOS cases, thus the modified clustering channel model parameters for these two cases were additionally calculated as presented in Table 4-1.

Table 4-1 presents the comparison of determined channel model parameters with Cramer’s results [Cra02] and with those presented by Chong et al. [Cho05b], Saleh et al. [Sal87] and Spencer et al. [Spe00]. Although a visual inspection of [Cho03a] possibly gives rise to unlike distributions of multipath cluster parameters and samples of cluster distributions evaluated in this study are less than those in Cramer’s results, clustering parameters compare relatively well to those obtained by Cramer. With regard to parameters obtained from all data, estimated clustering parameters are very similar as those of Cramer’s results. Furthermore, both results clearly show that \( \lambda >> \Lambda \). This is consistent with the typical characteristic of MPCs where there are several ray arrivals in each cluster.

However, some of the results suggest a gap for \( \gamma \) between this work and Cramer’s results (55.59ns and 84.1ns respectively). This is because the different process used for selecting the multipath clusters; therefore, the numbers of clusters, cluster arrival times, rays and ray arrival times were not exactly identical. This also resulted in a slight difference in \( 1/\lambda \) parameters, 39.17ns and 45.5ns respectively. Furthermore, as far as \( \Gamma_{\text{mean}} \) and \( \gamma_{\text{mean}} \) of the LOS case are concerned, a cluster becomes like a single ray; therefore two decay factors should have low values, \( \Gamma_{\text{mean}} = 5.573 \) ns and \( \gamma_{\text{mean}} = 2.667 \) ns.
4.6 Spatial Correlation Analysis on Propagation Channel Conditions

According to spatial correlation computation process and results as previously mentioned in Section 4.2.2, some noticeable characteristics are drawn here. In regard to the temporal domain, the first arriving signals in the first bins are coming from the direct path, thus signal phases do not relatively change due to clutter; therefore the highest values of correlation coefficients can be found. On the other hand, correlation coefficients of subsequent MPCs, which are taken into account from the second bin, consistently decay with increasing excess delays; however traces of high correlation values can also be observed at some of the subsequent bins. This is also the reason for less numbers of estimated parameters leading to TOAs and AOAs computed from the proposed technique are rarely obtained despite arriving of MPCs at later time delays.

![Diagram showing spatial correlation coefficients](image_url)

**Figure 4-11:** Numbers of spatial correlation coefficients calculated for all array positions in each direction (column, row, upward and downward directions).

When considering distributions of numbers of all spatial correlation coefficients in all directions with time delays, some time delays are clearly observed and identified as TOAs of multipath clusters as presented in Section 4.4. Variations of spatial correlation coefficients computed between each sensor and neighbouring ones in row, column, upward and downward...
diagonal directions express the same decreasing trend as illustrated in Figure 4-11 where a normalized distribution of significantly correlated values over time delay was plotted. Furthermore when employing the least-squares fitting of these distributions, variations of decrease in rate of change of correlation decreasing rates related to distances and scenarios of propagation paths could be determined.

Slopes ‘m’ of linear fitting lines present correlation decreasing rates, or decorrelation rate, of multipath signals detected at each scenario. Since the decreasing trends of correlation values in most scenario cases are nearly similar to others, in order to present the results clearly, only results of LOS1 and NLOS2 are plotted in this figure. From Figure 4-11, it can be noted that the decorrelation rate of multipath signals in LOS1 scenario is slower than in NLOS2 scenarios, corresponding to the lower decreasing rate of \( m = 0.0059 \text{ ns}^{-1} \) in LOS1 case (D=9.5 m.) and the higher decreasing rate of \( m = 0.0091 \text{ ns}^{-1} \) in NLOS2 case (D=13.5m.). This is due to obstructions and propagation through walls for the latter case that caused random fluctuation in received PDPs. For other propagation cases, their correlation decreasing rates are depicted in the next figure where decreasing rates of correlation are assumed to vary depending on distances between the transmitter and the receiver.

Nevertheless, distances from the transmitter to receiving locations have a particular effect on spatial correlation coefficients, that is, the longer the distance the lower the correlation values and, also, the higher decorrelation rate. This assumption can be observed in Figure 4-12 and Figure 4-13, which represent the correlation decreasing rates, which were averaged over all directions, calculated from UWB signals received at various distances from the transmitter. Figure 4-12 (a) and Figure 4-12 (b) describe the variation of decorrelation rate of the data set of all spatial correlation values \((-1 \leq R_{i,k} \leq 1)\) and only high spatial correlation values \((R_{i,k} \geq 0.5)\) respectively. At the same distance (D=5.5 m.), the decorrelation rate in LOS is lower than the rate in NLOS. Comparing between these two figures, the lower decorrelation rate can be obtained in Figure 4-12 (b) due to its smaller ranges of correlation values. In addition, the same decreasing trend is also obtained when correlation decreasing rate being classified in each direction as shown in Figure 4-13 where \( R_{i,k} \geq 0.5 \) are taken into account.
Figure 4-12: Variation of rate of change of average correlation with distances
4.7 Conclusion

This study explored the use of spatial correlation based signal array processing for determining TOAs and AOAs of UWB multipath clusters in dense channels. Spatial correlation coefficients of each PDP bin calculated between all adjacent array points revealed that when considering averaged correlation values at each array sensor, especially in a LOS case, traces of a high degree of correlation could be observed along array positions. It was found that high correlation values exist at the same MPC time delays. These corresponded to all results either when considering average correlation values over each array point, or when considering only high correlation values. These time delays were considered as the arrival time of multipath clusters. Accordingly, when this computed arrival time was compared with the TOA estimated by the Sensor-CLEAN algorithm, a strong relation was found. However, some inconsistency could be found for TOA estimation especially at later time delays. This is due to the considerable limitation of high correlation coefficients that were obtained at most of the initial temporal period. Furthermore, the estimation of multipath clusters using the presented methodology was consistent.
with and resulted to similar characteristics of typical UWB cluster modelling parameters obtained by previous work.

Finally, the spatial correlation characteristics related to scenarios and distances of propagation channels could be analysed and discussed. The correlation rates of multipath clusters received at NLOS locations decreased more rapidly than ones received at LOS locations. Moreover, the rate of decreasing correlation gradually increased with the increasing of distances between the transmitter and the array sensors. The methodology and results described in this chapter are published in [Mak06], [Mak07a], [Mak07b]. A new method to estimate AOA will be considered in the next chapter.
Chapter 5

5 Complex Correlation Analysis based Linear Array AOA Estimation Measurement

According to the results in Chapter 4, although spatial correlation between each array and its surrounding arrays could gain TOA of multipath signals, results of AOA estimation were not included. Therefore, other algorithms should be cooperatively considered to investigate AOAs. Regarding information of TOA obtained from correlation analysis, spatial correlation of received signal between each considering array is assumed to be the important role for determining AOAs as firstly proposed by Kieburz [Kie67] and Lee [Lee94]. However, these researchers reported the method related to complex correlation analysis which was not considered in the previous chapter. Only the magnitude of UWB database was obtained and was used in post-processing. Consequently, complex signal measurement data should be exploited in order to gain more phase information possibly leading to AOA information. This will be presented in this chapter.

Moreover, in this research, UWB linear array measurements are used instead of planar array systems in order to simply clarify the proposed AOA estimation technique. Remarkably, since the objective of this chapter is to determine AOAs using another high accuracy method, TOA estimation is not repeatedly investigated. Only the methodology of AOA estimation and its evaluation will be described.

5.1 Introduction

There are some differences between the array signal processing techniques of narrowband and UWB signals. In general, narrowband signals are assumed to arrive at all sensors and to maintain constant amplitudes during the observation period. On the other hand, there is no exact assumption that UWB signals can arrive at all arrays at the same time. Their amplitudes cannot be maintained constantly during the interaction period at all sensors. Despite this, there are various works investigating UWB signals in the time domain by dividing the temporal axis of a PDP into a small time bin related to the time resolution of signals [Has93], [Cas02], [Cho05b]. However, insignificant correlation coefficients between each time bin were obtained without any further
information for practical usage. This is possibly due to variation of differential averaged PDP bins caused by multiple diffracted and reflected contributions resulting in more randomness and smaller correlation. Those characteristics are also caused by the assumption that UWB multipaths are not well modelled as identical in phase-shifts [Cra00].

Some works reported the multipath effects on array system performances by using spatial correlation analysis [Kyr03], [Kyr00]. The performance of antenna arrays depends on spatial correlation between each element which its correlation values are based on the angular energy distribution [Sal94], [Tsa02]. In addition, Loyka et al [Loy02], [Zha04a] presented the cluster model of incoming multipath signals arriving at linear antenna arrays and spatial correlation coefficients between each array element. Spatial correlation equations were formulated corresponding to the angular spread of the incoming multipaths. Analytical results shown that correlation values significantly depended on power distributions of multipath clusters. Since multipath components arriving in different clusters were assumed to be uncorrelated, a time series of correlation values was presented as a series of cluster terms.

As a result, papers [Pre02], [Kar98], [Cho03a], [Cho03b] reported the essential results of characterising the spatial correlation of UWB signals. Multipath signals scattering around antenna arrays were examined. Additionally, the previous chapter suggested the modified spatial correlation technique calculated between each sensor array and its surrounding sensors at each time bin [Mak06]. Results from this research presented the significant coefficients which could be used to determine TOAs of MPCs [Mak07a]. Those UWB parameters obtained from this technique were consistent with the standard IEEE UWB characterisation [Mol04]. Consequently, in this work, spatial correlation at each time bin is mainly considered as representation of individual contributions of clusters between each antenna. However, because there is no carrier involved in UWB transmission systems, giving rise to no reference source, some research reported that the phase information is not an important parameter in the UWB channel model [Opp04]. Therefore, phase statistics are rarely used in UWB channel models and there is not utilisation of phase information in data post-processing.

Nevertheless, some research presented interesting results of exploiting phase data to approach direction of arrival or AOA of incoming signals. Kyritsi et al. presented the correlation analysis on MIMO channels to show that clustering of the complex correlation coefficients represented constant phase terms of received signals within propagation environments [Kyr03]. Alternatively, Haneda et al. considered phase differences associated with the wavefront and frequency sub-band models to determine the directional information [Han06a]. However, the measurement conducted in this work was limited by detecting the single wave in only a LOS with the fixed directions of signal arrivals. Also, spatial correlation to present cluster distributions was not included in this work.

This chapter, therefore, proposes the utilisation of phase parameters for investigating
time-variant UWB multipath arrivals. Nonetheless, phase information studied in this work is not directly analysed from the measurement system but it is a significant part originated from complex correlation calculations. Considering the cluster model and UWB pulse dispersion due to distortion effects, the modified spatial correlation technique will be implemented at each multipath bin by performing the complex correlation computation. Thus, both amplitude and phase information are taken into account regarding time and spatial domains. Closed-form expression of correlation coefficients derived from a cluster model will be derived. Apart from gaining the clustering of a complex correlation distribution as described in [Kyr03], the relative phase difference method is co-processed in this work in order to determine AOAs at each individual observation time. To verify this technique performs well in UWB channel sounding, the detected signals measured in the chamber room are analysed concerning the realistic environment. Thus, data computed in this work consists of wave incidents from both LOS and multipath reflections from various directions; leading to the reliability and numbers of received signal samples to be used in the computational process.

5.2 Spatial Cluster Model and Linear Array Phase Difference Method

5.2.1 Spatial Cluster Model for Uniform Linear Array

To derive the fading correlation function of uniform linear arrays, it is necessary to consider the vector channel model that its analytical expressions are derived applicably to any number of elements. The geometry of uniform linear array is presented in Figure 5-1(a) where $\phi$ is the azimuth angle of the incoming signal. It is assumed that various independent multipath signals arriving at the array within $\pm \Delta$ of the reference azimuth AOA, $\phi_{\text{ref}}$, with the Laplacian distribution of AOA probability density function. This reference AOA is determined by geometry position of a dominant cluster regarding the transmitter and the array receiver positions. Consequently, the array manifold vector $V(\phi)$ for a $N$-uniform linear array, where $N=5$, can be expressed as

$$V(\phi) = \begin{bmatrix} V_1(\phi) \\ V_2(\phi) \\ V_3(\phi) \\ V_4(\phi) \\ V_5(\phi) \end{bmatrix} = \begin{bmatrix} \exp(-j2\pi(d_{1,\text{ref}})/\lambda \sin \theta \cos(\phi - \phi_1)) \\ \exp(-j2\pi(d_{2,\text{ref}})/\lambda \sin \theta \cos(\phi - \phi_2)) \\ \exp(-j2\pi(d_{3,\text{ref}})/\lambda \sin \theta \cos(\phi - \phi_3)) \\ \exp(-j2\pi(d_{4,\text{ref}})/\lambda \sin \theta \cos(\phi - \phi_4)) \\ \exp(-j2\pi(d_{5,\text{ref}})/\lambda \sin \theta \cos(\phi - \phi_5)) \end{bmatrix} \quad (5-1)$$

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\(d_{n, ref}\) is the gap distance between each \(n^{th}\) antenna and the reference antenna. \(\phi_n\) is the azimuth AOA of the \(n^{th}\) antenna. For any element, the array spatial correlation between the \(m^{th}\) and \(n^{th}\) antennas can be derived by (5-2).

\[
\rho_{n,m} = \mathbb{E}\{V_n(\phi)V_m(\phi)\} = \int_{\phi} V_n(\phi)V_m(\phi)^*p(\phi)d\phi
\]  

(5-2)

The superscript * denotes the complex conjugate. \(p(\phi)\) is the Laplacian AOA distribution,
\(p(\phi)=1/(\sqrt{2}\sigma_\phi)\exp(-\sqrt{2}/\phi/\sigma_\phi)\) where \(\sigma_\phi\) is the angular spread of \(\phi\) and \(\zeta - \Delta \leq \phi \leq \zeta + \Delta\) [Tsa02], [Zha04a]. For simplicity of spatial correlation model, elevation AOAs at all elements are assumed to be equal. Thus, (5-2) can be defined by

\[
\rho_{n,m} = \frac{1}{\sqrt{2}\sigma_\phi} \int_{-\Delta}^{+\Delta} \exp\left(-\frac{j2\pi(n-m)d\sin\theta}{\lambda}\left\{\cos(\phi-\phi_n) - \cos(\phi-\phi_m)\right\}\right) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right)d\phi
\]

\[
= \frac{1}{\sqrt{2}\sigma_\phi} \int_{-\Delta}^{+\Delta} \exp\left(-\frac{j2\pi(n-m)d\sin\theta}{\lambda}\left\{\cos\phi\cos\phi_n + \sin\phi\sin\phi_n\right\}\right) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right)d\phi
\]

\[
= \frac{1}{\sqrt{2}\sigma_\phi} \int_{-\Delta}^{+\Delta} \exp\left(-\frac{j2\pi(n-m)d\sin\theta}{\lambda}\left\{\frac{\cos\phi_n - \cos\phi_m}{\kappa_1}\cos\phi + \sin\phi\sin\phi_n\right\}\right) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right)d\phi
\]

\[
= \frac{1}{\sqrt{2}\sigma_\phi} \int_{-\Delta}^{+\Delta} \exp\left(-\frac{j2\pi(n-m)d\sin\theta}{\lambda}\left\{\kappa_1\cos\phi + \kappa_2\sin\phi\right\}\right) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right)d\phi
\]

(5-3)

Give \(\sin\alpha = \kappa_1/\sqrt{\kappa_1^2 + \kappa_2^2}\), \(\cos\alpha = \kappa_2/\sqrt{\kappa_1^2 + \kappa_2^2}\), \(Z = \sqrt{\kappa_1^2 + \kappa_2^2}\) and \(\beta = 2\pi(n-m)d/\lambda\sin\theta\), the exponential term of (5-3) becomes \(\exp(-j\beta Z(\sin\alpha\cos\phi + \cos\alpha\sin\phi)) = \exp(-j\beta Z(\sin(\alpha + \phi)))\). Consequently, using the first-kind Bessel function, the real and imaginary parts of the spatial correlation can be described as follows:
Complex Correlation Analysis based Linear Array AOA Estimation Measurement

\[
\text{Re}\{\rho_{s,n,m}\} = \frac{1}{\sqrt{2}\sigma_\phi} \int_{-A}^{+A} \cos(\beta Z \sin(\alpha + \phi)) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right) d\phi
\]
\[\quad = \frac{1}{\sqrt{2}\sigma_\phi} \int_{-A}^{+A} \left\{ J_0(\beta Z) + 2 \sum_{k=1}^{\infty} J_{2k}(\beta Z) \cos(2k(\alpha + \phi)) \right\} \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right) d\phi
\]
\[\quad = 2 \left\{ J_0(\beta Z) + 2 \sum_{k=1}^{\infty} \frac{2/\sigma_\phi^2}{2/\sigma_\phi^2 + 4k^2} \cdot J_{2k}(\beta Z) \cos(2k(\alpha + \phi)) \right\}
\]
\[
\text{Im}\{\rho_{s,n,m}\} = \frac{1}{\sqrt{2}\sigma_\phi} \int_{-A}^{+A} \sin(\beta Z \sin(\alpha + \phi)) \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right) d\phi
\]
\[\quad = \frac{1}{\sqrt{2}\sigma_\phi} \int_{-A}^{+A} \left\{ 2 \sum_{k=1}^{\infty} J_{2k+1}(\beta Z) \sin((2k+1)(\alpha + \phi)) \right\} \exp\left(-\frac{\sqrt{2}\phi}{\sigma_\phi}\right) d\phi
\]
\[\quad = \frac{4}{\sqrt{2}\sigma_\phi} \sum_{k=1}^{\infty} \frac{\sqrt{2/\sigma_\phi^2} \left(1 + \exp(-\sqrt{2}\pi/\sigma_\phi)\right)}{2/\sigma_\phi^2 + (2k+1)^2} J_{2k+1}(\beta Z) \sin((2k+1)(\alpha + \phi))
\]

Both terms of the real and imaginary parts of spatial correlation between the \(n^{th}\) and \(m^{th}\) elements will be taken into account for relative phase difference method, where the implication of spatial correlation between any array and its reference (or the centre array), \(\rho_{s_n,ref}\), will be described in the following sections.

\[\text{Figure 5-1: Geometry of multipath arrivals to the uniform linear array antenna} \]

(a) Configuration of incoming multipath signals within \(\pm A\) of \(\phi_{ref}\)

(b) Phase difference mechanism for 5 uniform linear array

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5.2.2 Linear Array Phase Difference Method

Considering a regularly-spaced linear array of sensors with a plane wave incident to a broadside, an azimuth angle can be determined. If a planar array was used, both azimuth angle, $\phi$, and elevation angle, $\theta$, of the arriving signals can be resolved. However, in order to clarify the application of a phase difference method on AOA estimation, only one-dimensional-linear antenna arrays in azimuthal plane, $xy$-plane, are considered in this paper as depicted in Figure 5-1(b). The phase difference mechanism is shown for 5 uniform linear arrays with the centre reference element at the 3rd sensor and $dx$ is the gap distance between each element. The wave front of the signal is incident at an elevation angle $\theta_0$ and azimuth angle $\phi_0$. Generally, signal components between antenna arrays are correlated; only differing in the relative phases between each element can be observed either within a small gap or a $\lambda/2$ gap distance between each element. Information of the relative phase differences between two array elements on a horizontal baseline can define the principal azimuthal AOA of the incident signal. The incoming direction can be easily specified when the transmitting signal consists of a single plane wave and arrives within the main lobe of the array pattern.

When transmitted signals are propagating from a long-distance-source, the paths of each antenna element are virtually parallel with each other, thus the difference in path length between each element is determined by the AOAs of the incident signals. Figure 5-2 presents the MPCs detected at each element of five linear antenna arrays. Signals arrive at the 5th array first, then the 4th, 3rd, 2nd and 1st arrays respectively. The difference in time delay of arriving signals detected at each sensor is measured in nanoseconds. These time delay propagation differences are related to the phase difference between each array and its reference, the centre element. The relative phase difference between these partial arrays (the 2nd and the 4th sensors) and the reference array (the 3rd sensor) can be given by (5-6) where $t_2$, $t_3$, $t_4$ or $t_{\text{ref}}$ are TOAs of signals at each sensor respectively.

$$\Delta \Phi_{2,\text{ref}} = \frac{2\pi}{\lambda} c(t_{\text{ref}} - t_2)$$
$$\Delta \Phi_{4,\text{ref}} = \frac{2\pi}{\lambda} c(t_4 - t_{\text{ref}})$$

(5-6)

$\Delta \Phi_{n,\text{ref}}$ is the relative phase difference between the $n^{th}$ element and the reference element. $c=3\times10^8$ m/s and $\lambda$ is the centre frequency wavelength of transmitted signals. According to the geometrical propagation diagram of linear arrays as depicted in Figure 5-1(b), the excess distance relative to the reference array, $c(t_{\text{ref}}-t_{\text{ref}})$, can be described by $d_x \sin \phi \sin \theta$. Since the array baseline is located in the broadside direction of the incident plane wave where $\theta = 90^\circ$, therefore only an
azimuthal AOA, $\phi$, is taken into account for the relative displacement between two elements. Assuming that $\theta = 90^\circ$, $\sin \theta = 1$, (5-6) can be simplified by (5-7).

$$\Delta \Phi_{n,\text{ref}} = \frac{2\pi}{\lambda} d_x \sin \phi$$

Thus, the azimuthal AOA of the incident signal can be calculated based on the trigonometry configuration as given by

$$\phi = \sin^{-1} \left( \frac{\lambda}{2\pi d_x} \Delta \Phi_{n,\text{ref}} \right)$$

Figure 5-2: Comparison of MPCs arriving at different sensors
5.3 Correlation Analysis and Its Application based Phase Difference Method

5.3.1 Complex Correlation Analysis

Although small scale characteristics of UWB signals do not vulnerably change between each sensor array, some difference exists in received PDP due to multiple diffracted and reflected contributions when considering a whole period of observation time. Furthermore, since UWB signals have frequency-dependent distortion by scatterers along propagation paths, these result in more randomness and smaller coefficients of spatial correlation between signals detected at different sensors and in low temporal correlation between each time bin over the observation period [Has93], [Cas02], [Cho05b].

The computation technique performed in this study is based on the resolvable time bins of UWB pulse signals by assuming that the directions of MPCs arriving at each sensor are different and independent of each delay bin. Hence it can be implied that the time variation of any one-bin-correlation coefficient can be clearly observed. Equation (5-9) describes the complex-form of spatial correlation calculated at each time bin from which the time variant information of magnitude and phase of UWB impulses can be determined. Considering (5-2)-(5-3), the mathematic general formula of complex correlation can be written by

\[
\rho_{n,\text{ref}}(k) = \frac{E[R_{n,k} \cdot R_{\text{ref},k}^*] - E[R_{n,k}] \cdot E[R_{\text{ref},k}^*]}{\sqrt{E[|R_{n,k}|^2] - |E[R_{n,k}]|^2} \cdot (E[|R_{\text{ref},k}|^2] - |E[R_{\text{ref},k}]|^2)}
\]

(5-9)

where \( R_{n,k} \) is a MPC signal received at the \( n^{th} \) receive antenna and \( R_{\text{ref},k} \) is a MPC signal received at the reference sensor. This formula computes the correlation coefficient at the \( k^{th} \) time bin denoting the asterisk (*) for the complex conjugate operation. \( \phi_{n,\text{ref},k} \) is the phase term of the correlation value.

5.3.2 AOA Estimation Using Phase Difference Method

This study presents the determination of instantaneous AOs from the instantaneous phase differences. The probability density of the relative phase difference, \( \Delta \phi_{n,\text{ref},k} \) between signals received at the \( n^{th} \) antenna and the reference antenna at \( k^{th} \) time bin can be calculated in the form of correlation coefficients as described in (5-10) [Kie67], [Lee94].
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\[ p(\Delta \Phi_{n, \text{ref}, k}) = \frac{1 - \left| \rho_{n, \text{ref}}(k) \right|^2}{2\pi} \left( \frac{(1 - \beta^2) \frac{1}{2} + \beta(\pi - \cos^{-1} \beta)}{(1 - \beta^2)^{3/2}} \right) \]  

(5-10)

where \( \beta = \text{Re}(\rho_{n, \text{ref}}(k)) \cos \Delta \Phi_{n, \text{ref}, k} + \text{Im}(\rho_{n, \text{ref}}(k)) \sin \Delta \Phi_{n, \text{ref}, k} \). Consequently, the mean angle of arrival relative to the incident plane wave normal to the array baseline is given by

\[ \sin \phi_{n, \text{ref}, k} = \frac{\lambda}{2\pi d_X} \tan^{-1} \left( \frac{\text{Im}(\rho_{n, \text{ref}}(k))}{\text{Re}(\rho_{n, \text{ref}}(k))} \right) = \frac{\lambda}{2\pi d_X} \Delta \Phi_{n, \text{ref}, k} \]  

(5-11)

Thus according to (5-9) and (5-11),

\[ \Delta \Phi_{n, \text{ref}, k} = \tan^{-1} \left( \frac{\text{Im}(\rho_{n, \text{ref}}(k))}{\text{Re}(\rho_{n, \text{ref}}(k))} \right) = \phi_{n, \text{ref}, k} \]  

(5-12)

Consequently, the AOAs at each \( k^{th} \) time bin can be estimated as follows where \( N \) is the number of array sensors and \( K \) is the total measuring bin numbers. The average value of AOAs extracted by all relative phase difference is

\[ \bar{\phi}_k = \left[ \frac{1}{N-1} \sum_{n=1}^{N-1} \phi_{n, \text{ref}, k} \right] \]  

(5-13)

5.4 UWB Measurement Systems and Computation Scheme

5.4.1 Measurement Campaign

The UWB measurement is performed in the frequency-domain using a VNA with 1601 sweeping points over the 3-6GHz frequency range. IF bandwidth is set to 1 kHz and the frequency step is 1.875MHz. This allows the maximum excess delay of 533ns and maximum distance about 160m to be measured with suitable dynamic range. This measurement campaign observes UWB propagation channels under the same environment condition with 21 orientations of the receiver antenna. 3-6GHz radio frequency is transmitted from the 1-10GHz double ridge waveguide horn antenna. At the receiving site, a 5-linear-array circular disc antenna (25mm. diameter) [Yan03] is located on a turntable which is controlled by HP VEE software in order to rotate the receiver
direction every 18° from 0°-360°.

The configuration of the receiving antenna array is depicted in Figure 5-3. Each monopole circular disc antenna has 2.5cm diameter with a λ/2 gap distance of 4.29cm. The construction of antenna arrays used for this measurement are characterised in Appendix B. Figure 5-4 illustrates the measuring propagation paths conducted in a chamber room. Two positions of a transmitter are located, the centre position (Tx I) and the corner position (Tx II). The transmitting antenna at Tx II is located directionally to the location of an aluminum foil board. In order to synthesise real indoor building environments, reflections from a foil board and a floor are taken into account. Consequently, two sets of measurement are taken based on the location of the transmitter which includes reflections from scatterers created. Each set contains 21 orientations of the receiver arrays for analysis. Several samples of measured signals obtained from the measurement system can examine the accuracy of the phase difference method. This will be described in Section 5.5.
Furthermore, since UWB channel sounding is operated by VNA frequency domain measurements, perfect time domain information cannot be obtained by a simple IFFT. Moreover, frequency dependent effects on hardware components are also included in the measured data leading to signal dispersions. In order to reduce possible inaccuracies occurring due to this insufficient stability of FD measurement, extraction of UWB real time signals from the frequency domain requires implementation of the calibration process. The channel transfer function should be isolated from all hardware effects by measurement calibration process. The time reference points are moved from the VNA ports to the cable ends or at the antenna connectors during calibrating, hence, only the propagation delays are included in the delay profiles. Since the antenna element size is very small, the delay due to antenna itself is considered insignificant. Then, all measurement signals are scaled with these measured calibration data for data analysis and post-processing.

Figure 5-5 (a) and (b) depict the regenerated reference signal with the transmitter located at the middle of the chamber room, Tx I, and at the corner of the room, Tx II. The LOS signals are detected by the centre array with 0° azimuth angle and 90° elevation angle. Accordingly, reference propagation delays of approximately 10.73 ns and 9.9 ns can be estimated respectively as a consequence of Tx I-to-Rx distance = 3.22 m. and Tx II-to-Rx distance = 2.97 m. Time bin resolution of approximately 1 ns contains a whole significant multipath. This bin width will be employed in the post processing and computation technique. Moreover, reference propagation delays measured by all 5 sensors are shown in Figure 5-6. These reference LOS propagation delays are defined for each antenna at 21 orientations from 0°~360° in the horizontal plane. All values of propagation delays are used as the TOAs of the first multipath component. In figure (a) receiving from Tx I, the figure apparently shows that variations of propagation delays of MPCs arriving at each sensor are maximum at 0° orientation (broadside direction to the array baseline) relative to the reference propagation delay, and the variations are minimum at ±90° orientation. These two latter orientations present the endfire directions of incident multipath signals leading to no propagation delay which can be determined due to non-extent or weak signals. Alternatively, it might be implied that no signal higher than the threshold level is received at the endfire directions. On the other hand, when transmitting from Tx II at the corner of the chamber is considered, the broadside and endfire directions of the incident plane wave based on the array receiver are at -18° orientation and -108° / 72° orientation respectively as described in figure (b).
Figure 5-5: Regenerated signals and specification of a temporal bin for post-processing

Figure 5-6: Reference LOS propagation delays for all 21 orientations
5.4.2 Post Processing and Computation Technique

Firstly, to extract the channel parameters of interest from the frequency-domain measurement data, the cable losses and other factors are calibrated out, so only the path loss is measured. Therefore, the hardware effects of the measurement system, such as from amplifiers, cables and antennas, are excluded from the original measured data. Then the frequency-domain CTF are windowed by a Hamming function and transformed to the time-domain CIR using the real passband IFT. The CIRs, which have PDPS that exceed the threshold level, 6dB above the average noise floor level ≈ -25dB, are extracted to be used for the complex correlation in the next process. CIRs are divided into small bins with a 1ns temporal width. As the computational interval of each bin, the unwanted noise spike is eliminated, hence only the effective multipaths are analysed. Despite more than one path arriving at each bin, the content of each 1-ns-bin is referred to as one path. This bin width compromises with the high temporal resolution of the received signals (approximately 1ns pulse duration); as presented in Figure 5-5. In the computation process, the complex form spatial correlation coefficients are calculated to take into account all information of both magnitude and phase at each array sensor. Regarding the signal received at the reference 3rd sensor at the kth time bin, the matrix of all spatial correlation coefficients, $R_{n,ref,k}$ can be given by (5-14).

$$R_{n,ref,k} = \begin{bmatrix} \rho_{1,ref}(k) & e^{j\rho_{2,ref,k}} & e^{j\rho_{3,ref,k}} & \cdots \\ \rho_{2,ref}(k) & e^{j\rho_{3,ref,k}} & e^{j\rho_{4,ref,k}} & \cdots \\ \rho_{3,ref}(k) & e^{j\rho_{4,ref,k}} & e^{j\rho_{5,ref,k}} & \cdots \\ \vdots & \vdots & \vdots & \ddots \end{bmatrix}$$  (5-14)

From (5-8) and (5-11), the AOAs related to the reference array can be obtained by

$$\phi_{n,ref,k} = \sin^{-1}\left[ \frac{\lambda\phi_{1,ref,k}}{4\pi d_x} \frac{\lambda\phi_{2,ref,k}}{2\pi d_x} \frac{\lambda\phi_{3,ref,k}}{2\pi(0)} \frac{\lambda\phi_{4,ref,k}}{2\pi d_x} \frac{\lambda\phi_{5,ref,k}}{4\pi d_x} \right]$$  (5-15)

Since this method is computed relative to the reference antenna, $\phi_{3,ref,k}$ of the third term cannot be determined, as $d_x=0$. In consequence, only 4 values of estimated $\phi_{n,ref,k}$ are considered to obtain the average AOA as shown in (5-16).

$$\overline{\phi}_k = \left[ \frac{1}{4} \sum_{n=1}^{4} \phi_{n,ref,k} \right]$$  (5-16)
Based on the linear array configuration, the relative phase difference method for $I \times J$ planar arrays can be written by (5-17) and (5-18).

$$\bar{\phi}_k = \left[ \frac{1}{I \times J} \bar{\phi}_k \right]$$  \hfill (5-17)

$$\bar{\phi}_k = \sin^{-1} \left[ \frac{\lambda \phi_k}{2\pi \bar{d}^T} \right]$$  \hfill (5-18)

where at any time $k^{th}$, the phase terms of spatial correlation between each sensor and the reference array, $\phi_k = \tan^{-1} \left( \frac{\text{Im}(\rho(i,j)_{\text{ref}}(k))}{\text{Re}(\rho(i,j)_{\text{ref}}(k))} \right)$: $i = 1:I$ and $j = 1:J$, can be obtained by

$$\phi_k = \begin{bmatrix} \phi_{(1,1),\text{ref}} & \phi_{(1,2),\text{ref}} & \phi_{(1,3),\text{ref}} & \ldots & \phi_{(1,J),\text{ref}} \\ \phi_{(2,1),\text{ref}} & \phi_{(2,2),\text{ref}} & \phi_{(2,3),\text{ref}} & \ldots & \phi_{(2,J),\text{ref}} \\ \phi_{(3,1),\text{ref}} & \phi_{(3,2),\text{ref}} & \phi_{(3,3),\text{ref}} & \ldots & \phi_{(3,J),\text{ref}} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \phi_{(I,1),\text{ref}} & \phi_{(I,2),\text{ref}} & \phi_{(I,3),\text{ref}} & \ldots & \phi_{(I,J),\text{ref}} \end{bmatrix}$$  \hfill (5-19)

$$\bar{d} = \begin{bmatrix} \bar{d}_{(1,1),\text{ref}} & \bar{d}_{(1,2),\text{ref}} & \bar{d}_{(1,3),\text{ref}} & \ldots & \bar{d}_{(1,J),\text{ref}} \\ \bar{d}_{(2,1),\text{ref}} & \bar{d}_{(2,2),\text{ref}} & \bar{d}_{(2,3),\text{ref}} & \ldots & \bar{d}_{(2,J),\text{ref}} \\ \bar{d}_{(3,1),\text{ref}} & \bar{d}_{(3,2),\text{ref}} & \bar{d}_{(3,3),\text{ref}} & \ldots & \bar{d}_{(3,J),\text{ref}} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ \bar{d}_{(I,1),\text{ref}} & \bar{d}_{(I,2),\text{ref}} & \bar{d}_{(I,3),\text{ref}} & \ldots & \bar{d}_{(I,J),\text{ref}} \end{bmatrix}$$  \hfill (5-20)
where $d_{ij,ref}$ is the gap distance between the $(i,j)$ element and the reference array. According to (5-17)-(5-20), the computation for symmetric $I\times J$ antenna arrays is taken into account relatively to the reference array $((i+1)/2,(j+1)/2)$, the centre element.

Because the objective of this study is to introduce the AOA estimation method based on spatial correlation analysis, the simple linear array antenna measurement and the linear-array relative phase difference method are implemented. However, if AOA estimation method is analysed using another perspective which determines phase difference of each array relatively to its surrounding array elements, new results can be obtained. Thus, alternatively, complex correlation coefficients and relative phase difference are not computed regarding only the centre element of the array, in contrast, all elements can be employed as the reference to be correlated with their surrounding elements. This computation is based on previous publications of the author in [Mak06], [Mak07a] as described in (4-3)-(4-6) which consider its complex computation regarding to the spatial correlation, $R_{ij,k}$ between surrounding array elements. All subsets of complex spatial correlation are calculated with different reference values depending on the element of interest. When computing spatial correlation on linear array measured data in this different way (between each array element and its adjacent positions), the results of averaging correlation values among surrounding $(i\pm m,j\pm n)^{th}$ positions, $R_{ij,k}$, can gain consistence with using the original relative phase method, $R_{\text{rel}}$. This is caused by configuration of 5 linear arrays operated in this measurement; spatial correlation between only two adjacent arrays are computed leading to $R_{ij,k} \approx R_{\text{rel}}$ as depicted in Figure 5-7.

Figure 5-7 shows complex correlation results of received UWB signals propagating from Tx 1 with LOS components and multipath reflections from a foil board and a floor. Signals from all 21 array orientations are measured and calculated at the first bin, $k=1$, regarding the reference array (Figure 5-7 (a)), $R_{\text{rel}}=[\rho_{1,3} \rho_{2,3} \rho_{3,3} \rho_{5,3}]$. Figure 5-7 (b) presents complex correlation results calculated between each array and its surroundings arrays, $R_{ij,k}=[\rho_{1,2} \rho_{2,3} \rho_{3,4} \rho_{5,4}]$. It can be seen that two out of four complex correlation results of both sets are the same in particular $\rho_{2,3}$ and $\rho_{3,3}$ (or $\rho_{3,4}$), and some consistent results can be observed for $\rho_{1,3}$, $\rho_{5,3}$ and $\rho_{1,2}$, $\rho_{4,5}$. As a consequence, at each orientation, some consistent results of phase terms $\phi_{k=1}$ estimated from these two methods can be obtained. However, some significant differences can be seen when considering more elements of antenna arrays.

Additionally, planar array measurement can gain more significant results which can be further investigated due to increasing numbers of correlation results. This is because the planar array computation contains correlation values between each array and its more surrounding sensors, not just only side-by-side sensors in the broadside direction as considered in the linear array antenna. Nevertheless, apart from conducting planar array measurements, data processing
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becomes more complicated since there are various terms of complex correlation (magnitude and phase) to be considered corresponding to the reference element in each group of surrounding arrays. Further information of using this computation technique for AOA estimation is proposed in Chapter 7 as a future work.

Figure 5-7: Comparison of complex spatial correlation relative to the centre array reference and relative to the adjacent sensors
5.5 Results and Discussion

5.5.1 Experimental and Computational Results

According to the measurement systems, classified by 2 locations of the transmitter, Tx I and Tx II, two groups of results were analysed and presented in this section. Furthermore, since the receiver array was rotated on a turntable every 18° from 0° to 360°, 21 propagation paths were analysed for each scenario. Thus, this experimental process can generate various samples of multipath signal arrivals in all directions. An orientation diagram of multipath arrivals in the measurement system is illustrated in Figure 5-8, where Tx I, Tx II and the aluminium foil board are located at 0°, -18° and 22° relative to the normal respectively. Figures of the measurement in real environment are presented in Figure B-4 in Appendix B. Reflection from an aluminium foil board and a floor was provided in order to construct realistic indoor propagation environments. All 3 multipath scenarios, i.e. reflection from a foil board and a floor, reflection from a foil board only, and reflection from a floor only, were carried out for both transmitter locations. With the 21 orientations, 126 propagation paths were measured and computed using the proposed phase difference method for determining AOAs.
Figure 5-9 presents regenerated transformed time-domain signals received by the reference sensor at 0° orientation measured by the 3rd array antenna including all path conditions (LOS, reflection from a foil board and reflection from a floor). Figure 5-9 (a)–(c) presents received MPCs propagating from Tx I, and Figure 5-9 (d)–(f) presents received MPCs propagating from Tx II. The strongest signals can be seen in all figures representing the LOS signals followed by MPCs reflecting from a foil board, as shown in figure (a), (c), (d) and (f). The weakest signals observed in figure (b), (e) and (f) describe MPCs reflecting from a floor. For transmitting signals from Tx II, it is evident that MPCs reflecting from a floor are negligibly lower than the cut-off threshold. Therefore, only LOS and foil-reflecting-MPCs will be taken into account in a computation process for this propagation case. Regenerated time-domain signals are divided into a small time bin with 1ns temporal width. This time bin contains the signal with which its PDP is higher than -25dB. According to Figure 5-9, time periods of LOS and MPC signals scattering from all obstacles are approximately 5ns, hence 5 time bins are considered in the phase difference computation process as illustrated.
Examples of complex correlation coefficients of MPCs received from Tx I at 0° orientation are presented in Figure 5-10 where the magnitude $|\rho_{\text{ref}(k)}|$ and phase angle $\phi_{k}$ are shown. Results of complex correlation calculated at each time bin between each sensor and the 3rd sensor are plotted. High correlation coefficients with constant phases are clearly observed in the first bin corresponding to the arrival of LOS component. There are some scattering complex correlation coefficients in the latter two bins, but most coefficients are located around 330° or -30° from the reference orientation. These two bins are probably dominated by reflection from the foil board multipaths. MPCs of reflection from a floor can be presented in the 4th and the 5th multipath bins where most of the magnitude of high complex correlation and phase angles are located at 0°, especially at the 4th bin. In Figure 5-10 (f), some partial MPCs arriving at the 6th time bin are ignored due to their less than -25 dB PDPs and low coefficient values ($|\rho_{\text{ref}(k)}| \leq 0.5$). Hence, there exists a correlation value, $\rho_{13}$, only between the first sensor and the reference one in this time bin.
Figure 5-11: MPC estimation using the phase difference method
Applying relative phase difference on all orientations of the array receiver and on all 6 scenario cases, azimuthal AOAs, $\phi_k$, can be obtained as presented in Figure 5-11. Results display clusters of LOS and multipaths received at 0° orientation. On account of variance of estimated parameters due to several phase difference results from all sensors with 21 orientations, two minimum and maximum computed values are removed from the phase-data interests, $\phi_k$. In Figure 5-11, examples of the clusters of the TOAs and AOAs of LOS and multipath arrivals (reflecting multipaths from a foil board and a floor) measured at 0° orientation are depicted. Figure 5-11 (a) and (c) shows the determined TOAs and AOAs, $\phi_k$, for the transmitter located at Tx I. LOS components are detected between 10.53–11.10 ns with azimuthal AOAs varying from $-0.35^\circ$ to $0.87^\circ$. Its standard deviation is $\sigma_{\phi,LOS,Tx I} = 0.3675$. Estimated TOAs and AOAs of MPCs reflected from a foil board can be seen from 11.61–13.21 ns with AOAs of $-3.23^\circ$ to $5.11^\circ$. The standard deviation of dominant angles close to the mean AOA of this case, $\sigma_{\phi,foil,Tx I} = 2.82$ is less than one of the foil board reflecting case, $\sigma_{\phi,foil,Tx II} = 4.28$. Alternatively, when considering propagation channels from Tx II as presented in Figure 5-11 (b) and (d), AOAs of LOS components are clearly observed in ranges of 9.72–10.25 ns and $-21.12^\circ$ to $-20.42^\circ$. The standard deviation of estimated AOAs for LOS components is $\sigma_{\phi,LOS,Tx II} = 0.381$, and the standard deviation for a foil-board reflection is $\sigma_{\phi,foil,Tx II} = 3.52$ less than Tx I case since the main beam of the transmitter at Tx II is in the direction toward a foil board. Nonetheless, the AOAs of multipaths reflected from a floor are rarely obtained due to their low signal strengths. This also leads to some disappearances of AOA results in some environment cases.

### 5.5.2 Evaluation of the Phase Difference Method

Apart from the several data sets from repeated measurements of all scenario conditions as mentioned in Section 5.5.1, the accuracy of estimated multipath parameters computed by the relative phase difference method can be evaluated by using the sensor CLEAN algorithm [Cra00], [Cra02]. Essentially, since the analysis of this research relies on individual computation at each multipath bin, the CLEAN algorithm is suitable for comparing between results obtained by the proposed phase difference method. As a result of the pattern of a regenerated time-domain pulse signal as depicted in Figure 5-5, the 14th derivative of a Gaussian pulse is used in the CLEAN algorithm differently from the commonly-used second derivative one [Cra00], [Ben04].

Results compared between using the phase difference method and the sensor CLEAN algorithm are illustrated in Figure 5-12 where the sample case of 180° orientation is analysed. In this figure, although there are different units between these two methods, only TOAs and AOAs are considered for the comparison. Firstly, transmitting from Tx I case in figure (a) and (b), estimated
AOAs of LOS and reflection from a floor are located approximately at 180° relative to the direction of the receiver. Moreover, AOAs of multipaths reflected by a foil board can be observed at around 220°–230°. Figure 5-12 (c) and (d) present results when transmitting from Tx II. Most LOS components can be observed at 165°, and MPCs reflecting from a foil board can be seen at around 200°. Multipath clusters using the phase difference method (figure (a) and (c)) are more clearly observed than multipath clusters estimated by the CLEAN algorithm (figure (b) and (d)). Additionally, it can be notably seen that there are more dense estimated AOA results using the proposed phase difference method than results using the CLEAN algorithm. This is caused by individual computation in each temporal multipath bin in the relative phase difference process, hence all phase information of all multipath bins are determined. In contrast, $\theta=90°$ elevation angle is specified in the CLEAN algorithm process to estimate the azimuth angle; therefore TOAs and azimuthal AOAs of possible multipaths arriving with only 90° elevation angle could be obtained. This is different from the proposed method where all TOAs and azimuth angles can be estimated without any information of elevation angles.

Figure 5-12: Comparison between MPCs estimated by phase difference and sensor CLEAN method
Figure 5-13: Relative AOAs of MPCs using phase difference method

Figure 5-14: Relative AOAs of MPCs using sensor CLEAN method
In particular, computed AOA ranges from all environments, including both LOS and multipath incident waves, are variously different due to all 21 orientations of the array receivers. Thus average AOAs, $\overline{\phi}_n$, are presented in the relative values instead regarding to each orientation degree in order to clarify the overall incoming directions of LOS and reflected paths. Results in Figure 5-13 and Figure 5-14 correspond with the multipath arrival diagram in Figure 5-8. Results of MPCs are examined corresponding to all 21 array orientations and directions of signal arrivals. According to Figure 5-13, significant clusters of relative AOAs in LOS and multipath components are observed. In contrast, relative values of AOAs describing the arrivals of LOS and multipath components computed by the CLEAN algorithm (Figure 5-14) do not form as clusters compared to the previous results in Figure 5-13. In particular, results of MPCs reflected from a foil board and a floor in both cases (transmitting from Tx I and Tx II) are scattering. However, it is seen that relative AOAs of multipaths reflected from a foil board in case (b) are clustered similarly at the same angle (18°) but with different amplitudes.

Finally, azimuthal AOAs estimated from both methods are compared by different errors regarding a priori of direction of arrivals. Since the measurement systems were carried out within a chamber room, positions of obstructions and conditions of the environment in each case were controlled. Directions of propagation paths can be geometrically calculated and used as the reference AOAs, $\phi_{ref}$. The errors between average azimuthal AOAs, $\overline{\phi}$, and the reference angles in each scenario case are determined; error = $\phi_{ref} - \overline{\phi}$ as seen in Figure 5-15. Azimuthal errors of AOAs estimated by both techniques are within the ±10°. The RMSE of using the phase difference method are less than ones of using the CLEAN algorithm except in the reflection from floor case with Tx II. Estimated AOAs of LOS components could gain higher accuracy than others. However, there is some considerable variance of calculated errors in MPCs reflected from a foil board and a floor. Furthermore, some absence of estimated errors or even the high fluctuating errors can be observed as the results of endfire directions of the array receiver. This degrading accuracy of estimation in endfire directions was also found in [Han06a]. For instance, in the Tx I case, these events can be observed at the ±90° orientations for LOS and reflection from a floor, and at around -68° and 112° orientations for reflection from a foil board. On the other hand in the Tx II case, estimated errors are noticeably fluctuating or, even, not able to be identified at endfire directions of -108° and 72° orientations for LOS and reflection from a floor cases, and at -65° and 115° orientations for reflection from a foil board case.

Although the sounding measurement conducted in this research is operated in a chamber room as the ideal sparse channels for acquiring explicit results of the proposed AOA estimation method, the AOAs can be also estimated when employing this proposed relative phase difference method in practical dense propagation channels. Since the phase difference method is computed
based on correlation coefficients between each antenna and the reference antenna, the distribution of complex phase terms is considered relatively to distorted signals received at the reference antenna. Thus, variations of received signals propagating through any channel environments, also in multipath-rich channel environments, can be detected and represented as scattering of phase terms of complex correlations. However, there is limitation for utilising this method in severely dense multipath channels due to some ambiguities of received signals affected by signal dispersion and attenuation, in particular at late time bins where scattering and insignificant low correlation values ($|p_{ref}(k)| \leq 0.5$) are located. Further investigation for practical dense channels is considered as the future work including of the implementation of planar array measurement.

Figure 5-15: Comparison of azimuthal estimation errors between using the phase difference method and the CLEAN algorithm
5.6 Conclusion

This chapter presented the application of the relative phase difference method related to the spatial cluster model of array antennas to determine azimuthal AOAs of UWB multipath arrivals. The measurement systems of 6 scenario cases were performed with 5 linear antenna arrays. The estimated AOA results were compared with the results obtained by the sensor CLEAN algorithm. Smaller errors indicated by RMSE were obtained to verify the proposed method for LOS detection in particular. However, it should be noted that this phase difference method requires prior information of impinging AOAs at the centre element to determine relative arrival angles at other arrays. Additionally, this method cannot be employed in some cases due to the endfire orientations of the array receiver which could be improved using a planar array. Thus, direction finding in another aspect can be measured to avoid the minimum effective size of antenna aperture. The planar array measurement can also investigate elevation angles as future work. Furthermore, despite its moderate estimation results, the spatial correlation computation could gain significant knowledge of incident signals and phase information.

Although this proposed technique might be less advantageous than other methods such as the CLEAN algorithm or the SAGE technique, which can determine both TOAs and AOAs more accurately, the spatial correlation computation could gain significant information of incident signals. Using this information can lead to novel application on UWB time synchronisation and RAKE receiver system as will be explained in Chapter 6.
Chapter 6

6 Spatio-Temporal Applications on UWB Receiver Systems

In general, the UWB received multipath signal contains the superimposition of several attenuated, delayed, and distorted replicas of a transmitted waveform. Different replicas of the same transmitted pulse can be overlapped at the receiver when the corresponding inter-arrival time is less than pulse duration. Thus, signal amplitude observed at an instant in time is affected by the previous or next diffracted and reflected contributions. The number of independent paths at the receiver depends on the pulse duration: the smaller pulse width, the higher the number of independent receiving MPCs. In free space condition, it is assumed that all UWB multipath contributions are non-overlapping due to its very short pulse width of a few nanoseconds. Consequently, UWB systems can gain the advantage of temporal diversity of multipath contributions by combining a large number of different and independent duplicates of the same transmitted pulse. This can improve the receiver performance of the decision process in UWB receiver systems.

6.1 Introduction

Due to employing the extremely large bandwidth, a large number of resolved MPCs exists. The transmitted power is distributed over the large bandwidth, thus the power in each of these individual paths is considerably low. In general, there are several significant issues that need to be considered regarding the problems of multipath delays and received UWB pulses. Firstly, the large number of MPCs can lead to a huge amount of parameter estimation. Secondly, the very low power makes the low SNR in each individual MPC. Hence the accurate estimation becomes difficult. The third issue is the time-variation of the channel, which requires continuous updating of parameters. Finally, to perform fast and accurate channel parameter estimation, training bit transmission is required. However, to improve the system performance, the training bits have to be increased; this can cause degradation of data rates and channel capacity. As a result, MPCs
should be combined for more reliable data decisions. To combine desired path components, RAKE receiver systems are employed, but they inevitably contend with the intersymbol interference problem as the length of RAKE fingers is increased [Ben04]. Additionally, due to the short pulse width of UWB signals, a slight shifting in the delay estimation can lead to significant degradation of link performances so the receiver may not collect sufficient energy for signal demodulation. To combat these problems, wise strategies should be considered. One typical solution for solving this problem is the significant MPC tracking using time delay synchronisation.

Time synchronisation can be explained in general as the time reference providing process from which the receiver searches for the correct timing to synchronise with the transmitter. Furthermore, tracking different multipath contributions is based on correlation measurements. UWB synchronisation is different and more difficult than narrowband systems. Fast and accurate acquisition with low cost is required. However, demodulation and data detection cannot be possible if timing synchronisation is not correct [Ars06]. In direct spread spectrum systems, synchronisation typically performs in two stages [Pet95], [Sim85]. The first stage is responsible for achieving coarse synchronisation involving detection of signal existence and aligning the receiver with the correct transmitted pulse and symbol sequence. This stage is known as the acquisition stage. The second stage, or tracking stage, achieves fine synchronisation and allows the receiver gain the correct pulse timing. The synchronisation is locked through clock drifts occurring in the transmitter and the receiver. For the idealistic RAKE receiver, all RAKE receiver is selected to combine all of the resolved MPCs with unlimited resources (taps or correlators) [Win00b]. Since the RAKE receiver is required to lock into the individual MPCs, positioning of proper fingers is calculated during the synchronisation process [Ars06]. Nevertheless, in dense multipath channels or NLOS channels, resolvable MPC numbers increase with the spreading bandwidth; there are increasing loads of a number of appropriate path positions to be tracked. As a result, MPC numbers that can be utilised in a typical RAKE combiner are limited by power consumption, system complexity and the channel estimation. Moreover, the synchronisation becomes more complex due to an increase in multiple finger locations to estimate.

Consequently, in order to improve UWB receptions, this chapter mainly characterises UWB array receiver systems by using spatio-temporal processing in the multipath selection process. The application of spatio-temporal correlation analysis will be investigated to cope with those main issues especially with regard to estimating parameters of the large number of MPCs and detecting very weak MPC signals. Furthermore, useful parameters obtained from the spatio-temporal results as described in the previous chapters are employed to determine the weighting factors in a combiner module. Their implementation in other space-time array receiver techniques
is also presented herein for the purposes of distortion compensation and selecting significant multipath contributions.

6.2 UWB RAKE Receiver Systems

Win et al. reported that UWB multipath signals consist of several independent components and UWB impulse radio systems can take advantage of multipath propagation by gathering a large number of different and independent replicas of the same transmitted pulse [Win98a]. To collect all contributions, the temporal diversity of MPCs will be considered for implementing a generic UWB receiver. Consequently, in UWB receiver systems, different independent duplicates of the original transmitted pulse will be analysed separately by a correlator receiver, and eventually outputs from all correlators are weighted and combined before flowing to the decision process [Ben04].

In practice, there are three different strategies for exploiting temporal diversity in receiver systems depending on the knowledge of the channel tap gain in both amplitude and phase: Selection diversity (SD), equal gain combining (EGC), and maximal ratio combining (MRC). The receiver with the SD method selects and operates only the strongest multipath signal or the first arrival with the highest instantaneous SNR. Another method which can increase SNR by combining all multipath signals rather than selecting the best path is the EGC method. Different multipath signals are aligned in time domain and then added without any particular weighting. This method is different from the final one, MRC, in which different multipath signals are weighted before the combinations. The strongest components are amplified whereas the weak ones are attenuated. The optimised MPCs carry a significant percentage of the overall signal energy, hence the SNR is maximised before the decision process. Therefore, in a single user communication system, this MRC technique can achieve the best performance [Mal08b].

6.2.1 Generic UWB Receivers

A conventional UWB receiver is based on correlation or a matched filter receiver. The correlation circuit consists of a multiplier which multiplies the received signal with the template waveform. The template waveform which exactly matches the received waveform is required for maximising SNR and minimising the noise component. According to the receiver diagram as illustrated in Figure 6-1, after the band pass filter (BPF), UWB received signals are sampled at two locations. The first point is digital and software-defined radio-based receiver sampling before the multiplier unit. To support a very large dynamic range and to resolve received signals from narrowband interference, the extremely high sampling at or above the Nyquist rate is required.
Figure 6-1: Configuration of generic UWB correlator receiver

The second sampling is located after the integrator; its sampling rate is possibly low after received signals passing through analogue multiplier and integrator circuits. The multiplier and integrator of the correlating module should be sufficiently fast enough for processing each received short pulse. However, as the popular approach is to decrease hardware complexity and high speed operation, simple single correlator receivers that correlate received signals with a local template are implemented [Cho02].

More importantly, in the perspective of low transmission power of UWB systems, although UWB power is distributed over extremely large bandwidths causing very low interference to other narrowband users, the power itself in each of these UWB individual paths becomes very low [Ars06]. This can lead to unreliable estimation of timing information and weakness of time delay resolution. UWB receiver schemes coarsely approximate the pulse shape which results in the degradation of received SNR since the limited total power is distributed over many MPCs. Furthermore, in addition to distortion that causes the receiver not to have the exact knowledge of the received pulse waveform, a long period propagating through multipath channels can also lead to a large delay spread at the receiver. Thus, UWB receiver might synchronise to more than one possible incoming MPC for detecting all significant multipath energies. As a consequence, the UWB receiver requires an additional complexity scheme with high potential to effectively eliminate these problems. Then this will end up being a RAKE reception. In RAKE receiver systems, the single correlator structure, as shown in Figure 6-1, is constructed in each parallel branch, where these finger branches are combined after the integrator unit. The number of fingers required in RAKE systems in order to collect optimum multipath energy is considerable. Further information about RAKE receiver systems will be presented in section 6.2.2.
In contrast, in less coherent schemes such as the transmitted reference (TR) [Zha03], [Que04], the differential detector [Dur04] and the energy detector [Dub05], the receiver does not need to lock into the individual MPCs but lock into the cluster region of MPCs instead. These noncoherent receivers do not require a local template and the pulse shape estimation. Hence, a small error in cluster region pointing does not cause significant performance degradation. The system becomes more robust to synchronisation errors, and less complex synchronisation algorithms can be further exploited. Recently, there has been rapid interest in performance improvement of the TR scheme with remaining the reasonable complexity. Groups of unmodulated reference pulses are transmitted along the modulated data pulses with a delay between each group less than a coherence time. The TR-receiver scheme block diagram is depicted in Figure 6-2. As a consequence, without the necessity of fine time estimating, the TR receiver employs the reference pulse with the delay version as the template in the correlating unit instead of correlating with a local template generated as used in a common correlator receiver. Thus, the TR-UWB receiver has simple timing and more immunity to timing errors. A channel estimator is not required and received pulse waveforms do not need to be estimated [Aed05].

However, the BER performance of the TR receiver is worse than the correlator receiver due to energy wasted on transmitted reference pulses and detecting both noise and signals over a considered window. The latter is caused by tracking incoming signals without fine timing, hence there is undesired noise containing in some signal samples and the integrator does not collect the optimum energy. Consequently, in order to solve this problem, some research proposed the adaptation process that can control the integrator interval to collect only desired MPCs [Fra04]. This proposed system collects energies only from received samples where MPCs are located. Thus the locations of MPCs should be known. As a consequence, Section 6.4.1 proposes the adaptive multipath searching unit that TOA estimation for tracking MPCs can be applied in
integrator interval adaptation unit. This processing can also be implemented in the correlator receiver to enhance the UWB receiver performance.

6.2.2 Conventional RAKE Receiver Systems

A RAKE receiver is realised as a time domain process in UWB reception. There are several correlators, known as fingers, which are delayed differently to adjust incoming signals into individual MPCs. Each correlator or finger is decoded independently at the initial stage, but all fingers are combined at the later stage to take into account different transmission characteristics from all fingers. Furthermore, when the combiner is constructed by SD or MRC methods, amplitudes of MPCs should be known to adjust the weighting factors.

Figure 6-3: Structure of RAKE receiver

Figure 6-3 presents a basic RAKE receiver consisting of a parallel bank of $Q$-correlators, followed by a weighted combining which determines the decision variable on the transmitted symbol. Each correlator is locked on one of the different replicas of the transmitted signal. Thus the knowledge of the received pulse shape is required in each $q$th branch with the correlator mask $m_q(\tau)$ which is aligned in time with the $\tau_q$ delayed replica of the transmitted symbol. A different set of combining weight factors, $\omega = \{\omega_1, \omega_2, ..., \omega_Q\}$, is used to combine the outputs of the correlators, depending on the diversity method. The weight factors are equal to zero in the SD method, except for the factor on the branch where the highest amplitude signal is detected. In the EGC case, the combiner simply adds all outputs of the correlators without applying any weighting; all factors are equal to 1. On the other hand, the output of each branch is multiplied by...
a weight factor proportional to the signal amplitude on that branch in the MRC method. There is the alternative RAKE receiver using the time delay unit or time shift elements. All multipath contributions are aligned in time. Hence, the conventional RAKE receiver can be simplified by adopting only the same correlator mask \( m_g(\tau) = m(\tau) \) for all RAKE fingers. This alternative less complex RAKE receiver is depicted in Figure 6-4, where the equivalent implementation of time delay unit RAKE receiver based on discrete-time channel models also consists of parallel correlators and time delay units. The different MPCs are divided into a multiple of time bin duration, \( \Delta \tau \). The correlator integrates the product between the receiver template \( m(\tau) \) and the differently-delayed received signal \( R(\tau) \). Next, the correlator output is sampled with a period \( \Delta \tau \) and then passed through a weighting unit and a combiner [Ars06].

The ideal RAKE receiver or all RAKE captures all of the received signal power by utilising a number of fingers equal to the number of MPCs. However, the implementation of all RAKE is not possible due to the fact that this approach requires a very large number of RAKE branches, which means a very large number of correlators leading to complexity of a receiver. The complexity of a RAKE receiver considerably increases with the number of MPCs analysed and combined in the process. In order to reduce the receiver complexity, the number of MPCs used to be processed should be decreased, but without losing high percentage of total energy captured by the receiver. More practical RAKE receiver implementation is a selective RAKE or
S-RAKE. The S-RAKE only uses the $L_r$ strongest propagation paths, thus the SNR is maximised. The complexity of the S-RAKE receiver is considerably reduced relative to the A-RAKE since it selects only those significant magnitudes of MPCs. Another possibility, the partial RAKE receiver or P-RAKE, is a simplified approximation to the S-RAKE combining that the P-RAKE selects only the first $L_r$ propagation paths without operating any selection among all other MPCs. Although, the first MPCs are typically the strongest and contain most of the received signal power, some of those first MPCs are not necessarily the strongest ones thus optimum performance might not be achieved [Win99].

6.2.3 RAKE Receiver Link Performances for Classified Multipath Clusters

Remarkably, it can be noted that in typical RAKE structures, the time distribution composing all incident multipath waveforms should be required. In order to allow the receiver to achieve time alignment of all MPCs synchronously with the transmitter, the multipath searching scheme consisting of the channel estimator and correlation processing units is proposed in this study. To be able to combine MPCs in the RAKE receiver effectively, the channel parameters such as delays and attenuations of MPCs are required [Ben06]. Estimated channel parameters are derived from the information-bearing signal rather than from isolated pulses. Thus, the matched template synchronised with incoming MPCs can be determined for each RAKE finger. However, this processor will be clearly demonstrated for UWB array systems in section 6.4. According to Figure 6-3 which describes the equivalent correlator RAKE receiver with the channel estimator and adaptive multipath searching units, perfect knowledge of CIR coefficients and time alignment synchronisation of received signal are assumed. Link performances of this RAKE receiver system is simulated and presented in the following results.

Link performance of RAKE receiver when propagating through multipath channels can be evaluated by assuming a specific channel model to gain channel parameters related to the considering environment. Moreover, performance of RAKE receiver can be evaluated by computing the bit error probability, $P_{b}$ as a function of $E_b/N_0$ ratio. Figure 6-5 and Figure 6-6 present the RAKE receiver performances of indoor UWB propagation channels based upon cases and parameters corresponding to realistic propagation conditions and classification of multipath clusters as simulated in Chapter 3. According to simulated results of CIRs in Chapter 3, since the best performances of propagation cases (LOS Case-A, NLOS Case-B, NLOS Case-C (both Class-II and Class-III clusters) and NLOS Case-D) allocated in different frequency sub-bands, CIRs simulated at 4GHz, 5GHz, 4GHz, 6GHz and 10GHz respectively are selected and analysed in this chapter. In the simulation process, the PPM-TH-UWB transmitted signals are generated with conveying 100,000 bits through 100,000 pulses. One pulse is transmitted for each bit, thus the ISI is not taken into account. The chip time is set to be equal to the 1-ns-bin width, $T_c=1$ns. Pulse
repetition period is $T_r = 60$ ns. Accordingly, propagation of PPM-TH-UWB signals are simulated and determined by receiving performances corresponding to CIRs of each scenario. MRC is selected as the strategy for diversity combining. Figure 6-5 shows the performance of all scenario cases with the ideal All RAKE receiver scheme. For multipath cluster class-III, the performance of LOS Case-A (MPC numbers = 6) present the best performance followed by NLOS Case-B (MPC numbers = 23) and Case C (MPC numbers = 10) respectively. When propagating through NLOS Case-C and Case-D where obstruction clusters class-II are dominant, receiving performances are consistent with propagation conditions. That is, since UWB transmitted signals are blocked by many obstructions along propagation paths in longer distances between rooms leading to lower captured total energies, higher BER of NLOS Case-C (MPC numbers= 61) and Case-D (MPC numbers= 47) can be measured. In multipath dense channels, since the RAKE combines many MPCs to obtain the best link performance, the complexity of All RAKE receiver employs a large number of fingers causing it to become impractical for real applications. Consequently, to solve the receiver becomes less complex, using S-RAKE and P-RAKE types by combining only a subset of resolvable MPCs are alternatively considered despite facing substantially worse performance than the All RAKE [Opp04].

![Figure 6-5: All RAKE receiver performance of classified UWB multipath channels](image-url)
Figure 6-6: S-RAKE and P-RAKE receiver performances of classified UWB multipath channels

Figure 6-6 (a) and (b) demonstrate link performances of S-RAKE receiver and P-RAKE receiver respectively. The number of RAKE fingers is relative to the number of MPCs of each scenario case. Link performances of using different numbers of fingers are resulted in this figure. In Figure 6-6, since there are 6 MPCs detected in LOS-Case-A, numbers of 2, 4 and 5 fingers are used to be compared. It can be found that the link performances of both S-RAKE and P-RAKE
become low when using 2 fingers. Higher performances can be gained when using more fingers. But there is no significant difference of link performances between using 4 and 5 fingers. In NLOS-Case-B, five numbers of finger or correlator (5, 10, 12, 15 and 20) are used for S-RAKE and P-RAKE. Low link performances can be seen when 5 and 10 numbers of fingers are occupied. Receiver system performs well when using 12, 15 and 20 fingers. However, it can be observed that no further better performance can be obtained when using fingers more than 15 corresponding to numbers of MPC in this scenario (MPC=23). Implementing more fingers cannot improve any higher performance. This is because energy signals captured from these NLOS propagation cases are considerably low.

In order to evaluate UWB RAKE receivers in realistic conditions, distortion effects should be considered in the performance computational process. Due to the frequency dependence of the UWB channels, received pulse waveforms are distorted and are different from transmitted pulse signals depending on propagation paths. The per-path impulse response is taken into consideration to observe pulse distortion for each individual path. The previous results describe the link performance of UWB channels with the presence of multipath but without the distortion effects; hence the receiver performances achieve the conventional RAKE receiver performance. However, when considering impacts of pulse distortion on the system performance, the mismatched distorted receiving pulses can greatly degrade the system performance. Figure 6-5 also illustrates comparison of A-RAKE receiver performances between UWB undistorted channels and UWB distorted channels. It is found that distortion effects degrade system performance as resulted in higher probability of bit error rate. In addition, there is dramatic decreasing of UWB performances due to distortion effects especially in P-RAKE and S-RAKE receiver systems.

Received signals selected by RAKE fingers are distorted leading to small values of total captured energy. Consequently, one of the important issues for the UWB receiver design is the necessity of recovering or compensating the signal energy dispersed over severe multipath effects, i.e. distortion. It should be noted that this RAKE improvement should be aimed with keeping the receiver complexity low. Next section will describe the solution for degradation of link performance due to UWB multipath distortions.

### 6.3 Distortion Compensation using Time Reversal Mirror Technique

The objective of this section is to demonstrate the solution of distortion effect that leads to deterioration in the UWB link performance. When UWB is propagating through an inhomogeneous medium, apart from the delay effect dominated by reflection, refraction,
diffraction and multiple scattering, the spatial and temporal shape of the waveform is also
distorted. Since all individual received signals are required to be considered to gain the optimum
reception, the technique that can solve the distortion effect should be investigated without
increasing any difficulty or complexity on the system.

6.3.1 Theory of Time Reversal Mirror Technique

Qiu et al. [Qiu06a], [Guo07] reported about using the time-reversal mirror technique
(TRM) with multiple input single output (MISO) UWB communications to compensate distortion
effects and reduce complexity of receiver systems. Using temporal convolution on MPCs, this
technique exploits and gains usefulness of the appearance of MPCs and transferring the difficulty
of receiver systems into the transmitter side instead. In the low-cost and low-power sensor, simple
non-coherent receivers can be approached by implementing this technique. Moreover,
convolution or correlation on MPCs can greatly reduce ISI for high data rates. In addition,
because of the sharp peak of multipath convolution, the anti-jamming capability of the system can
also be increased.

Referring to Chapter 3, the received signal at a receiver is \( R(\tau) = y(\tau) + n(\tau) \) where
\( y(\tau) = W(\tau)^* h(\tau) \), \( W(\tau) \) is the transmitted pulse and \( n(\tau) \) is an AWGN. To approach the receiving
optimisation, the filter of template that is matched with the received signal \( y(\tau) \) is taken into
consideration [Qiu04], [Qiu06b]; this filter term will be used to be convoluted with the received
signals. Since convolution and correlation are identical when the filter is symmetric [Pro01], thus
by convoluting \( R(\tau) \) with the matched filter \( y(\tau) \), \( y(\tau)^* R(\tau) \), the output of the optimum receiver
can be written into the correlation expressions, \( R_{yy}(\tau) \) as shown in (6-1).

\[
R_{yy}(\tau) = R_{yy}(\tau) + R_{ym}(\tau) \tag{6-1}
\]

\( R_{ym}(\tau) \) is a new Gaussian random variable and \( R_{yy}(\tau) = y(\tau)^* y(\tau) = [W(\tau)^* h(\tau)]^* [W(\tau)^* h(\tau)] \).
According to convolution properties [Kre00], \( R_{yy}(\tau) \) can be described by \( R_{yy}(\tau) = R_{WB}(\tau)^* R_{md}(\tau) \).
Correlations \( R_{yy}(\tau) \) and \( R_{ym}(\tau) \) are symmetric with maximum at \( \tau = 0 \), therefore the term \( R_{md}(\tau) \) can
be reformed based on the mathematical justification for time reversal communications [Qiu06a],
[Guo07] as shown below.
\[ R_{hh}(\tau) = h(\tau) \ast h(-\tau) \]
\[ = \sum_{n=1}^{N} a_n^2 \{ [h_n(\tau) \ast h_n(-\tau)] \} + \]
\[ \sum_{l=1}^{N} \sum_{l \neq k} \sum_{k=1}^{N} a_l a_k \{ [h_l(\tau) \ast h_k(-\tau)] \} \delta[\tau - (\tau_k - \tau_l)] \] (6-2)

\( N \) is the number of all generalised multipaths with amplitude \( a_n \), delay \( \tau_n \) and per-path impulse response \( h_n(\tau) \). It can be noted that CIR \( h_n(\tau) \) has unit energy, thus the total energy of CIR at \( \tau = 0 \) is \( R_{hh}(0) = \sum_{n=1}^{N} a_n^2 \). Correlations are symmetric with maximum value at this time instant which locates the sampling point for the optimum receiver. All energies of \( N \) MPCs are accumulated coherently leading to the compensation of multipath distortion. However, at other time instant \( \tau \neq 0 \), CIR energies decrease as the second term of (6-2), \( h_l(\tau) \ast h_k(-\tau) \), are added destructively and noise-like spikes are generated instead. In practical UWB communications, realisation of \( R_{yy}(\tau) \) is remarkably difficult and complex at the receiver side. As a result, instead of the receiver side, exploiting time reversal mirror of \( R_{yy}(\tau) \) at the transmitter side is considered as its less complex than at the receiver side.

### 6.3.2 Distortion Compensation Results

In practical aspect, knowledge of the CIR is a main key to the time reversal technique. Accordingly, the receiver sends a short reference pulse \( W(\tau) \) through CIR \( h_{Rxx}(\tau) \) to the transmitter. Then, the transmitter can record and recognise the received signal \( y(\tau) = W(\tau) \ast h_{Rxx}(\tau) \). Next, for the transmitting process, the transmitting information bit signals, \( W'(\tau) \), are precoded with the time reversal \( y(-\tau) \) and then coded pulse signals \( y'(\tau) = W'(\tau) \ast y(-\tau) \) are retransmitted over the CIR \( h_{txR}(\tau) \) to the receiver. Regarding to the channel reciprocity, it can be implied that the Tx-to-Rx CIR is identical to the Rx-to-Tx CIR, thus, \( h_{Rtx}(\tau) = h_{Rxt}(\tau) = h(\tau) \). Finally, at the receiver, receiving signal \( R(\tau) = y'(\tau) \ast h(\tau) + n(\tau) \) can be calculated. In order to regenerate data signals, autocorrelation terms of \( R_{wW}(\tau) \) and \( R_{nn}(\tau) \) are taken into account. The diagram of the time reversal process is described in Figure 6-7.
Figure 6-7: Block diagram of time reversal mirror process

Figure 6-8 shows the comparison between the normal CIR and the CIR occupying the transverse reversal mirror technique. In order to clarify this methodology, CIRs over the AWGN channel are not exemplified in this figure; only the presence of the distortion effect is illustrated without white Gaussian noise. Figure 6-8 (a) presents a common CIR when transmitting signal through the scenario LOS Case-A. Using the time reversal mirror technique, the symmetrical CIR can be obtained as seen in Figure 6-8 (b). Since maximum total energies are located at \( R_{yy}(\tau = 0) \), the symmetric function of \( R_{yy}(0) \) is very sharp leading to greatly reduction in the ISI. Consequently, to achieve optimum receiving performance, a matched filter that is matched to \( R_{yy}(\tau) \) should be used at the receiver to gain the optimum energy signal.

Figure 6-8: Comparison of general UWB CIR and TRM-UWB CIR
A semi-analytical approach is adopted to evaluate BER using the time reversal mirror technique for a single user with a single input single output (SISO) system. One pulse transmitted for representing one symbol is operated for simplification. Figure illustrates UWB link performances of various channel conditions when the time reversal mirror is implemented. Comparison between new performances (solid line) and the conventional All RAKE, S-RAKE and P-RAKE receiver performances (dash line with * symbol) can be observed in Figure 6-9 (a), (b) and (c) respectively. More effective performances can be seen for the time reversal implemented channels. The results clearly suggest that for UWB distortion compensation, this method can effectively support particularly for P-RAKE, S-RAKE and All RAKE receiver systems respectively. P-RAKE performances can be considerably improved. For instance, at BER=10^{-3}, an energy capturing loss can be compensated nearly 30% for LOS case and 24% for the severe NLOS case. Comparison of RAKE performances with TRM implementation for all scenario cases is presented in Table 6-1. The energy compensation is calculated relative to the performance of conventional systems. As a result, to reduce complexity in the reception, the time reversal technique can be proposed applying together with the P-RAKE receiver system. The reception does not need to build a complicated hardware by constructing all multipath fingers to achieve the optimum receiver.

![Figure 6-9: RAKE performance comparison between TRM compensation and distorted systems](image-url)
Figure 6-9 (cont.): RAKE performance comparison between TRM compensation and distorted systems
6.4 Spatio-Temporal Processing on UWB Array RAKE Receiver Systems

The study of UWB array RAKE receiver systems is presented in this work where the application of spatio-temporal analysis is effectively applied herein to improve the performance especially the time synchronisation of incoming multipaths. Furthermore, since various indoor scenarios are investigated, this work also highlights which particular RAKE architecture is the most suitable for each scenario condition. PPM-TH-UWB pulse signals are taken into account in link performance simulation for 3x3-UWB planar array receiver systems. System performance is evaluated for a single user in various conditions regarding classified scenario cases, RAKE receiver structures, and aiding techniques.

6.4.1 Adaptive Multipath Searching Unit in UWB Array Receiver

This section proposes the adaptive multipath searching unit implemented in the UWB array RAKE receiver structure. The proposed scheme can improve the knowledge of time distribution by enhancing alternative capability of scanning CIR, tracking, and adjusting MPC delays in each receiver branch. It can also reduce the complexity of the searching algorithm. Furthermore, in order to solve the complexity of signal acquisition and temporal synchronisation, the solution that optimises the system simplicity should be investigated. There is one simple solution based on the pilot pulse sequence transmission or the training bit transmission or the transmitted reference receiver as described previously [Zha03], [Que04] [Ben04]. At the receiver side, the correlator which is matched with this sequence is implemented. But this technique can decrease the SNR of data signals due to spending energy for transmitting reference pulses. To avoid the low SNR problem, a new multipath searching technique is investigated based on the application of spatial correlation technique reported in Chapter 4. This algorithm synchronises the

<table>
<thead>
<tr>
<th>UWB Distorted Channels</th>
<th>Percentage of Distorted Compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>All RAKE</td>
</tr>
<tr>
<td>LOS-Case-A-III</td>
<td>10.41</td>
</tr>
<tr>
<td>NLOS-Case-B-III</td>
<td>15.20</td>
</tr>
<tr>
<td>NLOS-Case-C-III</td>
<td>8.66</td>
</tr>
<tr>
<td>NLOS-Case-C-II</td>
<td>10.49</td>
</tr>
<tr>
<td>NLOS-Case-D-II</td>
<td>4.06</td>
</tr>
</tbody>
</table>

Table 6-1: TRM energy compensation of distorted multipath channels
initial time that can reduce the average acquisition time for finding the existence of UWB multipath profiles. The starting point of MPCs, or cluster arrival time, can be determined. The spatial correlation peaks between each antenna and its surrounding antennas can align the temporal sequence of incoming pulse signals with the receiver time reference. This multipath searching unit can be implemented in common UWB receivers to improve tracking of arriving MPCs as illustrated in Figure 6-1-Figure 6-4.

According to the proposed array RAKE receiver block diagram as illustrated in Figure 6-10, the channel estimator and adaptive multipath searching units work cooperatively together. The developed channel estimator unit has an implicit timing synchronisation capability that can estimate delays and gains of individual multipath components relative to the reference time. In the adaptive multipath searching unit, the channel estimator estimates the tapped delay positions \( \tau_n' \) and gains \( \alpha_n' \) to synthesise matching discrete channels \( \sum_{n=1}^{N} \alpha_n' \delta(t - \tau_n') \), where \( N \) is the total
number of resolved time bins. The time positions with high peak of average spatial correlation are distinguished as the estimated delays \( \tau_{\text{est}} = \tau_i = \arg \max_k \{ R_{\text{th}}(i,j,m,n) \} \). The relative values of the spatial correlation peaks corresponding to these estimated delays are provided as the tap weighting coefficients for their individual array elements \( \{ m_{i,j} \} \). As a consequence, using these estimated parameters, UWB channels are synthesised which are matched with the actual channel as expressed in (4-2). These knowledge parameters are used for the template generator.

Nonetheless, parameters estimated from multipath selection present several other time positions that also locate high correlation values. In such conditions, signal detection can be deteriorated by these several false alarms leading to mismatched time synchronisation and correlated template. Consequently, in the multipath searching unit, the capturing energy threshold value is proposed to solve this problem. Energy captured from detected MPCs at estimated tracking times should be more than 60% of total energies of the considering received signals \( R_{i,j,k}(\tau) [\text{Cas02a}] \).

### 6.4.2 Time-Reversal Mirror Technique for Array UWB systems

Consequently, when employing the time reversal technique together in the proposed antenna array systems, the performance can be improved. Distorted pulses are regenerated, thus energies captured become increasing. The reference signals \( W(\tau) \) are sent from each antenna array receiver to the transmitter to sound the particular propagation channels. The transmitter records channel characteristics \( h_{i,j}(\tau) \) and transmits precoded signals convoluted with the reverse version \( W(\tau) * y_{i,j}(\tau) \) corresponding to individual propagation paths in each receiver. This strategy results in spatial focusing on individual propagating paths of each array \( (i,j) \) from Rx-to-Tx and Tx-to-Rx paths. Thus the adaptive multipath searching unit can estimate the TOAs of MPCs from tracking the highest correlation peaks of received signals \( R_{i,j}(\tau) = R_{y,y}(\tau) = R_{\text{th}}(\tau) \) directly. At each antenna, the first TOA of the strongest MPC can be determined by (6-3). Then the delay version of \( R_{i,j}(\tau) \) corresponding to the estimated previous tapped delay is subtracted from the sequence of \( R_{i,j}(\tau) \) to remove the correlation peak tracked in the previous process. The computation is repeated to estimate the tapped delay of the next strongest path and so on. This technique is called as the successive cancellation algorithm [Ami02], [Mol05].

\[
\tau'_{\text{est}} = \arg \max_\tau \{ R_{i,j}(\tau) \} \quad (6-3)
\]
Despite the set of peak amplitude of \(|R_{ij}(\tau_{\text{en}})|\) at the estimated times that can be used for compensating distortion, the complex spatial correlation analysis \(\rho_{ij}(k)\) of \(R_{ij}(\tau = \tau_{\text{en}})\) is also implemented to investigate distortion effect. This methodology is based on one presented in Chapter 5 and proposed to be processed in the multipath searching unit. Together with the TRM technique, all time periods of received signals do not need to be computed, only the peak-located times (\(\tau_{\text{en}}\)) are considered. Contrary to the technique in 5.3, the array which is used as the reference is the array \((ij)\)-element of interest not the centre element of the array antenna. At the \(k\)th tapped delay, with time \(\tau_{\text{en}}\), complex spatial correlation of \(R_{ij}(\tau_{\text{en}})\) between interested element and its surrounding elements is computed, and then the vector of the phase term, \(\varphi\), can be determined by (6-4). According to (4-4)-(4-6) and (5-12), at any \((ij)\) element, the phase term of complex correlation can be given by

\[
\varphi = \varphi_{(i,j),(i\pm m, j\pm n),k} = \tan^{-1}\left(\frac{\text{Im}(\rho_{R(i,j),(i\pm m, j\pm n),k}(k))}{\text{Re}(\rho_{R(i,j),(i\pm m, j\pm n),k}(k))}\right); \quad m, n = \{0, 1\} \mid m \neq n \neq 0 \quad (6-4)
\]

This phase term vector, \(\varphi\), includes the relative phase difference when considering array element \((ij)\) and other surrounding elements. Hence \(\varphi = \Delta \varphi_{(i,j),(i\pm m, j\pm n),k}\) where \(m, n = \{0, 1\}\) and \(m \neq n \neq 0\). The degree of distortion effects can be measured by the relative phase difference \(\Delta \varphi_{(i,j),(i\pm m, j\pm n),k}\) and the magnitude term of the complex correlation coefficient. At each array, the estimated tap bin \(k = \tau_{\text{str}}\) which locates the scattering of \(\Delta \varphi_{(i,j),(i\pm m, j\pm n),k}\) is implied as an existent time of distortion when its standard deviation \(\sigma(\text{mag}(\rho_{R(i,j),(i\pm m, j\pm n),k}(k))|\Delta \varphi_{(i,j),(i\pm m, j\pm n),k}) \geq 1\). The weighting vector, \(w_{i,j,k}\), is implemented individually for each corresponding \(k\)th tapped delay and \((ij)\)-array element to provide the synchronisation and the optimum SNR.

Since the new TRM channel typically contains a strong peak with weak sidelobes, the link performance simulation process is much easier to handle. But the computational loads still exist when considering idealistic all RAKE performance. Since TRM performance is verified already that can be used for counterbalancing the distortion effects on UWB pulse waveform, thus the comparison of link performances in array systems between TRM implementation and without TRM is not described here.

According to simulation processing for the array case using all RAKE structure, all contributions are calculated and consuming high period of computing time compared to the single receiver case. Although all RAKE receivers can achieve the best performance, it requires massive computational complexity. The next section presents the less complex RAKE finger selection
schemes for array systems. The quality of link performances can be obtained equivalently to or slightly less than one when using All RAKE receiver.

6.4.3 Two-Dimensional Finger Selection for Space-Time UWB RAKE Receiver Systems

Based on the structure of 3x3-planar array receiver systems exploited in this chapter, both space and time domains should be considered in received signal processing. Thus, RAKE finger selection and correlated signal combining have to be processed in a spatio-temporal structure. The finger selection scheme studied here is based on the S-RAKE receiver where maximum captured energies are detected. Differently from the conventional S-RAKE receiver for a single element which considers only one-dimensional time domain selection strategy, RAKE finger selection strategy for array elements are considered in both time and space domains. According to Chapter 2, a typical single user PPM-TH-UWB signal can be expressed by [Hu04],

$$W(\tau) = \sqrt{\frac{E_b}{N_s}} \sum_{j=0}^{N_s-1} g(\tau - jT_s - c_pT_c - \epsilon)$$

(6-5)

where $g(\tau)$ is the energy-normalised Gaussian pulse, $E_b$ is the bit energy, $N_s$ is the pulse numbers for transmitting one data bit, $T_s$ is the average pulse repetition period, thus the bit duration $T_b = N_sT_s$. $T_c$ is the TH chip period which satisfies $N_sT_c \leq T_s$, where $N_a$ is the number of hops. $C_j$ represents the pseudorandom TH code of the $j^{th}$ pulse and $\epsilon$ is the PPM time shift. Assuming that UWB signals received by each array antenna experience independent channel fading, hence the equivalent baseband TDL channel model for a single user to the $p^{th}$ antenna can be given by (6-6) where $p=1,2,...,P$ is the numbers of array antenna and $L$ is the resolvable path numbers. $\delta(\tau)$ is the Dirac delta function with delay time $\tau \in [0,T_c]$. Consequently, when the transmitted PPM-TH signal, $W(\tau)$, propagates through the channel model $h_p(\tau)$, the received PPM-TH-UWB signal at the $p^{th}$ array can be written by (6-7) including with the zero mean and variance white Gaussian noise $n_p(\tau)$. The noise at different antennas is assumed to be mutually independent [Aed05], [Cha06].

$$h_p(\tau) = \sum_{l=0}^{L-1} a_{p,l} \delta(\tau - lT_c - \tau_d)$$

(6-6)
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\[ R_p(\tau) = \sum_{l=0}^{L-1} a_{p,l} W_r(\tau - IT_c - \tau_d) + n_p(\tau) \]  
(6-7)

where \( W_r(\tau) = \sum_{l=0}^{L-1} \varphi(\tau - IT_c - \tau_d); \varphi(\tau) = \delta(\tau) * g(\tau) \)

According to Figure 6-10, there are \( P \) array antennas which comprise \( Q \) RAKE fingers in each antenna, thus, leading to \( PQ \) correlators in the proposed array RAKE receiver systems. The output signal of the \( q \)th correlator at the \( p \)th antenna during the \( n \)th symbol can be given by (6-8) where \( k(p,q) \) is the path number selected by the \( q \)th correlator at the \( p \)th antenna [Cha06]. Since the incoming signals are tracked by the multipath searching unit, the received multipaths are aligned synchronously to be processed in each branch. Hence, the propagation delay term, \( \tau_n \), is not taken into consideration as shown in the following expression.

\[ r_{p,q}[n] = \sum_{j=0}^{(n+1)N_s-1} a_{p,j} W_r(\tau - jN_s T_c - k(p,q)T_c) \]  
(6-8)

To obtain perfect channel estimation, \( Q \) spatial combiners are constructed in the systems followed by one temporal combiner. Thus, at each \( q \)th spatial combiner, the input vector \( r_q[n] = [r_{1,q}[n], r_{2,q}[n], ..., r_{q,q}[n]]^T \) will be weighted by weighting vector \( w_q \). Thus the output of the \( q \)th spatial combiner is \( r^*_q[n] = w_q r_q[n] \). Remarkably, the spatial combiners are controlled by the multipath searching unit for the spatial weighting estimation. The signal from each antenna is individually weighted regarding to computation of complex spatial correlation among its surrounding antennas as described previously in 6.4.2. The spatial weighting vector, \( w_q \), for each branch consists of the set of weighting coefficients for each \((i,j)\) antenna at the individual \( k \)th tapped delay. Each weighting coefficient corresponds with inverse relation to the standard deviation of spatial complex correlation \( \rho_{\text{spatial}}(i,j;i',j';k) \). Thus the weighting vector can be described by (6-9)

\[ w_q = \omega_q \frac{E_b}{N_s} R_q^{-1} h_q \]  
(6-9)

where spatial weighting coefficient \( \omega_q = [\omega_{i,j,k}]_{q=1}^Q \) subject to \( \rho_{\text{spatial}}(i,j;i,m,n,k) \), \( m,n = \{0,1\} \) and \( m \neq n \neq 0 \). The weighting coefficient is 0 when the considered time delay is not equal to \( \tau_c \). That
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means $k_q \in \{\tau_{\text{est}}\}$; delays in all branches are tapped at the same time as the estimated delay $\tau_{\text{est}}$. But it is not necessary to construct all $Q$ numbers of branch at every single estimated time. Both impulse response $h_q$ and array correlation matrix of $r_q[n]$ can be determined by a sampled average of training signals. In contrast, without individual weighting coefficients, input terms of the temporal combiner $r'_q[n]$ from all finger branches, $r'_q[n] = [r'_1[n], r'_2[n], ..., r'_Q[n]]^T$, are combined with the temporal weighting $w = RR'h$. Note that $R'$ is the array correlation matrix of $r'[n]$. Both $h$ and $R'$ can be estimated during the pulse training phase. Hence the output of temporal combiner before sending to the decision process is $R_{\text{tot}} = w^T r'[n]$. 

When selecting RAKE finger based array elements, the two-dimensional space-time selection is considered [Cha06]. The finger selection process needs to consider temporal varying of all signals in any path number $l(p,q)$ captured by all antennas $1 \leq p \leq P$ and all correlator fingers $1 \leq q \leq Q$. In idealistic method, to achieve the optimum finger selection which can gain the best performance, all path numbers are selected. However, since all RAKE fingers from all antennas are analysed, the finger selection process becomes exhaustive computational load; thus, this all RAKE structure is not practical for implementation. Accordingly, other space-time finger selection methods are considered. These new feasible strategies rely on the maximum energy of the desired UWB signals. The first method is focused on maximum energy in each array or array-based energy selection; the path number $l(p,q)_{q=1, ..., Q}$ that contains the maximum energy is selected.

$$\{l(p,q)\}_{\text{ABE}} = \arg \max_{\{l(p,q)\}} \left\{ \sum_{p=1}^{P} \sum_{q=1}^{Q} |a_{p,l(p,q)}|^2 \right\} \quad (6-10)$$

Alternatively, if the selected path number is considered depending on the correlator branch, the constrained correlator-based energy selection is suggested here. At each $q$ receiver finger, the process searches for the maximum energy summation captured by the path number $l(p,q)_{q=1, ..., Q}$ from all $P$ antennas. The searching process is constrained by (6-11) where at each $q$ correlator, selected path numbers can be obtained from more than one different antenna, $l(p,q) = l(p',q)$ for all $p \neq p'$, $1 \leq q \leq Q$.

$$\{l(p,q)\}_{\text{CBE}} = \arg \max_{\{l(p,q)\}} \left\{ \sum_{q=1}^{Q} \sum_{p=1}^{P} |a_{p,l(p,q)}|^2 \right\} \quad (6-11)$$

$$\{l(p,q)\}_{\text{CBB}} = \arg \max_{\{l(p,q)\}} \left\{ \sum_{q=1}^{Q} \sum_{p=1}^{P} |a_{p,l(p,q)}|^2 \right\}$$
Figure 6-11 shows the selected samples of signals regarding the space-time finger selection algorithm. Signals stored in selected path numbers are combined and weighted by spatio-temporal weighting units. Three cases of finger selection are described accordingly to the space-time finger selection process. In addition, since distortion effects are not excluded from the investigation, the time reversal mirror technique is also employed leading to the symmetry pattern of received signals. Furthermore, comparison between capturing signal through the sparse channel (LOS case) and through the dense channel (NLOS case) is also presented. Number of antennas, $P$, and correlators, $Q$, are not very large in general, so this space-time finger selection is practical and feasible to be implemented.

![Figure 6-11: Examples of signal energy collection by finger selection strategy process](image)

(a) All RAKE
(b) Array-based energy selection
(c) Correlator-based energy selection
(d) All RAKE
(e) Array-based energy selection
(f) Correlator-based energy selection

Figure 6-11: Examples of signal energy collection by finger selection strategy process
(a)-(c) for LOS case and (d-f) for NLOS case
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Figure 6-12: Link performance of array UWB RAKE receiver systems with RAKE finger selection strategy and time-reversal mirror techniques

As a consequence, Figure 6-12 shows the link performance of UWB array RAKE receiver systems implemented by all supporting spatio-temporal techniques, i.e. the adaptive multipath searching unit, the time-reversal mirror technique and the space-time finger selection technique. The same channel parameters and transmitted signals as used in previous sections are simulated based four environment scenarios without considering narrowband interference case. It can be seen from the figure that the consistent performances can be obtained for each finger selection technique comparing to the ideal one (all RAKE). Employing array-based energy selection and correlator-based energy selection can gain similar performances which greatly reduce the complexity of using the all RAKE structure.

However, the correlator-based energy scheme seems to gain better performances than using the array-based energy one especially in dense channels, NLOS Case-C and NLOS Case-D. This is due to propagating through obstructions leading to low energy signals diffracting to some arrays. When using the correlator-based energy selection, at each finger, path numbers that contain maximum energies are selected. Thus only one or more antennas capturing high summation of energy signals are taken into account; the receiver system gains the advantage of spatial diversity. In contrast, for the array-based energy selection, some path numbers \( l(p,q) \) from all antennas are selected. Strongest signals among their correlator sets \( \{l(p,q)\}_{m=1,\ldots,q} \), but might be
weaker comparing to signals detected by other antennas, are taken into account. Therefore, insignificant signals selected from some arrays are included into the link performance simulation leading to lower system performances.

6.5 Conclusion

This chapter presented the application of spatio-temporal array processing for the UWB array receiver. The complex spatial correlation analysis described in Chapter 4 and Chapter 5 was mainly implemented in the multipath searching unit. The estimated time delays, magnitude and phase terms of complex correlation coefficients were employed to gain receiving optimisation. The UWB receiver was constructed based on RAKE receiver systems. The preliminary results presented link performances of conventional RAKE receivers in various propagation channels. Since the received signals that used to be simulated were classified regarding to multipath cluster groups, low link performances were drawn due to different distortion effects related to classification of multipath clusters. Consequently, implementing time reverse technique was suggested into the system in order to compensate distorted captured signals. The technique could be viewed as space-time precoding at the transmitter, hence, it could be implied that detected signals were spatially and temporally focused. Incorporating this TRM technique with generic RAKE structure could improve link performances especially in P-RAKE system.

When analysing UWB array receivers, the adaptive multipath searching unit including spatio-temporal array processing mainly played the important role in the systems. TOAs of MPCs could be determined when conditional spatial correlation coefficient $\rho_{ij}$ achieved the maximum values. Estimated channel parameters with conditional spatial $(i,j)$ positions and $k^{th}$ tapped delays could reduce numbers of multipath contributions to be considered. Only significant path constraints were taken into account. Concerning the complexity of this spatio-temporal processing, the multipath selection did not require many sources for operating but the weighting vectors, in particular, for all RAKE fingers spent extreme computational loads. According to the simulating process, the computational complexity mainly appeared in the spatial weighting estimation. The simulation required $II$ iterations ($IxJ$ elements) for determining individual weighting coefficients corresponding to the complex spatial correlation. However, computational loads of spatio-temporal processing in multipath searching unit were fairly acceptable since only $IxJ$ repetition searching for TOAs and complex correlation terms were processed once during the TRM pulse training period. All estimated parameters were recorded as the reference parameters occupied in other units such as in the channel estimator, template generator, weighting combiners etc.
Finally, the 2-D space-time finger selection was proposed for UWB array receiver systems to reduce the extreme complexity of using all RAKE receivers for obtaining the optimum receiver. The array-based energy selection and the correlator-based one were functioned corresponding to the maximum energies captured from any $q^{th}$ branches in each antenna or scanning maximum energies from any $p^{th}$ antennas in each considering correlator branch respectively. Results show the consistence of link performances using these two strategies as well as performing by all RAKE receivers with slightly better results in the correlator-based energy selection for UWB dense channels.
Chapter 7

7 Conclusion and Future Works

7.1 Conclusion

In this thesis, the characteristics of spatio-temporal UWB array systems have been investigated and the following four new research perspectives have been achieved.

1) Modified spatio-temporal simulation of UWB discrete channel impulse responses based on classifications of multipath cluster and pulse distortion effects was generated. CIR parameters and distortion effects corresponding to multipath cluster classifications could be found in particular. In addition, frequency dependent characteristics were also included in the simulation. Thus it was advantageous that the proposed modified simulation could specify these UWB properties in addition to amplitude, TOA and AOA information, which were commonly generated by a conventional CIR simulation.

2) A novel spatial correlation computation based array structure was presented. Spatial correlation analysis was operated at each time bin, such that spatial correlations which varied during excess time delays were temporally characterised. Thus distortion effects were assumed to be included in the characterisation process. Furthermore, the application of this computation process proposed the new technique of TOA estimation of multipath clusters and their estimated results were consistent with standard UWB characteristics.

3) Based on the spatial correlation analysis, this work presented the AOA estimation technique by combining the relative phase difference technique in the linear array aspect. Measurements using a linear array were also carried out to evaluate estimated AOAs. In addition, the complex spatial correlation analysis could also determine the distortion effects at particular time bins. This proposed analysis initiates an advance algorithm for using in planar, or even any, array structures by considering any elements as the references.

4) Finally, spatio-temporal computation was proposed in this research, which could be applied to UWB array receiver systems known as the adaptive multipath
searching unit. This unit consisted of the multipath searching and distortion selection processors, which could support RAKE selection and weight estimator functions. Together with finger selection strategies and TRM techniques, this proposed unit could take advantage of its implementation in array RAKE receiver systems which could gain improvement of UWB array link performances with less complexity.

The main objective of this thesis is to study characteristics of spatio-temporal UWB array systems. Space-time processing computation in UWB propagation based array receivers is the main focus. The research investigation can be applied for future UWB communication systems especially for UWB smart antennas and UWB-MIMO technologies. According to the literature review, several research publications reported utilisation of multi-antenna and smart antenna technologies; therefore, both spatial and temporal characteristics become the major topic of interest to be investigated. The investigation outputs can lead to the enhancement of transmission technologies and space-time signal processing principles.

7.1.1 Review of Modified Spatio-Temporal Simulation and Distorted Channels

Modified simulation of UWB spatio-temporal CIRs was proposed in Chapter 3 for further investigation in distortion channels and frequency dependence. In the simulation, propagation scenarios were classified depending on the obstructions nearby the transmitter and the receiver. Furthermore, physics-based distortion effects were also generalised and functioned in the simulation algorithm. Consequently, the simulation results presented both CIRs and distorted pulse regarding to specific knowledge of TOAs, AODs, AOAs, classified multipath clusters and dominant distortion effects for each particular propagation path.

A simulation model was generated based on 3x3 array antennas over 10 frequency sub-bands (2-11 GHz) with 9 types of obstruction material; therefore, frequency dependent characteristics of UWB propagation through various environment conditions, both LOS and NLOS cases, were studied. Variation of energy distribution corresponding to multipath cluster classes, frequency sub-bands, distortion mechanisms and obstruction materials were characterised.

Further investigation of quantification of distortion effects in each simulated propagation channel was reported in terms of probability of bit error rate and ranging error. Finally, all simulated UWB signals and channel parameters were used in Chapter 6 as the captured signals in order to calculate link performances in various scenario cases.
7.1.2 Review of Spatial Correlation Technique based Array Processing for Multipath Cluster Estimation

The primary computation process applied in this research was proposed here. Spatial correlation analysis based array processing was computed at each time bin, thus this bin width was assumed to be able to include each distorted pulse waveform. Spatial correlation between each array element and its surrounding elements were computed to examine multipath arrivals scattering from all direction around antenna arrays. Traces of average high correlation values were considered as TOAs of MPCs.

Consistent results with the sensor-CLEAN algorithm and standard UWB cluster modelling could evaluate the accuracy of estimated TOAs obtained by the proposed computation. However, AOA estimation derived by time slices of spatial correlation could not bring the consistent results especially at the later time delays. The time variations of groups of arrays where high spatial correlation coefficients are located are for consideration as a future work.

Spatial correlation rates were characterised corresponding to propagation cases, computed subgroups in different directions and distances. Results have shown the agreement with reasonable hypotheses that correlation rates in NLOS decreased more rapidly than ones in LOS, and longer distances between both ends leaded to increasing of decorrelation rates.

7.1.3 Review of Complex Correlation Analysis based Linear Array AOA Estimation Measurement

Further effort to determine AOAs of MPCs was still investigated in Chapter 5 using the relative phase difference technique based spatial complex correlation. To simplify this method, the UWB sounding measurement was conducted in an anechoic chamber with the 5 linear array receiver. Direction of AODs, AOAs and obstruction positions were controlled to gain the reference information to be compared with the calculated results.

Differently from the spatial correlation technique proposed in Chapter 4, an AOA estimation process was computed based on spatial correlation between each element and the reference element. Since measurement data was simply conducted by 5-linear-array antenna, implementing the computation technique proposed in Chapter 4, $R_{ijk,b}$ gained the similar results with using the correlation relative to the reference antenna $R_{m,n,b}$. 

Limitation of this technique is the requirement of prior knowledge of incident AOAs at the reference element in order to determine relative AOAs for other elements.
Accordingly, consideration of element of interest for planar arrays, either on each element or only reference element regarding to both spatial correlation techniques, will be proposed as the alternative method for further investigation.

7.1.4 Spatio-Temporal Applications on UWB Receiver Systems

The last contribution highlighted the application of array spatio-temporal analysis for UWB receiver systems. The possibility for the receiver to detect and estimate the correct position of multipath could be improved by the proposed adaptive multipath searching unit based on complex correlation analysis. The estimated channel parameters such as time delays, magnitude and phase terms of complex correlations were used to gain weighting for the optimum receiver.

The magnitude term of complex spatial correlation analysis was employed in the multipath searching unit. On the other hand, the phase term of complex spatial correlation was taken into account for tracking the suitable times when the time reversal technique should be operated.

Furthermore, implementation of TRM technique, could improve receiver performance from distortion channels particularly for P-RAKE systems. Combining weighting was executed just only at phase variance tapped delay corresponding to variation of complex correlation phase term.

2D space-time finger selection strategies, array-based energy and correlator-based energy selection, were exploited in array RAKE systems. The latter selection could perform better link performance than the array-based one particularly in dense channels. Constructing these space-time RAKE finger selections with the proposed adaptive multipath searching unit and TRM technique could obtain the semi-All-RAKE performance but with much lower complexity.

7.2 Future Works

7.2.1 Spatio-Temporal Application on UWB RAKE Multiuser Systems

Link performance results presented in Chapter 6 were computed based on single user communication. In order to validate that the proposed spatio-temporal process can also be applied in the multiple access strategy, The current framework should be extended to the multiuser case. Since multiuser interference is also taken into consideration, investigating alternative methods that provide multiuser interference cancellation, faster multipath searching and lower complexity is still an open issue. Any study that examines whether the proposed or adaptive complex spatial
correlation technique can gain the ability of separating the desired signals from interferers should be further investigated.

7.2.2 Application of Complex Correlation Analysis for Smart Antenna

TOAs and AOAs estimated by complex correlation coefficient at particular $k^{th}$ time delays, can be used as the adaptive information for beam steering technique in the UWB smart antenna. It is challenging to determine the smart algorithm that can gain steering vectors and achieve a set of perfect delay lines in order to scan for the main beam direction sequentially around surrounding space. The steering delays are usually calculated based on the information of AOAs and array steering vectors. The distortion selection phase terms can be used as the information for an adaptive array approach that the beam pattern can be adjusted to null the direction of incoming distorted signals. Furthermore, the novel algorithm should be further investigated for determining the weighting vectors which are dynamically updated according to the environment.

7.2.3 Further Measurement and Spatial Correlation Analysis of Planar Arrays

Although the UWB array signals analysed in Chapter 4 could determine TOAs of arriving multipaths using spatial correlation technique, only the magnitude terms were taken into account for the computing process. However, this data set cannot be used to compute for relative phase difference since data phase term does not exist. Thus planar array measurement should be carried out by processing both magnitude and phase term data. In addition, repeating the previous measurement campaign in Chapter 5 for planar array receiver can improve AOA estimation algorithm based on the relative phase difference technique. This is because the scale of considering phase difference between each element relative to the reference array (the centre position), $R_{\phi_{r,ref}}$, can be scoped to finer scale, $R_{\phi_{r,k}}$ instead as presented in Figure 7-1.
Complex correlation coefficients and relative phase difference are not computed regarding only the centre element of the array, but by the fact that all elements can be employed as the reference to be correlated with their surrounding elements as shown in the diagram. Consequently, there are various subgroups of interested antennas; gaining various terms of magnitude, phase and estimated AOAs in each group. Modified algorithm for determining finer scale and higher accuracy of AOAs should be explored with comparison with the results obtained by the centre reference technique.

### 7.2.4 Further Investigation of Space-Time Correlation Characteristics

According to the spatial correlation computation, slice diagrams of high spatial correlation values between each antenna can be obtained. The diagrams for the time bins located high correlation coefficients are taken into consideration. There is the presence of changing locations of high degree correlation coefficients with time varying. An example of this issue is depicted in Figure 7-2. This remarkable observation motivates the idea of using other adaptive space-time processing to investigate array UWB channels. Thus, the independent component analysis (ICA) [Che05], [Saw05] is the proposed algorithm for future study which continues examining spatial correlation computation.
Figure 7-2: Spatial correlation sliced diagram with changing of traces of high correlation coefficients

Figure 7-3: Example of 3-D space-time correlation feature
The space-time ICA technique is a blind source separation technique that relies on the separation of individual signals of interest (from mixture of unknown sources) into spatially independent components at each time bin. According to a stacked 3D-spatial correlation over time as drawn in Figure 7-3, the summation of the space-time independent components, ICs, can be presented by (7-1). The best features distinguished from the spatio-temporal ICA can be applied to improve the source classification, i.e. various incoming directions of AOAs scattering from obstructions, and multiple source localisation [Saw05].

\[
R_{ij} = \sum_{n=1}^{N} a_n \cdot IC_n(i, j, k) \quad ; a_n \text{ is the weight of } IC_n \tag{7-1}
\]

Furthermore, due to the cause that all arrays are taken into account, the adaptive space-time ICA technique should be modified to include the matter of different scale sizes of correlation subgroups into the algorithm.
Appendix A

Method of equal areas (MEA) has been originally introduced in [Pat96] to model the classical Jakes/Clark Doppler spectrum and the Gaussian Doppler spectrum. In this thesis, MEA is applied to maintain low computational complexity with high performance in channel simulation. The deterministic channel simulators using the summation of sinusoids can be used to generate the AOD, AOA and TOA profiles. In the following, the methodology of MEA is applied to derive the closed-form solutions for these model parameters.

\( f(q) \) is defined as a spectral density function which based on the properties:

1) Symmetry condition: \( f(q) = f(-q) \)

2) Unit power condition: \( \int_{-\infty}^{\infty} f(-q) dq = 1 \)

3) Limited bandwidth condition: \( f(q) = 0 \) for \( |q| > \varphi_{\text{max}} \)

Define a set of \( N \) discrete parameters, or discrete frequencies in the original works, \( \{\varphi_n\} \) in the range of \( 0 < \varphi_{n-1} < \varphi < \varphi_n < \varphi_{\text{max}} \). Thus the area \( A_n \) under the power spectral density function \( f(q) \) is equal to \( 1/(2N) \) for all \( n=1,2,...,N \).

\[
A_n = \int_{\varphi_{n-1}}^{\varphi_n} f(q) dq = \frac{1}{2N} , n=1,2,...,N \tag{A-1}
\]

where \( \varphi_0=0 \). To determine the discrete parameters \( \varphi_n \) regarding to Equation (A-1) and MEA properties, function \( F(\varphi_n) \) is introduced as described in Equation (A-2).

\[
F(\varphi_n) = \int_{-\infty}^{\varphi_n} f(q) dq \tag{A-2}
\]

\[
= \frac{1}{2} + \sum_{\nu=1}^{n} \int_{\varphi_{\nu-1}}^{\varphi_{\nu}} f(q) dq = \frac{1}{2} + \frac{n}{2N}
\]
Thus, if there exists the inverse function of \( F(F') \), discrete parameters can be determined by

\[
\phi_n = F^{-1}(F(\phi_n)) = F^{-1}\left(\frac{1}{2} + \frac{n}{2N}\right) \tag{A-3}
\]

Considering MEA to be applied to PAS of AOAs which agrees well with the zero-mean Laplacian probability density function [Cho03a], azimuth angles \((\phi_h)\), also elevation angles, can be obtained.

\[
f(\phi_n) = \frac{1}{\sqrt{2\pi} \sigma_\phi} \exp\left(-\frac{\phi_n^2}{2\sigma_\phi^2}\right) \tag{A-4}
\]

\[
F(\phi_n) = \int_{-\infty}^{\phi_n} f(\phi) d\phi = \int_{-\infty}^{\phi_n} \frac{1}{\sqrt{2\pi} \sigma_\phi} \exp\left(-\frac{\phi_n^2}{2\sigma_\phi^2}\right) d\phi 
\]

\[
= \frac{1}{2} \exp\left(-\frac{\phi_n^2}{2\sigma_\phi^2}\right) \tag{A-5}
\]

where \( \sigma_\phi \) is the standard deviation of \( \phi \). Thus from Equation (A-2),

\[
\frac{1}{2} - \frac{1}{2} \exp\left(-\frac{\phi_n^2}{2\sigma_\phi^2}\right) \approx \frac{1}{2} + \frac{n}{2N} \tag{A-6}
\]

\[
\therefore \quad \phi_n = \frac{\sigma_\phi}{\sqrt{2}} \ln\left(\frac{-2n}{N}\right)
\]

In a similar manner, MEA can determine the discrete TOAs in such a way that the area under the PDS, \( f(\tau) = 1/\sigma_\tau \cdot \exp(-\tau/\sigma_\tau) \), over the delay interval \([\tau_{n-1}, \tau_n]\) is equal to \(1/N\):

\[
\int_{\tau_{n-1}}^{\tau_n} f(\tau) d\tau = \frac{1}{N} \tag{A-7}
\]

Considering the cumulative distribution function (cdf) of \( f(\tau) \),
\[ F(r) = \int_{r_0}^{r_n} f_r(r) \, dr = \sum_{r=1}^{n} \int_{r_{n-1}}^{r_n} f_r(r) \, dr \]
\[ = \frac{n}{N} , n = 1, 2, \ldots, N - 1 \]  \hspace{1cm} (A-8)

Thus

\[ \int_{t_0}^{t_n} f_t(r) \, dr = 1 - \exp \left( -\frac{t_n}{\sigma_t} \right) = \frac{n}{N} \]
\[ \therefore \tau_n = -\sigma_t \ln \left( 1 - \frac{n}{N} \right) \]  \hspace{1cm} (A-9)
Appendix B

The 5-linear-array antenna is constructed using a 2.5cm. circular copper disc with a gap distance between each array is 4.29cm. avoiding the mutual coupling effect between each array element. All arrays are fed via 0.0625cm. SMA connectors mounted on 8.58x30cm. ground plane as illustrated in Figure B-1. When conducting the measurement at each element, only the measured array is connected with the receiving cable to the port2 of network analyser, whereas other arrays are terminated with the termination loads. Radiation patterns for all elements are nearly constant with frequency over 2-8GHz presenting the omnidirectional pattern.

Figure B-2 (a) and Figure B-2 (b) present the radiation patterns in H-plane, $E_\phi (\phi, \theta = 90^\circ)$, and in E-plane, $E_\theta (\theta, \phi = 0^\circ)$, respectively. These radiation patterns are measured at the reference antenna, the third array sensor. The maximum gain is observed to be approximately 1.2 dBi at 0° and 180° orientations for H-plane pattern, whereas some degradation can be seen at 90° and 270° orientations for E-plane pattern.

Figure B-1: 5 Linear array antenna
Figure B-2: Radiation pattern of the array antenna measured at the reference element.
Actually, the UWB antenna should not cause any electromagnetic interference or can reject an interference with existing nearby wireless network communication systems such as IEEE 802.11a in the U.S. (5.15-5.35GHz, 5.725-5.825GHz) and HIPERLAN/2 in Europe (5.15-5.35GHz, 5.47-5.725GHz) [Ker03]. Thus, UWB antennas with notched characteristics in these allocated frequency bands should be designed. However, since all measurements carried out in this study are operated in a chamber room without any interference from other wireless network communications, simple wideband bandpass antenna is characterised and constructed as described below.

Return loss ($S_{11}$) between the transmitter and the received array antenna positioned for maximum gain with vertical polarisation is measured and shown in Figure B-3. The result shows that the good matching ($S_{11} \leq -10$dB) of this array antenna is between 2.65-8GHz with a resonance at 6GHz probably related to the physical size of the circular disc antenna.

![Figure B-3: Return loss of the array antenna](image)
This linear array antenna is employed to measure UWB signal under realistic condition with building reflection from a foil board and a floor along the propagation path. Figure B-4 (a) and (b) show the example of the measurement in real situation which the signal transmitted from the transmitter Tx-II with 0° orientation of the receiver.

Figure B-4: Example of UWB channel measurement in a chamber room
References


References


References


References


References


References


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References


