

An Holistic Approach to Optimal Ultra-Wideband Wireless Communications System Design

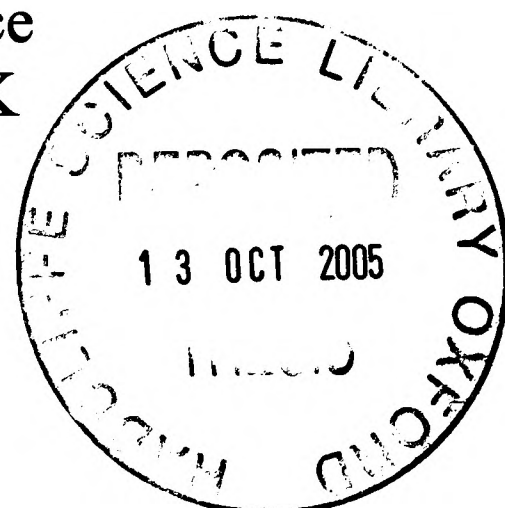
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ABSTRACT

Ultra-wideband (UWB) wireless systems rely on signals spanning very wide bandwidths, typically several gigahertz, for information transmission. The distinguishing feature of UWB communications technology is the unrivalled data-rates it provides, with other benefits such as fade resistance and spectral reusability. These characteristics render UWB the technology of choice for a gamut of modern wireless communications applications, including multimedia transmission, personal- and body-area networks, imaging devices, and sensor networks. The use of wide bandwidth signals, however, leads to many complications that necessitate specialised design considerations. The propagation channel and system components acquire frequency-selective characteristics, and their nonlinear and dispersive nature, usually innocuous in a conventional setting, causes signal distortion and erroneous detection. This thesis analyses various aspects of the indoor channel and the distortion to a UWB signal propagating through it. The performance of transmitter and receiver sub-systems is evaluated, with an emphasis on the challenges posed by the large operating bandwidth. The significance of incorporating this knowledge into the system design process is demonstrated, and a novel framework for optimising the performance–complexity tradeoff is presented.

The following are the contributions of this thesis to the state of the art in UWB communications.

- Experimental characterisation of the indoor UWB channel spanning the FCC band (3.1–10.6 GHz)
- Demonstration of the variability of propagation characteristics in the spectral sub-bands
- Assessment of frequency-dependent pathloss and the consequent signal waveform distortion
- Polarimetric analysis of the temporal, spectral and angular channel evolution
- Evaluation of rake receiver performance and its dependence on various channel conditions
- Investigation of the effect of antenna angular-spectral distortion on signal propagation
- A technique for the normalisation of UWB link aberration due to antennas
- Performance evaluation of diversity and spatial multiplexing with multiple-antenna systems
- Design of gigabit wireless links for high data-rate applications or high user density scenarios
- A novel holistic framework for the design of an optimal UWB communications system

*To the radio communications engineer who is my source of inspiration,
my father*

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1. INTRODUCTION

1.1. Ultra-Wideband Communications

Ultra-wideband (UWB) technology offers a fundamentally different approach to wireless communications from the narrower bandwidth systems presently in vogue. The use of low power transmissions at very wide bandwidth, typically spanning several gigahertz, enables spectral reuse and coexistence while providing very high capacity wireless links at short ranges [1]. With other advantages such as resistance to multipath interference and significantly lower spatial and temporal fading, UWB holds the promise of solving critical propagation problems that have plagued conventional radio systems.

Formally, current definitions regard a radio communications system as UWB if its absolute bandwidth exceeds 500 MHz or its fractional bandwidth is larger than 20%, where the fractional bandwidth is the ratio of the bandwidth, Δf , to the centre frequency, f_c , i.e.,

$$f_{fb} = \frac{\Delta f}{f_c} = \frac{2(f_h - f_l)}{f_h + f_l}, \quad (1.1)$$

with f_h and f_l denoting the high- and low-end frequencies.

The Federal Communications Commission (FCC) in USA has allocated the 3.1–10.6 GHz frequency band for the operation of UWB devices for indoor wireless communications [2]. As shown in Figure 1.1, this emission level coincides with the Part 15 regulation for unintentional emitters in that band, which effectively defines the noise floor [2]. The power spectral density of UWB signals is therefore on the order of the noise power spectral density, rendering them noise-like in character and lowering the likelihood of interference and unintended reception.

The concept of UWB communications originated when radio was nascent. The spark-gap transmitter used by Hertz for radio wave generation in 1887 was a crude implementation of a UWB pulse generator, followed by Marconi in 1901 who used the technology for long-distance wireless communications. This transmission technique, however, was abandoned shortly afterwards in favour of narrowband, tuned communications that used modulated carriers to facilitate spectrum sharing.

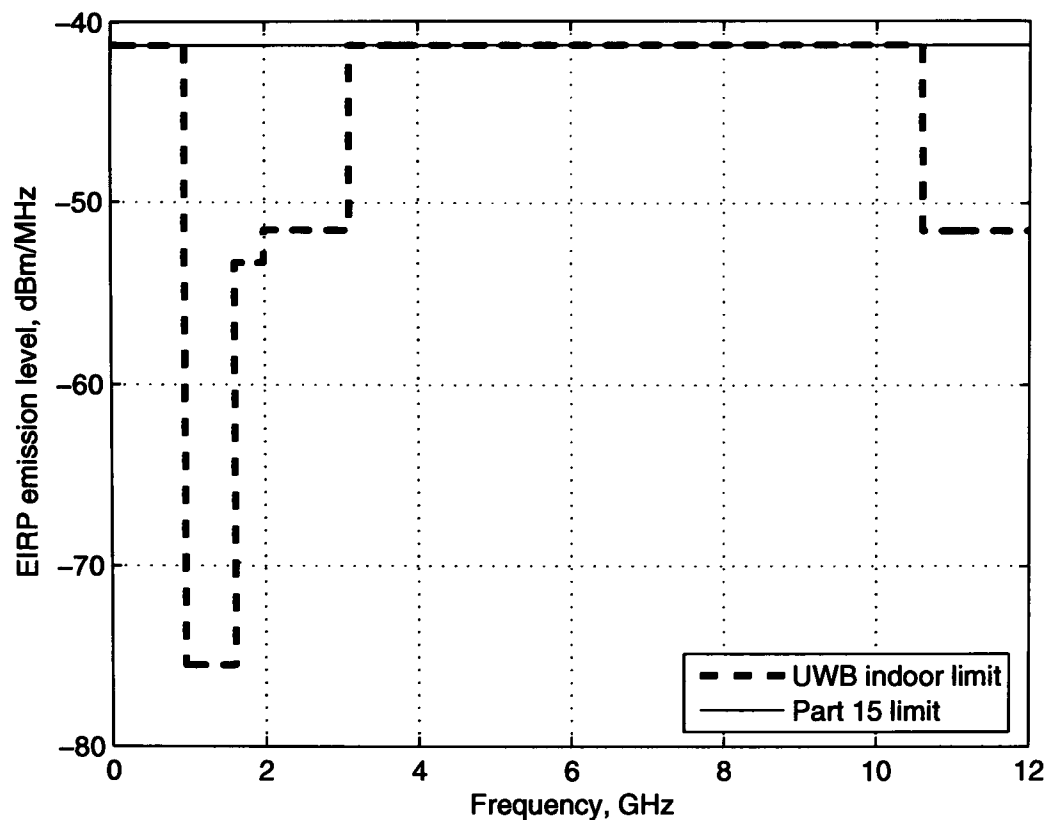


Figure 1.1 The indoor UWB communications EIRP radiation mask defined by the FCC, USA

More recently, research into transient time-domain electromagnetics for radars in the 1960s led to the exploration of pulsed baseband transmission systems [1, 3]. The first patent relating to UWB communications was awarded in 1973 [4] for concepts related to nanosecond impulses and matched filtering. Shortly afterwards, a UWB ground-penetrating radar was developed, followed by other subsurface radar designs. The proposed paradigm shift [5] from the established techniques of carrier-based radio propagation to baseband transmission was initially met with the usual inertia [6, 7] but soon gave way to vigorous exploration as UWB's extensive application became apparent [8]. Much of the subsequent research in this domain was undertaken by the military, owing to its interest in UWB covert communications systems and radars with low probability of interception and detection.

UWB research received renewed attention in the 1990s due to the market imperative for high data-rate communications and the dearth of radio spectrum. The Time-Hopping Impulse Radio system, employing picosecond duration pulses with low duty-cycles such as those shown in Figure 1.2, was proposed as a practical mechanism for implementing a carrierless UWB system with multi-user operation [9, 10]. Following this renewed impetus, multicarrier UWB approaches based on direct sequence spread-spectrum, orthogonal frequency division multiplexing (OFDM) and other signalling

formats were also presented [11]. As these multicarrier techniques provided a natural evolution from well-understood narrower-band approaches, they quickly gained popularity and are currently seen as the key contenders for future high data-rate UWB devices in wireless personal area network (WPAN) applications. Impulse-based UWB has, however, managed to find its niche in sensor networks for low data-rate communications, localisation and ranging applications [12].

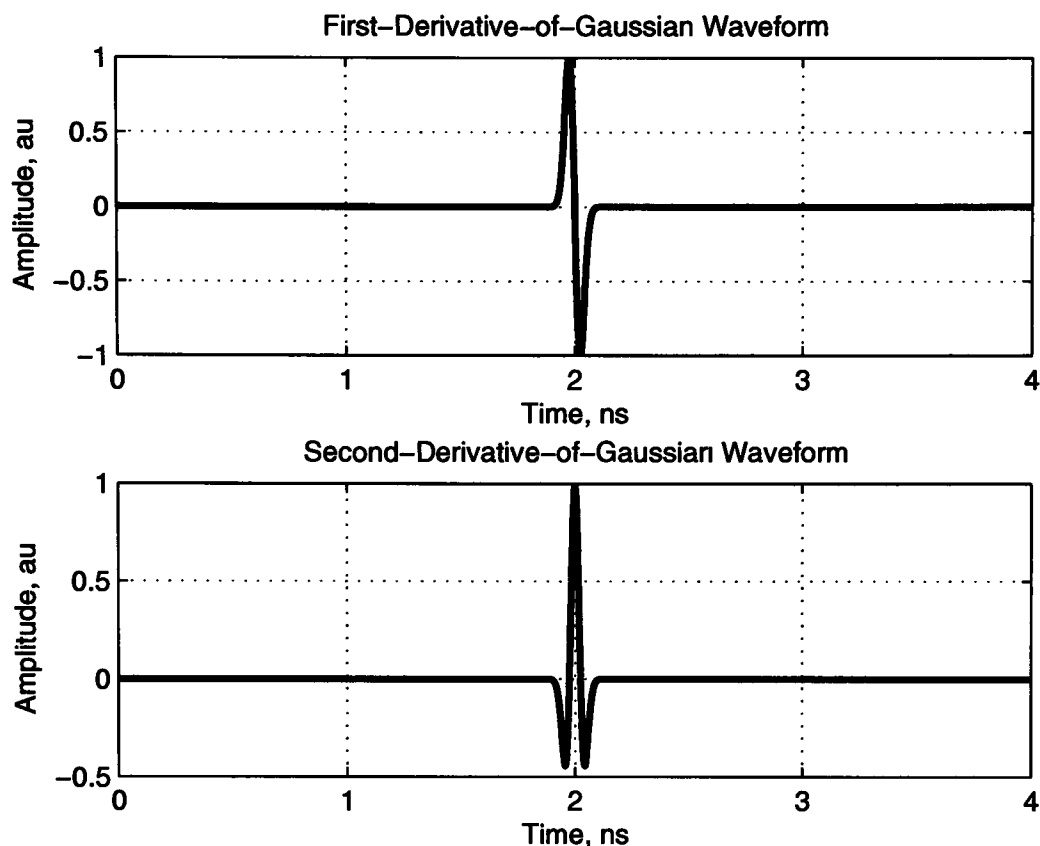


Figure 1.2 Signal waveforms used in pulsed UWB communications systems

1.2. Applications

Due to its high spatial resolution, ability to penetrate a range of materials and undetectable operation, UWB has enjoyed wide application in radars, an arena where its popularity can be reasonably expected to hold its ground in the future. A number of communications applications of UWB can also be foreseen. Indoor communications and measurement systems are expected to account for more than 88% of all UWB units [13]. Exciting new applications of short-range high data-rate communications are emerging in the multimedia home entertainment arena. Wireless networked consumer products such as portable data transfer devices, wireless interfaces, display panels, audio and video receivers, DVD players, video game consoles and joysticks are expected to be in wide domestic use in the near future. Especially resource-hungry among these would be applications requiring high-

definition audio-visual transmission, such as 3-D gaming equipment. There is also a potential market for set-top boxes that would serve as the wireless gateway for the home or office subnet, providing connectivity to a broadband cable or ADSL service provider. The demand for high-rate computer communications will give rise to multiple-access networks with high user density. A similar scenario will be created by wirelessly connected electronic devices pervading the future home and office environments. Some specialised situations would place even higher demands on the connectivity, such as the concurrent delivery of several multimedia datastreams to hundreds of nodes in a conference hall or a passenger aeroplane cabin. With the seamless coexistence, low electromagnetic interference and high spatial density that accompany UWB's high data-rates, it is arguably the only technology that can empower the high-performance short-range wireless applications of the future.

1.3. Research Challenges

UWB communications technology is still in its infancy and a multitude of practical issues require resolution before its deployment. Prominent among those are the challenges that surface due to the large-bandwidth regime, many of which remain uncharted territory. While the advantages of wideband signals are readily recognised, a complete understanding of the fundamental propagation mechanisms of UWB signals in a variety of environments, and the underlying physical phenomena, has hitherto been elusive [14]. As the performance of a communications system depends on the accuracy of channel modelling and prediction, there is an urgent requirement for reliable and comprehensive UWB channel characterisation. The performance of many system components, such as antennas, amplifiers and detectors, is also sensitive to frequency, and varies under wideband signals. Another important issue is UWB's interference and coexistence with other electronic equipment, which is an aggressively debated topic at present [15]. The capacity limits in multi-user environments and picocellular network architectures is a practical consideration in networked application. Hardware constraints related to pulse generation and analogue-to-digital conversion limit the achievable data rates. These and several other outstanding issues merit further research before UWB systems can be commercially produced and their true performance potential realised.

1.4. Problem Statement

The importance of an investigation into the physical layer aspects of UWB wireless communications with an aim to understand the propagation channel and sub-system characteristics leading to efficient system design is apparent in the light of the above discussion, and is the goal of this thesis. The first part of the thesis undertakes a study of the UWB channel characteristics, emphasising its novel attributes due to the high frequency selectivity. The spectral, spatio-temporal, angular and polarisation characteristics of the UWB channel are characterised statistically. The second part evaluates the performance of various system components, such as antennas, rake receivers and multiple-antenna arrays, under UWB signals. Comprehensive experimental validation of the ideas developed is provided throughout. Ultimately, the channel knowledge gained in the initial chapters is coupled with the systems study to obtain a novel framework for the design of optimised communications systems with high performance and low complexity.

1.5. Thesis Organisation

The rest of this thesis is organised as follows. Chapter 2 presents a full-band characterisation of the UWB channel, with a focus on multipath propagation and small-scale fading, based on channel measurements and deterministic modelling. Chapter 3 establishes the variability of channel characteristics with frequency and quantifies its impact on signal propagation. Chapter 4 undertakes a polarimetric study of the channel with a view to evaluate the extent of depolarisation experienced by a UWB signal. The performance of a UWB rake receiver is related to the signal bandwidth and number of resolved multipaths using measurement results in Chapter 5. Chapter 6 investigates the signal dispersion caused by the frequency-dependent behaviour of antennas, and presents a technique to remove the link aberration. Chapter 7 analyses the performance of multiple-antenna UWB systems for diversity and spatial multiplexing. A framework for optimal system design taking into account the channel and sub-system characteristics is presented in Chapter 8, along with a set of illustrative case studies derived from the preceding characterisation. The thesis concludes with Chapter 9.

2. THE ULTRA-WIDEBAND PROPAGATION CHANNEL

2.1. Introduction

A thorough understanding of the propagation channel is vital for the design of an efficient wireless communications system. The accuracy of the propagation model directly influences the quality of service of the system designed on the basis of that model. The key objective of channel modelling is to determine the coverage and capacity limitations in various propagation environments without system implementation. This facilitates the selection of appropriate signalling schemes and receiver designs.

The flat Rayleigh-fading channel model [16] is used to represent systems with bandwidths narrow enough to cause the superposition of a large number of multipath components, as the central limit theorem leads to Rayleigh amplitude statistics for the combined signal envelope. Wideband systems with bandwidths up to several MHz can, however, resolve some of the multipaths, but Rayleigh or Rician amplitude statistics still apply in cluttered multipath channels. Individual multipath resolution is much more likely at ultra-wide bandwidths, as a 7.5 GHz wide UWB signal can resolve pathlength differences up to 4 cm, or time-of-arrival differences of 133 ps. Due to the smaller number of overlapping multipaths in each time-bin, the central limit theorem no longer applies, and the complex amplitude of the received signal cannot be modelled with a two-dimensional Gaussian random variable. Furthermore, at sufficiently high bandwidths and low scattering, the channel impulse response can deviate from a continuum of exponentially decaying incident power, and can indeed be more accurately represented by a sparse tapped delay line in some environments. In addition, individual waveform distortion is observed in the multipath components due to frequency selectivity of the channel [17]. This often results in pulse broadening, causing adjacent channel taps to be influenced by a single physical multipath arrival.

As a consequence of these phenomena, the fading and dispersion properties of the channel differ greatly from narrowband channels. While several empirical and deterministic analyses of the UWB channel have been carried out, some of the fundamental questions of UWB propagation have not been satisfactorily answered to date as acknowledged in [14]. These open issues range from the

investigation of system performance across the full band to the quantification of multipath propagation characteristics in the UWB regime. This vacuum in the current understanding of UWB channels provides the motivation for the work reported in this chapter, which begins by highlighting the important characteristics of the UWB channel. Key results from some of the recent UWB channel studies and models are summarised, and their shortcomings are pointed out. Empirical channel characterisation techniques are discussed. Experimental results pertaining to channel fading and dispersion are obtained, and the empirical distributions of delay spread and coherence bandwidth are analysed. A probabilistic and information theoretic treatment of multipath propagation is presented. An efficient technique for deterministic channel modelling based on geometric optics is also proposed.

2.2. The UWB Channel: A Review

It is well known that the channel characteristics of wideband channels can be summarised by parameters like the delay spread and pathloss [18, 19]. These parameters are analysed statistically from a large number of measurements performed in a variety of propagation environments for channel characterisation. While several measurement-based UWB channel experiments have been reported in recent years, nearly all of them have characterised only a portion of the UWB band. Very few studies have presented results that span the FCC-allocated UWB band. This is especially true for time-domain analyses, as the generation and reception of ultra-short pulses is difficult. Furthermore, higher-frequency signal components undergo greater pathloss compared to the lower-frequency components, requiring higher receiver efficiency in the upper frequency bands. In addition, CMOS technology, which is appealing for high-volume production, is available only for frequencies up to 5 GHz at the time of this writing. Therefore, in view of immediate commercial interest, most of the recent UWB propagation research has focussed on the 3–5 GHz band [14]. This section summarises the key results from some of the recent UWB channel characterisation efforts.

Initial time- and angle-of-arrival modelling results using UWB channel measurements were reported in [17], which also postulated individual pulse shape distortion of the multipaths. The measurements in [20] showed that the first arriving multipath is not always the strongest, as the

interference of multiple paths can sometimes produce a dominant composite multipath. The robustness of UWB systems to multipath was shown by means of the low fade range of 5–6 dB [17, 20, 21], compared to 35 dB for conventional narrowband indoor systems [22]. Various studies have reported good fits for the multipath amplitude and power statistics with various probability distributions, as summarised in Table 2.2. It was shown in [23] that the blockage due to the human body creates a deep null of about 24 dB in a low multipath environment, such as a large auditorium, but this null reduces to 7 dB in a dense multipath environment such as an office.

Table 2.1 UWB channel statistics reported in literature

Channel Parameter	Reported Value
Small-scale fading range	5–6 dB [17, 20, 21]
Pathloss exponent	1.91 (LOS) and 3.66 (NLOS) [24]; 2.4 [25]; 1.63 (LOS office), 1.79 (LOS residential), 3.07 (NLOS office) and 4.58 (NLOS residential) [26]; 1 (LOS corridor), 1.2 (LOS industrial), 1.5–2 (LOS office and residential), 2–2.5 (NLOS industrial), 3–4 (NLOS office and residential) and 4–7 (NLOS harsh) [27];
Multipath power distribution	Gamma [28]
Multipath amplitude distribution	Rician (LOS) and gamma (NLOS) [29]; Nakagami [25]; lognormal [30]; Rayleigh (industrial) [31]
Time-decay constant distribution	Lognormal [28]
Maximum excess delay	70 ns [32]
Rms delay spread	4.7 (LOS) and 8.2 (NLOS) [32]
Number of multipaths	60–80 (NLOS) [32]
Multipath correlation coefficient	0.2 [28]
Number of multipath clusters	1–2 [32, 33]; 3 [34]; 1–4 [35]

Pathloss exponents ranging from 1 to 7 have been reported, as listed in detail in the table. The pathloss exponent was modelled as a Gaussian random variable with a large standard deviation in [32]. A double-slope pathloss model was proposed based on measurement results in [36] and on theoretical analysis in [37, 38]. The rms delay spread was found to increase exponentially with transmitter–receiver separation, d , as $d^{0.26}$ and $d^{0.36}$ in LOS and NLOS environments respectively [32]. Modelled as a Gaussian random variable, it was shown to increase with pathloss [32, 39]. Similarly, a linear dependence of the rms delay spread on distance was observed in [36], with a slope of 3 ns/m.

Uncorrelated scattering and independent multipath fading was reported in [25]. Clustered multipath arrival has been observed, as closely arriving multipaths originating from a single object can often be resolved with UWB. The number of clusters depends on the propagation environment and measurement bandwidth. A relatively small number of clusters have been reported in the indoor environment. It was found that industrial environments present a dense multipath channel even at the full UWB bandwidth of 7.5 GHz [31], while [34] suggested that residential environments can be modelled with a sparse tapped delay line.

The IEEE 802.15.3a working group has adopted a channel model for high-data rate, short-range UWB communications, an overview of which has been presented in [40]. In this model, the time-of-arrival statistics are derived from the clustered Saleh-Valenzuela channel model [41]. A complex tapped delay line model was not used, as complex baseband representation of carrierless, real UWB signals is non-intuitive. The model distinguishes between cluster arrivals and ray arrivals within those clusters. The clusters and ray arrival rates are Poisson processes, while the powers follow a double exponential decay. The path amplitudes are lognormally distributed, and a shadowing term accounts for the received power loss due to LOS blockage. The freespace pathloss expression, computed at the centre frequency, is used, and antenna gain is assumed to be independent of frequency. Mathematically, the channel impulse response is given by

$$h(t) = X \sum_{l=0}^L \sum_{k=0}^K \alpha_{k,l} \delta(t - T_l - \tau_{k,l}) \quad (2.1)$$

where X is the lognormal shadowing, $\alpha_{k,l}$ are the multipath gains, T_l is the delay of the l^{th} cluster, $\tau_{k,l}$ is the delay of the k^{th} multipath relative to the l^{th} cluster, and δ is the Dirac delta function. Four distinct

propagation scenarios have been identified: CM1 refers to an LOS channel with 0–4 m antenna separation, CM2 is an NLOS channel with 0–4 m separation, CM3 is an NLOS channel with 4–10 m separation, and CM4 is a heavy multipath environment. Some of the significant channel parameters from this model have been listed in Table 2.2 [40].

Table 2.2 Channel fading and dispersion statistics from the IEEE 802.15.3a channel model

Channel Parameter	Channel Type			
	CM1	CM2	CM3	CM4
No. of multipaths $L_{85\%}$	20.8	33.9	64.7	123.3
Rms delay spread, ns	5	8	15	25
Mean delay, ns	5.0	9.9	15.9	30.1

2.3. Empirical Channel Characterisation

Wireless channel models are based on large sets of statistics gathered from channel measurements as described above. In comparison with analytical and simulation models, measurement-based investigations provide actual channel propagation information free of any potentially erroneous assumptions. Thus channel measurement is considered more credible than the alternative approaches, and is employed for the ultimate validation of a model. A variety of techniques can be employed for channel measurement. Time-domain pulse transmission and reception enables the direct measurement of the channel response to the pulse. The generation of ultra-short pulses is, however, quite challenging, but this technique has been used to perform UWB channel measurements in [17, 25]. On the other hand, frequency-domain channel sounding with a vector network analyser (VNA) involves sweeping across a frequency range with a chirped continuous-wave signal and recording the complex transfer function of the channel [42].

The results in this thesis are based entirely on channel measurements, unless otherwise specified. Coherent channel measurements in the UWB band were carried out using a VNA arrangement in a

typical indoor office environment as shown in Figure 2.1. The $S_{21}(f)$ complex transmission response of the channel was measured. The FCC-approved frequency range with bandwidth $B=7.5$ GHz between $f_l=3.1$ GHz and $f_h=10.6$ GHz [2] was sounded at $n_f=1601$ discrete, uniformly-spaced frequency points, so that the frequency resolution,

$$\Delta f = \frac{f_h - f_l}{n_f - 1}, \quad (2.2)$$

was 4.68 MHz. The output port of the VNA was connected to the transmitting antenna, and the input port to the receiving antenna through a low-noise amplifier (LNA) that provided 30 dB gain across the UWB band. In this experiment, the channel was the device-under-test (DUT). Thus prior to the measurement, the DUT was removed, i.e. the two ends of the cables were disconnected from the antennas and joined together for calibration. This procedure allowed the frequency-dependent attenuation and phase distortion of the cables, connectors and LNA to be removed from the measurement data. The propagation environment consisted of a 6m x 6m room with block walls, glass windows, and several wooden and metallic objects. A large number of reflective surfaces provided a rich scattering environment. Both line-of-sight (LOS) and non-line-of-sight (NLOS) scenarios were measured. The blockage in the case of the NLOS scenario was created by placing a large aluminium sheet in the direct propagation path, grounded to prevent secondary wave generation. The stationarity of the environment was maintained throughout the duration of the measurement by ensuring that there were no human or other movement in the room.

Two identical discone antennas were used to transmit and receive the signal [43]. The antennas were linearly polarised and nominally omni-directional in the azimuth plane. The return loss of the antennas in the UWB band was well below -10 dB. They were placed at a 1.5m height above the floor such that their electric fields were vertically polarised. The transmitting antenna was fixed at a location close to one corner of the room. A computer-controlled horizontal positioning grid was used to control the location of the receiving antenna within a 1m^2 area. The distance between the transmitting antenna and the centre of the grid was 4.5 m. The measurement was conducted at $n_x=n_y=100$ positions in each direction of the square grid, each point being $\Delta x=\Delta y=0.01\text{m}$ apart. Thus a total of $n_x n_y=10,000$

complex transfer functions were measured for each propagation scenario. The resolution of the grid points was chosen to meet the Nyquist criterion for the smallest wavelength, i.e.,

$$\Delta i \leq \frac{c}{2f_h}, \quad i = \{x, y\}, \quad (2.3)$$

where $c = 3 \times 10^8$ m/s is the speed of light in freespace. The size of the grid aperture was over ten times larger than the largest signal wavelength,

$$n_i \Delta i \geq 10 \frac{c}{f_l}, \quad i = \{x, y\}, \quad (2.4)$$

in order to obtain independent and statistically reliable channel data.

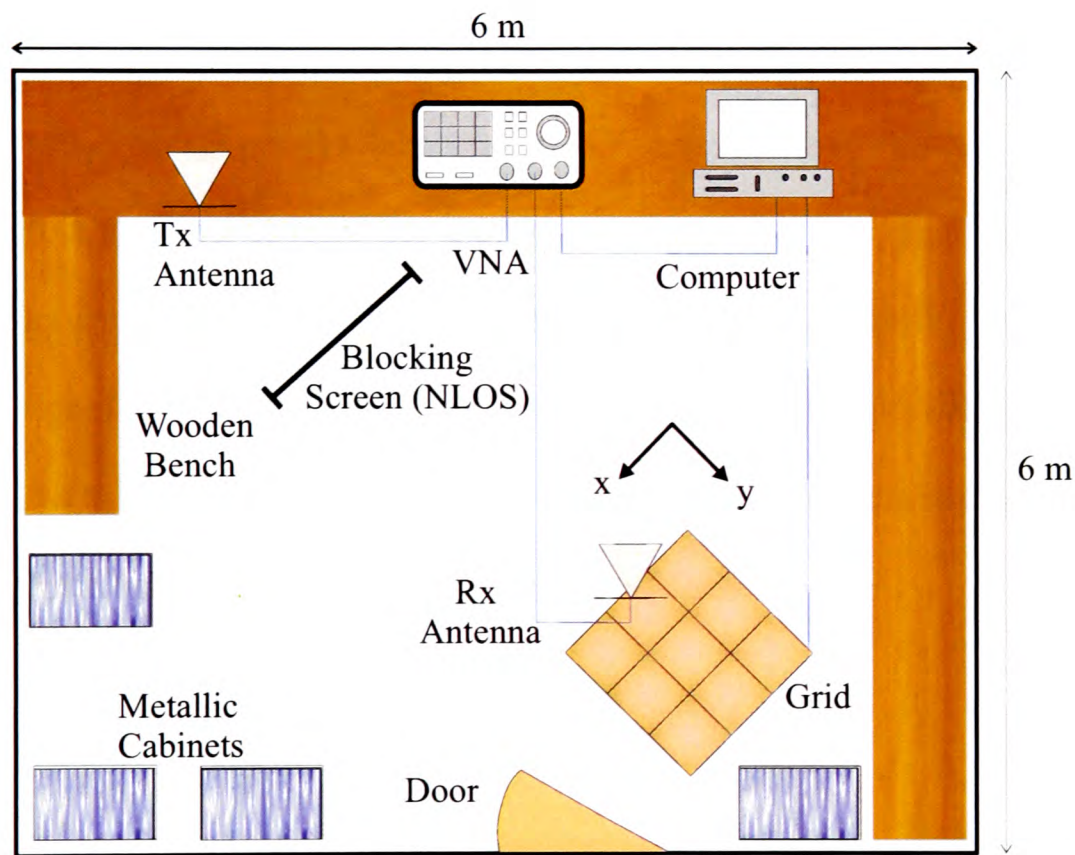


Figure 2.1 Experimental configuration for VNA-based synthetic aperture channel measurement

The complex channel transfer function thus measured can be represented mathematically by

$$G_{x,y}(f) = \sum_{k=0}^{n_f-1} A_{x,y,k} e^{j\psi_{x,y,k}} \delta(f - k\Delta f), \quad (2.5)$$

where $A_{x,y,k}$ and $\psi_{x,y,k}$ are the magnitude and phase responses at the k^{th} frequency component measured at receiver location (x,y) . The corresponding complex impulse response (CIR) of the channel is related to its frequency transfer function through the inverse discrete Fourier transform, and is evaluated as

$$g_{x,y}(\tau) = \sum_{k=0}^{K-1} a_{x,y,k} e^{j\varphi_{x,y,k}} \delta(\tau - k\Delta\tau) \quad (2.6)$$

where $a_{x,y,k}$ and $\varphi_{x,y,k}$ are the magnitude and phase responses in the k^{th} time-bin, and

$$\Delta\tau = \frac{1}{n_f \Delta f} \quad (2.7)$$

is the time resolution. The time delay variable τ represents the time referenced to the time-of-arrival of the first multipath component, in order to remove the propagation delay. The CIR can then optionally be power-normalised by its Frobenius norm to remove signal attenuation due to large- and small-scale fading, if it is not relevant to the study, to obtain

$$h_{x,y}(\tau) = g_{x,y}(\tau) \left(\sum_{k=0}^{K-1} a_{x,y,k}^2 \right)^{-2} = \sum_{k=0}^{K-1} \alpha_{x,y,k} e^{j\varphi_{x,y,k}} \delta(\tau - k\Delta\tau). \quad (2.8)$$

From the inspection of individual CIRs, it is found that the received signal reaches a constant level at about 30 dB below the amplitude of the strongest multipath, and no significant multipath arrivals can be detected afterwards. Thus to remove the noise and residual multipath signals with an adequate margin, the peak signal level in the CIR is taken as reference and $c_{th}=25$ dB threshold is used. It is verified that with this threshold, the time-of-arrival of the first path is estimated correctly for over 99% of the CIRs. The signal threshold level for each CIR is thus found as

$$\alpha_{x,y,th} = \frac{\max_k(\alpha_{x,y,k})}{10^{c_{th}/20}}. \quad (2.9)$$

After removing the signals that fall below this threshold, the multipath components are identified through a local maxima detection algorithm. The magnitude, phase and delay for each multipath in the measured CIRs is extracted for channel characterisation and system performance prediction.

2.4. Channel Transfer Function and Impulse Response Analysis

In view of the shortcomings of the channel characterisation efforts to date mentioned in the previous sections, some results from UWB channel measurements are presented here. The analysis is based on the full-band channel sounding described in the previous section.

Figure 2.2 shows the magnitude and phase response of the LOS channel at one measurement location extracted from the measured $G_{x,y}(f)$ as a representative example. A number of magnitude nulls and phase discontinuities can be observed in the frequency response, so that the channel exhibits frequency-selective fading behaviour. This is one of the defining characteristics of indoor UWB channels with a large number of incident multipaths.

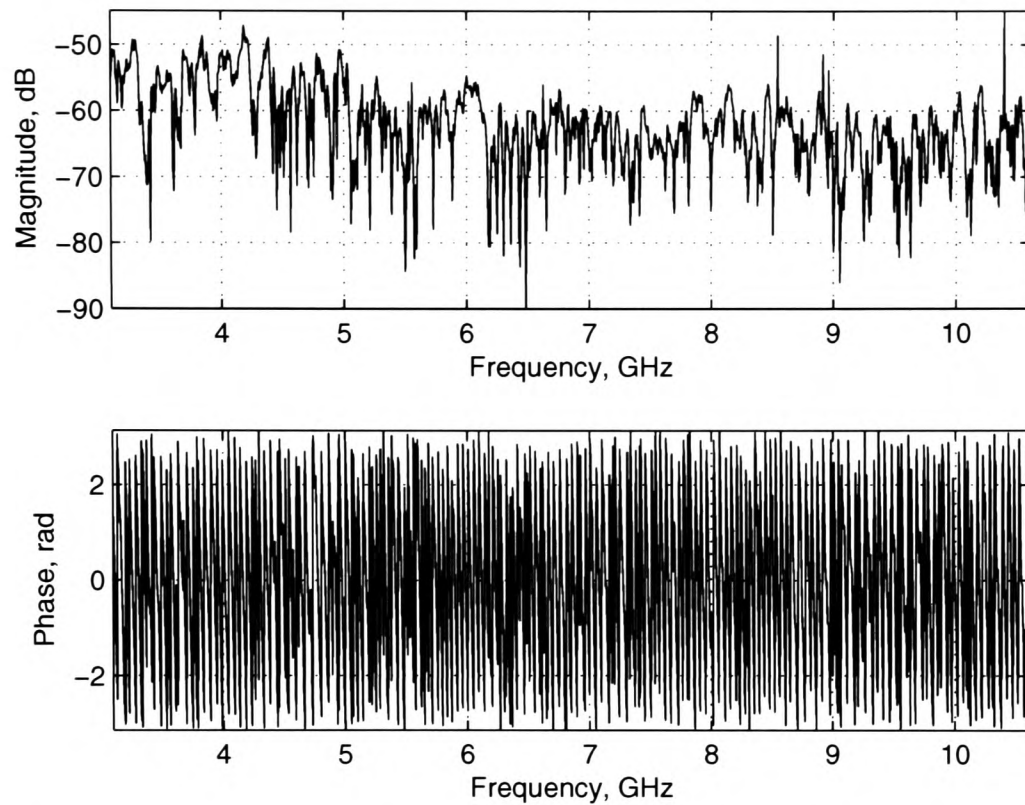


Figure 2.2 The complex channel transfer function of an indoor LOS UWB channel, measured at a single location, as a typical example

Figure 2.3 shows the CIR for this LOS UWB channel normalised to the peak power level with the propagation delay removed. A number of clustered multipaths are apparent. In this example, the direct path is not the strongest path, providing evidence of multipath overlapping in some time-delay bins. The noise floor is reached after approximately 65 ns.

The average power delay profile (PDP), derived from the ensemble average of the data measured on the grid, is shown in Figure 2.4. The LOS and NLOS PDPs correspond closely, with the only difference being the blockage and attenuation of the direct path in the NLOS PDP. The delays and powers of the dominant multipath clusters, as well as the time-decay profile, bear great similarity.

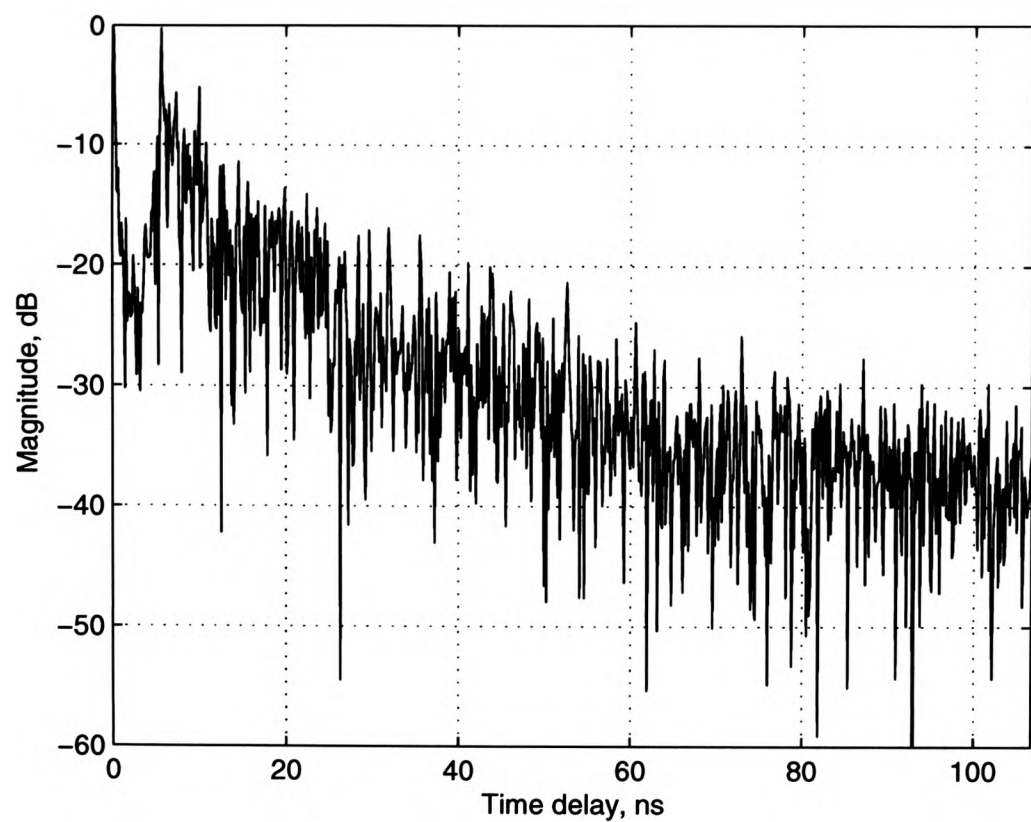


Figure 2.3 The peak-normalised impulse response of the above channel without propagation delay

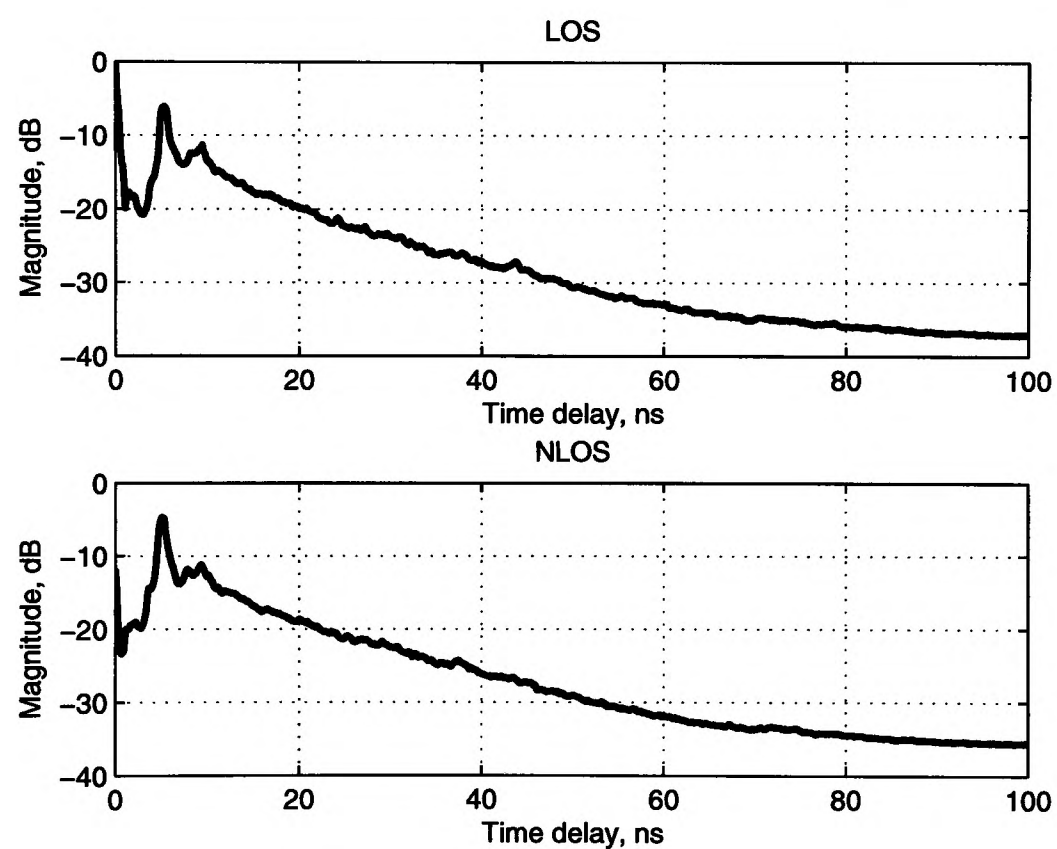


Figure 2.4 Average power delay profile normalised to the LOS peak power

2.5. Coherence Bandwidth

The coherence bandwidth of the channel is closely related to the frequency auto-correlation function or the spaced-frequency correlation function [44]. It can be evaluated from

$$r(\Delta f) = E[G(t_1, f)G^*(t_1, f - \Delta f)], \quad (2.10)$$

where $G(t_1, f)$ is the channel transfer function at time t_1 , $(.)^*$ denotes complex conjugation, $E[.]$ denotes expectation, and r is the frequency-domain channel correlation. Then the coherence bandwidth, B_c , satisfies the condition

$$r(B_c) \geq C_{cor}, \quad (2.11)$$

where C_{cor} is the correlation threshold. Since the assessment of coherence and distortion is subjective, there is no universally accepted value for C_{cor} . Some choices of values popular in literature for C_{cor} are 0.6, 0.7 and 0.9. With $C_{cor} = 0.6$, the UWB coherence bandwidth comes out to be 29 MHz and 31 MHz for LOS and NLOS channels respectively. Thus the coherence bandwidth of the UWB channel is only 0.4% of the full bandwidth, imparting it strongly frequency-selective characteristics. However, signals that occupy bandwidths smaller than 30 MHz, such as the sub-carriers in multiband-OFDM UWB systems [45], experience flat fading.

2.6. Time Dispersion and Small-Scale Fading

The time dispersion of a channel is characterised in terms of the time delay spread, which depends on the location and composition of scatterers and the overall geometry of the propagation environment. It provides information about frequency-selective fading and limits the highest achievable data-rates. The delay parameters are extracted directly from the PDP. The mean excess delay is the first central moment of the PDP and is defined in terms of the multipath magnitudes and delays as [16]

$$\tau_m = \frac{\sum_{k=0}^{K-1} \alpha_k^2 \tau_k}{\sum_{k=0}^{K-1} \alpha_k^2} \quad (2.12)$$

The α_k in (2.12) are the $\alpha_{x,y,k}$ in (2.8). The rms delay spread is the second central moment of the PDP,

$$\tau_{rms} = \sqrt{\tau^2 - \tau_m^2}. \quad (2.13)$$

The maximum excess delay τ_{max} is the delay of the last resolved multipath above the noise threshold,

$$\tau_{max} = \sup_k(\tau_k). \quad (2.14)$$

For the sake of stochastic analysis, the delay parameters extracted from band-limited measurements can be treated as discrete random variables. With a large number of CIRs, however, they can be approximated with continuous random variables. The probability density functions (PDFs) of τ_m , τ_{rms} and τ_{max} are estimated from the LOS and NLOS channel measurements. A Gaussian window is applied for the smoothing of the probability estimates. The PDFs of τ_m are shown in Figure 2.5. The NLOS channel experiences greater multipath dispersion, as obvious from the higher mean and standard deviation of the NLOS τ_m PDF in Table 2.3. The mean value of τ_m is 8 ns for LOS and 10 ns for NLOS. From inspection, the mean excess delay can be well approximated with a Gaussian random variable.

A similar trend is observed in the rms delay spread, with the NLOS channel exhibiting a higher standard deviation. The τ_{rms} can also be approximated with a Gaussian variable with a mean of 7.8 ns. The distributions of τ_{max} , however, exhibit long tails. This is especially true in the NLOS case, where the excess delay appears to be as high as 105 ns for a small proportion of the CIRs. The mean τ_{max} is 52 ns. The probability density estimates of τ_{rms} and τ_{max} are shown in Figure 2.6 and Figure 2.7 respectively.

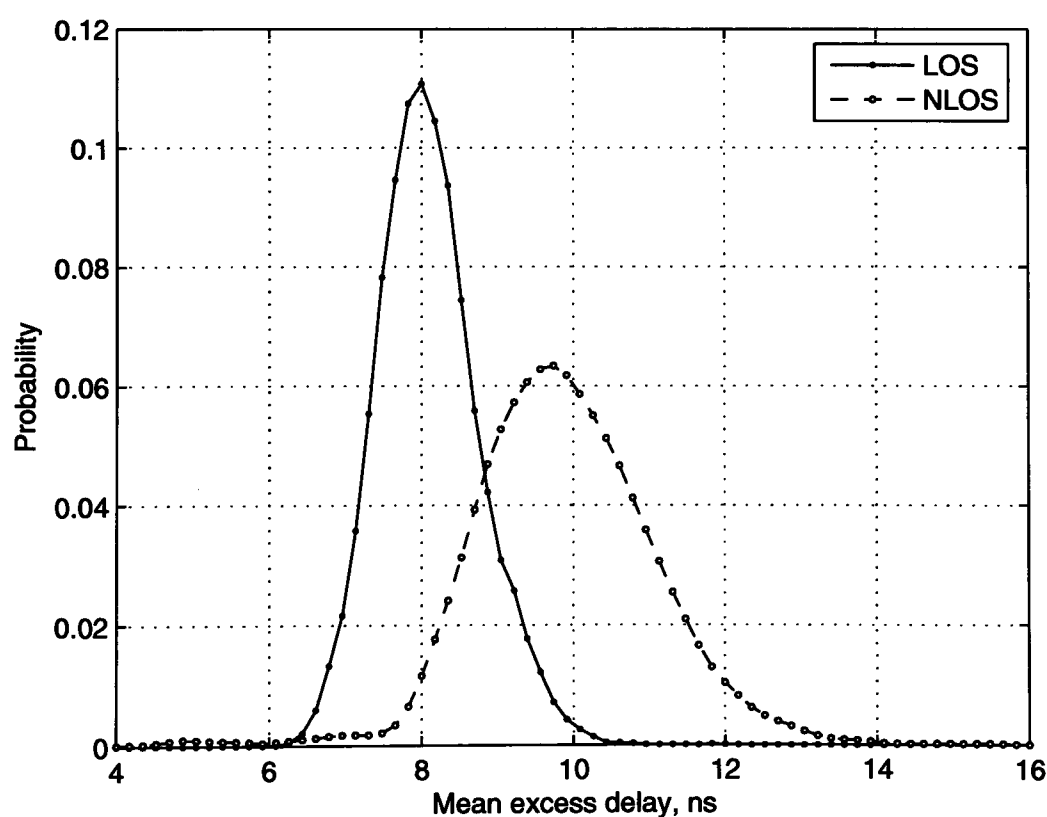


Figure 2.5 Probability density estimate of the mean excess delay

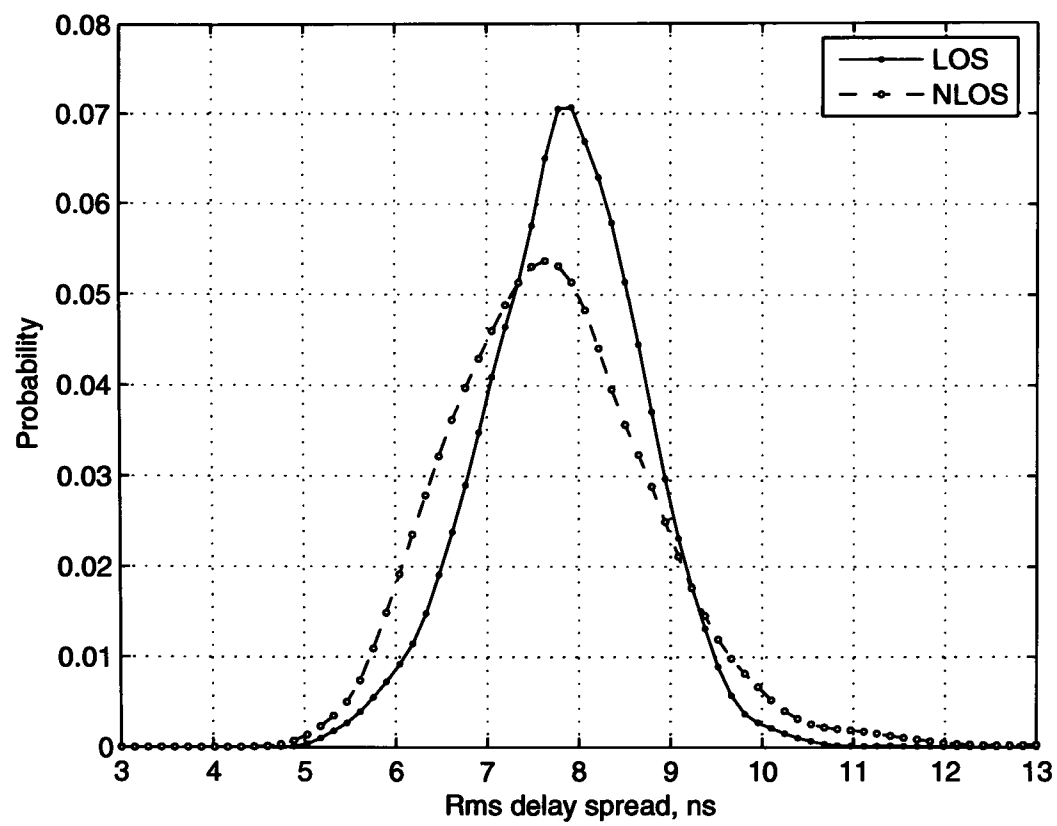


Figure 2.6 Probability density estimate of the rms delay spread

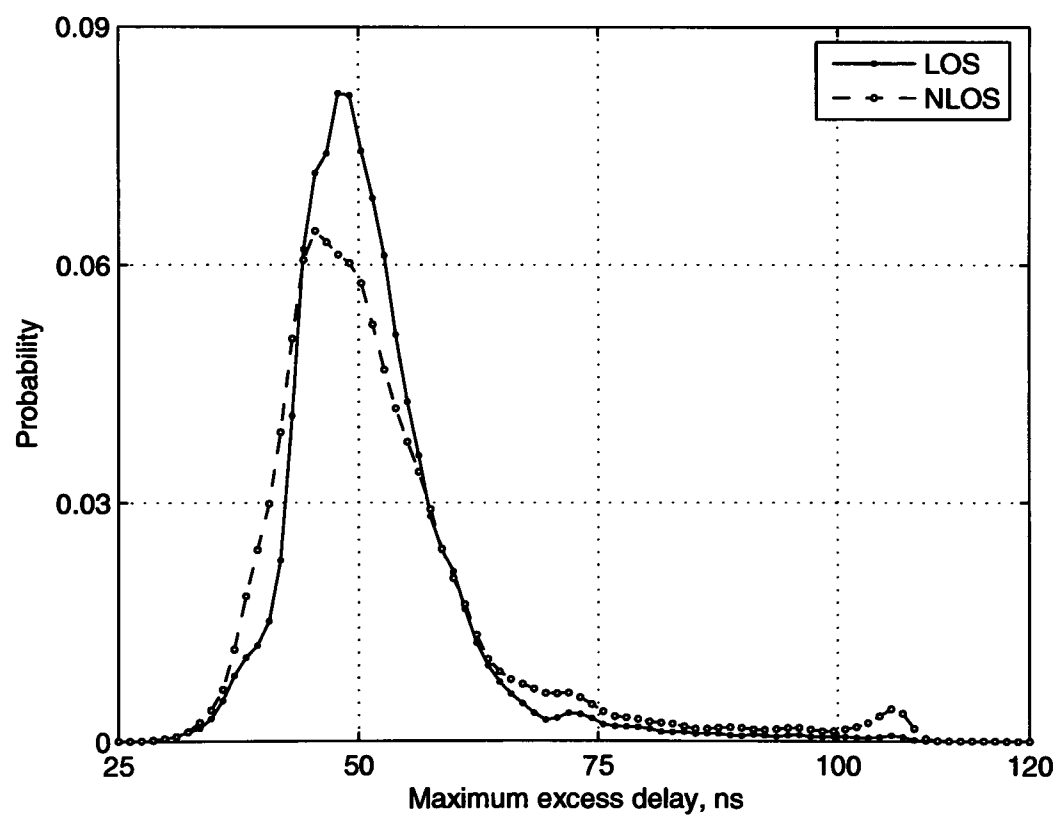


Figure 2.7 Probability density estimate of the maximum excess delay

The first order statistics of the mean delay, rms delay spread and maximum excess delay are summarised in Table 2.3. This characterisation is based on the full UWB band and is thus superior to

that in Table 2.2, which is based on measurements spanning, in most cases, the lower portion of the FCC-allocated UWB band.

Table 2.3 Measured full-band UWB channel dispersion statistics with 25 dB noise threshold

Delay Parameter	LOS		NLOS	
	Mean	Std.	Mean	Std.
Mean excess delay, ns	8.10	0.68	10.02	2.18
Rms delay spread, ns	7.85	0.93	7.84	1.78
Max. excess delay, ns	51.31	8.91	52.61	12.62

2.7. Multipath Propagation

Multipath propagation in wireless channels often leads to fading unless specialised signal processing techniques are applied. Antenna diversity can be used effectively to reduce fading and extend the coverage range of a UWB communications system. The optimal MIMO antenna separation and the consequent improvement in link quality depends, among other factors, on the system bandwidth, operating frequency, and multipath. This raises questions about the impact of wide bandwidth on the design and performance of antenna array systems. Similarly, the SNR gain of a rake receiver depends on the number of incident and resolved multipaths, their amplitude distributions and mutual independence. These are in turn determined by the multipath environment and bandwidth, and a clear understanding of the role of these is important for the design of UWB systems that can achieve the optimal performance in dense multipath channels. Therefore in this section, the multipath properties of the UWB channel are analysed using measured channel data.

The CIR yields the multipath profile

$$m_{x,y,l} = \sum_{l=1}^L \alpha_{x,y,l} e^{j\varphi_{x,y,l}} \delta_{\tau,\tau_{x,y,l}} \quad (2.15)$$

where l is the time-bin where a multipath is detected and L is the total number of multipaths in the CIR. The mean multipath profile provides an alternative representation to the average PDP. It is the ensemble mean

$$\bar{m}_l = \frac{1}{n_x n_y} \sum_{x,y=1}^{n_x, n_y} m_{x,y,l}, \quad (2.16)$$

where l is the multipath index.

Figure 2.8 shows $\bar{m}_l$ for the LOS and NLOS indoor UWB channels for $L=100$, normalised by the global peak signal level given by

$$\bar{m}_{\max} = \sup_l (\bar{m}_l). \quad (2.17)$$

As observed earlier in the average PDP, the direct path in LOS is much stronger than that in NLOS because of the presence of the obstruction in the latter. The rest of the scatterers appear to be largely unaffected by the NLOS blockage, as was also noted earlier. Most noticeably, the strongest path has the same power and delay in both types of channels. The LOS multipath profile decays more rapidly than the NLOS profile. Strong multipath clustering can also be observed in both scenarios.

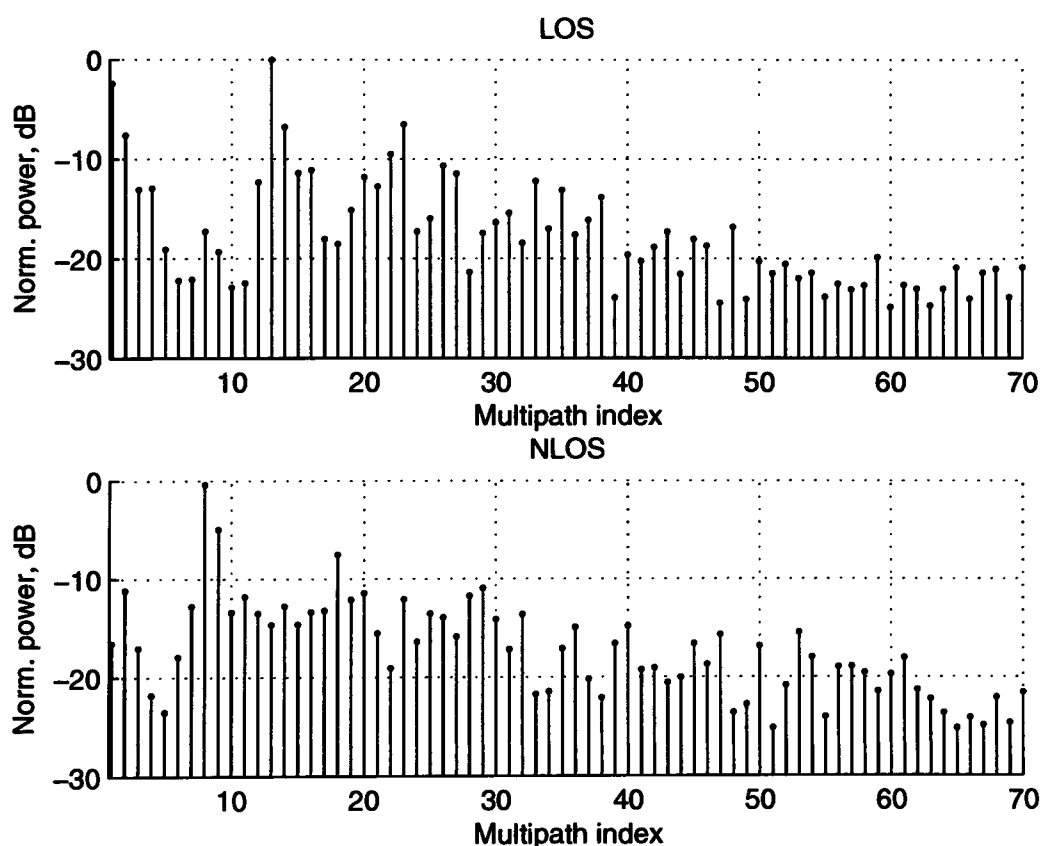


Figure 2.8 Measured average multipath powers of LOS and NLOS channels, normalised to the power of the dominant multipath component

The reciprocal relationship between the time-bin width and bandwidth, B , suggests a monotonic increase in L with B due to the increased time resolution, with constant centre frequency $f_c=6.85$ GHz. From experimental analysis, L is indeed found to vary approximately linearly with B , as shown in Figure 2.9, with slopes of 10.85 and 10.98 for LOS and NLOS channels. The errorbars indicate that the standard deviation increases with the number of resolved multipaths and bandwidth. This relationship can be used in choosing the required rake receiver diversity order for a given UWB system bandwidth.

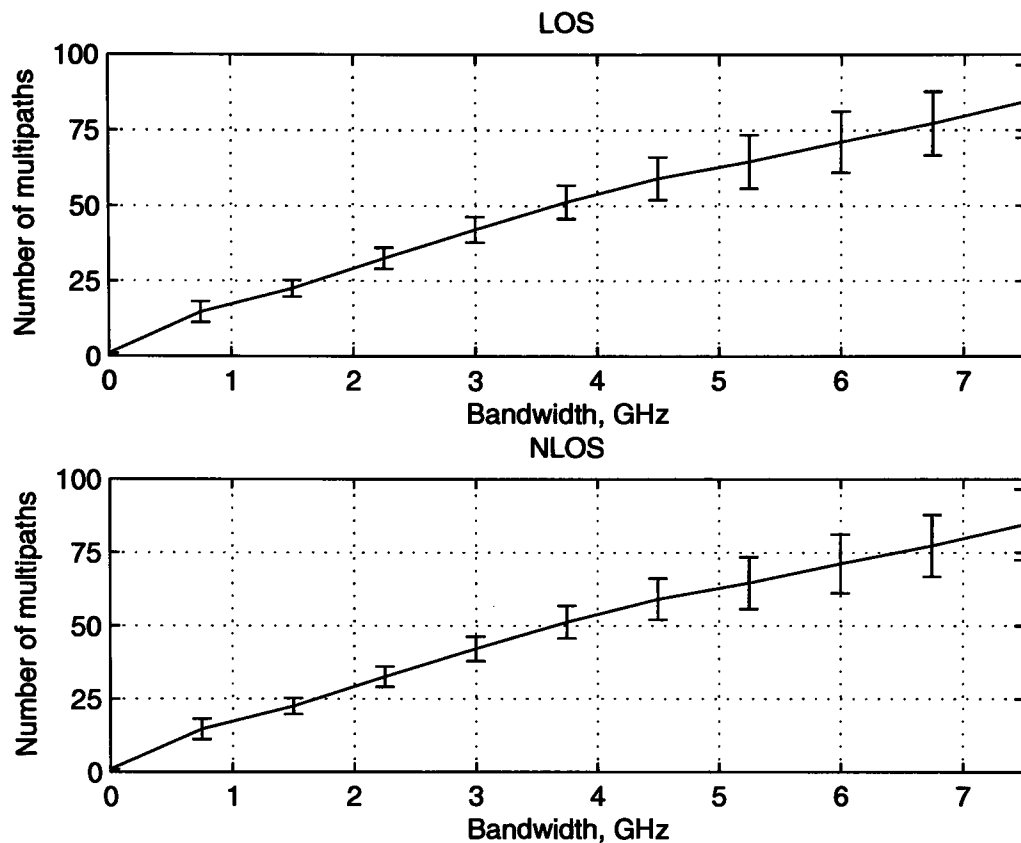


Figure 2.9 Variation of number of resolved multipaths with system bandwidth

Considering that the multipath arrivals can be represented by discrete random variables with finite probabilities, a measure of multipath diversity can be obtained in the form of multipath entropy. The entropy corresponding to the multipath profile at receiver location (x, y) can be calculated as [46]

$$\mathcal{H}_{x,y} = -\sum_{l=1}^L m_{x,y,l}^2 \log_L m_{x,y,l}^2, \quad (2.18)$$

and can be used to derive the effective number of multipaths, or the multipath degrees of freedom, by the relation

$$\hat{L}_{x,y} = L_{x,y}^{\mathcal{H}_{x,y}}. \quad (2.19)$$

The CDFs of $\hat{L}_{x,y}$ for the measured LOS and NLOS channels are shown in Figure 2.10. The mean $\hat{L}_{x,y}$ is 28. This is a high multipath diversity order, and an appropriate combining scheme can use it to reduce the outage probability to a large extent. Also, the multipath diversity is offered effectively, on average, by only a fraction of the received multipaths. This information can be used to obtain an estimate for the diversity order of a sub-optimal rake receiver as discussed in Chapter 5. From Figure 2.10, the number of effective multipaths is much smaller than the number of resolved multipaths, with a mean percentage of about 30%. It can be inferred that the multipaths have finite fading correlation, and the usual assumption of uncorrelated scattering in indoor UWB channels is only an approximation. Also, this result indicates that from the perspective of reduction in the probability of outage, an H-S/MRC rake receiver is more suitable than an MRC rake that, by definition, combines all resolved multipaths when the performance–complexity tradeoff is a consideration (see Chapter 5).

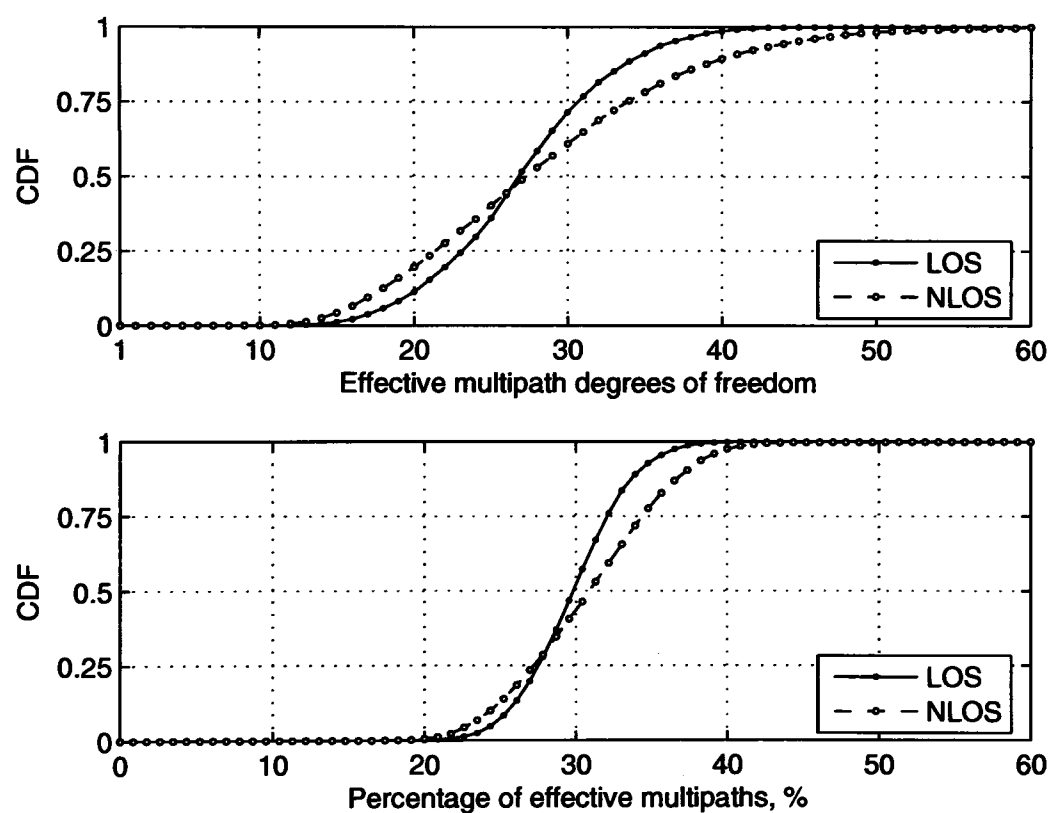


Figure 2.10 CDFs of multipath degrees of freedom and percentage of effective multipaths

The multipath power gains in a channel profile determine the distribution of power across the multipaths. This provides a measure of the performance of a rake receiver and predicts the bounds on its power capture capability. The power gains, therefore, are analysed next in order to determine the achievable multipath diversity gain. The multipath covariance matrix is given by

$$\mathbf{K}_{x,y} = E \left[m_{x,y,l} m_{x,y,l}^H \right], \quad (2.20)$$

where $(.)^H$ denotes the Hermitian transpose and $m_{x,y,l}$ is the multipath profile of the channel. The empirical covariance matrix for the measured frequency-selective channel is

$$\mathbf{K}_{x,y} = \frac{1}{n_x n_y} \sum_{l=1}^L (m_{x,y,l} - \bar{m}_{x,y,l}) (m_{x,y,l} - \bar{m}_{x,y,l})^H, \quad (2.21)$$

where L is upper bounded to a fixed value, ignoring the weak, late-arriving multipaths for computational efficiency. The eigenvalues, λ_i , of $\mathbf{K}$ are computed and normalised with the criterion $\sum_i \lambda_i = 1$. The significant eigenvalues can be identified by using a power sum threshold s , such that

$$\sum_{i=1}^{L_s} \lambda_i \leq s, \quad 0 \leq s \leq 1, \quad (2.22)$$

provides an estimate of the number of significant eigenvalues. The threshold, s , is usually taken as 90% or 98%. The 90% L_s for LOS and NLOS come out to be 9 and 13 respectively, while the 98% values are 56 and 62 respectively. The mean eigenvalues for some of the most dominant modes are shown in Figure 2.11, illustrating the percentage of the total power that is contained in each.

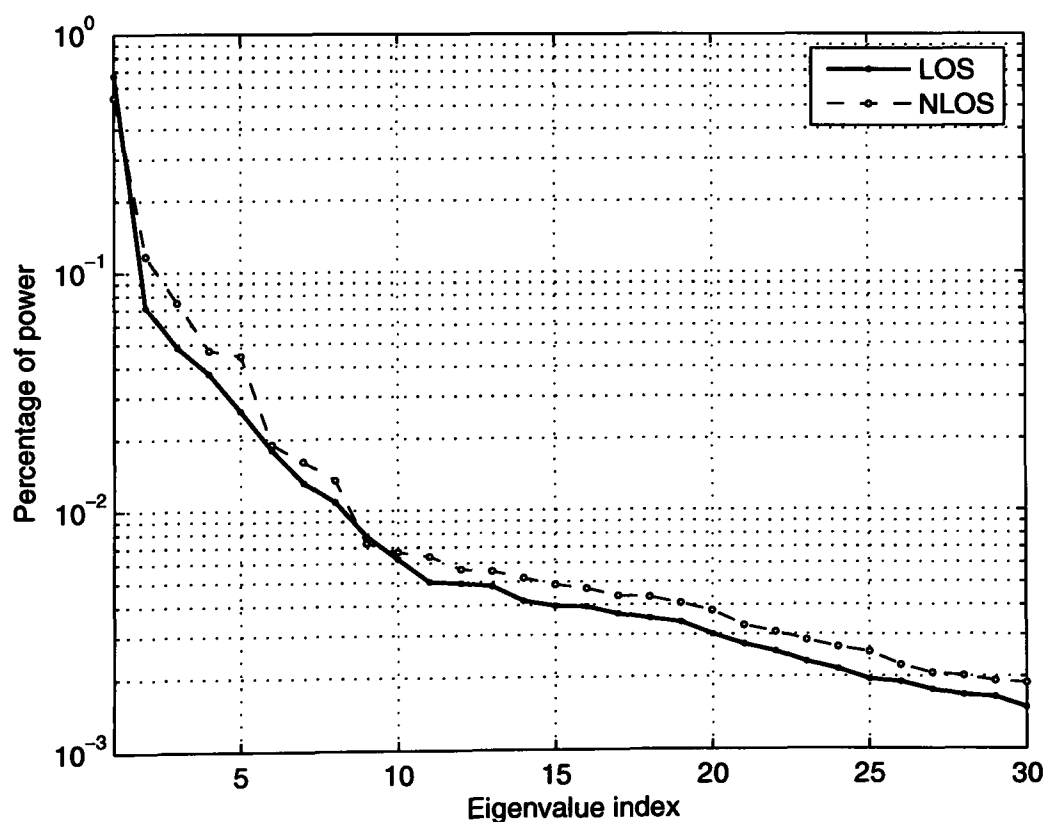


Figure 2.11 Significant eigenvalues of the multipath covariance matrix, indicating the power gains

The first LOS eigenvalue is larger than the first NLOS eigenvalue, while the later NLOS eigenvalues are larger than the LOS ones. The results of this section are summarised in Table 2.4.

Table 2.4 Multipath statistics from full-band channel measurements

Multipath Parameter	LOS		NLOS	
	Mean	Std.	Mean	Std.
No. of multipaths	90.0	11.6	89.3	15.0
No. of multipaths vs. bandwidth, GHz ⁻¹	10.85	–	10.98	–
Degrees of freedom	27.7	4.9	28.9	7.4
No. of eigenvalues (90%)	9	–	13	–
No. of eigenvalues (98%)	56	–	62	–

2.8. Deterministic Channel Modelling with Ray-Tracing

Optical ray-tracing is frequently employed for deterministic channel modelling and simulation useful in situations where the geometry of the propagation environment is known. It is based on the solution of Maxwell’s equations with high-frequency approximations and image theory [47]. Ray-tracing uses the knowledge of the location and the electromagnetic properties of scattering objects to calculate the resultant electric field at the receiver.

At microwave frequencies, electromagnetic waves conform to ray behaviour [48]. Non-specular components can be neglected, for instance, for a brick wall at frequencies lower than 20 GHz as an approximation [49]. Thus simple indoor radio propagation models usually include purely specular reflection. Some researchers have also considered wall textures in order to include the effect of scattering in the model [50]. From the Fraunhofer criterion for flat surfaces, however, scattering is negligible across the FCC UWB frequency range and is important only for millimeter waves [49]. Diffraction is also taken into account by some ray-tracing models [51] but it can also be ignored at

microwave frequencies in indoor channels [52]. Conventional ray-tracing, which approximates sinusoidal plane waves as rays, is only applicable for narrowband signals. The technique was extended to UWB by superposition of ray-tracing for the individual frequency components [53]. An alternative time-domain approach, based on the addition of the impulse responses of individual rays, was presented in [54].

Exhaustive ray-tracing methods provide an accurate estimate of the channel in complex environments, but are computationally intensive and thus often impracticable. In recursive ray-launching, or shoot-and-bounce, schemes [55], rays are launched from the transmitter in all directions, intersecting surrounding objects and undergoing transmission or reflection, and a point of reflection is considered a secondary source. The more efficient virtual source (VS) techniques, based on the method of images, treat objects as mirrors and determine the locations of transmitter images behind them which are then treated as secondary sources. A hierarchical structure of sources is created with the transmitter at its root and each level representing a reflection order. Thus the presence of a ray-path is determined for a given receiver location. Preprocessing of location visibility trees [56] or illumination zones for a given transmitter position [55-57] can improve run-time efficiency.

A simple virtual-source ray-tracing approach based on the use of a cuboidal room-tiling model is presented in this section. It provides an efficient means of simulating the indoor UWB channel although at the cost of accuracy in the deterministic, site-specific modelling of complex environments such as the microcellular channel. In this method, virtual rooms extend in an infinite three-dimensional space containing virtual sources that are images of the actual source reflected across the room's surfaces. The high computational efficiency of this technique is achieved at the cost of the ability to specify irregular room geometry, locations of obstructing objects, and non-homogeneous material properties. Typical wireless personal area network architectures involve vertically polarised antennas and azimuthal signal propagation [16]. For this reason, the model can be simplified to two-dimensional, or rectangular, tiling without a catastrophic impact on its channel prediction capability. The concept is illustrated in Figure 2.12, where the central rectangular region is the actual room and the surrounding extensions contain the virtual transmitting sources perceived by the receiver that arise from the wall reflections.

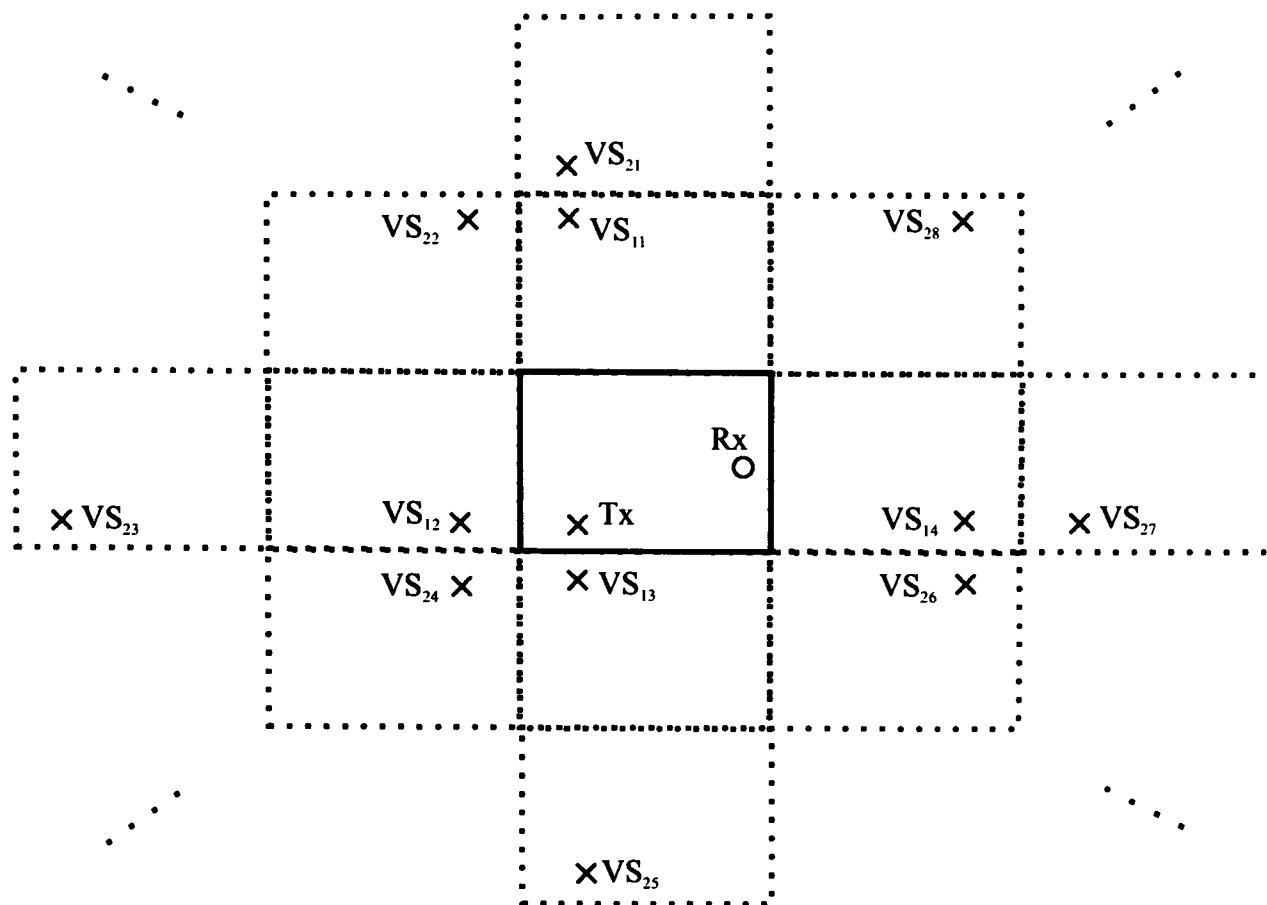


Figure 2.12 A two-dimensional ray-tracing model for the simulation of indoor radio propagation based on the method of images. The virtual sources are represented by VS_{ab} , where b is the index to the virtual sources corresponding to the a^{th} reflection order

For reasons discussed earlier, specular reflection is considered while diffraction and scattering are neglected, so that every wave component traverses one or more freespace propagation paths between the transmitter and receiver. The model can simulate both LOS and NLOS scenarios, the latter by artificially attenuating the first arrival to simulate the LOS blockage. For simplicity, however, the results here will be limited to the more common indoor scenario of LOS propagation. The proposed model is parametric and allows the specification of the probe signal type, transmitter location, receiver location, room dimensions, wall reflection coefficient, maximum reflection order, and noise variance. Other factors that affect signal propagation such as signal polarisation, antenna gains and radiation patterns, and dependence of the reflection coefficient on incidence angle can be easily added. It can also be used to simulate antenna arrays, so that information about the multipath angles of departure and arrival can be extracted and the performance of multiple-antenna systems can be predicted.

The reflection order, K , refers to the number of reflections considered in the model. It increases the complexity of a ray-tracing model but a low order suffices for most channels. It was shown in [58] that $K=2$ provides sufficient information about large-scale channel variation, while [29] suggested that

$K=3$ is sufficient for modelling indoor UWB propagation. Furthermore, [59] showed that the power level prediction improvement is lower than 3 dB beyond $K=3$. The proposed two-dimensional model generates a total of N virtual sources for order K . From the method of images, N can be calculated by

$$N = \ell + \sum_{k=1}^K 4k, \quad (2.23)$$

where $\ell=0$ for NLOS and 1 for LOS. The complex electric field incident at the receiver due to a single impinging wave is a function of the total distance traversed, d , as well as the number of reflections p . This received signal due to the n^{th} wave can be mathematically represented as

$$E_{r,n} = E_0 A(d) \prod_p R_p e^{-j2\pi d/\lambda_c} \quad (2.24)$$

where E_0 is the transmitted field, R_p is the wall reflection coefficient, and

$$A(d) = \frac{\lambda_c}{4\pi d} \quad (2.25)$$

is the free-space signal loss. For simplicity, the wavelength, λ_c , corresponding to the centre frequency is used here. Also, vertical, uniformly reflecting walls are assumed. The fractional power transmission is assumed to be independent of the angle of incidence. As all material interactions are now identical, (2.24) reduces to

$$E_{r,n} = E_0 A(d) R^p e^{-j2\pi d/\lambda_c}. \quad (2.26)$$

Thus the number of ray-wall intersections is important and not their coordinates. The vector superposition of the N individual ray contributions provides the received signal, i.e.,

$$E_r(t) = \sum_{n=1}^N E_{r,n}, \quad (2.27)$$

where time dependence has also been introduced. In terms of the wave amplitude α_n , time delay τ_n and phase shift ϕ_n , and after the removal of the propagation delay, this takes the familiar form of a series of complex-weighted Dirac delta functions, i.e.,

$$r(\tau) = \sum_{n=1}^N \alpha_n(\tau) \delta(\tau - \tau_n) e^{j\phi_n}. \quad (2.28)$$

The rest of this section presents channel simulation results obtained using the proposed ray-tracing model. Perfectly reflecting walls are assumed so that there is no signal leakage from the environment, and $K=5$ is used in order to simulate a dense multipath channel. A Gaussian monocycle signal of the functional form

$$s(t) = A \frac{t}{\tau} e^{-\left(\frac{t}{\tau}\right)^2}, \quad (2.29)$$

where τ is the decay constant and A is the amplitude, is used as the probing signal. With a single time-domain zero-crossing and 3 dB bandwidth spanning 3.1–10.6 GHz, this first derivative of Gaussian waveform is used in UWB Impulse Radio [9] and thus simulates actual operating conditions in this model. The sampling frequency, $f_s=40$ GHz, is chosen to be over twice the highest signal frequency. The reflection coefficient is set as $\Gamma=0.707$, assuming walls with 3 dB transmission and no dissipation. A room with dimensions similar to the one described earlier in the context of empirical characterisation is specified, with transmitter and receiver locations also chosen to match the experimental setup. A 50x50 receiving location grid is simulated with 0.04 m resolution.

The channel impulse response is obtained from the received signal by the deconvolution [60],

$$h(t) = \int_{-f_s/2}^{f_s/2} \frac{R(f)}{S(f)} e^{j2\pi ft} df, \quad (2.30)$$

where $R(f)$ and $S(f)$ are the spectra of $r(t)$ and $s(t)$ respectively and f_s is the sampling frequency. A 20 dB noise threshold is used for the PDP. The simulation results are then verified against those derived from physical measurements. Figure 2.13 shows the power-normalised CIR obtained by ray-tracing. Clustered multipaths can be observed similar to [41] and [20], with a double-exponential decay and closely arriving multipaths some of which are likely to overlap in the absence of high-rate sampling. The multipath clusters are produced due to the minor differences in the pathlengths and flight times of the paths, with the inter-multipath delays determined by the room geometry and the location of the transmitter and receiver.

The multipath fading parameters of the channel can be used to evaluate the ray-tracing model accuracy [61]. The first order statistics of the simulated channel's delay spread are listed in Table 2.5.

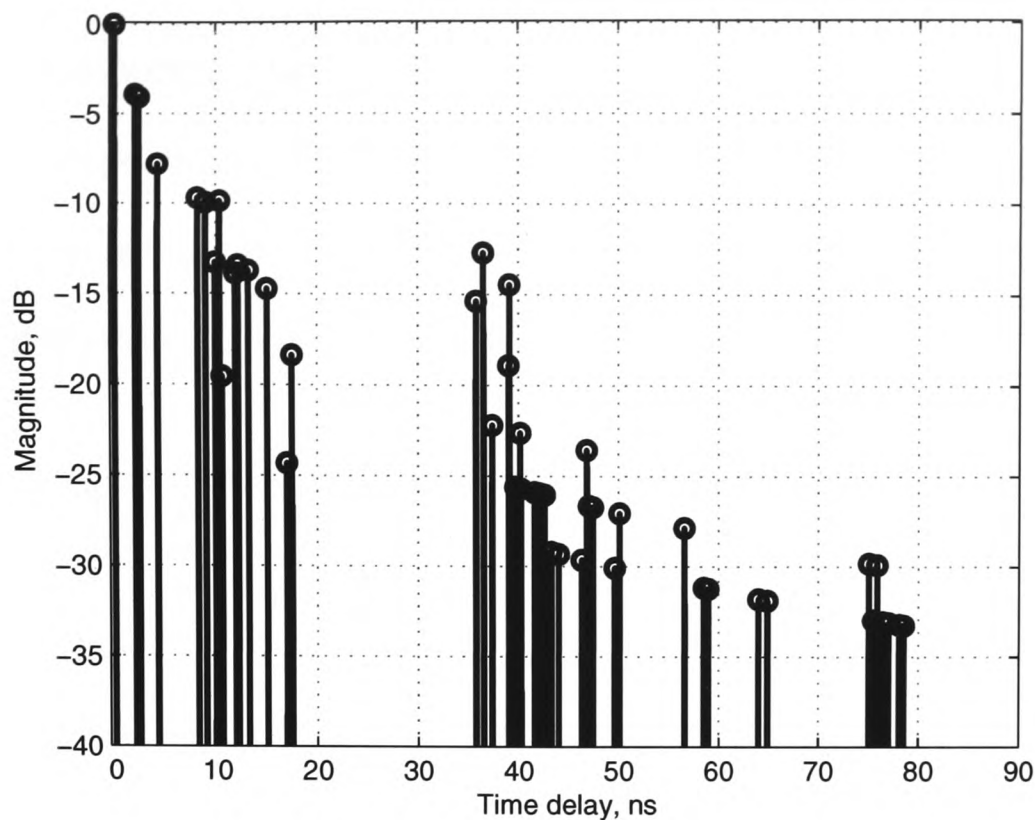


Figure 2.13 Peak-normalised magnitude of the LOS UWB channel’s complex impulse response, generated by ray-tracing at one location in an indoor environment, as a representative example

As the environment analysed through the ray-tracing technique does not contain any reflecting objects or structures other than walls, it is similar to an uncluttered multipath channel. The channel statistics obtained through ray-tracing correspond well to those obtained through measurement as discussed earlier in this chapter within experimental error. Furthermore, these are also comparable to the results reported in [32, 62]. The mean excess delay is the only delay parameter that shows a significant deviation from the measured value, underlining the point that while the proposed ray-tracing model is not perfect, it does provide a reliable first approximation of the indoor channel.

Table 2.5 A comparison of the mean values of delay statistics obtained from ray-tracing and measurement in a LOS indoor UWB environment

Channel Parameter	Simulated	Measured
Mean excess delay, ns	4.09	8.10
Rms delay spread, ns	7.06	7.85
Max. excess delay, ns	39.27	51.31

Due to the Friis free-space transmission equation, the fractional power gain with increasing K follows the law of diminishing returns, as higher order multipaths have greater pathlengths and lower power. It is found that power gain converges to within a dB with increase in K beyond $K=3$ for $\Gamma=0.7$ and $K=6$ for $\Gamma=1$. This is illustrated in Figure 2.14. Thus with imperfect reflection, 3–4 reflections are sufficient to model the indoor UWB channel.

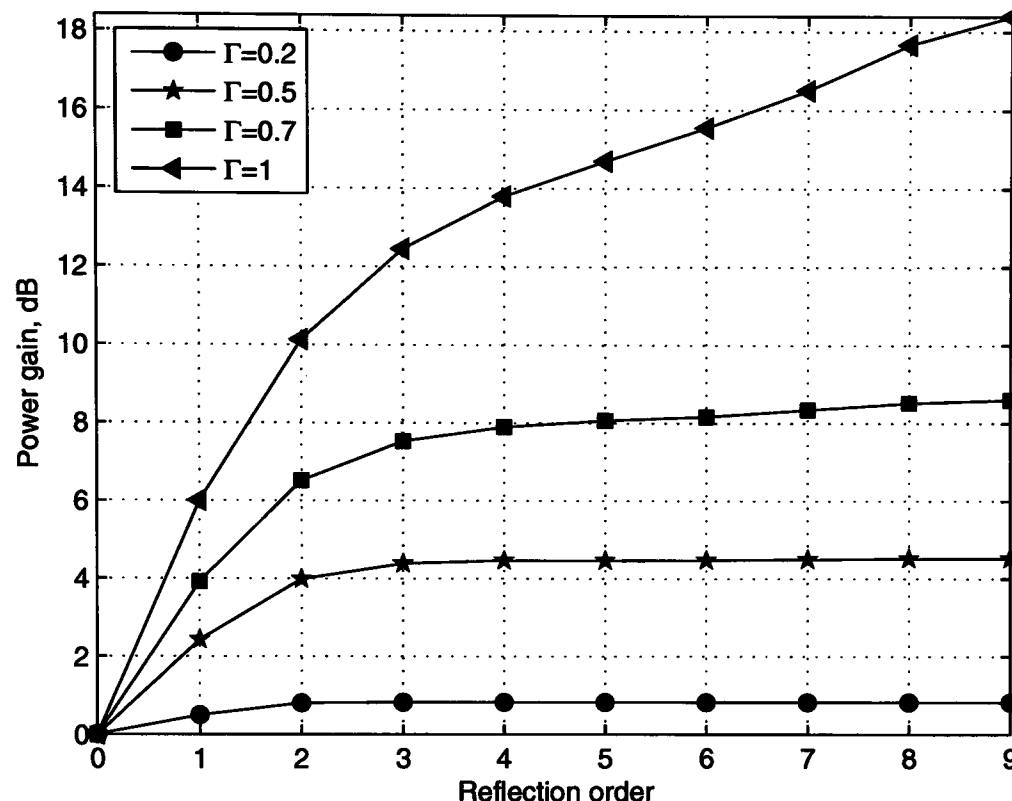


Figure 2.14 Convergence of received power with reflection order for various reflection coefficients

2.9. Summary

The distinctive attributes of the UWB channel have been identified in this chapter. A review of the existing UWB channel models has been presented and their shortcomings identified. Novel experimental results with full-band UWB channel measurements have been obtained in LOS and NLOS conditions. The channel transfer function and impulse response have been studied to characterise dispersion and small-scale fading. The multipath delay spread has been statistically analysed. The coherence bandwidth of the channel is found to be about 30 MHz. The number of resolved multipaths is shown to increase linearly with bandwidth with a slope of 11. It has been shown that the multipath degrees of freedom are much smaller than the number of resolved multipaths and only about 30% of the resolved multipaths exhibit uncorrelated fading. This provides an estimate for

the required diversity order for a sub-optimal rake to obtain near-optimal diversity gain. A simple parametric ray-tracing technique based on the method of images has been presented for the simulation of indoor UWB propagation. The results of this flexible and computationally efficient deterministic technique are comparable with channel measurements. This chapter has characterised the channel statistics over the FCC-allocated UWB band, while the next chapter explores the variation of these statistics within the band and provides an insight into the effects of the channel's frequency selectivity.

3. FREQUENCY DEPENDENCE OF PROPAGATION

3.1. Introduction

The existing UWB channel models rely on narrowband channel characterisation approaches and largely ignore the frequency dependent behaviour arising due to the large bandwidth. The composite behaviour of the full frequency band is analysed with little insight into any variations at the sub-band level [20, 32, 39, 40]. This approximation is questionable over an ultra-wide frequency band for a number of reasons. Practical system components such as antennas and filters do not provide a perfectly flat response over a very wide frequency range. Also, wave propagation modes, such as scattering, diffraction and material penetration, are influenced by frequency [63, 64], rendering the multipath characteristics of the channel sensitive to bandwidth and operating frequency.

The variation in the channel propagation properties directly influences its multipath profile and in turn the channel fading statistics such as pathloss and delay spread. Such non-uniform spectral behaviour can severely deteriorate the performance of both single- and multi-band UWB systems unless adequately incorporated in system design. For a multi-band UWB system, the system performance in one sub-band can differ considerably from that at another depending on its spectral location within the UWB band. Pulsed UWB systems can suffer waveform distortion and erroneous signal detection. An adaptive system can provide improved performance by pre- or post-compensation, especially with the recent shift towards flexible spectral control techniques such as soft-spectrum adaptation and cognitive radio [65]. The frequency-dependence of the channel and system components must, however, be adequately characterised before such sophisticated techniques can be employed.

In denial of frequency-selective propagation effects, [66] claimed that UWB channel statistics such as pathloss are frequency-independent. On the other hand, [67] studied the frequency dependence of material properties and its consequences for indoor UWB signal propagation, reporting a finite support of the impulse response of a layer of material that results in pulse distortion and spreading. Similarly, [68] argued that the frequency dependence of material reflection coefficients causes signal reflection from walls at lower frequencies and penetration at higher frequencies, leading to an increase

in the pathloss exponent with frequency in LOS scenarios and a decrease in NLOS environments. However, it neither provided a complete statistical characterisation of the pathloss exponent's variation with frequency, nor analysed the frequency dependence of other descriptors such as the delay spread.

This chapter thus aims to fill the vacuum in the understanding of frequency-dependent propagation and its effect on UWB system performance. The variation of channel fading and dispersion statistics between sub-bands within the UWB spectrum is characterised experimentally. The sub-band impulse response disparity is quantified in terms of the number of multipaths, their gains and delays. Synthetic aperture analysis of the multipath directions-of-arrival establishes frequency-selective scattering in the environment. The dispersion index is introduced as a measure of signal waveform dispersion due to frequency-dependent pathloss. Its first order statistics and spatial variation are estimated, and its dependence on signal polarisation and propagation environment is studied. The sensitivity of single-band and multi-band UWB system performance to frequency-dependent propagation effects is highlighted and relevant system design techniques are discussed.

3.2. Analysis Procedure

The characterisation of the frequency effects on signal propagation is based on measured channel information obtained through the experimental procedure described in Chapter 2. A complex impulse response, $h_{x,y}(\tau)$, is obtained for each measurement location (x,y) , given by

$$h_{x,y}(\tau) = \sum_{k=0}^{K-1} \alpha_{x,y,k} e^{j\varphi_{x,y,k}} \delta(\tau - k\Delta\tau). \quad (3.1)$$

where the same notation is used as in Chapter 2. Linearly polarised antennas are used, and the effect of polarisation is highlighted where important.

3.3. Sub-Band Non-Uniformity

To investigate the variation of small-scale fading with frequency, the UWB spectrum, measured at $n_f=1601$ frequency points is divided into sub-bands $b=\{1,\dots,5\}$, each 1.5 GHz wide. The sub-band

transfer functions, each now comprising $(n_f-1)/5$ frequency points, are zero-padded to regain the n_f point length. The sub-band CIRs,

$$h_{x,y,b}(\tau) = \sum_{k=0}^{K_b-1} \alpha_{x,y,b,k} e^{j\varphi_{x,y,b,k}} \delta(\tau - k\Delta\tau), \quad (3.2)$$

are calculated, where $\alpha_{x,b,k}$ and $\varphi_{x,b,k}$ are the magnitude and phase responses for the b^{th} sub-band, and K_b is the number of its time-delay bins. Figure 3.1 shows the sub-band CIRs for the vertically polarised LOS channel at one location as a representative example. The sub-band CIRs are remarkably dissimilar, as evident from inspection in terms of the initial multipath amplitudes and delays, establishing strong dependence of the channel response on the frequency sub-band. It also illustrates that not all multipaths contain the full frequency content, a situation that can significantly affect the link performance across the spectrum.

With this preliminary evidence, the relation of channel fading statistics with frequency is investigated in order to quantify the channel variation across the band. For each measurement point, the sub-band statistics are derived using the absolute noise level of the corresponding full-band CIR as the threshold. The sub-band pathloss, delay spread and number of multipaths are characterised.

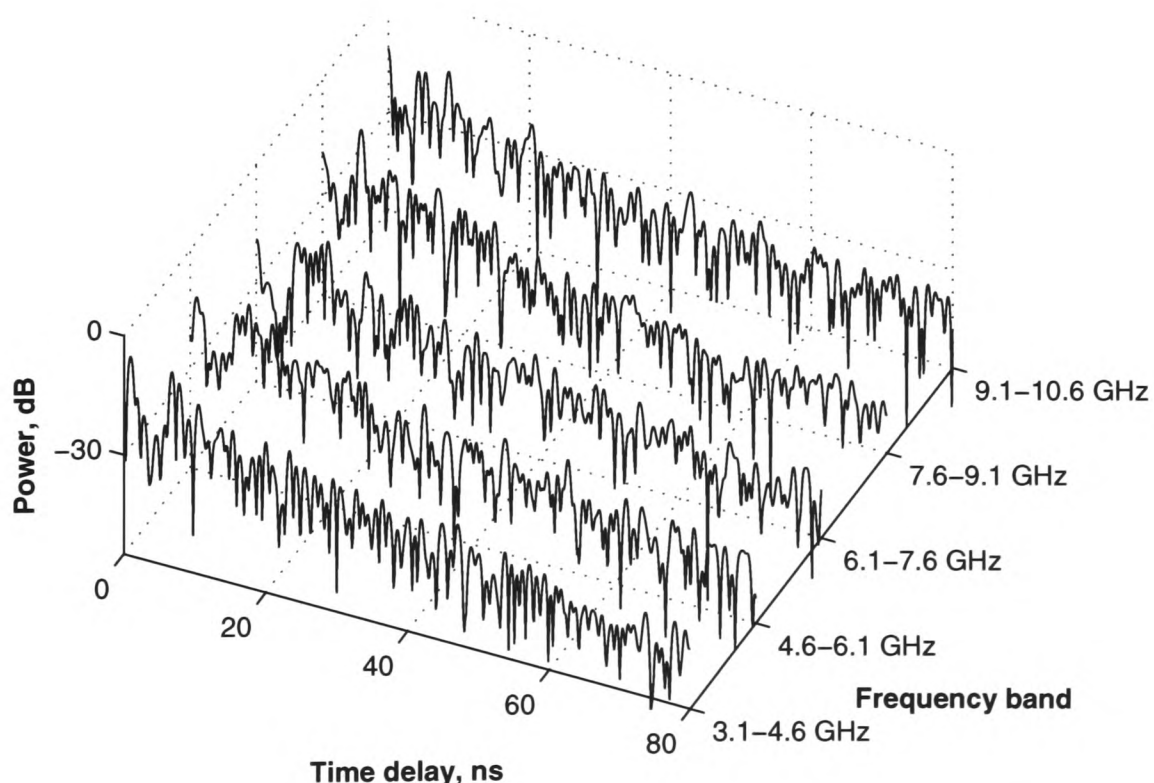


Figure 3.1 Complex impulse responses of the LOS UWB channel sub-bands derived from a single measured full-band channel transfer function. The time delay is with reference to the first arrival in the full-band CIR, and the powers are normalised to the full-band CIR's peak power

The pathloss determines the amount of power received and thus affects the link budget and quality of the radio link. Accurate pathloss characterisation is especially important for power-limited UWB systems. The pathloss dependence on frequency f arises from the antenna power density and gain variation with f . Thus the pathloss of a frequency-selective channel, when omni-directional antennas are used, is given by [69]

$$PL(f, d)_{[\text{dB}]} = PL(f_0, d_0)_{[\text{dB}]} + 10n \log_{10} \left(\frac{d}{d_0} \right) + 10m \log_{10} \left(\frac{f}{f_0} \right), \quad (3.3)$$

where d_0 and f_0 represent the reference distance and frequency respectively, while n and m denote the distance- and frequency-dependent pathloss exponents. The inclusion of frequency in (3.3) indicates that the sub-bands within the UWB band experience non-uniform pathloss. To further investigate frequency-dependent pathloss in the UWB sub-bands, the wideband power received in sub-band b at location x is determined as

$$P_{x,y,b} = \int_{f_{b,l}}^{f_{b,h}} |H_{x,y}(f)|^2 df, \quad (3.4)$$

where $f_{b,l}$ and $f_{b,h}$ refer to the low and high frequency boundaries of the sub-band. In order to reduce the effect of noise through thresholding, the power calculation is performed in the time domain. Using the Parseval relation [69] and introducing discrete notation, the sub-band power is the Riemann sum

$$P_{x,y,b} = \sum_{\tau=0}^{T_s} p_{x,y,b}(\tau) \Delta \tau, \quad (3.5)$$

where $p_{x,y,b}(\tau)$ represents the power delay profile (PDP) of the b^{th} sub-band and T_s is the time-dispersed symbol duration. The power of each sub-band is normalised with respect to the full-band power at that location in order to obtain a reference pathloss figure, so that the normalised power is

$$\hat{P}_{x,y,b} = P_{x,y,b} \left(\sum_{\tau=0}^{T_s} p_{x,y}(\tau) \Delta \tau \right)^{-1}, \quad (3.6)$$

where

$$p_{x,y}(\tau) = |h_{x,y}(\tau)|^2 = \sum_{k=0}^{K-1} \alpha_{x,y,k}^2 \delta(\tau - k\Delta\tau). \quad (3.7)$$

The probability distribution of received power for each sub-band in LOS and NLOS scenarios is estimated. From a comparison of the empirical CDFs of pathloss, plotted in Figure 3.2, a difference of up to 11 dB is observed between the pathloss of the highest and lowest sub-bands, increasing monotonically with frequency. Also, the mean pathloss of the mid-band range is about 14 dB higher than the full-band pathloss. This points to the inaccuracy caused by approximating the pathloss of the whole band with that of the centre frequency f_c , as assumed in some simple channel models, and emphasises that the individual pathloss of each sub-band must be taken into account during system design. The NLOS channel experiences a greater variation in sub-band pathloss with frequency. From Table 3.1, the higher sub-bands experience larger pathloss standard deviation. This observation can be related to Figure 3.1, where significant variation in the amplitudes of the dominant multipaths in the various sub-bands is obvious.

The time dispersion of a channel can be expressed in terms of the time delay spread, which depends on the location and composition of scatterers and the overall geometry of the propagation environment. It provides information about frequency-selective fading and limits the achievable data-rates. The time delay parameters are extracted directly from the PDP. Here, three delay parameters, namely the mean excess delay, rms delay spread, and maximum excess delay are analysed separately for each sub-band, using the definitions in Chapter 2.

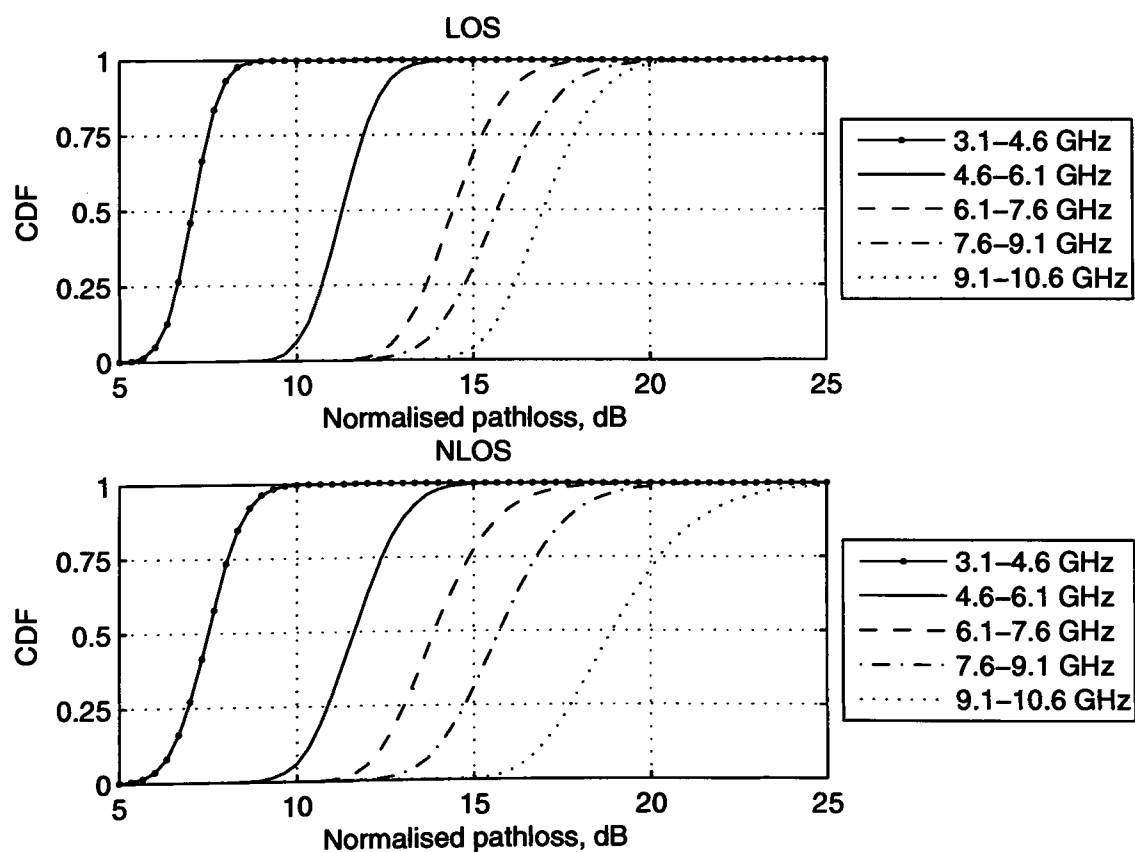


Figure 3.2 CDFs of pathloss in the sub-bands normalised to the full-band pathloss

From Figure 3.3, there is a variation of 3–4 ns in τ_{rms} between the sub-bands, while the full-band τ_{rms} is consistently higher than the sub-band τ_{rms} . Table 3.1 shows that the sub-band τ_m varies on average by 3–5 ns. The difference in the sub-band τ_{max} is about 20-25 ns. The lowest sub-band's τ_{max} is 10-15 ns lower than that of the full band, and the rest follow. The lower delay spread statistics of the higher sub-bands are due to higher attenuation, causing the multipaths to rapidly fall below the PDP threshold. The variation of these statistics with the centre frequency does not appear to be monotonic. Some of the highest sub-band NLOS PDPs contain little signal, indicated by a finite probability of $\tau_{max}=0$.

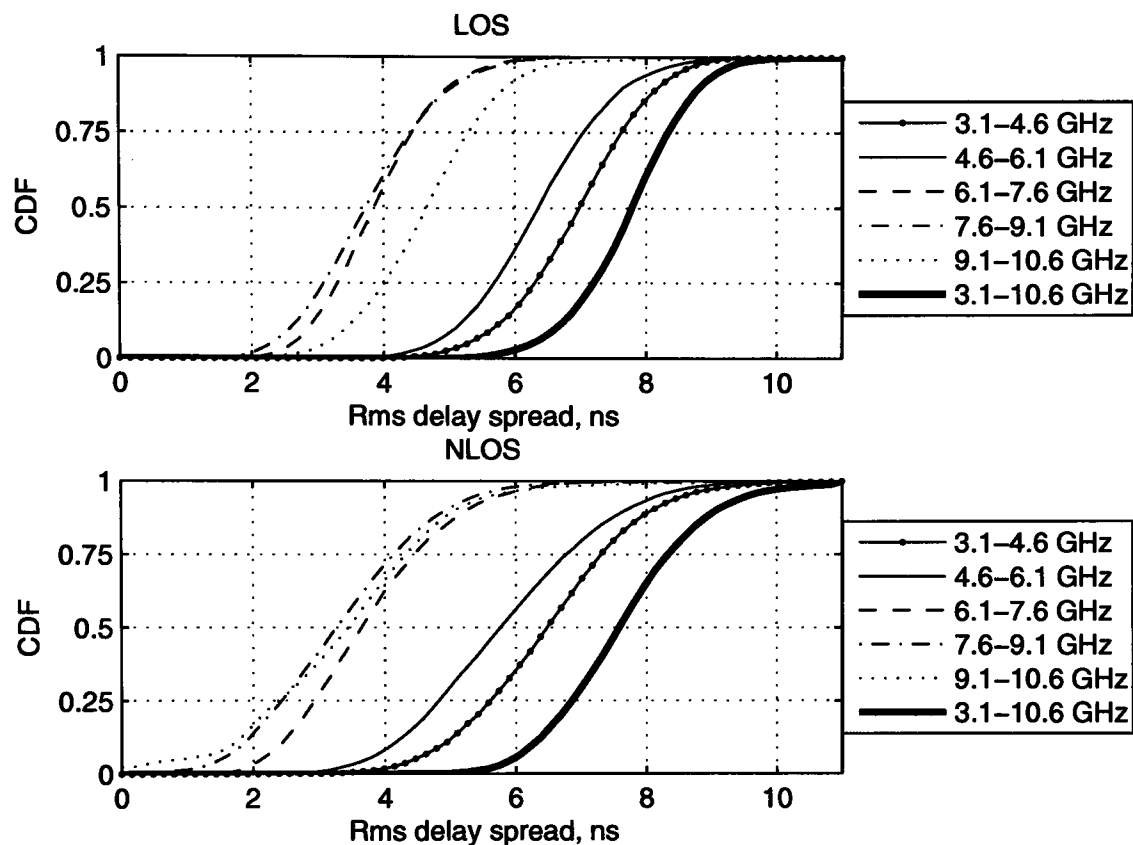


Figure 3.3 CDFs of rms delay spread in the sub-bands

Multipath resolution is an important factor in the small-scale fading behaviour of a channel, and high multipath resolution leads to lower fade depths. The bandwidth determines the time resolution and thus the number of resolved multipaths, L . In the context of this analysis, it is important to remember that the sub-bands have identical bandwidth but different centre frequencies. Table 3.1 lists the first order statistics of L_b for each sub-band, and also of L for the full band. The latter is almost five times larger owing to its five-fold bandwidth. The sub-bands, however, have progressively decreasing L_b with frequency, as observed in Figure 3.4. This is not only due to the dropping power levels as discussed in the context of Figure 3.1, but also the frequency selectivity of the channel which causes distortion to the received signal and alters its frequency content at the individual multipath

level. The range of L_b across the band indicates that the lowest sub-band receives 20 additional multipaths as compared to the highest sub-band, which is significant when the total number of multipaths received in a sub-band is considered.

Table 3.1 First order statistics for fading and time dispersion in the full- and sub-bands

Frequency Band, GHz	Propagation Scenario	Normalised		Rms Delay		Mean Excess		Max. Excess		Number of	
		Pathloss, dB		Spread, ns		Delay, ns		Delay, ns		Multipaths	
		Mean	Std.	Mean	Std.	Mean	Std.	Mean	Std.	Mean	Std.
3.1–10.6	LOS	0	0	7.83	0.86	8.21	0.87	50.65	6.74	90.12	11.82
	NLOS	0	0	7.75	1.22	9.96	1.70	50.91	8.22	89.80	16.70
3.1–4.6	LOS	7.23	0.63	7.03	0.97	8.33	1.04	37.80	5.82	30.15	4.92
	NLOS	7.67	0.84	6.57	1.28	8.71	1.74	35.80	7.01	26.61	6.41
4.6–6.1	LOS	11.48	0.87	6.45	1.01	8.38	1.19	29.36	6.35	20.72	4.33
	NLOS	11.83	1.12	5.92	1.37	9.29	1.73	28.72	7.60	18.85	5.64
6.1–7.6	LOS	14.67	1.19	3.98	0.83	7.76	0.86	19.55	4.66	12.11	3.11
	NLOS	14.19	1.42	3.84	1.16	5.54	2.49	17.41	7.07	11.85	4.19
7.6–9.1	LOS	15.91	1.44	3.86	0.96	6.89	0.94	16.92	4.24	9.57	3.04
	NLOS	15.99	1.59	3.43	1.39	3.62	2.18	13.14	6.20	8.26	4.20
9.1–10.6	LOS	17.21	1.25	4.76	1.02	5.51	1.16	16.24	5.13	8.60	3.09
	NLOS	19.32	2.81	3.50	1.65	3.74	2.36	11.39	6.89	6.51	4.48

This analysis of fading and dispersion in the sub-bands establishes that the full UWB channel cannot be treated in a uniform fashion, and sub-band level treatment is essential for reliable channel characterisation. In particular, the results establish the dependence of the small-scale fading statistics of the UWB channel on the centre frequency.

The next section analyses the directional properties of the channel with the help of the angular power spectra and investigates the channel frequency selectivity in the angular domain.

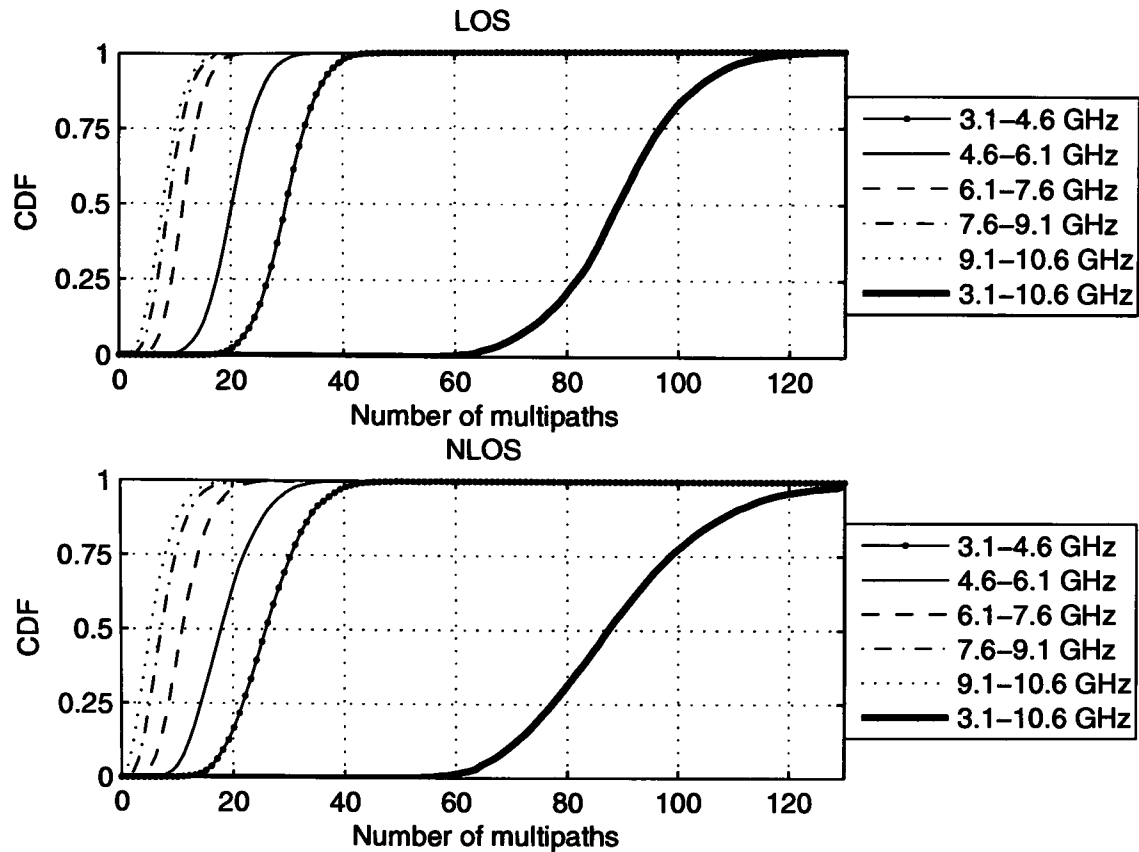


Figure 3.4 CDFs of the number of received multipaths in each of the sub-bands

3.4. Frequency-Selective Angular Scattering

The multipath directions-of-arrival (DoAs) identify the angular locations of the scatterers in the propagation environment. Synthetic aperture back-propagation can be used to produce an image of the multipath propagation environment in the (θ, φ) plane [70]. If each element in a uniform rectangular antenna array with $n_x \times n_y$ elements and inter-element spacing d radiates a wideband signal $\tilde{s}(x, y, f)$, the composite signal reaching a point R in space is given by

$$\tilde{\zeta}(\theta, \varphi, f) = \sum_{x=0}^{n_x-1} \sum_{y=0}^{n_y-1} \tilde{s}(x, y, f) e^{-jk(xd \sin \theta \cos \varphi + yd \sin \theta \sin \varphi)}, \quad (3.8)$$

where k represents the wavenumber corresponding to f , and the normalisation

$$\tilde{s}(x, y, f) = H(x, y, f) f \quad (3.9)$$

compensates for the frequency-dependent attenuation of the omni-directional antennas. Now, (3.8) is a two-dimensional Fourier transform, so that the fast Fourier transform can be used to form an image of the scatterers in the (k_x, k_y) plane, i.e.,

$$\tilde{\kappa}(k_x, k_y, f) = \frac{1}{n_x n_y} \sum_{x=0}^{n_x-1} \sum_{y=0}^{n_y-1} \tilde{s}(x, y, f) e^{-j2\pi \left(\frac{xk_x}{n_x} + \frac{yk_y}{n_y} \right)}. \quad (3.10)$$

The angular power spectrum $\tilde{A}(\theta, \varphi, f)$ is obtained from (3.10) by the spherical coordinate transformation given by

$$\theta = \sin^{-1} \left(\frac{\lambda}{n_x d} \sqrt{k_x^2 + k_y^2} \right) \quad (3.11)$$

and

$$\varphi = \tan^{-1} \left(\frac{k_y}{k_x} \right). \quad (3.12)$$

As the elevation field-of-view of the aperture is determined by f , there is little angular information available at the low frequencies about high squint angles. Also, in order to obtain the angular spectrum as a function of uniformly spaced θ and φ , cubic interpolation gridding is used here, and points on or near the convex hull incorrectly estimated by the gridding algorithm are corrected.

Now, the Rayleigh resolution for azimuth angle [70] is

$$\Delta\varphi = 70 \frac{\lambda}{D} \quad (3.13)$$

where $\Delta\varphi$ is in degrees, λ is the wavelength and D is the aperture size which is equivalent to the dimensions of a side of the square grid when the wave approaches it normal to that face, such as for the LOS path in our configuration. When the incoming wave has oblique incidence, it perceives a wider aperture and greater resolution is obtained. Thus $\Delta\varphi$ also varies with φ : the highest resolution is obtained at $\varphi = n\pi/4$, $n = \{1, 3, 5, 7\}$, and the lowest at $\varphi = n\pi/2$, $n = \{0, 1, 2, 3\}$. For a wave whose Poynting vector is normal to the aperture, higher frequencies provide greater angular resolution, and $\Delta\varphi = 2^\circ$ at $f_h = 10.6$ GHz for our 1 m^2 measurement grid. On the other hand, the angular spectrum suffers from low elevation resolution, decreasing with θ , and an up-down ambiguity due to the horizontal plane measurement. Consequently, $\tilde{A}(\theta, \varphi, f)$ is restricted to $0 < \theta \leq \pi/2$, and an elliptical distortion is present in θ , a problem that can be overcome with a synthetic volume aperture. The angular resolution also depends on the far-field condition being satisfied. The Fraunhofer boundary is given by [71]

$$d_f = \frac{2D^2}{\lambda} = \frac{2D^2 f}{c} \quad (3.14)$$

where c is the speed of light in freespace. For our aperture and range of frequencies, d_f varies from 20.67 m to 70.67 m, which is longer than the distance of the receiver antenna from most scatterers. The Fourier transform assumes that all scatterers are located an infinite distance away, which can be approximated with the far field. Since that condition is not strictly met in our scenario, Fresnel diffraction takes place leading to defocusing and blurring of incoming beams in the (θ, φ) plane, causing lesser resolution than predicted by (3.13). A full-wave treatment using double-convolution rectifies this inaccuracy but is computationally intense [70].

In order to investigate the spectral evolution of multipath directions, Figure 3.5 shows $\tilde{A}(\theta, \varphi, f)$ for the LOS channel at $f = \{3.1, 5.6, 8.1, 10.6\}$ GHz. All images are intensity-normalised globally and thresholded to 25 dB below that level. The transmitter direction is $(\theta, \varphi) = (90^\circ, -90^\circ)$, from the synthetic aperture orientation (see Chapter 2). The direct component is seen at all frequencies, albeit with different strengths and resolutions. With greater angular resolution, however, comes enhanced noise suppression, as apparent from the increased power spectrum contrast at high frequencies in which some of the weaker multipaths seen at low frequencies are suppressed. Most importantly, however, the cluster and multipath arrival patterns are not identical. The room features, i.e. walls and corners, can be identified at the lower end of the spectrum but not at the upper end, indicating that some of the signal from high elevations only contains low frequency components. Similarly the signal from some other scatterers is observed only at particular frequencies, providing evidence of frequency-selective scattering. This is further analysed by inspecting the two-dimensional cross-correlation of the angular power spectra across the frequency band. The zero-angular displacement correlation is calculated using the centre frequency's angular power spectrum as the reference, i.e.,

$$\rho_{\tilde{A}}(\theta', \varphi', f) = \int_0^{2\pi} \int_0^{\pi/2} \tilde{A}(\theta, \varphi, f_c) \tilde{A}(\theta + \theta', \varphi + \varphi', f) d\theta d\varphi, \quad f = [3.1, 10.6] \text{ GHz}, \quad (3.15)$$

and the result is shown in Figure 3.6 for LOS and NLOS channels for $\theta' = \varphi' = 0$.

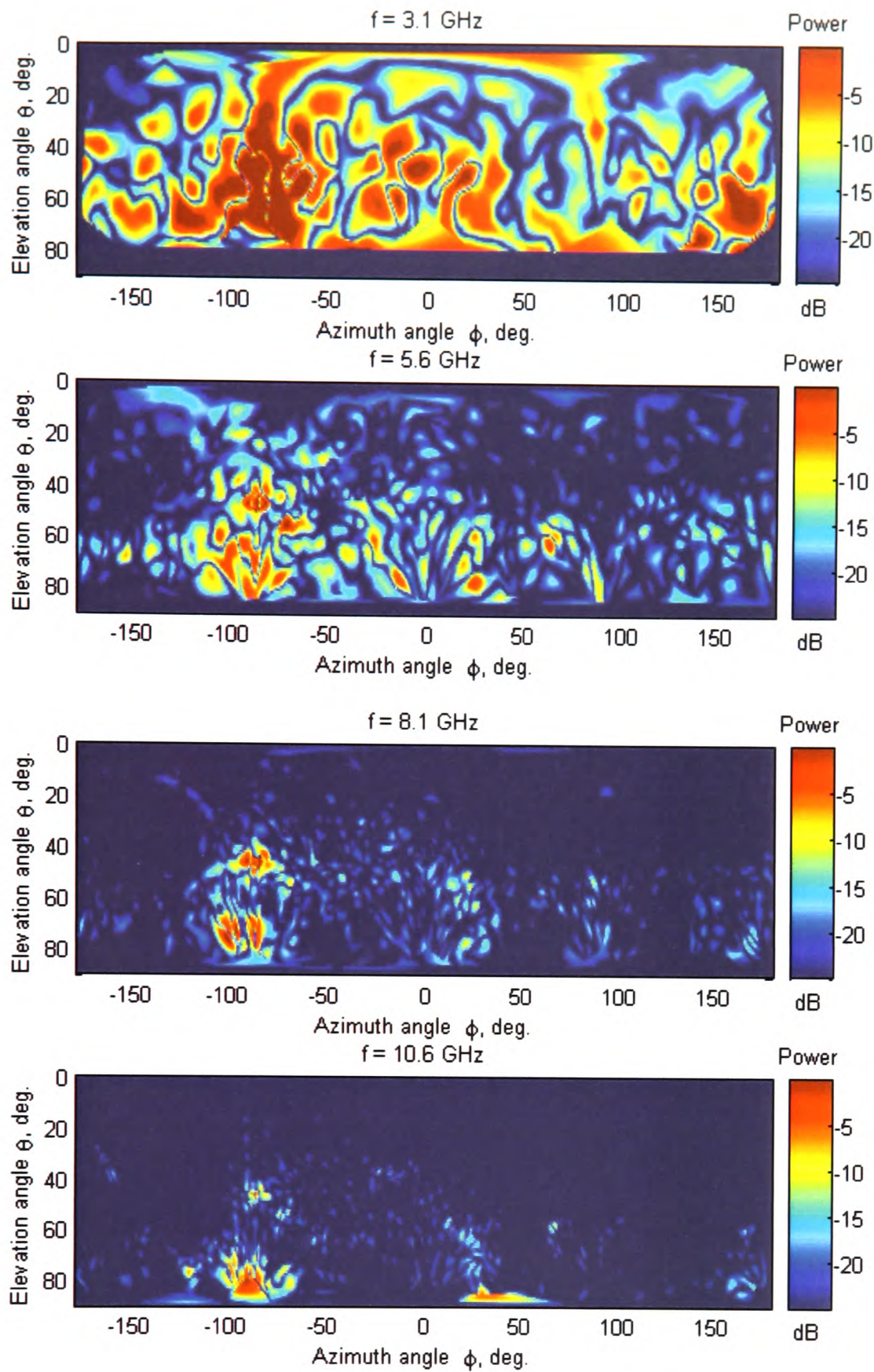


Figure 3.5 Angular power spectra at four equidistant frequencies normalised to the global peak intensity

The correlation coefficient is lower than 0.6 for nearly all of the frequency components except those closest to the reference image. At frequencies within the correlation bandwidth in the neighbourhood of the centre frequency, the angular spectra cross-correlation approaches the auto-correlation and is thus high. A large-scale spectral decay and a small-scale fluctuation can both be observed. The low spectral correlation establishes the frequency-selective nature of the angular power spectra and the variation of the directional multipath profiles with frequency.

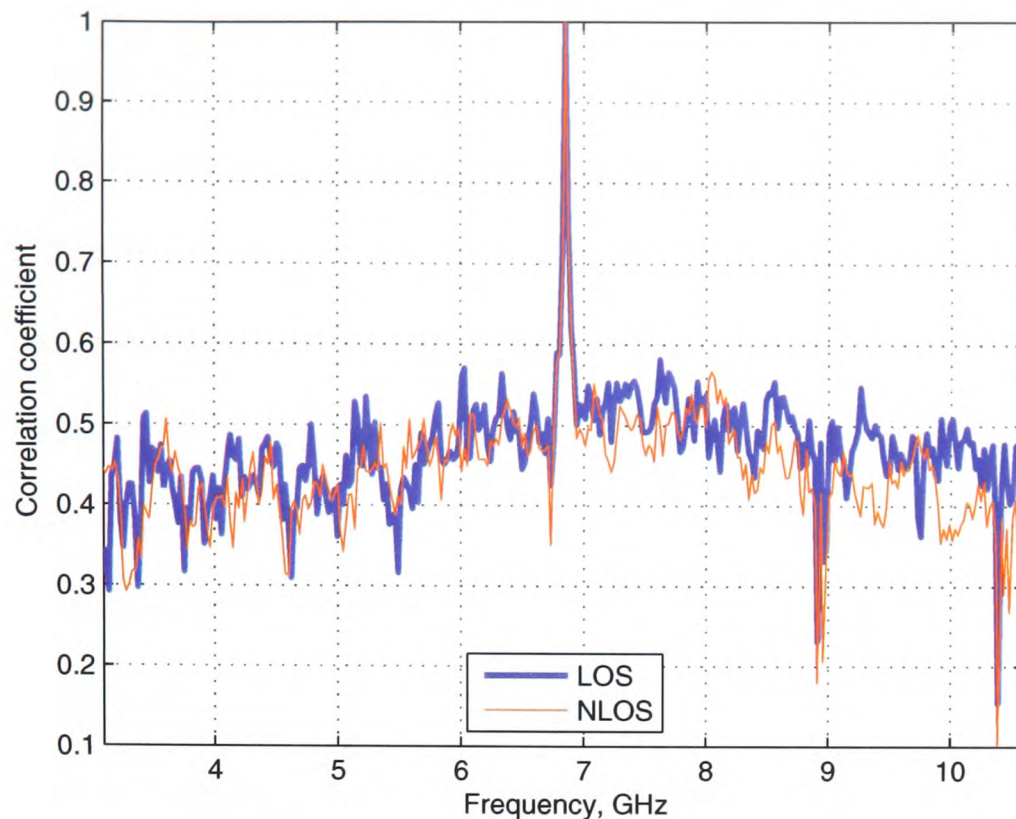


Figure 3.6 Correlation of the angular power spectra across the band with that at the centre frequency

3.5. Frequency-Dependent Pathloss

The non-flat frequency response of the channel, antennas, amplifiers, and other equipment is a key design parameter ignored in narrowband communications but significant in UWB systems. From the link budget perspective, this channel behaviour manifests itself in the form of frequency-dependent pathloss (FDP), causing non-uniform signal transmission levels across the band. Non-flat spectral gain results in time-domain waveform distortion and can severely degrade the performance of a communications receiver, as analysed later in this chapter. Pre-distortion filtering at the transmitter or frequency-domain automatic gain control at the receiver can partially compensate for this distortion, even though such techniques may degrade the SNR and data-rate due to noise enhancement. The first step, however, is the characterisation of the channel's FDP, which is undertaken in this section.

In most situations, the dominant cause of FDP is the antenna pair used in the link. The linear but non-uniform power density of an omni-directional antenna across the band appears in the form of FDP [72]. In addition, some components of the propagation channel can cause frequency-selective scattering, due to several reasons. Metallic objects of specific geometric shapes and sizes such as railings and gratings are sources of frequency-selective reflection. Textured wall surfaces can lead to

partial specular reflection and partial scattering of a signal comprising a wide frequency range [16]. As the reflection coefficients of materials used in walls and other objects also vary with frequency [73], NLOS channels that have a higher probability of material penetration are generally expected to suffer from greater FDP. Finally, diffraction is another propagation mechanism that is frequency dependent [74]. As the indoor multipath channel experiences strong multipath effects [20], there is a finite probability of significant FDP due to the aforementioned wave propagation phenomena.

Acknowledging the importance of FDP characterisation, [75] modelled it with an exponential decay factor with mean values of 1.01 and 1.36 in LOS and NLOS respectively. On the other hand, [29] used a power-law model and extracted the frequency transfer function amplitude decay exponent with 1 GHz as reference. It reported exponents between 0.8 and 1.4 with variations by up to a factor of 2 over small distances in an office environment. A further refinement of this model with the exclusion of antenna effects led to values between -1.4 in industrial and 1.5 in residential environments [26].

The approach adopted in this section is simpler and more natural as it relates directly to the propagation mechanism by estimating the frequency decay constant of the power spectrum. The proposed FDP model has the same functional form as the conventional distance-dependent pathloss, allowing a convenient formulation for the expression of the composite, bivariate pathloss $PL(d, f)$. The extraction of the FDP model parameter is based on the measured channel transfer function between a single transmit and receive antenna pair. The probabilistic variation of the FDP index, henceforth referred to as the dispersion index and denoted by ξ , is characterised and its small-scale spatial variation is analysed. With omni-directional antennas, the Friis freespace pathloss is nominally a quadratic function of frequency. In terms of the channel transfer function,

$$|H_{freespace}(f)| \propto f^{-\xi/2}, \quad \xi = 2. \quad (3.16)$$

Any departure from this ideal behaviour can be attributed to either multipath or frequency-selective scattering. For a channel with temporal and spatial stationarity, the exponent ξ that governs the dependence of the received power on frequency can be incorporated in the pathloss equation (3.3) by assuming the distance to be constant. Thus if C is the pathloss at f_i ,

$$PL(f)_{[\text{dB}]} = 10\xi \log_{10} f + C. \quad (3.17)$$

For the free-space channel with constant-gain (omni-directional) antennas, $\zeta=2$, which corresponds to 6 dB/decade pathloss [72]. With two constant aperture (directional) antennas, $\zeta=-2$, while for an asymmetric antenna pair, $\zeta=0$. However, system imperfections and channel frequency-selectivity can lead to a deviation from this ideal behaviour. The regression of FDP on frequency can be obtained from the gradient of the least-squares best-fit line that provides an estimate of ζ for the empirical data. Figure 3.7 shows a measured UWB transfer function as an example, along with the regression line.

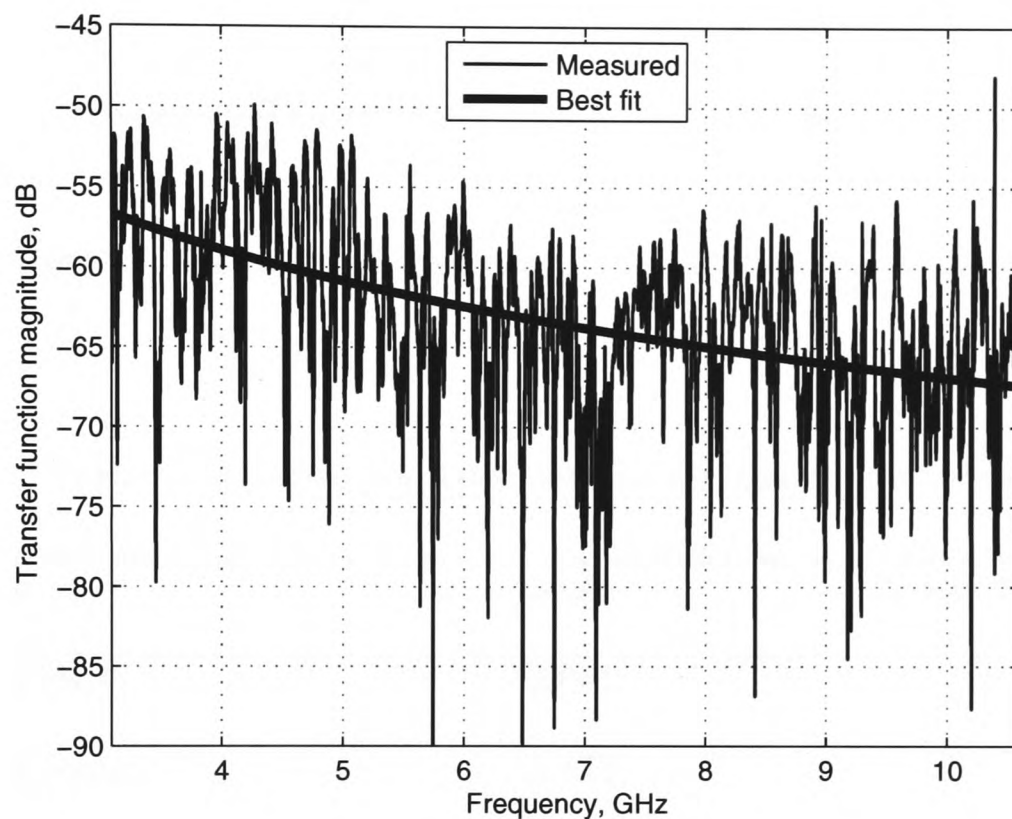


Figure 3.7 A measured channel transfer function in the indoor LOS channel and its best fit regression line from which the value of the frequency-dependent pathloss index, or dispersion index, is estimated

Next, ζ is characterised statistically using all of the measured channel responses in LOS and NLOS environments with vertical and horizontal polarisation. Figure 3.8 shows the empirical CDFs of ζ for all of these combinations. It can be observed that the mean ζ is 2 only in the vertically polarised LOS channel, denoted by V_{LOS} , and NLOS channels generally have higher ζ than LOS channels. Significant deviations of nearly the same order from the mean are observed in all cases, with a standard deviation of 0.2. The first order statistics for the dispersion index estimated from the measured data are summarised in Table 3.2.

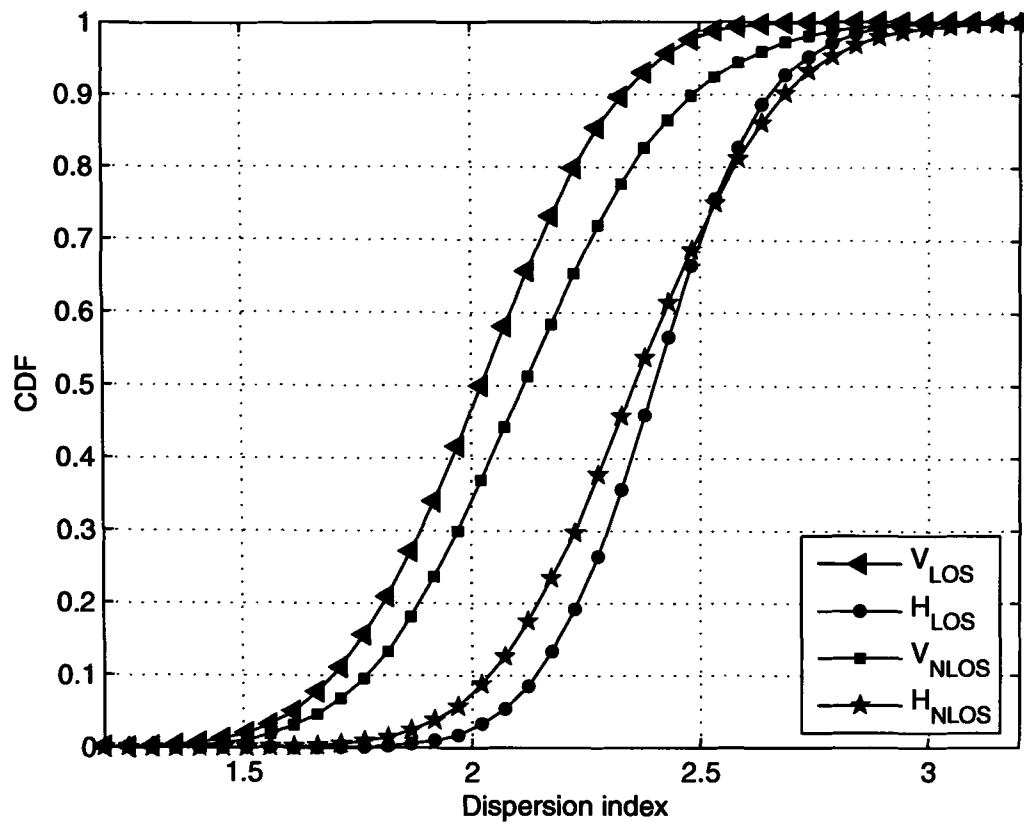


Figure 3.8 CDFs of the dispersion index in a variety of channel conditions

Table 3.2 Measured dispersion index for the indoor UWB channel

Scenario	Mean	Standard Deviation
LOS, vertical	2.043	0.245
LOS, horizontal	2.424	0.202
NLOS, vertical	2.148	0.283
NLOS, horizontal	2.386	0.255

Figure 3.9 shows the estimated values of ζ on the 1 m^2 planar measurement area, and reconstructs the interference patterns of ζ present in that region. The vertical channels experience lower ζ than the horizontal channels measured at the same positions on the grid, in general. This observation points to the inherent disparity in polarised wave propagation, analysed in greater detail later in this thesis. For the vertical channels, distinct interference fringes are observed. A closer examination and comparison with the measurement environment and grid position (see Chapter 2) reveals that the interference pattern is aligned with the orientations of the walls, indicating that the specific dispersion index

experienced at a spatial location is a function of the combination of multipath arrivals from nearby reflecting objects. Furthermore, the LOS and NLOS channels at a given polarisation exhibit similar behaviour in terms of the spatial variation of ζ , suggesting that it is not influenced to a great degree by the direct path alone. The horizontally polarised channels, on the other hand, have a more uniform distribution in addition to a higher mean ζ . The observation of greater large-scale variation in horizontal pathloss extends earlier results for narrowband channels [76]. The rapid fluctuation in ζ with only a short receiver displacement can lead to unpredictable performance. The vertical and horizontal polarisations offer a tradeoff, as one offers a low but rapidly fluctuating ζ and the other presents greater but more uniform dispersion.

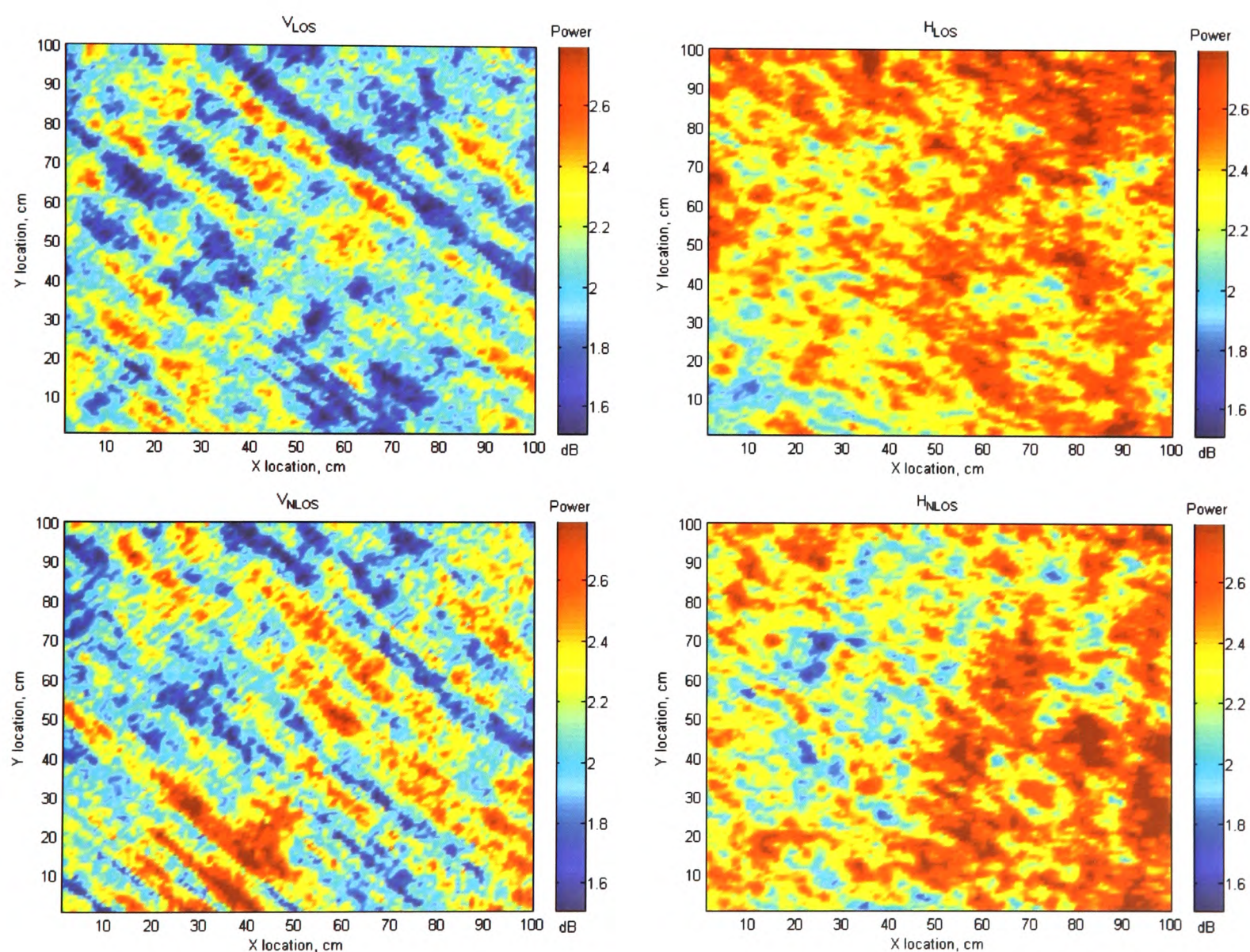


Figure 3.9 Spatial variation of dispersion index over a 1 m² horizontal plane area, obtained from a synthetic aperture grid measurement, in LOS and NLOS indoor environments

3.6. System Design Implications

A single-band pulsed UWB architecture, such as the Impulse Radio [9], typically uses a correlation receiver with carefully constructed signal and template waveforms. Power and pathloss variation with frequency results in spectral distortion and thus modification of the pulse shape during transmission, degrading the correlation receiver output considerably in the absence of a transmitted reference or adaptive template estimation scheme. Higher attenuation at some frequency ranges results in a reduced operating bandwidth for the UWB system, affecting its frequency diversity, increasing the outage probability, and lowering the achievable capacity. The dissimilar spectra of multipaths cause individual waveform distortion in each received multipath component as discussed in [64], lowering its correlation with the template waveform at the receiver in an unpredictable manner in the absence of suitable channel compensation.

The variation of delay spread with frequency restrains the achievable data-rate. For some intermediate values of the maximum excess delay spread, inter-symbol interference occurs for some frequency components but not others at high data-rates, complicating signal detection. A large rms delay spread indicates a dispersive channel and the distribution of incident power across a large number of paths. In the absence of a sufficiently complex rake receiver, this can lead to high bit-error rates. With a non-uniform delay spread and number of paths across the band, the performance of a fixed-complexity sub-optimal rake receiver would vary across the spectrum. The rake output would then have a relatively strong contribution from only a subset of the frequencies in the transmitted signal, limiting the captured power and capacity, and reducing the inherent frequency diversity.

Depending on the specific signalling mechanism and receiver implementation, sub-band pathloss variation may be less catastrophic in multi-band systems than in single-band systems. The variation of time dispersion and number of paths, however, still poses problems. It is wasteful for a multi-band receiver to operate a rake with the highest required diversity order required for any band. Delay spread variation between sub-bands would lead to different data-rates. An adaptive multi-band system capable of suitably catering to the frequency-dependent fading would then yield better performance than one operating uniformly across the whole UWB spectrum.

If the dispersion index $\xi > 0$, as is the case with an omni-directional antenna pair, the high frequencies in the signal are de-emphasised, altering its time-domain pulse-shape. From Fourier theory, therefore, a linear increase in pathloss with frequency f is in essence equivalent to the integration of a signal, apart from phase rotation and scaling. To analyse the effect of this dispersion, let us assume that a signal $s(t)$ with Fourier transform $S(f)$ is distorted due to propagation through a channel with dispersion index ξ . The received signal is then given by

$$r(t) = \mathcal{F}^{-1} \left[f^{-\nu} S(f) \right] = j2\pi D^{-\nu} s(t), \quad \forall \quad \nu = \frac{\xi}{2} \geq 0, \quad (3.18)$$

where $\mathcal{F}[\cdot]$ denotes the Fourier transform operation. The $D^{-\nu}$ operator in (3.18) denotes the ν -fold integral. As $\xi \in \mathcal{R}$, $D^{-\nu}$ represents the fractional integral given by [77]

$$D^{-\nu} x(t) = \frac{1}{\Gamma(\nu)} \int_0^t (t-\varsigma)^{\nu-1} x(\varsigma) d\varsigma, \quad (3.19)$$

with $\Gamma(\nu)$ representing the gamma function.

If, on the other hand, $\xi < 0$, the fractional derivative of $s(t)$ is obtained instead [77]. In terms of waveform distortion, a conventional narrowband signal using a continuous-wave carrier only undergoes a phase variation as a consequence of such fractional calculus operations. A pulsed UWB signal, on the other hand, suffers from pulse-shape distortion due to FDP that changes the functional form of the signal.

The impact of this waveform distortion on link performance can be illustrated with the simple example of a pulse-based receiver, such as the Impulse Radio system [9]. A popular pulse waveform for IR is the second derivative of Gaussian function, also known as the Mexican hat or W function [11], shown in Figure 3.10. For the analysis here, it is assumed that this pulse, $s(t)$, provides the excitation to the transmitting antenna. Furthermore, the channel in this treatment is inclusive of the antennas, so that the overall signal dispersion due to the antennas and propagation is characterised.

The result of filtering the waveform in Figure 3.10 through dispersive channels with $\xi = \{-4, \dots, 4\}$ is shown in Figure 3.11, where the signals $r(t)$ obtained from the receiving antennas are plotted. The effect of phase variation is also shown by plotting both the real and imaginary components.

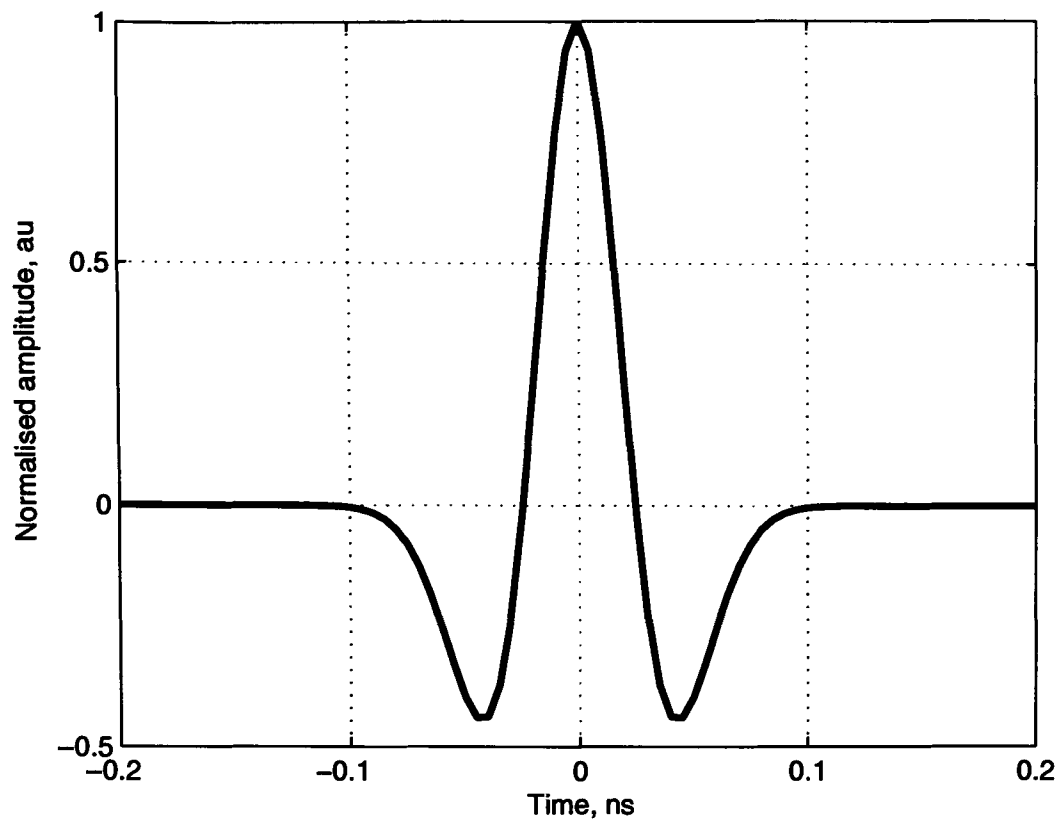


Figure 3.10 The second derivative of Gaussian pulse. The instantaneous amplitude is expressed in arbitrary units (au) after normalisation with the peak amplitude

It can be observed that a negative ζ transforms the original pulse waveform into higher order Gaussian pulses, an operation similar to differentiation, while a positive ζ results in signal waveform integration. The pulse shape is preserved only for $\zeta=0$, where no dispersion takes place. Higher order Gaussian pulses have smaller relative bandwidths and higher centre frequencies. Thus FDP results in the translation and scaling of the signal spectrum. Also, the signal oscillates between in-phase and quadrature states at even ζ . It is complex at odd ζ with equal power distribution between the real and imaginary parts. At fractional ζ , the signal assumes intermediate complex values. This modelling of the FDP helps evaluate the signal distortion that is caused due to the frequency-selective channel and the directional or omni-directional antennas. There is a variation in the degree of similarity between $s(t)$ and $r(t)$ with ζ , which can have a bearing on a correlator-based impulse radio.

Considering such a system operating in a dispersive UWB channel $h(t)$, the received signal is given by

$$r(t) = s(t) * h(t), \quad (3.20)$$

where $*$ signifies convolution, and the noise has been ignored for simplicity.

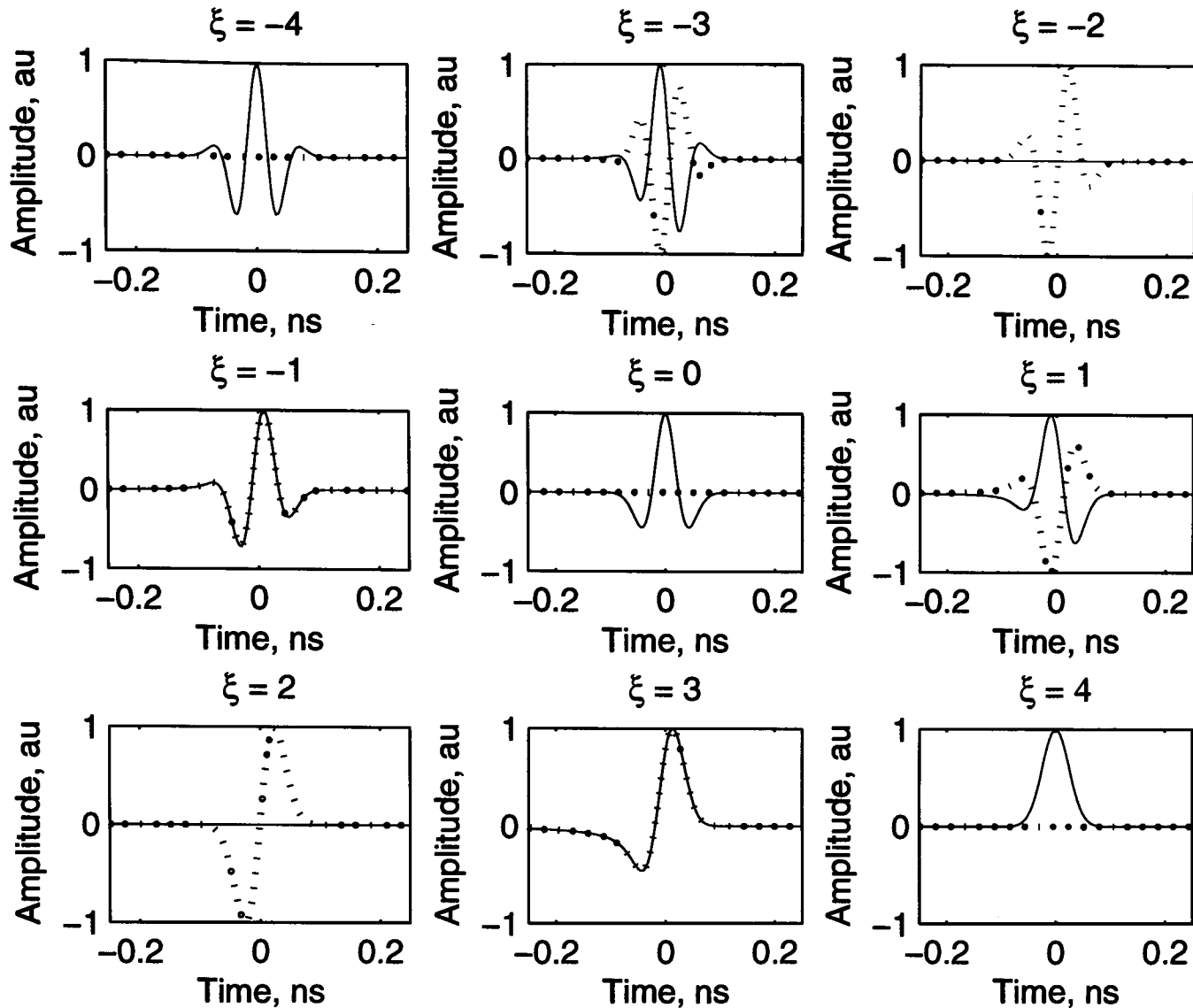


Figure 3.11 Pulse distortion as a function of dispersion index, ξ . The amplitudes are peak-normalised and expressed in arbitrary units. The solid blue line and the dotted red line show the real and imaginary parts

The receiver correlates $r(t)$ with the template waveform, which in the simple case is $s(t)$ itself, so that the correlator output is

$$r_c(t_j) = \int_0^{T_s} s(t) r^*(t + t_j) dt, \quad (3.21)$$

where t_j represents delay and accounts for the timing jitter, and T_s is the symbol time. For the sake of simplicity, multipath propagation and timing jitter are ignored, and the zero-delay correlation coefficient $r_c(0)$ is evaluated for the subsequent analysis. This correlation output, r_c , is then a scalar quantity that determines the receiver SNR and signal quality. Figure 3.12 shows the magnitude of the zero-delay complex correlation r_c in the dispersive UWB channel. From the plot, $\xi = \pm 2$ completely decorrelates the received signal from the template. It can be seen that the correlation is not an even

function due to the asymmetric time-domain waveforms of all but the even derivatives of Gaussian, as observed in Figure 3.11. Also shown in Figure 3.12 is the magnitude correlation, evaluated as

$$r_m(t_j) = \int_0^{T_s} |s(t)| |r(t+t_j)| dt. \quad (3.22)$$

From the plot, the zero-delay r_m can be approximated as a raised, damped sinusoidal function of ζ about $r_m=0.8$. The lowest correlation value, 0.66, is obtained at $\zeta=\pm 2$. Like r_c , r_m also does not exhibit even symmetry. Figure 3.12 also shows the SNR loss in a correlation receiver due to signal dispersion. The magnitude correlation suffers a loss of about 1.9 and 3.5 dB at $\zeta=\pm 1$ and ± 2 respectively, while the complex correlation falls by 3 dB at $\zeta=\pm 1$.

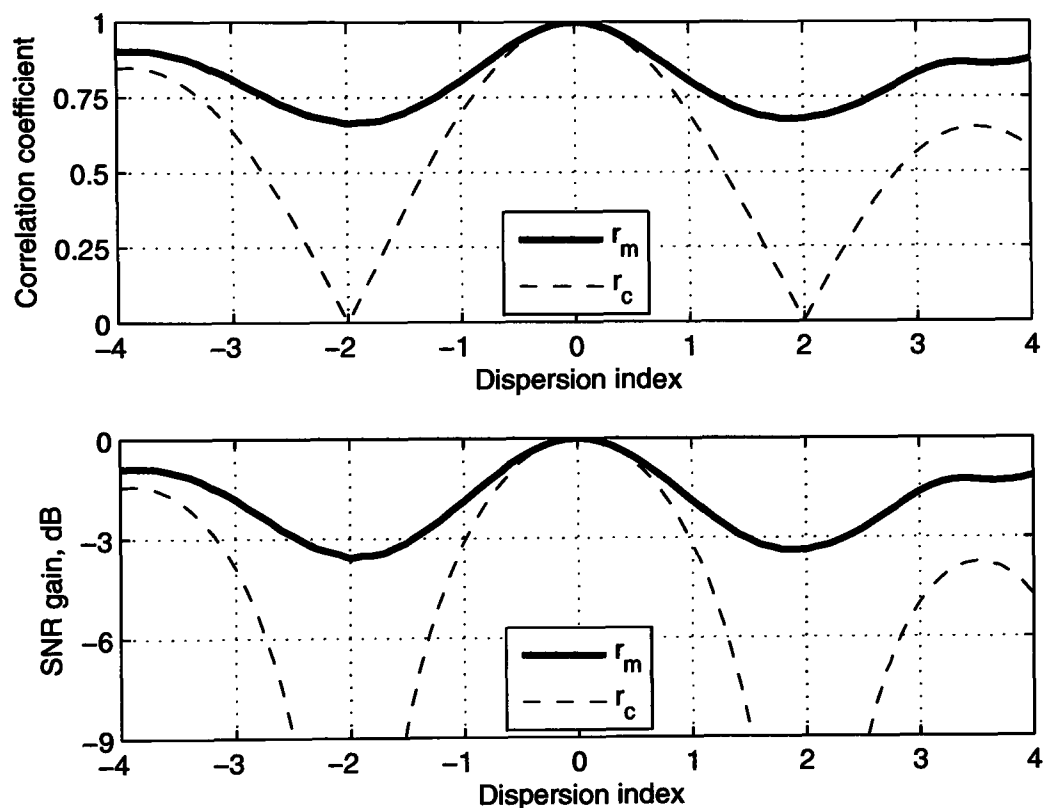


Figure 3.12 Correlation receiver performance in the presence of frequency-dependent pathloss and consequent waveform dispersion. r_m and r_c denote the magnitude and complex correlation coefficients

This analysis demonstrates the sensitivity of the communications link to the waveform dispersion caused by frequency-dependent pathloss. A correlation receiver operating in the UWB channels characterised in the earlier part of this chapter would be prone to detection errors. The damage due to the antenna-induced dispersion index can be mitigated with the use of a distortion-compensated template waveform at the correlation receiver. Thus for omni-directional antennas, the first derivative

of a Gaussian pulse can be used as template when the second derivative of Gaussian is transmitted. From Table 3.2 and Figure 3.12, this scheme will still suffer an SNR loss of over 0.5 dB in some propagation scenarios. The mitigation is further undermined by the fluctuation in the signal correlation arising from the variance of dispersion index due to the channel's spatial non-stationarity. The variation thus observed in ξ can result in a total loss of up to 1.2 dB when the antenna-compensated template is used. A possible solution is the adoption of a transmitted reference (TR) technique to compute the dispersion index periodically and use it for adaptive template generation. With this sophistication comes the requirement for channel state information at the receiver (CSI-R), whose cost is increased receiver complexity and data-rate reduction.

3.7. Summary

The UWB channel has been characterised in this chapter with respect to its frequency-dependent characteristics. Considerable variation of fading statistics such as pathloss and delay spread on the centre frequency has been established with the help of sub-band analysis. It has been shown that the higher sub-bands suffer greater attenuation, with power difference of up to 11 dB between the extreme sub-bands, due to spectrally non-uniform power density of omni-directional antennas. The number and powers of resolved multipaths vary between sub-bands, as some multipaths are absent at some frequencies due to frequency-selective fading. Synthetic aperture analysis of the multipath directions-of-arrival provides evidence of frequency-selective scattering. In order to quantify the frequency-dependent pathloss, the dispersion index, ξ , is introduced. Modelled with the power law, its first order statistics over the ensemble of measured channels are derived. Considerable deviation of the mean ξ from its nominal value of 2, induced by the antenna power spectral decay, is observed. Vertically and horizontally polarised signals experience a mean ξ of 2.1 and 2.4 respectively with a standard deviation of 0.2–0.3. The small-scale spatial behaviour of ξ over a planar area shows that horizontally polarised channel generally experiences higher but more uniform dispersion than the vertically polarised channel. The impact of this dispersion on link performance has been illustrated with the example of an impulse radio system using Gaussian signal waveforms. Fractional calculus has been used to model the

pulse shape transformation and consequent signal decorrelation. It is found that a correlation loss of up to 1.2 dB can occur even with a template designed to compensate for antenna dispersion. Such a situation can be frequently encountered in UWB multipath channels, posing a challenge for reliable signal detection and error-free communications. This work has demonstrated the importance of considering frequency-dependent propagation effects for robust system design that are usually ignored in the conventional modelling of UWB channels. The analysis so far has mostly assumed vertically polarised transmission and signal polarisation preservation during propagation. The next chapter analyses the polarisation characteristics of the UWB channel.

4. POLARIMETRIC CHANNEL CHARACTERISATION

4.1. Introduction

Conventional channel modelling involves the characterisation of various channel descriptors such as the pathloss, delay spread and directions-of-arrival [16, 20]. Another physical property of the channel that significantly affects signal propagation is its electromagnetic polarisation behaviour [48]. This consideration is important in rich multipath environments, such as the indoor channel, where significant signal depolarisation may occur due to multiple reflections from the walls and various scattering objects.

In case of co-polarised transmitting and receiving antennas, any channel-induced depolarisation leads to a polarisation mismatch between the receiving antenna and the wave incident on it. In the extreme case, such as in a reverberating chamber, complete depolarisation of the signal would result in the detection of only a small fraction of the incident power [78, 79]. As a perfect uni-polar antenna cannot sense an incident electric field at an orthogonal polarisation, a multiple-antenna receiver using polarisation diversity combining [80] that spans the entire polarisation basis is then required to extract the signal from all incident polarisations. Partial depolarisation occurs in less harsh scattering conditions, appearing as additional pathloss and affecting the link budget in a single polarisation receiver. If the channel is not stationary, the signal depolarisation and the consequent pathloss is unpredictable and can lead to outage. Such a situation can arise frequently in short- or medium-range wireless communications systems due to the high level of scattering and channel variability. In addition, the orientation, and thus polarisation, of a mobile station's antenna is often both variable and unknown at the base station, further compounding the unpredictability of polarisation-dependent signal loss. Accurate modelling of the channel polarisation characteristics is imperative for the elimination or mitigation of these channel impairments and provision of sustained quality of service.

This chapter investigates the polarisation characteristics of the UWB channel using dual-polar indoor LOS and NLOS channel measurements. The probabilistic behaviour of channel-induced polarisation rotation is characterised using an elliptical polarisation basis and the Stokes-Mueller

formulation. Signal polarisation purity is quantified in terms of the first order statistics for cross-polar discrimination (XPD), ellipticity angle and tilt angle. The probability distributions of XPD, degree of linear polarisation, and phase angle between orthogonal polarisations are estimated. The temporal and spectral evolution of XPD is determined. The dependence of XPD on the multipath direction-of-arrival (DoA) is explored through synthetic aperture angular spectra at various frequencies. Horizontal and vertical transmission link quality is compared in terms of pathloss and delay spread.

4.2. Narrowband Channel Polarisation Characteristics

The polarisation characteristics of narrowband, outdoor mobile channels have been studied in recent years, with the earliest research approaching the topic from the perspective of polarisation diversity [81]. Lower signal loss has been reported for vertically polarised narrowband signals than horizontally polarised signals in urban and suburban mobile channels in [81-84] and indoor channels in [76]. Narrowband horizontal polarisation was shown to experience greater incremental pathloss with distance than vertical polarisation in a hallway with a strong LOS component but not in an NLOS scenario inside a room [76, 85]. Contrary to these results, [86] reported that the received power with two horizontal antennas was 10 dB higher on average than that using two vertical antennas. The difference in performance of one polarisation over the other has been attributed to different transmissivities of walls and other surfaces for the two states of polarisation [16]. These studies demonstrate that the propagation characteristics of horizontally polarised fields differ considerably from those of vertically polarised fields in a given environment. In most wireless channels, vertically polarised signals suffer less depolarisation due to the vertical geometry of the dominant scattering objects, which are large buildings and walls in microcellular and picocellular communications channels respectively.

The XPD at the receiver provides a direct measure of the extent of depolarisation induced by the channel to a pure polarised transmission [87]. An inverse quantity, the cross-polar coupling, is also sometimes used to represent the same information [88], but the former is adopted here. In previous work, mean XPD values of 6-7 dB have been observed in an urban mobile environment [82, 83, 89].

The XPD increases with distance and falls with greater scattering, and sub-urban, indoor and NLOS channels experience relatively lower XPD [89, 90].

It is intuitively obvious that in a propagation environment with Ricean fading [16] and low depolarisation, the linearly polarised antennas at the two ends of the link should be co-polar for maximum power transfer [69]. A vertical transmitter and two receiver antennas oriented at $\pm 45^\circ$ were suggested to overcome differential fading statistics across the linear polarisation diversity branches in [91] but found to have little advantage over a traditional 0° – 90° receiver antenna pair in [89, 90].

4.3. Polarisation in the UWB Regime

The work cited above has addressed outdoor or narrowband channel depolarisation behaviour. The analysis for UWB channels is expected to be different in several ways. The indoor environment typically has a large number of electromagnetic scattering objects [41]. With UWB's superior spatial, temporal and angular resolution, individual scatterers can be resolved with greater precision. As signal depolarisation and entropy depend on the number of scattering events [92], it follows that the polarisation characteristics of the indoor UWB channel would differ from those of the narrowband or outdoor channels.

Furthermore, the depolarisation of the wave incident on a scatterer depends on the latter's reflection coefficient [48]. Frequency-variant depolarisation is therefore expected due to non-uniformities in scattering material properties across the UWB band [73]. This is further compounded by frequency-selective fading [69] and frequency-dependent pathloss, due to which the power spectral density (PSD) typically varies by over 10 dB over the UWB band (see Chapter 3), also causing the XPD to vary with frequency. Thus it is important to investigate the polarisation properties taking into account the effects of the wide bandwidth. Such a study has been performed in the domain of optical communications [93, 94], but not for radio channels.

The higher temporal resolution of UWB signals permits the analysis of the variation in polarisation characteristics over very short intervals typically on the order of a few picoseconds. This raises the question of how the polarisation characteristics of the channel affect the time-domain UWB

signal at the level of the individual pulses. The temporal evolution of polarisation properties for ultra-short optical signals has been discussed in [95]. With an indoor UWB system, the availability of this information would provide insight into the rate of decay of the multipath-dispersed, depolarised signal at the receiver in response to the transmission of an individual pulse representing a data symbol. For instance, a higher mean XPD corresponding to specific paths can be used to obtain a higher signal-to-noise ratio (SNR) by a selective rake receiver [96] with polarimetric capability.

Additional motivation can be drawn from the application of polarimetric analysis in synthetic aperture antenna array systems [70, 97] used for the determination of the multipath directions of arrival (DoAs) [98]. The signal processing techniques used for the generation and combination of co-polar and cross-polar angular spectra are of great interest in radar imaging [99] and beamforming communications systems [100, 101]. The variation of the polarised angular spectra over the large UWB frequency band therefore also merits an investigation.

4.4. Electromagnetic Wave Polarisation Representation

For the analysis of signal depolarisation during propagation, consider an electromagnetic wave with its Poynting vector, $\mathbf{P}$, oriented along the z-direction in the horizontal plane within the three-dimensional Cartesian space. Its electric field, present in the plane normal to the direction of propagation, can be expressed mathematically in terms of its complex orthogonal components as

$$\mathbf{E} = E_x \hat{\mathbf{e}}_x + E_y \hat{\mathbf{e}}_y = |E_x| e^{j(\omega t - kz + \delta_x)} \hat{\mathbf{e}}_x + |E_y| e^{j(\omega t - kz + \delta_y)} \hat{\mathbf{e}}_y \quad (4.1)$$

where $\hat{\mathbf{e}}_x$ and $\hat{\mathbf{e}}_y$ are the unit vectors along x and y directions, $k=2\pi/\lambda$ is the wavenumber, E_x and E_y are the magnitudes, and δ_x and δ_y the phase angles of the x and y components of the electric field respectively. The phase difference between the components is given by

$$\delta = \delta_x - \delta_y, \quad \delta \in [-\pi, \pi]. \quad (4.2)$$

The expression in (4.1) represents the general form of an elliptically polarised wave in the Ludwig III sense [87]. A special case of elliptical polarisation is linear polarisation, for which

$$\frac{|E_x|}{|E_y|} \in \{0, \infty\} \quad \text{or} \quad \delta = n\pi, n \in \{0, \pm 1\}, \quad (4.3)$$

while pure circular polarisation is governed by the condition

$$\frac{|E_x|}{|E_y|} = 1 \quad \text{and} \quad \delta = \pm \frac{\pi}{2}. \quad (4.4)$$

Thus the trace of the tip of a travelling wave defines an ellipse on a normal plane. The polarisation ellipse is specified by its axial ratio $1 \leq R \leq \infty$ and ellipticity angle $-\pi/4 \leq \chi \leq \pi/4$. The tilt angle of the ellipse, $-\pi/2 \leq \psi \leq \pi/2$, defines its orientation. With the decomposition of the wave into a basis with horizontal and vertical components E_h and E_v , the wave in (4.1) can be represented by

$$\mathbf{E} = E_h \hat{\mathbf{e}}_h + E_v \hat{\mathbf{e}}_v = |E_h| e^{j(\omega t - kz + \delta_h)} \hat{\mathbf{e}}_h + |E_v| e^{j(\omega t - kz + \delta_v)} \hat{\mathbf{e}}_v \quad (4.5)$$

with polarimetric phase angle

$$\delta = \varphi_h - \varphi_v, \quad \delta \in [-\pi, \pi]. \quad (4.6)$$

The polarisation ellipse in the horizontal-vertical basis is shown in Figure 4.1.

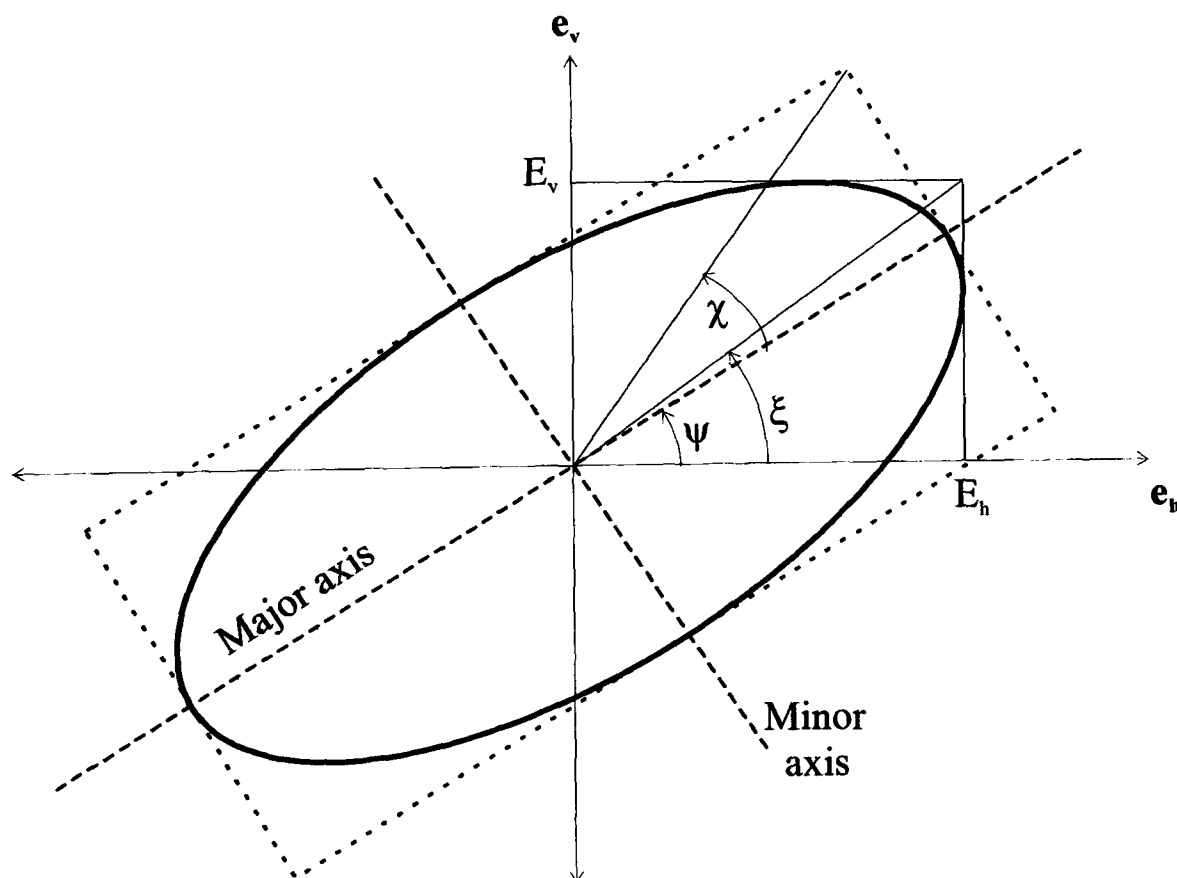


Figure 4.1 Electric field polarisation ellipse

The ellipticity angle of the incident field is given by

$$\chi = \frac{1}{2} \sin^{-1}(\sin 2\xi \sin \delta) \quad (4.7)$$

and the tilt angle is

$$\psi = \frac{1}{2} \tan^{-1}(\tan 2\xi \cos \delta) \quad (4.8)$$

where the inclination angle or auxiliary angle ξ is defined as the angle that the electric field makes with the x-axis

$$\xi = \tan^{-1} \left| \frac{E_h}{E_v} \right|. \quad (4.9)$$

Further explanation can be found, for instance, in [48].

4.5. Mueller Calculus

Mueller calculus is adopted here to evaluate the UWB channel's polarisation characteristics [87]. Channel depolarisation is formulated as the linear transformation of the incident wave's Stokes vector [102] into that of the scattered wave through the Mueller matrix [103] of the scattering medium, which in this case is the propagation channel. Polarisation entropy is also sometimes used to measure the degree of depolarisation caused by a system [104]. This involves the polarimetric decomposition of the scattering matrix into the coherency matrix through a set of complex orthonormal basis matrices (usually the Pauli spin matrices from quantum mechanics [105]) and then calculating the system's von Neumann entropy from the eigenvalues of the coherency matrix [106-108]. The coherency matrix and Mueller matrix provide alternative, interchangeable polarimetric representations [109]. The latter is adopted in this treatment to be conversant with the wider convention in communications literature.

The wave representation in Sec. 4.4 is used to construct the Jones matrix, or the polarimetric scattering matrix, $\mathbf{s} \in \mathbb{C}^{2 \times 2}$ that describes the polarisation response of the channel in terms of complex transmitted and received signals, given by [107, 108]

$$\mathbf{s} = \begin{bmatrix} s_{vv} & s_{vh} \\ s_{hv} & s_{hh} \end{bmatrix}. \quad (4.10)$$

Here, for instance, s_{vh} refers to a vertically transmitted, horizontally received signal. For the UWB channel, we can introduce frequency dependence to form the UWB scattering matrix,

$$\mathbf{s} = [\mathbf{s}^{(f)}], \quad f \in [f_l, f_h], \quad (4.11)$$

where $\mathbf{s}^{(f)}$ is the single-frequency scattering matrix corresponding to the frequency component f . The Stokes vector $\mathbf{S} \in \mathcal{R}^{4 \times 1}$ can represent the state of polarisation (SOP) of a partially polarised wave [102]

$$\mathbf{S} = \begin{bmatrix} S_0 \\ S_1 \\ S_2 \\ S_3 \end{bmatrix} = \begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} S_0 \\ S_0 \cos 2\psi \cos 2\chi \\ S_0 \sin 2\psi \cos 2\chi \\ S_0 \sin 2\chi \end{bmatrix} = \begin{bmatrix} \langle |E_v|^2 \rangle + \langle |E_h|^2 \rangle \\ \langle |E_v|^2 \rangle - \langle |E_h|^2 \rangle \\ 2\langle E_v E_h \cos \delta \rangle \\ 2\langle E_v E_h \sin \delta \rangle \end{bmatrix} = \begin{bmatrix} \langle |E_v|^2 \rangle + \langle |E_h|^2 \rangle \\ \langle |E_v|^2 \rangle - \langle |E_h|^2 \rangle \\ \langle E_v^* E_h + E_v E_h^* \rangle \\ \langle E_v^* E_h - E_v E_h^* \rangle \end{bmatrix} \quad (4.12)$$

The Stokes parameter, S_0 , in (4.12) represents the total power, S_1 the degree of linear horizontal polarisation, S_2 the linear polarisation along 45° , and S_3 the amount of circular polarisation. The Stokes formalism can also be used to derive a geometric representation of the SOP in the form of the Poincaré sphere [110]. The depolarising effect of a system on an electromagnetic wave can be represented by a linear transformation, which is equivalent to rotations on the Poincaré sphere, and is given by

$$\mathbf{S}_s = \mathbf{M} \mathbf{S}_i. \quad (4.13)$$

where the 4x4 Mueller matrix $\mathbf{M}$ [103] characterises the polarisation properties of the scattering medium. The subscripts i and s in (4.13) refer to the incident and scattered waves. Using this framework, a partially polarised signal can be expressed by

$$D^2 S_0^2 \geq S_1^2 + S_2^2 + S_3^2 \quad (4.14)$$

where D is the degree of polarisation. For the vertically polarised transmitted signal, the degree of polarisation of the received signal is given by

$$D_v = \frac{|E_v| - |E_h|}{|E_v| + |E_h|} = \frac{\sqrt{\sum_{n=1}^3 S_n^2}}{S_0}, \quad 0 \leq D_v \leq 1 \quad (4.15)$$

The degree of linear polarisation in the received signal can be calculated as

$$D_l = \frac{S_1}{S_0} \quad (4.16)$$

and the tilt angle is

$$\psi = \frac{1}{2} \tan^{-1} \left(\frac{S_2}{S_1} \right). \quad (4.17)$$

The XPD is the isolation between the received co-polar and cross-polar electric field components. For a vertically polarised wave incident on the scattering channel, the XPD is given by

$$X = 20 \log_{10} \frac{|E_{vv}|}{|E_{vh}|}. \quad (4.18)$$

4.6. Analysis Procedure

To understand the polarisation effects of the UWB channel, a VNA-assisted experimental evaluation of its properties was undertaken. The channel sounding process described in Chapter 2 was separately repeated for each of the four possible polarisation combinations, i.e., VV, VH, HV, and HH. Here V and H denote vertical and horizontal polarisations respectively, and the first letter refers to the polarisation of the transmitted signal. The four measurements were carried out sequentially so that there was only one antenna at each end to avoid antenna coupling. The rest of this chapter describes the analysis and observations following this experimental procedure.

4.7. Channel Depolarisation Statistics

This section provides a statistical treatment of the indoor UWB channel in terms of the polarisation properties defined above and using the measured channel data. The probability distributions of those quantities are estimated by treating them as continuous random variable, an approximation justified by the large ensemble size. The four possible cases defined by the vertical and horizontal transmissions in the LOS and NLOS channels are considered. The probability density

function (PDF) of the measured XPD, degree of linear polarisation, phase difference, ellipticity angle, and tilt angle for each polarisation case is estimated as

$$P_{m,n}(x), \quad m \in \{v, h\}, n \in \{LOS, NLOS\}, x \in \{D_l, \delta, X, \chi, \psi\}, \quad (4.19)$$

where m refers to the vertical or horizontal transmit polarisation, n refers to the LOS or NLOS propagation scenario, and x is the random variable examined over all spatial and frequency points.

The relationship between the magnitudes of the co-polarised and cross-polarised fields is explored through X , as shown in Figure 4.2. X provides information about the polarisation preservation and the coupling across orthogonal polarisations. A value of $X=\infty$ indicates perfect polarisation, $X=0$ represents complete and uniform depolarisation, while $X<0$ shows that there is high coupling with the orthogonal polarisation.

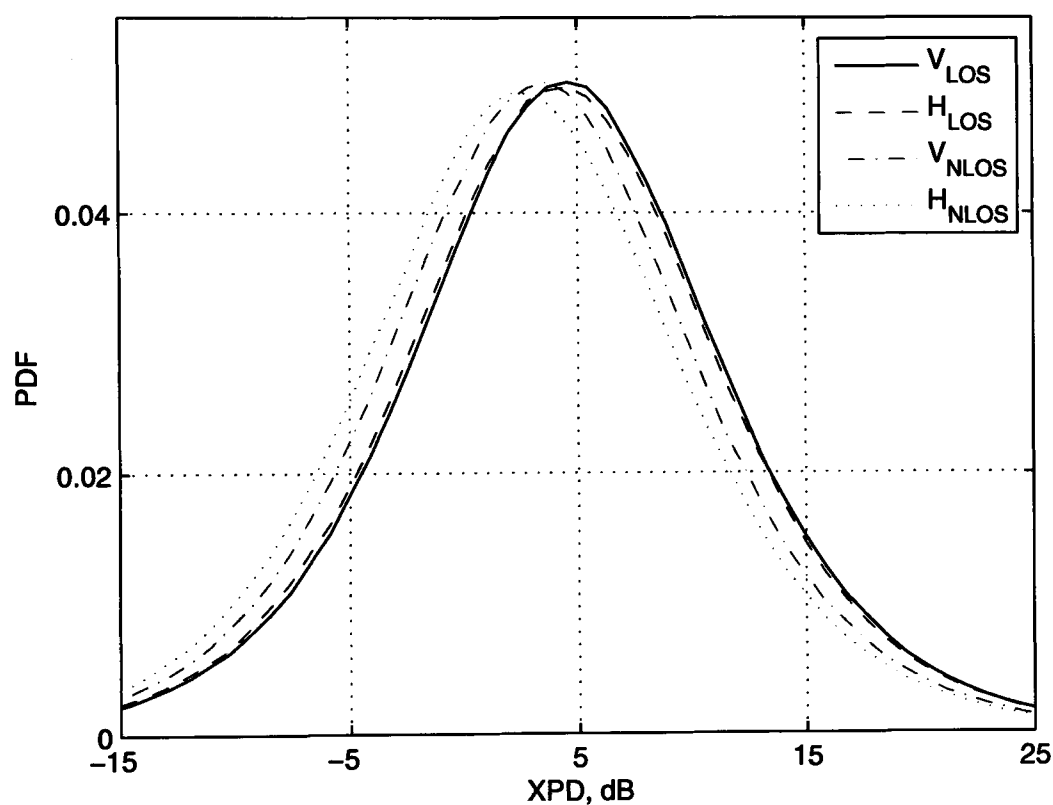


Figure 4.2 Probability density estimates of cross-polar discrimination (XPD)

From our measurement, the horizontally polarised signal in the NLOS channel appears to have the lowest mean X , while the LOS and NLOS vertically polarised signals enjoy the highest levels of mean X . From the shape of the PDFs, X can be well approximated with a Gaussian random variable for each propagation scenario. X has a large range of values, and there is non-zero probability of $X<0$. Thus cross-polar power can sometimes be higher than the co-polar power. The mean X for V_{LOS} is 4.5 dB,

while for H_{NLOS} it is 2.31 dB. The standard deviation is 8 dB for all four cases, indicating a large spread compared to the mean. The values of mean and standard deviation of X for the four cases are listed in Table 4.1.

The degree of signal polarisation D , with the receiving antenna polarisation matched to that of the transmitting antenna, provides a measure of the depolarisation introduced by the channel. A polarisation preserving medium yields a scattered wave such that $D = 1$, while smaller values of D indicate proportionate depolarisation. The four PDFs of D_l appear to have an exponential form, as shown in Figure 4.3. Most of the received signal is linearly polarised, as the peak of the PDF occurs at unity. There is, however, a non-zero probability of receiving a depolarised signal. From the figure, this is true for both polarisations and propagation scenarios, but there is relatively higher probability of $D_l = 1$ for the V_{LOS} channel compared to the other channels. The mean D_l for V_{LOS} is 0.81 and its standard deviation is 0.21. From Table 4.1, the corresponding values of the other three cases are also similar.

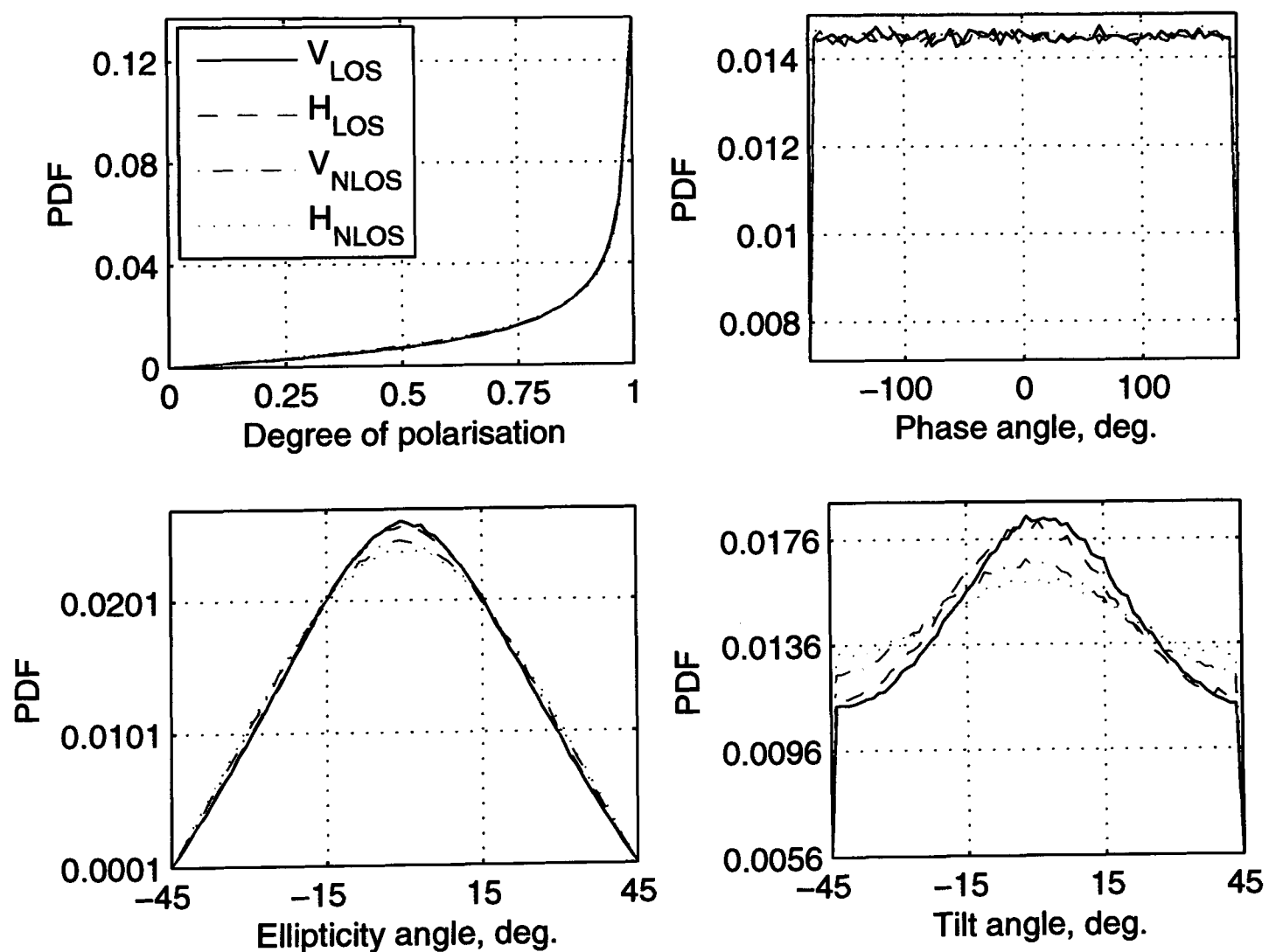


Figure 4.3 Empirical PDFs of the polarisation ellipse parameters: degree of linear polarisation, phase difference between orthogonal components, ellipticity angle and tilt angle

With the evidence that some polarisation rotation is taking place as shown by the finite X and the deviation of D from unity, the phase difference, δ , between the received vertical and horizontal field components is analysed. The PDFs of δ for the set of four linearly polarised channels are shown in Figure 4.3. It is found that the horizontal and vertical field components arrive with a uniformly distributed phase difference in the range $[-\pi, \pi]$, or there is no relationship between their phases. As listed in Table 4.1, δ is a zero-mean uniform random variable with a standard deviation of about 104° (or $7\pi/12$) in each case.

It is obvious from the analysis of D_l and δ that the incident wave is elliptically polarised, and cannot be represented by a linear polarisation state. The ellipticity angle χ is calculated next and is found to have a mean value of 0 and a standard deviation of 18° . Shown in Figure 4.3, the PDFs of χ are all of the same form with an approximately triangular shape. It can be inferred that the co-polar component has much higher power than the cross-polar component, yet the latter is non-zero.

Table 4.1 First order statistics for elliptical polarisation parameters

Polarisation	Median				Standard Deviation			
Parameter	V _{LOS}	H _{LOS}	V _{NLOS}	H _{NLOS}	V _{LOS}	H _{LOS}	V _{NLOS}	H _{NLOS}
DOP	0.81	0.80	0.81	0.79	0.21	0.21	0.21	0.22
XPD, dB	4.50	3.68	4.89	2.31	7.87	7.94	7.93	8.01
Ellipticity angle	0.19°	-1.23°	0.31°	-0.39°	18.32°	18.70°	18.13°	19.20°
Tilt angle	0.62°	0.11°	-0.02°	-0.14°	24.00°	24.49°	23.60°	25.22°
Phase difference	0.08°	0.09°	-0.07°	0.10°	103.93°	103.94°	103.95°	103.91°

Another parameter of interest is the polarisation tilt angle, ψ , which measures the shift in the tilt of the major axis of the polarisation ellipse from the transmitted linear polarisation. The probability densities of ψ are shown in Figure 4.3. Although the mean lies close to 0° , the range of values extends nearly all the way to the extrema. The shapes of the PDFs resemble truncated normal distributions. It can also be observed that $p_{H, NLOS}(\psi)$ comes closest to a uniform distribution as compared to the other

tilt angle PDFs, implying that in this case the polarisation ellipse would experience the widest range of shifts from its nominal tilt.

4.8. Temporal and Spectral Evolution

In this section, the variation of the channel depolarisation characteristics with time and frequency is investigated. The mean XPD is analysed, which now becomes $X(f)$ or $X(\tau)$, where $f \in [f_l, f_h]$ is the range of frequencies within the FCC UWB band, and τ is the time-delay. The average is taken over the measurement ensemble. The four analysis cases are the same as described in the previous section, i.e. V and H transmissions under LOS and NLOS propagation. The maximum excess delay of the indoor UWB channel is 50 ns (see Chapter 2). For the temporal analysis here, τ is restricted to a maximum of 20 ns, as most of the signal power is received within this delay window considering that the PDP follows an exponential decay [41].

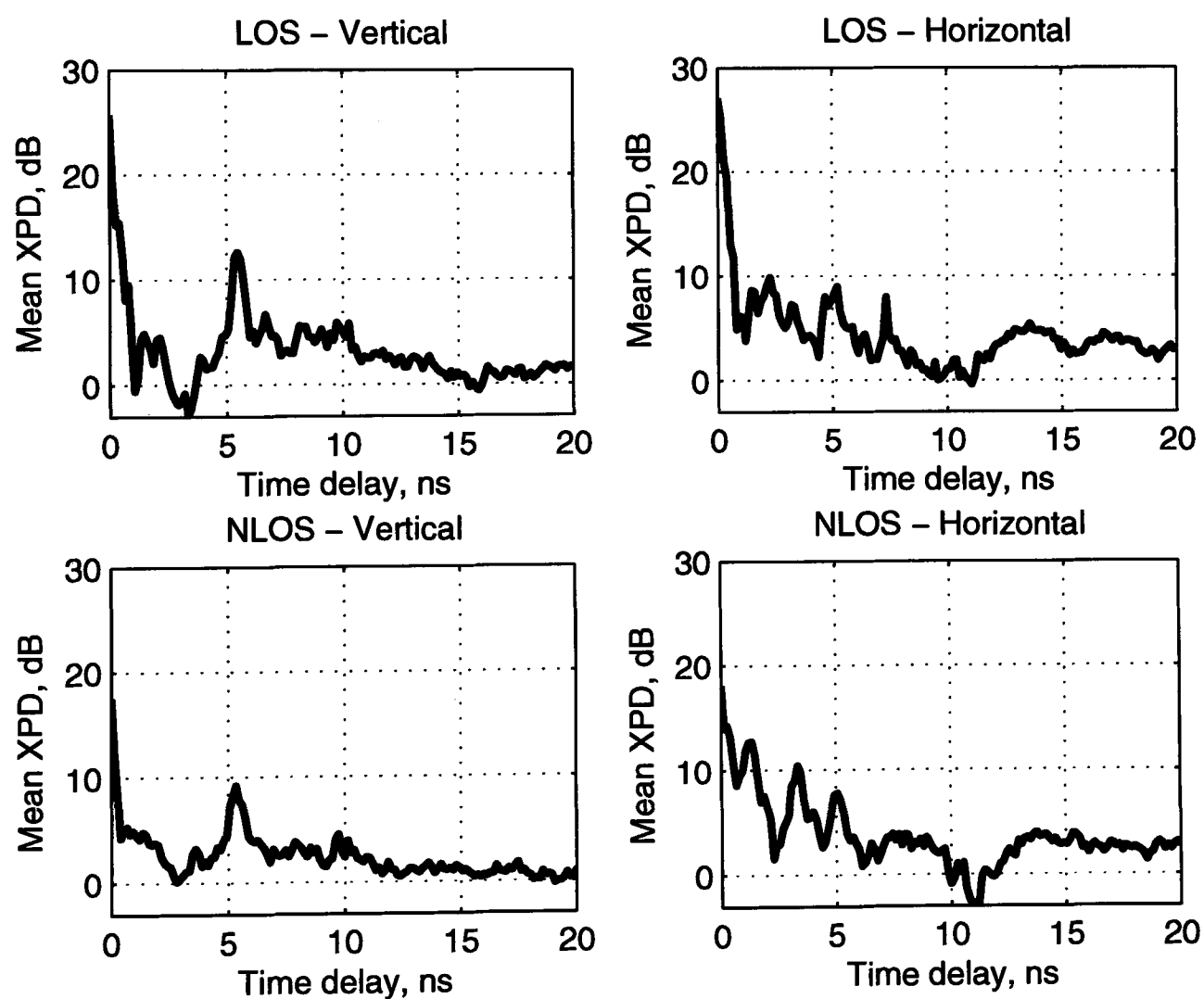


Figure 4.4 Temporal evolution of XPD with vertically or horizontally polarised signal transmission

The temporal evolution of the mean XPD is shown in Figure 4.4 for vertical and horizontal transmissions. The XPD is higher than 10 dB during the first 2 ns, after which it falls rapidly with a small number of oscillations. Thus the polarisation state is maintained by the signal arriving after up to a single reflection, while significant polarisation rotation takes place for the secondary and tertiary multipaths. The inspection of the LOS $X(\tau)$ reveals that the high XPD time-instants coincide with the strong, early multipath arrivals. Within 10–12 ns, a steady XPD value of 1–4 dB is achieved. For each polarisation, the LOS and NLOS $X(\tau)$ are similar, indicating robustness to direct path blockage.

The variation of the mean XPD across the UWB frequency band is depicted in Figure 4.5. There is strong similarity between the corresponding polarisations at LOS and NLOS. $X_h(f)$ and $X_h(f)$ differ in the same scenario, more so in NLOS. For instance, $X_h(f)$ is low in the 7 GHz region but $X_v(f)$ is high in that sub-band. This suggests that a UWB system with dual-polar diversity can provide an appreciable diversity gain, and provides the motivation for the analysis of UWB polarisation diversity in Chapter 7.

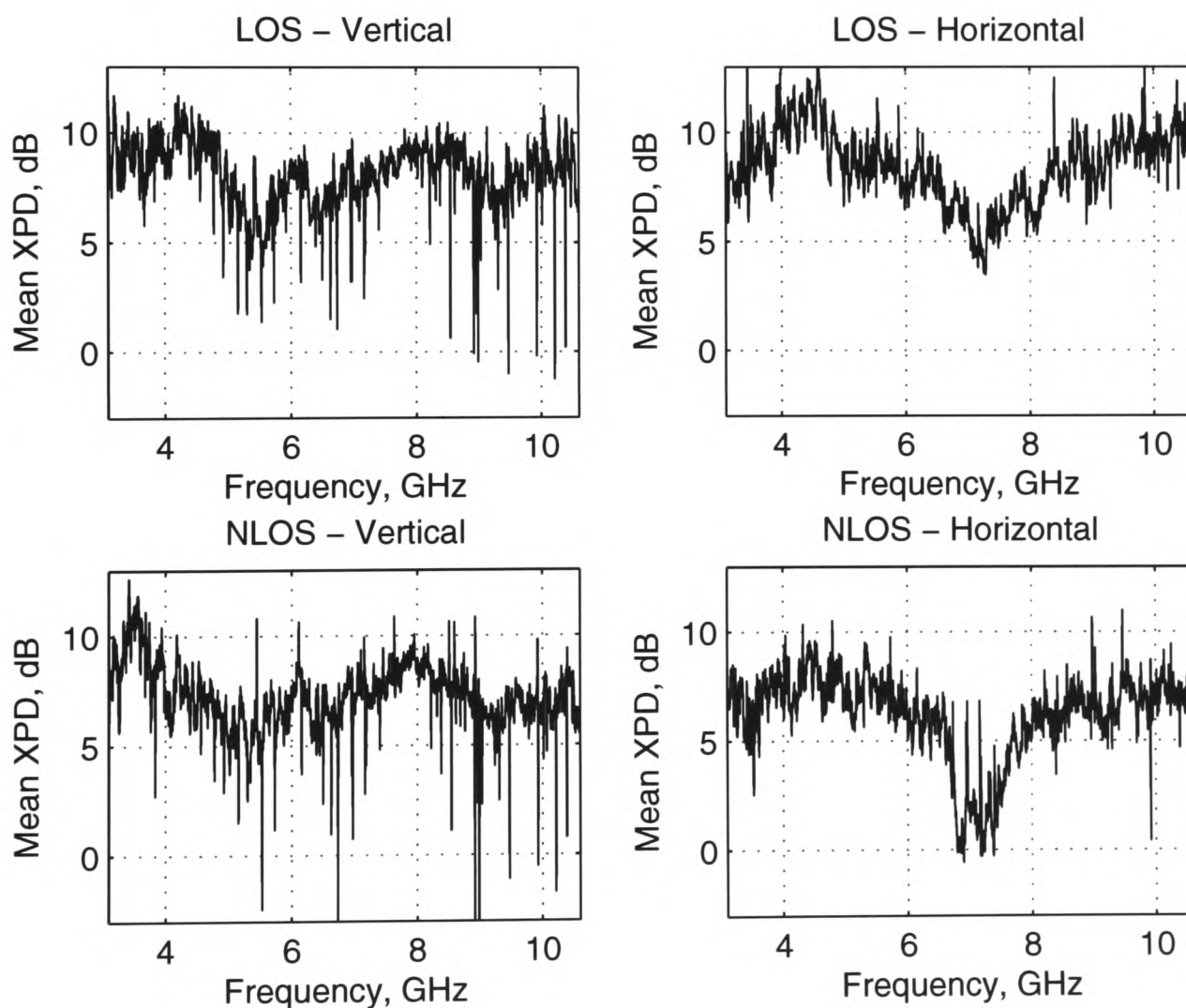


Figure 4.5 Frequency variation of XPD with vertically or horizontally polarised signals

4.9. Polarised Angular Power Spectra

In this section, the variation of the channel's polarisation properties with frequency is investigated in terms of the multipath DoAs calculated through the Fourier method. The analysis is based on single-frequency angular power spectra $A(\theta, \phi, f)$ corresponding to the elements of the Jones matrix in (4.10) for the LOS and NLOS channels. The angular spectra of the co- and cross-polar channels for a given frequency are normalised to the highest angular power level received at that frequency, in order to provide a good visual representation. The images are also thresholded at the -25 dB level to suppress noise. The angular spectra at the centre frequency, $f_c=6.85$ GHz, in the UWB band are shown in Figure 4.6 for the LOS channel.

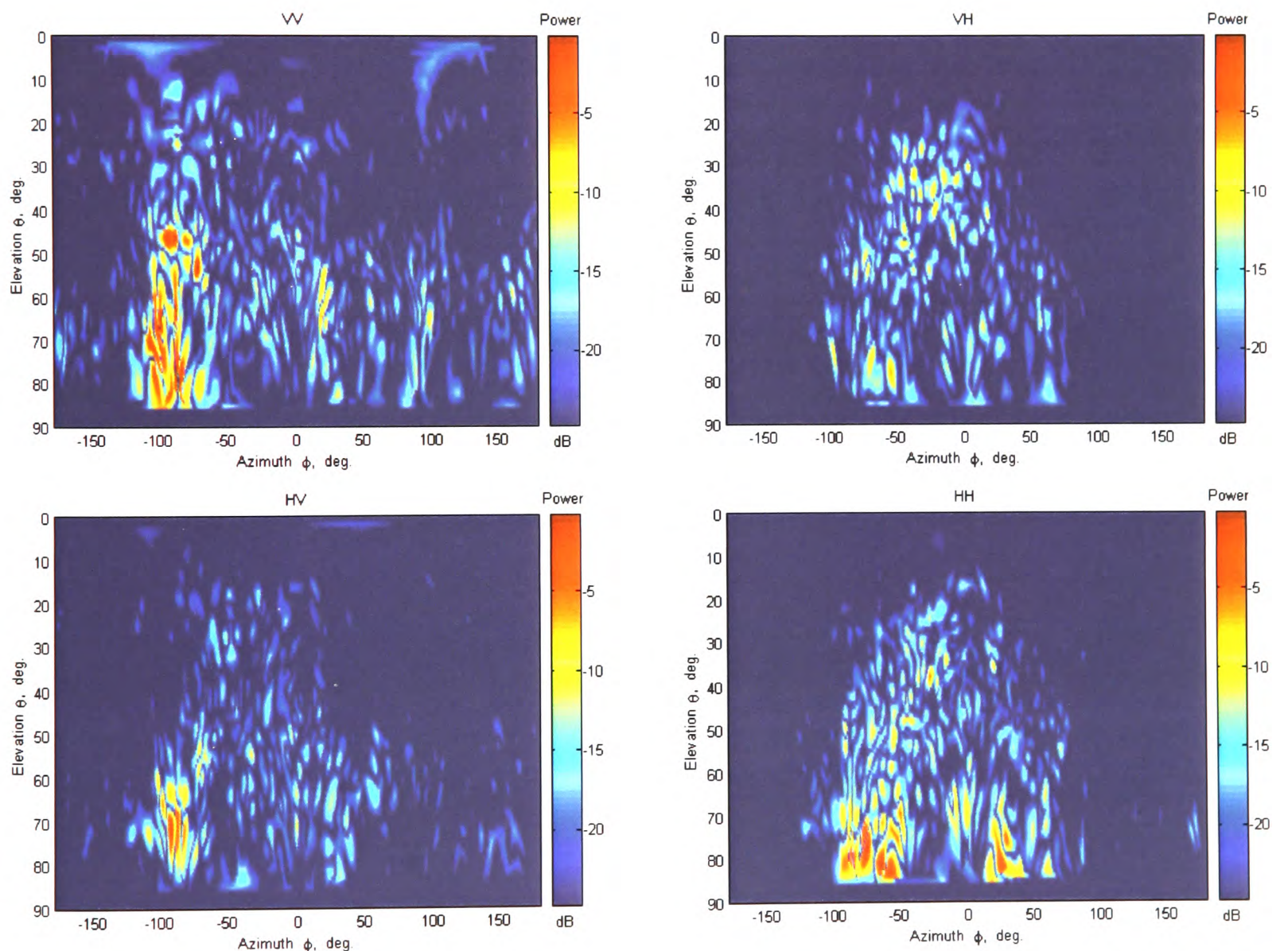


Figure 4.6 Polarimetric angular power spectra at the UWB centre frequency (6.85 GHz) for the LOS channel, normalised to the highest intensity across polarisations for comparison, with a -25 dB threshold

After the coordinate transformation, the axes are oriented such that the transmitter is located at $(\theta, \phi) = (90^\circ, -90^\circ)$. In the VV image, the reflection from the ceiling is also visible above the direct path,

as are the corners of the room. A large number of reflections can be observed close to the azimuthal plane, which some strong reflections are also visible. The other co-polar image corresponding to the HH channel shows an angular spectrum restricted to a short range of ϕ , due to the radiation pattern of the discone antennas [43]. Thus a consequence of using the horizontal polarisation is the qualification of the angular spectrum in a limited range of directions, which can be useful in situations that require low angular spread. Also, the strengths of the signals from some directions within the common field of view are not identical, indicating the directions of the scattering objects that reflect one polarisation more strongly than the other. This is true, for example, for the scatterer in the HH image at $(\theta, \phi) = (80^\circ, 40^\circ)$, which is missing in the VV image. The cross-polar spectra, VH and HV, also contain an appreciable amount of signal power. This polarisation rotation is caused by the scattering events. Also, the two cross-polar spectra are not identical, and less overall power is contained in the HV spectrum than the VH spectrum. This result demonstrates that the polarised channel scattering is not reciprocal as often assumed in the case of monostatic radars [101].

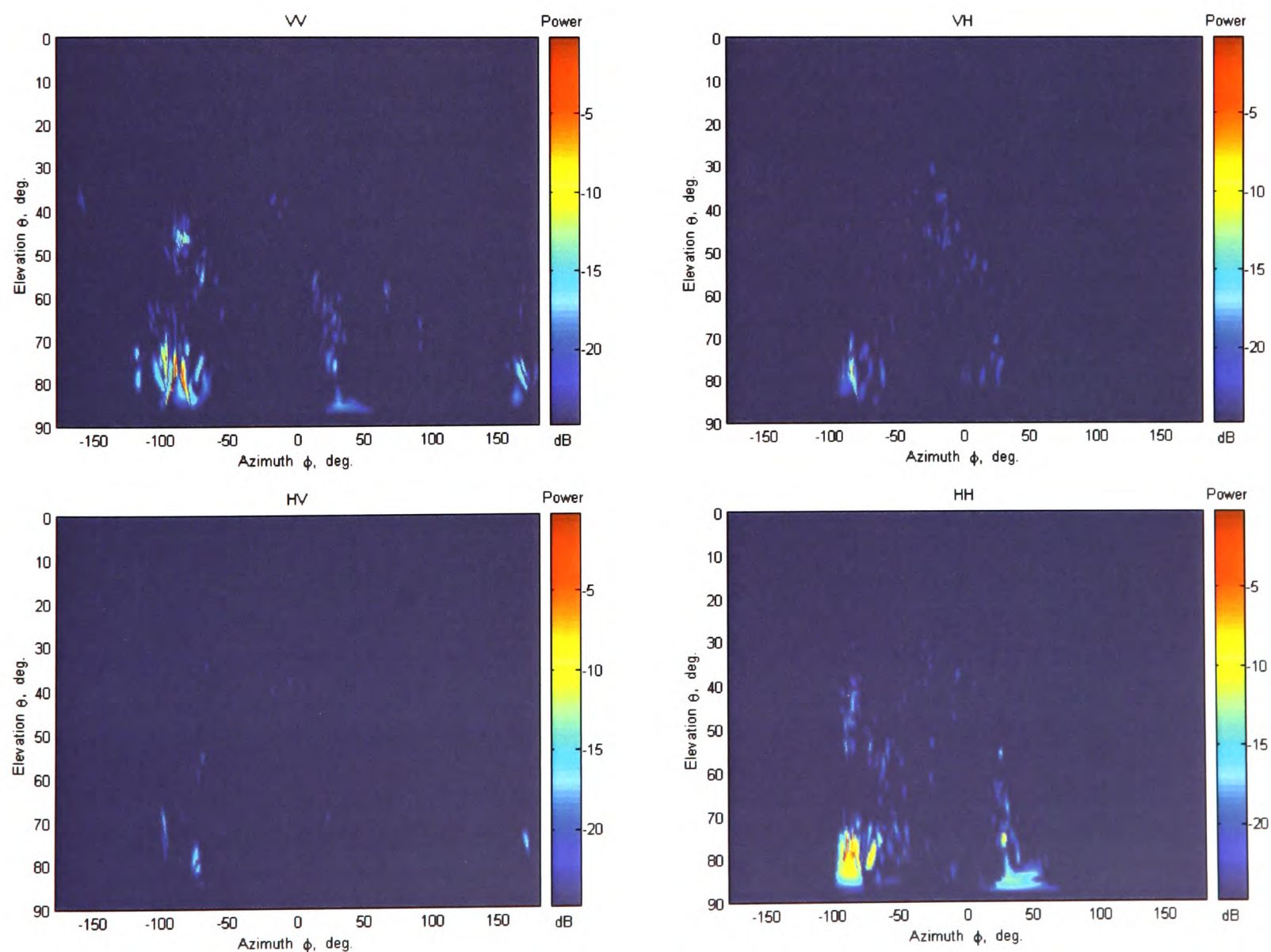


Figure 4.7 LOS angular spectra at the UWB highest frequency (10.6 GHz) with global normalisation

Due to the unique information present in the four images, a polarimetric radar, location-finding or imaging system can combine the power in all of these spectra and provide a better estimate of the direction, orientation and material composition of a target or the scattering environment. The angular spectra calculated at the highest frequency in the UWB band, $f_h=10.6$ GHz, are shown in Figure 4.7. The VV and HH images still contain some differences, which are made more obvious by the greater angular resolution, noise rejection and contrast of the images due to increase in frequency. There is, for instance, a scatterer on the far right perceived only with the vertical polarisation. There is lower polarisation cross-coupling at this frequency as shown by the VH and HV images, in comparison with the images at f_c in Figure 4.6. An analysis of the NLOS channel's angular spectra, shown in Figure 4.8 at f_c , leads to similar observations. In the absence of a dominant multipath, a larger number of incident signals lie above the noise threshold. The previous observations about the absence of polarimetric reciprocity and non-uniformity of the channel polarisation properties across the spectrum also hold true. Higher polarisation leakage is observed than for the LOS scenario.

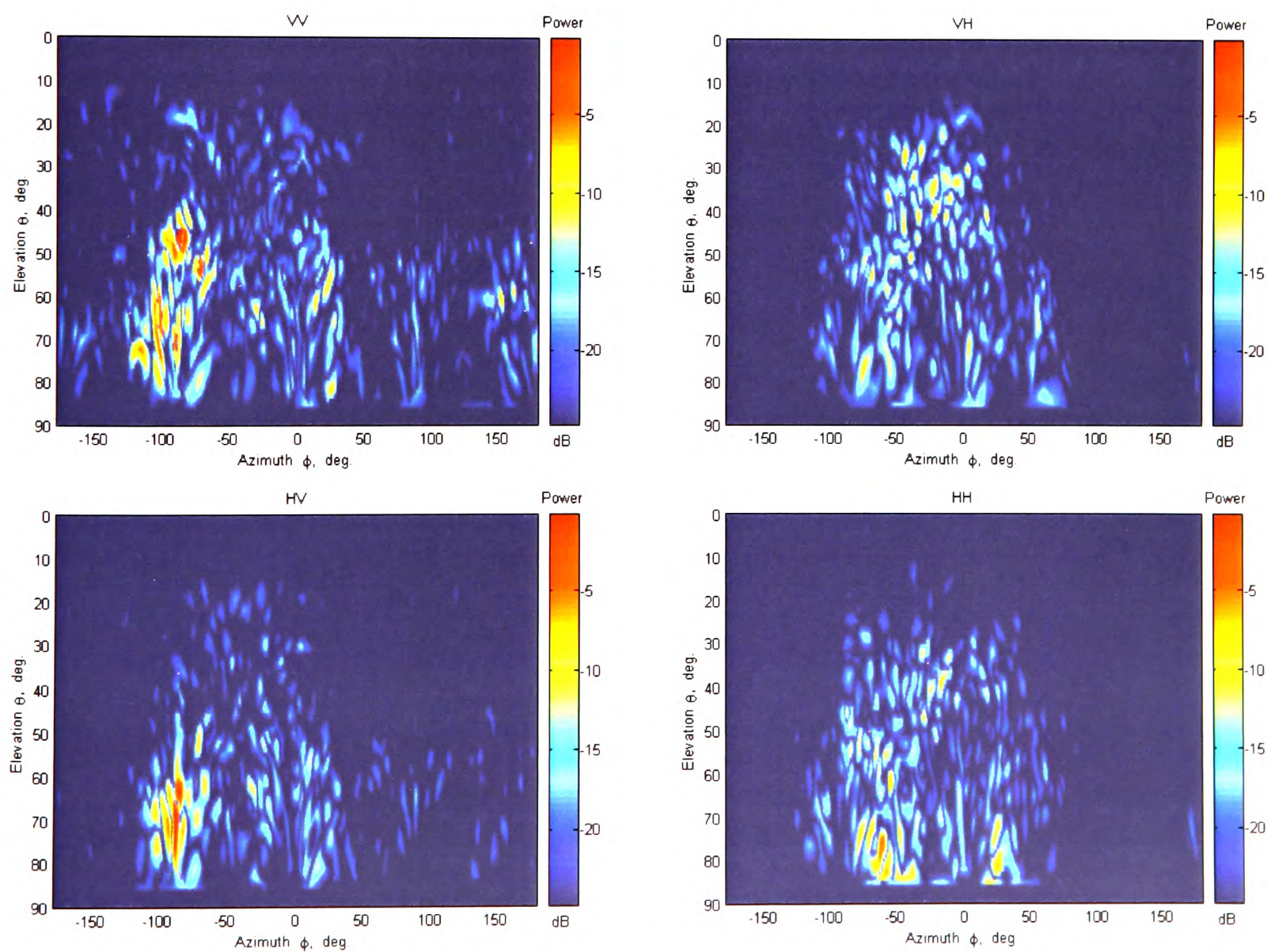


Figure 4.8 NLOS angular spectra at the UWB centre frequency with global power normalisation

4.10. Comparative Link Performance

The four polarised channels are analysed in this section in terms of classical channel description parameters, namely rms delay spread and small-scale pathloss [16]. These PDP parameters provide insight into the communications link performance and provide sufficient system design information for most purposes [19], as discussed in Chapter 2.

The τ_{rms} CDFs for each of the four polarised channels in both LOS and NLOS propagation scenarios from our measurements are shown in Figure 4.9. It can be seen from the figure that the VV impulse response in both LOS and NLOS cases undergoes the lowest dispersion. The other co-polar link, HH, displays a higher spread than VV, and the difference is much more pronounced in NLOS than LOS. The two cross-polar links suffer from a large rms delay spread, which indicates the likelihood of absence of dominant early multipaths and the division of power into a much larger number of paths. It is also seen that the HH link performance is very sensitive to the presence of line of sight. This is undesirable from the perspective of robustness and reliability, as it leads to severe performance degradation under LOS blockage. Finally, the two cross-polar links have distinct τ_{rms} CDFs, confirming earlier observations about non-reciprocity.

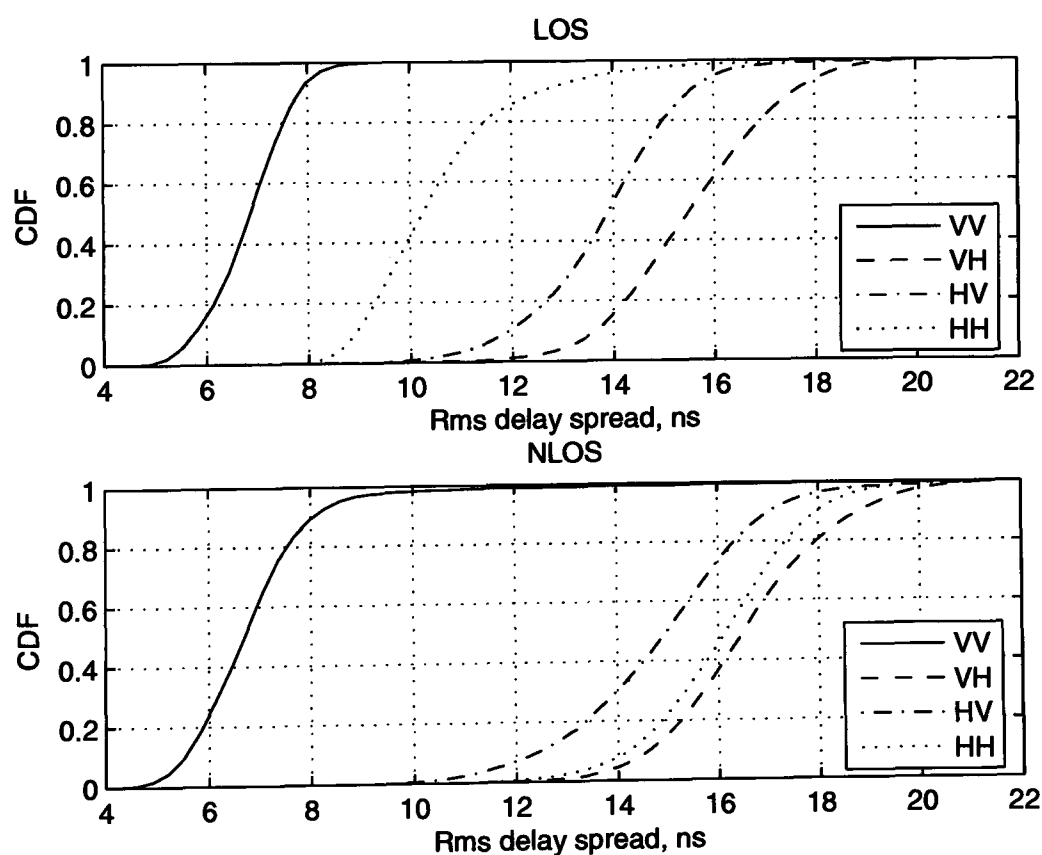


Figure 4.9 Empirical CDFs of the rms delay spread

The CDFs of pathloss, PL_{dB} , for the four polarised channels in LOS and NLOS scenarios are shown in Figure 4.10. For the sake of comparison, each measured pathloss value is normalised to the mean VV_{LOS} pathloss. The VV channel appears to suffer the least pathloss, carrying the most power to the receiver. However, PL_{dB} of the VV_{NLOS} channel is higher than that of the VV_{LOS} channel, as the energy of the direct path does not reach the receiver in the former case. HH_{NLOS} performs up to about 5 dB worse than VV_{LOS} , confirming that there is lower polarisation-preserving scattering for a horizontally polarised electric field. The two co-polar channels differ by only about 1–2 dB. On the basis of similar pathloss, they can be used effectively in MIMO communications [111], as discussed in Chapter 7.

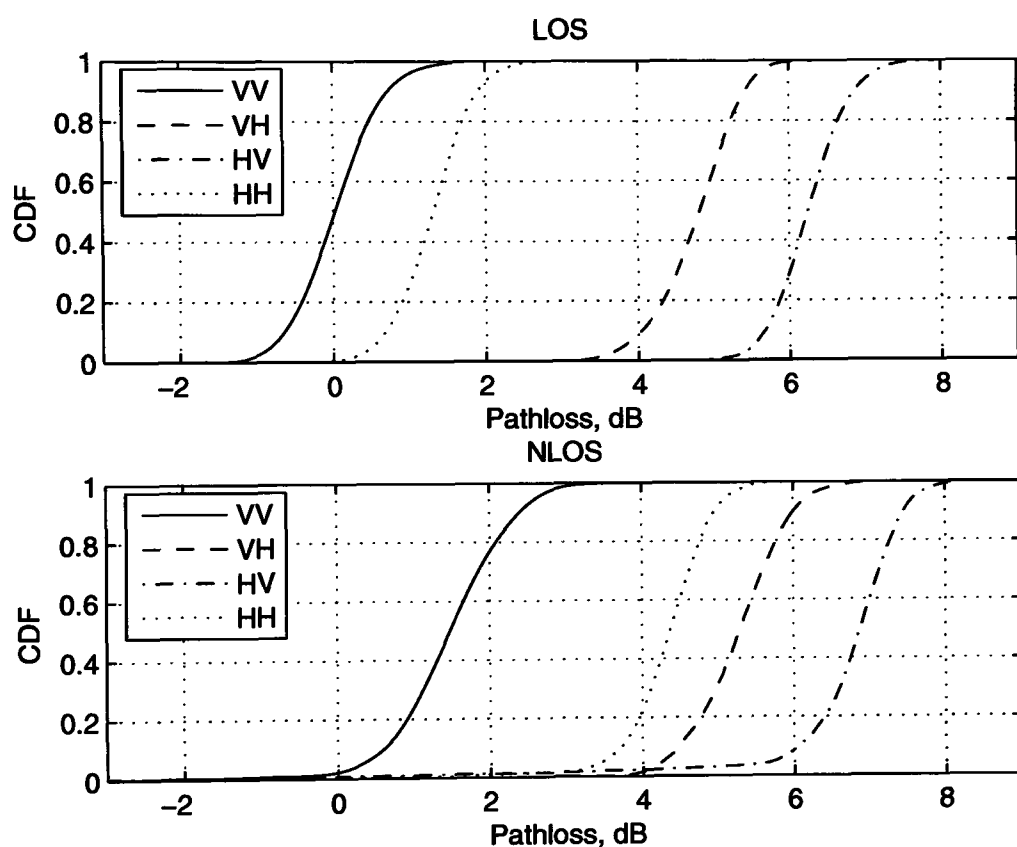


Figure 4.10 Empirical CDFs of the normalised pathloss

4.11. Summary

Signal depolarisation due to the UWB channel has been investigated using the XPD as the basic figure of merit, and elliptic polarisation parameters are also analysed. The mean XPD is found to lie between 2.5–4.5 dB, depending on the transmitted polarisation and the propagation scenario. This is consistent with previously reported figures for narrowband indoor channels. Larger spreads in XPD are, however, observed for all cases with a standard deviation of about 8 dB. The mean degree of linear

polarisation D_I of the received signal is found to be 0.81. The ellipticity angle and tilt angle also show large spreads. D_I , XPD and phase difference between orthogonal incident polarisations can be modelled with exponential, Gaussian and uniform random variables respectively. The highest XPD is achieved within 20 ns of the reception of the first multipath component at about 3 distinct time delays, after which a steady-state 6 dB mean XPD is maintained. Large XPD variations at certain frequencies are also found for both polarisations, establishing that polarised scattering in the UWB channel is frequency-selective. This information can be used in joint adaptive selection of the optimal polarisation and frequency sub-band in UWB communications. Polarised angular spectra indicate that polarimetric reciprocity is violated. In addition, different frequencies exhibit dissimilar polarisation leakage and noise suppression behaviour. The vertically polarised channel yields better performance than a horizontal polarisation in an indoor UWB channel due to its superior XPD, lower pathloss and smaller delay spread. This study extends the previously published results for narrowband mobile and indoor channels. It establishes the need to take frequency-selective polarisation characteristics of the UWB channel into account during system design, and also provides a framework for the comparison of single- and multi-polar UWB communications. The next chapter evaluates the performance of rake receivers to exploit the inherent multipath delay diversity in UWB systems.

5. RAKE RECEPTION

5.1. Introduction

In addition to reduced fading, the high temporal resolution of UWB systems enables the use of multipath combining with rake reception [112]. Rake receivers have been used extensively in the context of wideband systems to exploit multipath propagation in frequency-selective channels by capturing and combining the power contained in the multipath components. This is especially useful in power-limited UWB systems, where rake reception can boost the post-detection signal-to-noise ratio (SNR) at the receiver. This results in reduced bit error rates and outage probability, extended coverage range and increased system capacity. Thus the rake is an important component of a UWB communications receiver and is investigated in this chapter.

This chapter experimentally evaluates the performance of rake receivers with linear diversity combining in UWB channels. The power capture characteristics of various optimal and sub-optimal rake configurations are compared. The interdependence of diversity gain, bandwidth, multipath resolution, and system complexity is examined. The variation in the rake output as a function of bandwidth and number of resolved paths is assessed, and the optimal diversity order required to capture a given percentage of the incident power is determined. The resultant increase in the system's mutual information capacity is computed. The analysis provides ready-to-use metrics for the design of single- and multi-band UWB systems that are easily tailored to the operating conditions.

5.2. Rake Implementations

A rake receiver consists of multiple branches, or fingers, that track individual multipath components. The fingers outputs, often with unequal SNRs, are combined using linear or decision-oriented diversity. Two popular linear diversity combining techniques are selection combining (SC) and maximal-ratio combining (MRC) [80]. The optimal MRC rake that combines all resolved multipaths offers the highest SNR gain [69]. A full-band UWB system operating in a rich scattering

environment can typically resolve almost a hundred incident multipaths, as observed in Chapter 2. An MRC rake with such a high diversity order is, however, impracticable with the current device technology. The hybrid selection/maximal-ratio combining (H-S/MRC) rake, sometimes also referred to as a general selection combining (GSC) [113] or selective rake (SRake) [114], offers a flexible alternative as it allows a tradeoff between complexity and performance by combining only a subset of the resolved multipaths [96, 115]. MRC inherently emphasises the paths with the highest SNR, maximising the output SNR under the condition that the paths fade independently. The channel analysis in Chapter 2 has, however, shown that the multipaths degrees of freedom are typically far fewer than the number of resolved paths, indicating imperfect multipath orthogonality and finite fading correlation. This provides additional justification for the use of H-S/MRC from the perspective of reducing the outage probability to a sufficient degree with realistic system complexity.

Several other rake implementations have been presented recently with optimisations for specific channel conditions. Generalised rake combining with maximum likelihood (ML) [116], and optimum combining with MMSE [117] and eigencanceler [118] implementations were proposed for the reduction of narrowband or multiuser interference in wideband channels. Noncoherent rake techniques were presented for spread-spectrum systems in [119], while [120] proposed a noncoherent receiver for combining unresolvable multipaths. A hybrid adaptive MMSE rake and multiuser receiver for UWB communications was proposed in [121]. A scheme to tackle inter-symbol interference alongside rake reception for UWB channels was presented in [122]. Sub-optimal UWB rakes with square-law combining (SLC) and MRC were compared in [123]. The performance of a coherent rake was found to be similar to an adaptive differential detector in [124].

It was concluded in [125] that imperfect channel estimation does not incur an appreciable penalty on a coherent rake. Computationally efficient path selection algorithms for a minimum mean-square error (MMSE) rake were presented in [126]. The effect of the number of resolved paths and bandwidth on reduced-complexity rake symbol-error probability (SEP) was determined analytically in [96, 114, 127]. Semi-analytical evaluation of UWB rake performance was carried out using a statistical tapped-delay-line channel model in [128] and with a sparse channel assumption in [129]. It was suggested that

a certain optimal spreading bandwidth exists for a rake with a given complexity in a Rayleigh fading channel in [130] and a Nakagami channel in [131].

5.3. Analysis Procedure

The analysis in this chapter is based on LOS and NLOS channel data obtained from the measurement scheme described in Chapter 2. Channel sounding at $n_f=1601$ frequency points provides a large number of complex channel transfer functions. The transfer function at grid position (x, y) can be given by

$$H_{x,y}(f) = \sum_{k=0}^{n_f-1} A_{x,y,k} e^{j\psi_{x,y,k}} \delta(f - k\Delta f), \quad (5.1)$$

where $A_{x,y,k}$ and $\psi_{x,y,k}$ are the magnitude and phase responses at the k^{th} frequency component, and $\Delta f=4.7$ MHz is the frequency resolution. The complex impulse response (CIR), after removal of propagation delay and power-normalisation, is expressed in terms of the magnitude and phase response, $\alpha_{x,y,k}$ and $\varphi_{x,y,k}$, as

$$h_{x,y}(\tau) = \sum_{k=0}^{K-1} \alpha_{x,y,k} e^{j\varphi_{x,y,k}} \delta(\tau - k\Delta\tau), \quad (5.2)$$

where τ is the time delay and $\Delta\tau$ is the time resolution. After noise thresholding, the magnitude, phase and delay of individual multipaths in each CIR are used for rake performance prediction.

5.4. Rake Receiver Model

When a UWB signal waveform $s(t)$ is transmitted through a frequency-selective channel $h(t)$, the signal

$$r(\tau, t) = s(\tau, t) * h(\tau, t) + n(t) \quad (5.3)$$

reaches the receiver, where $n(t)$ is additive Gaussian noise. Assuming ideal antennas, this signal is fed to the rake receiver as $r_{in}(t)$. A linear-combining rake receiver, shown in Figure 5.1, integrates the

power from the resolved multipaths by appropriately weighting and combining the co-phased copies of the received signal in each of its fingers [69]. In the general form, its output can be represented by

$$r = \sum_{l=1}^L w_l \alpha_l, \quad 1 \leq L \leq N, \quad (5.4)$$

where w_l is the weighting factor for the l^{th} finger, α_l is the magnitude of the corresponding path, L is the number of rake fingers known as the rake's diversity order, and N is the number of incident multipaths. Here an impulse-based signal transmission has been assumed for simplicity, and the pulse waveform is not taken into account. The instantaneous SNRs of the multipath components in each CIR are represented by $\gamma_1, \dots, \gamma_n$. In selective configurations, the paths are ordered on the basis of SNR prior to combining, represented as $\gamma_{(1)}, \dots, \gamma_{(n)}$. A selection combining (SC) rake has one finger, or $L=1$. It processes all of the N resolved multipaths in the CIR and selects the one with the highest SNR, such that its instantaneous output SNR is

$$\gamma_{sc} = \gamma_{(1)}. \quad (5.5)$$

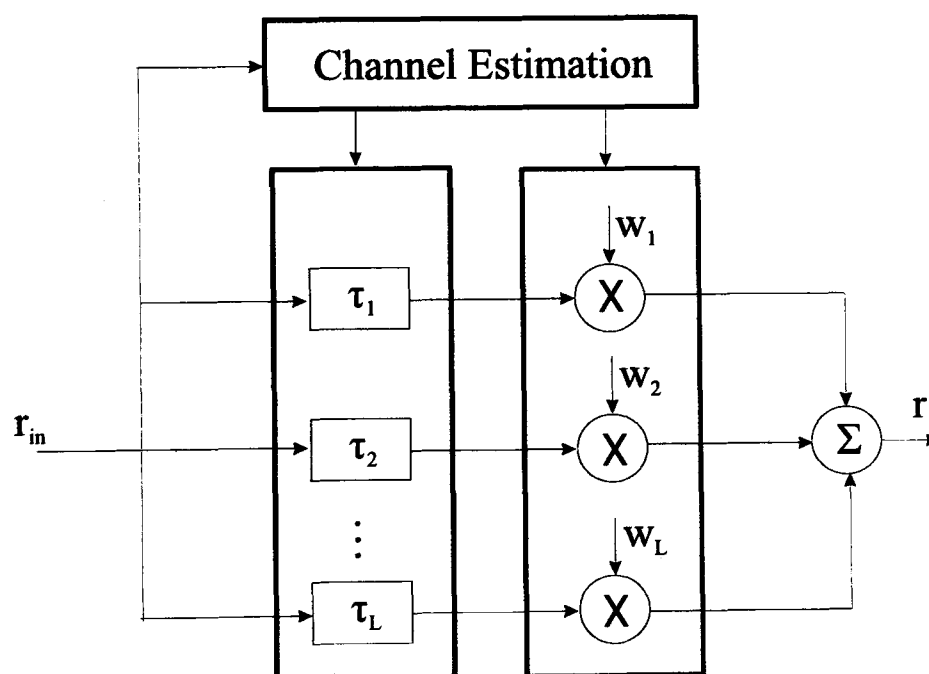


Figure 5.1 A coherent MRC rake receiver with L fingers

An MRC rake, with $L=N$, combines all the multipaths after weighting through the fingers with weighting factors

$$w_l \propto \gamma_l, \quad \text{such that} \quad \sum_{l=1}^L w_l = 1, \quad (5.6)$$

to yield a signal with instantaneous SNR given by

$$\gamma_{MRC} = \sum_{l=1}^L \gamma_l, \quad L = N. \quad (5.7)$$

Combining a large number of multipaths can result in an unnecessarily complex receiver. The complexity arises from the estimation of the channel parameters, namely the multipath delays and gains, which are not known *a priori* in a realistic channel [132]. A reduced-complexity rake of order L uses H-S/MRC to combine the L strongest multipaths and combines them using MRC. The instantaneous SNR of the rake output is then

$$\gamma_{H-S/MRC} = \sum_{l=1}^L \gamma_{(l)}, \quad 1 \leq L \leq N. \quad (5.8)$$

where $\gamma_{(l)}$ are the ordered set $\{\gamma_{(1)}, \dots, \gamma_{(n)}\}$. Another sub-optimal combining scheme with complexity lower than H-S/MRC is partial combining or PMRC. It also uses MRC but does not order the multipaths according to their SNRs, reducing the complexity associated with tracking the strongest L paths. Its instantaneous SNR is

$$\gamma_{PMRC} = \sum_{l=1}^L \gamma_l, \quad 1 \leq L \leq N. \quad (5.9)$$

The relative magnitudes of the multipath components determine the amount of additional power that can be captured from them, and their fading correlation determines the diversity improvement.

If the noise is uniformly distributed across all the paths, this comparative power analysis can be treated in terms of SNR. The receiver SNR influences the probability of outage, bit error rate, and channel capacity. In most practical situations, the SNR is replaced by the signal-plus-noise to noise ratio without any significant bearing on the results [16]. The same approach is adopted here.

5.5. Rake Diversity Gain

Following the nomenclature in [80], the diversity gain is defined in terms of the mean instantaneous increase in power. Thus with zero-mean Gaussian noise, the increase in SNR as a result of rake combining can be referred to as its diversity gain. The diversity gain can be determined directly from a comparison of the received signal power with and without rake combining. In this analysis, the

rake output is compared with the power of a single multipath component. This component is taken to be either the direct component or the strongest component, depending on the context. This methodology helps quantify the gain from the L -finger H-S/MRC rake and benchmark its performance.

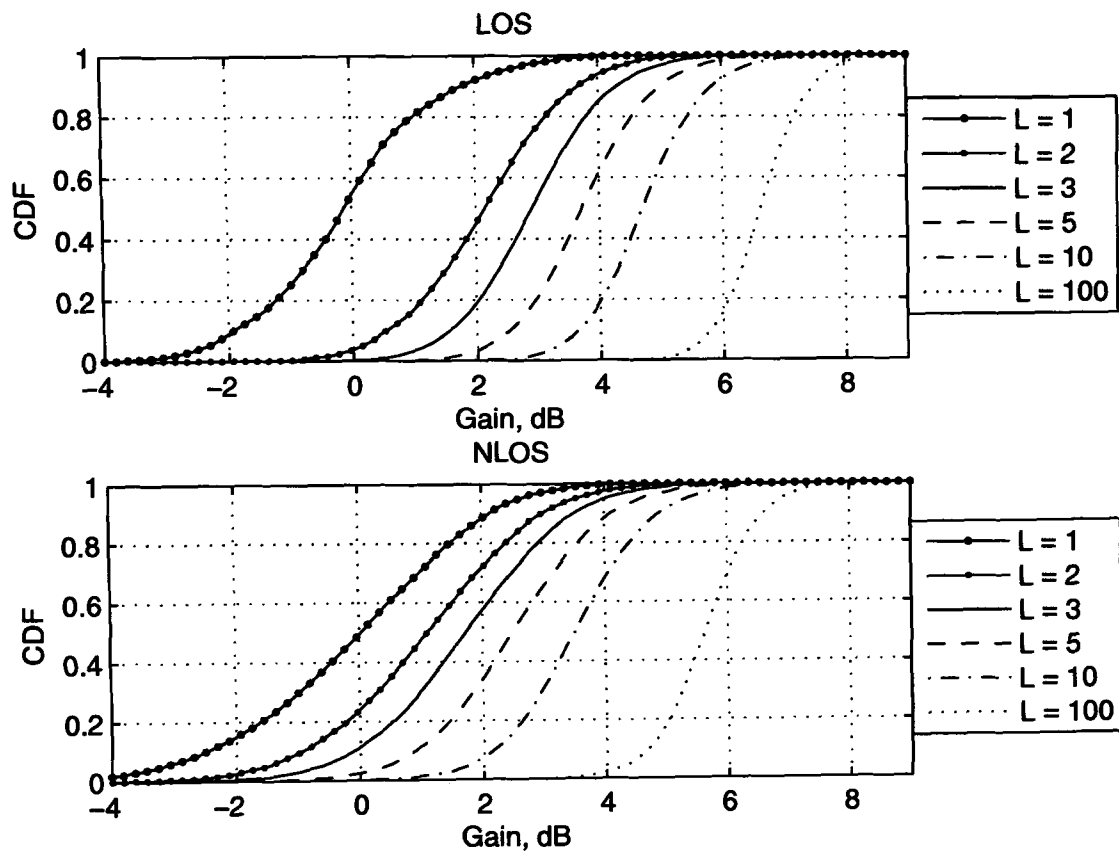


Figure 5.2 CDFs of H-S/MRC rake output SNR normalised to the SC rake ($L=1$) output

In order to calculate the diversity gain from the measured channel data, H-S/MRC rake reception is applied individually to each CIR, using the full bandwidth, i.e. $\Delta f_{max}=7.5$ GHz. Let γ_{out} be the output SNR of an L -finger H-S/MRC rake and $P(\gamma_{out})$ its cumulative distribution function (CDF). Figure 5.2 shows $P(\gamma_{out})$ for various diversity orders L , with each γ_{out} normalised to the mean output SNR of an SC rake, i.e. H-S/MRC with $L=1$. It is seen that an increase in the number of rake fingers in both LOS and NLOS channels leads to an increase in the rake output, but the advantage diminishes with the diversity order. The difference between $L=1$ and $L=2$ is about 2 dB in LOS and 1 dB in NLOS. The advantage of a low-order H-S/MRC rake is less dramatic in NLOS, and a more fingers are required to achieve the same diversity gain. This is due to the concentration of the incident power within a few dominant multipaths in LOS but relatively more uniform distribution in NLOS. Also, the variance of the instantaneous SNR decreases with the diversity order, as listed in Table 5.1, and the CDFs become

steeper. The reason is that higher order rakes combine a large percentage of the power, and the residual power is no longer so significant as to manifest its variation as appreciable small-scale fading.

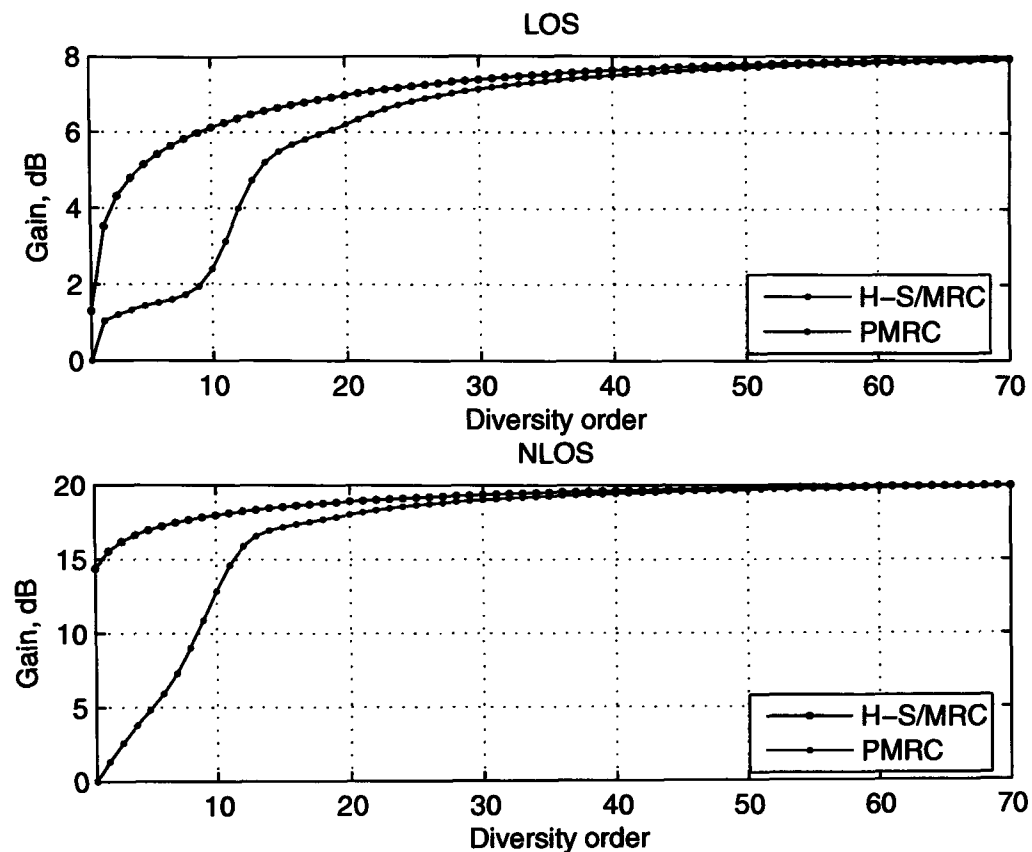


Figure 5.3 Mean H-S/MRC and PMRC rake diversity gain normalised to the direct component

Figure 5.3 shows the mean diversity gains of the H-S/MRC and PMRC rake receivers, normalised to the SNR of the direct path, as a function of the diversity order. This enables the quantification of the benefit from path selection on the basis of SNR. It is observed that an SC rake provides a 1.5 dB SNR gain in the LOS channel and 14 dB in the NLOS channel relative to the LOS or quasi-LOS path. Another finger increases the H-S/MRC rake SNR by about 2 dB in LOS, but further increase in the diversity order does not yield a proportionate gain. The decline in fractional gain is well known for diversity combining systems, due to which sub-optimal implementations have found favour in practical implementations [133]. This result confirms the previously reported conclusions drawn analytically for wideband Gaussian [96] and Rayleigh [134] fading channels. Compared to the H-S/MRC, the PMRC rake SNR is significantly smaller for low diversity order, but the performance curves converge at large L . The PMRC diversity gain reaches within 1 dB of the H-S/MRC gain with about 15 fingers. Thus PMRC can be used effectively only for a medium to large diversity order. Clustered scattering environments [135] can thus severely degrade the performance of a low-order PMRC rake, as some of

the late-arriving strong multipaths are not captured with such a system. Its performance, therefore, is strongly dependent on the scattering environment and the extent of multipath propagation. Also, from the figure, the asymptotic median diversity gain of an MRC rake is 8 dB in LOS and 20 dB in NLOS. The difference in the median diversity gain of SC and MRC is about 7 dB in LOS and 6 dB in NLOS. For small values of L , increasing the H-S/MRC diversity order in LOS causes a sharp rise in SNR while the NLOS gain increases more uniformly. These results are summarised in Table 5.1.

Table 5.1 The diversity gain of various selective rake configurations with respect to the direct path

Diversity Combining	Diversity	LOS		NLOS	
Scheme	Order	Median, dB	Std., dB	Median, dB	Std., dB
SC	1	1.3	1.4	14.4	1.8
H-S/MRC	2	3.5	1.2	15.5	1.5
H-S/MRC	5	5.1	1.0	17.0	1.3
H-S/MRC	50	7.8	0.7	19.8	0.8
MRC	∞	8.0	0.6	20.1	0.8

5.6. Dependence on Bandwidth

So far the rake output has been derived assuming a UWB signal covering the entire 3.1–10.6 GHz band. The effect of variation of bandwidth on rake receiver performance is now investigated. To achieve this, spectral sections from the measured channel data are used to generate lower-bandwidth channels. Such a reduced bandwidth channel measured at location x is given by

$$G_{x,y}(f) = \sum_{k=1}^{n_b} A_{x,y,k} e^{j\psi_{x,y,k}} \delta(f - k\Delta f), \quad 1 \leq n_b \leq n_f, \tag{5.10}$$

where n_b is the number of frequency points considered and Δf is the frequency resolution of the measurement. This gives a CIR with lower temporal resolution than the full-band CIR, represented by

$$h_{x,y}(\tau) = \sum_{k=1}^{n_b} \alpha_{x,y,k} e^{j\varphi_{x,y,k}} \delta(\tau - k\Delta\tau). \quad (5.11)$$

The consequence of lower temporal resolution is that the time-bins are dilated, since the time resolution is the reciprocal of bandwidth, i.e.

$$\Delta\tau = \frac{1}{n_b\Delta f}. \quad (5.12)$$

Some of the multipaths resolved by the full-band signal appear in the same time-bin when the bandwidth is reduced to $B=n_b\Delta f$, so that now the number of multipaths $N_b \leq N$. As fading increases at lower bandwidths, a change in B can strongly influence rake receiver performance. A detailed analysis of the interplay between B , N_b , L , and diversity gain follows.

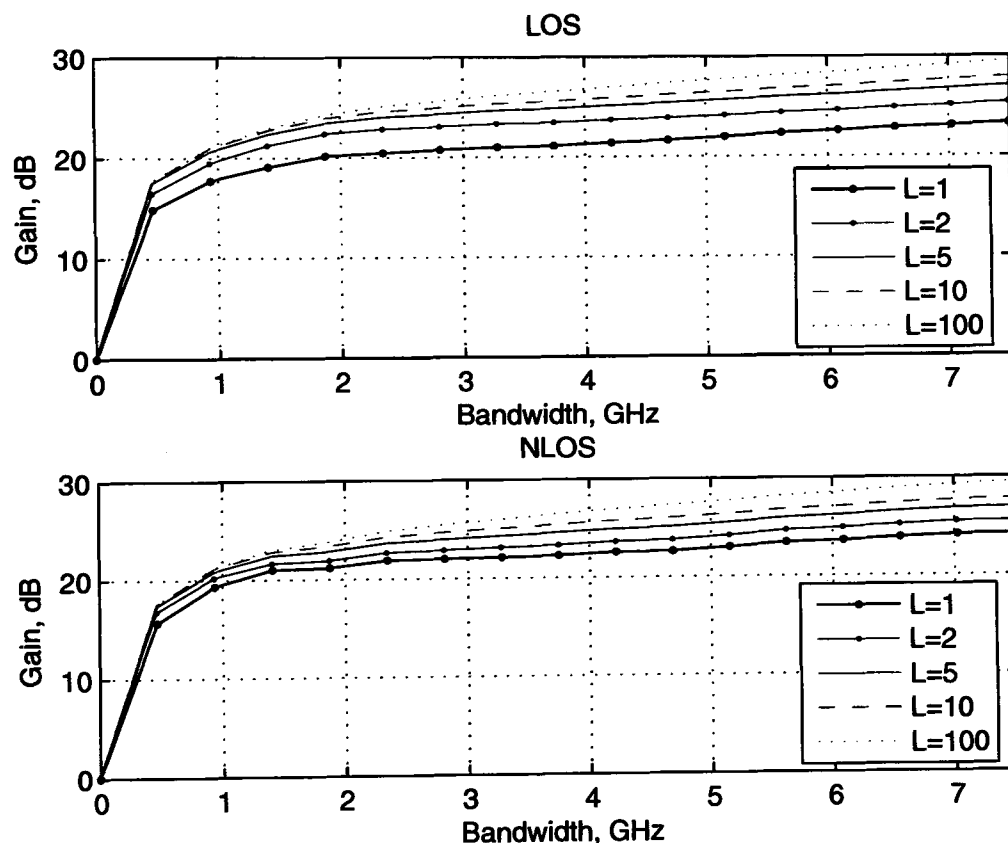


Figure 5.4 Mean H-S/MRC rake diversity gain as a function of bandwidth B for a fixed diversity order L , normalised to the gain at $B=0$

Figure 5.4 shows the variation of the rake diversity gain with B for various L . At vanishingly small B , the output is independent of L as there is little multipath resolution in that case. Such a narrowband signal is described by only one channel tap, and increasing the number of rake fingers beyond $L=1$ provides no diversity gain. Now, considering the characteristic of SNR for the $L=1$ rake, it can be observed that the SNR gain rises sharply with B initially but its fractional increase falls at larger B so

that it settles into an almost linear variation beyond $B=1.5$ GHz. The explanation is that as B increases, the multipath resolution also increases, the temporal widths of an observed multipath and a rake finger decrease, and consequently more signal is captured due to the correlation gain, increasing the SNR. As L is increased, there is an increase in the output SNR at a given B , but the overall characteristic of the relationship with bandwidth remains the same. It is also evident that the constant- L curves diverge at higher B . The behaviour of the LOS and NLOS channels is similar.

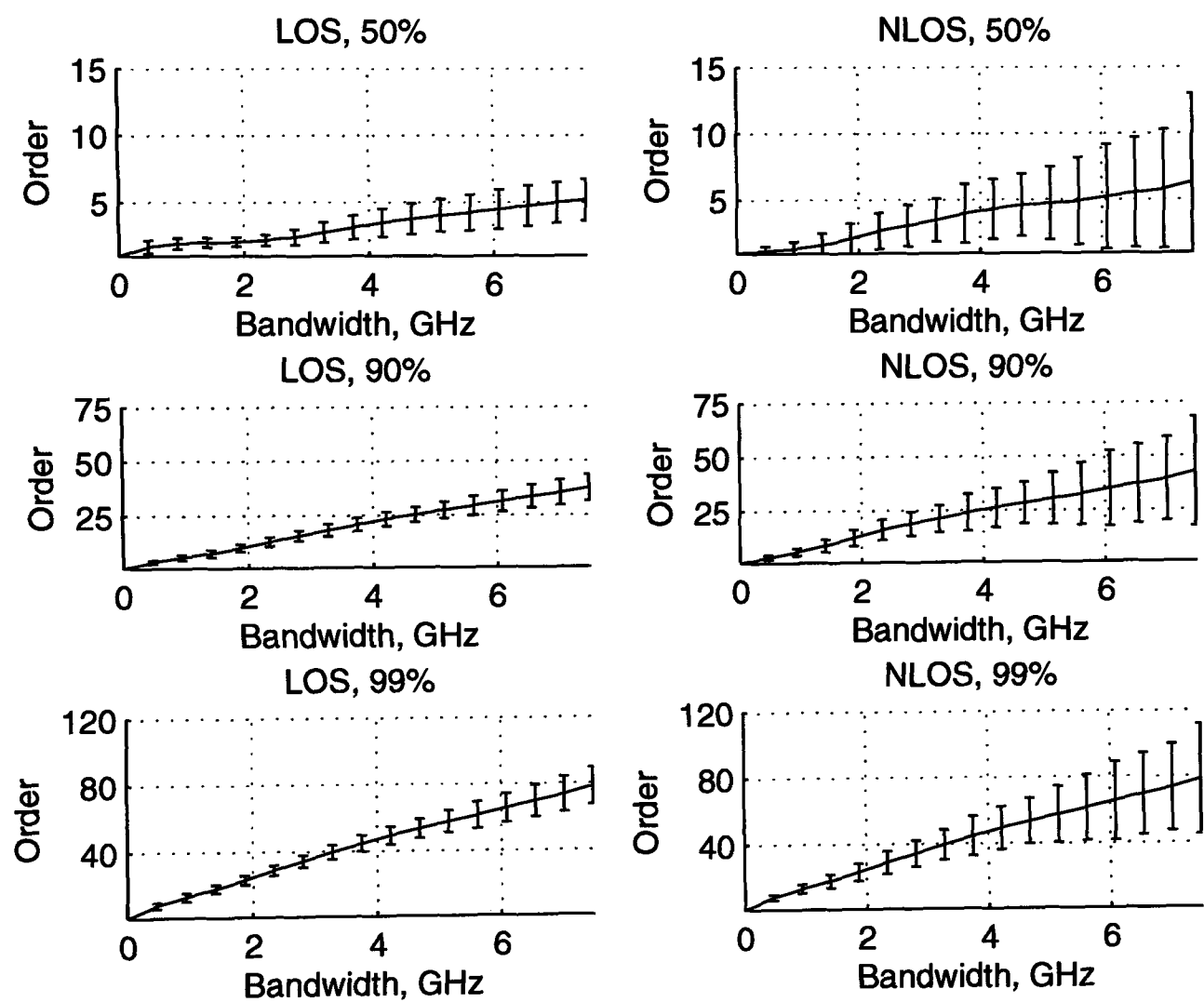


Figure 5.5 Diversity order required to capture a specific percentage of the incident power. The solid lines represent the mean $L_{X\%}$ and the errorbars indicate the standard deviation

Keeping this relationship in mind, the optimal diversity order for a given bandwidth can be found. The number of rake fingers, $L_{X\%}$, required to capture $X\%$ of the incident multipath power is derived and its relationship with B is investigated in Figure 5.5 for $X=\{50,90,99\}$. It can be observed that $L_{X\%}$ varies approximately linearly with B in each case. Thus with a 500 MHz subchannel, such as is used in many multiband UWB approaches, $L_{90\%}=4$ and $L_{99\%}=9$ fingers on average. For a single-band Impulse Radio UWB system occupying the full band, $L_{90\%}=40$ and $L_{99\%}=80$. Also, $L_{X\%}$ increases with B with

gradients of about 0.6, 5 and 10 fingers/GHz for $X=\{50,90,99\}$ respectively. The increasing slope indicates the requirement for a larger proportion of the resolved multipaths to be combined for greater power capture at higher B . The variance of $L_{X\%}$ increases with B and is larger in NLOS than in LOS.

5.7. Effect of Number of Multipaths

The relationship between the number of resolved multipaths N , the diversity order L , and the rake output power is evaluated here. For the analysis in this section, the bandwidth B is kept constant at 7.5 GHz. The variation in the normalised output power of a fixed diversity order rake with the number of paths is shown in Figure 5.6, where the scatter points show the observed values and the solid lines show the linear best-fit estimates. Four representative values of L are selected for the sake of comparison. In each case, as the number of multipaths in the CIR increases, the rake output power falls. With an SC rake, the resolution of every additional multipath brings down the power by 0.1 dB on average. High-order rakes are less prone to this variation, so that an $L=100$ rake suffers a power loss by only 0.02 dB with each additional path. The reason for this linear decrease can be traced to small-scale fading. At certain locations within the propagation environment, a few of the multipaths may fall in the same time-bin, some of them constructively, thus producing m strong composite multipath components in the CIR. A low order rake with L on the order of m performs well in such a scenario. At another location, the same multipaths may occupy separate time-bins, increasing the number of resolved multipaths but decreasing the power of the strongest m paths. The same rake can now integrate only a smaller amount of power, and its performance is degraded. This also explains the robustness at higher L observed in the figure. This leads to the characterisation of the required L as a function of N . From Figure 5.7, N also has a linear relationship with $L_{X\%}$. The regression, which is positive, provides an approximate figure for the percentage of resolved multipaths that must be combined. Thus on average about 10%, 45% or 90% of the resolved multipaths must be combined if it is desired to capture 50%, 90% or 99% of the power respectively within the 25 dB threshold of the UWB receiver. Some of the key results from the above analysis are summarised in Table 5.2.

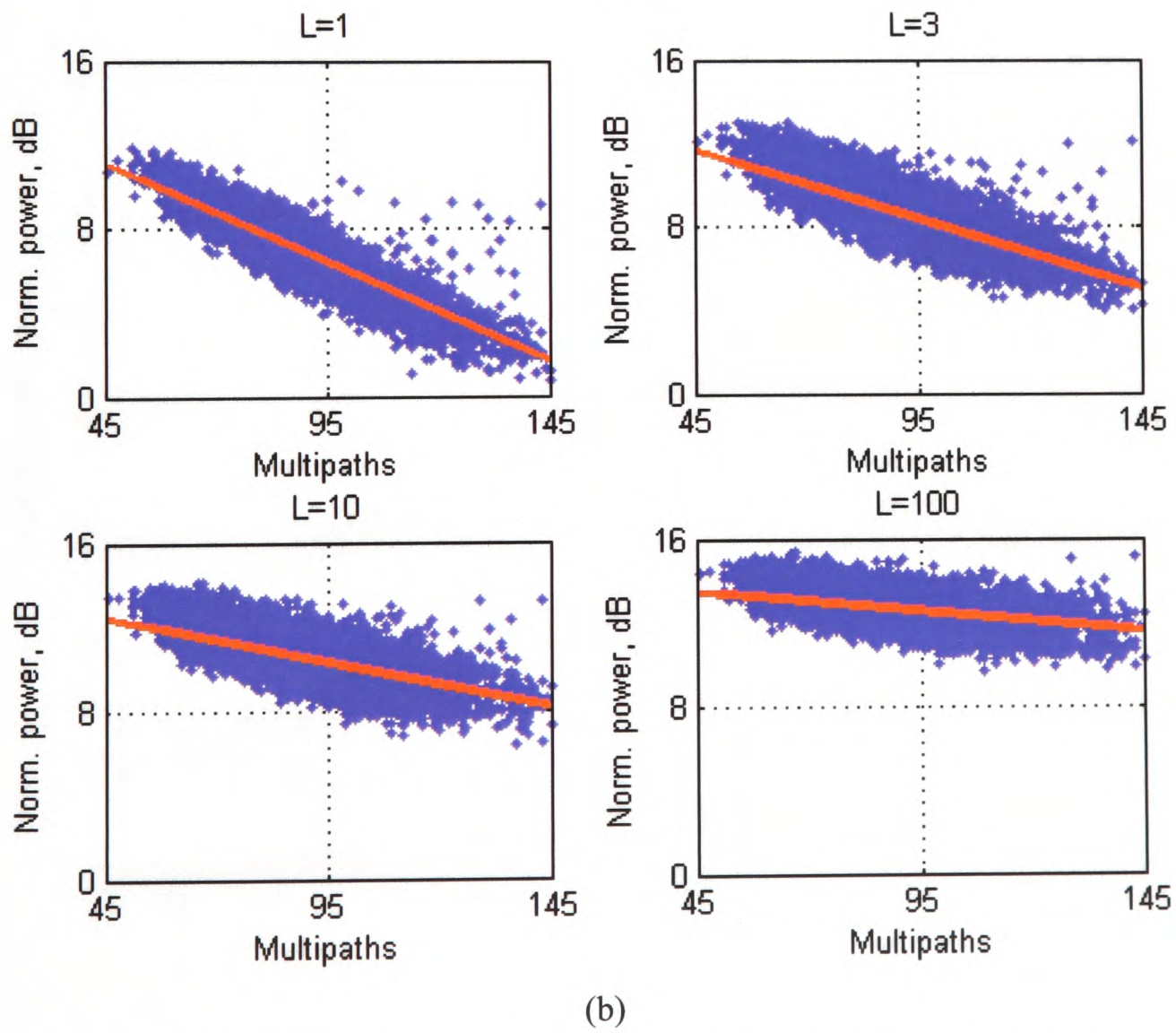
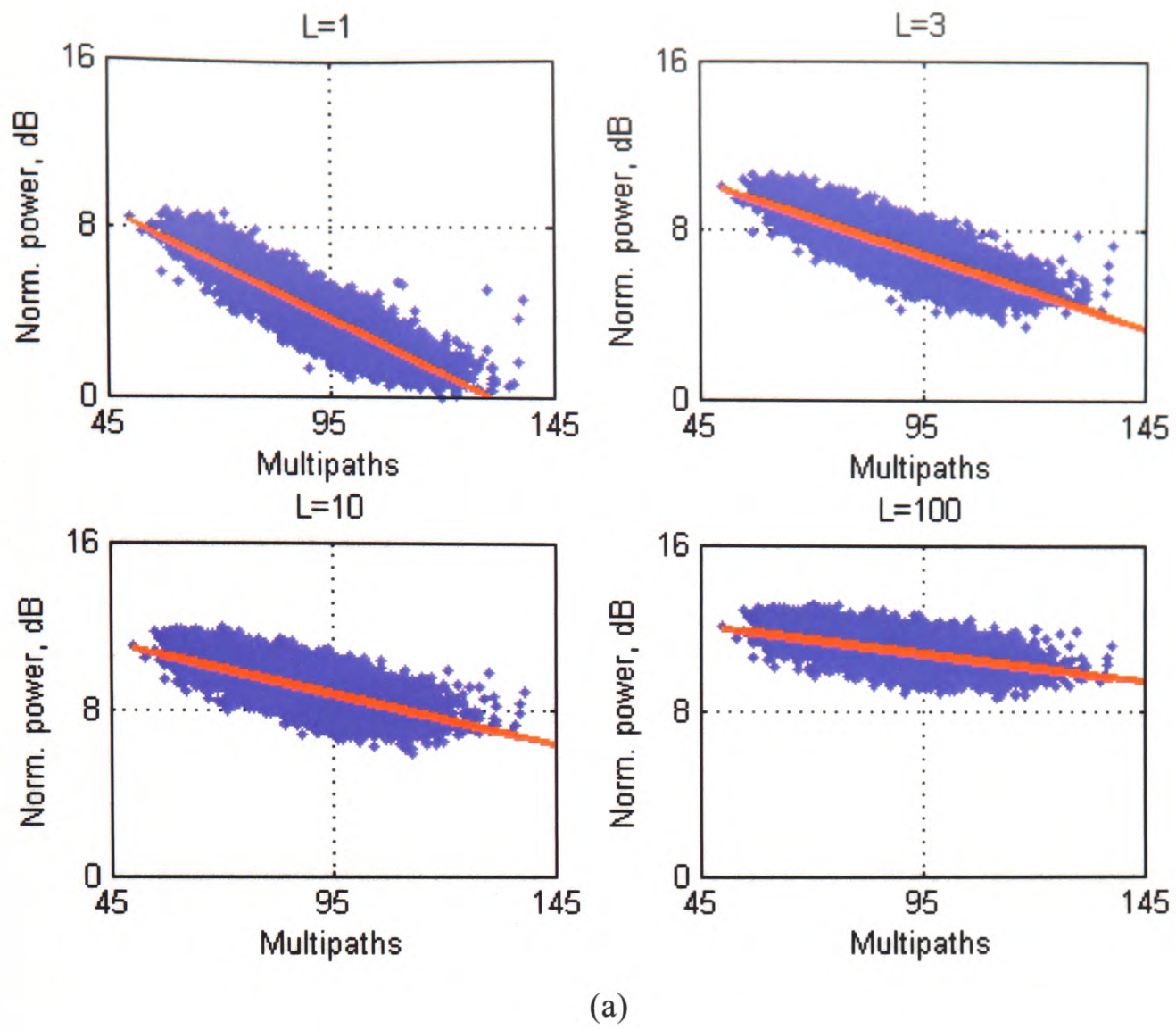
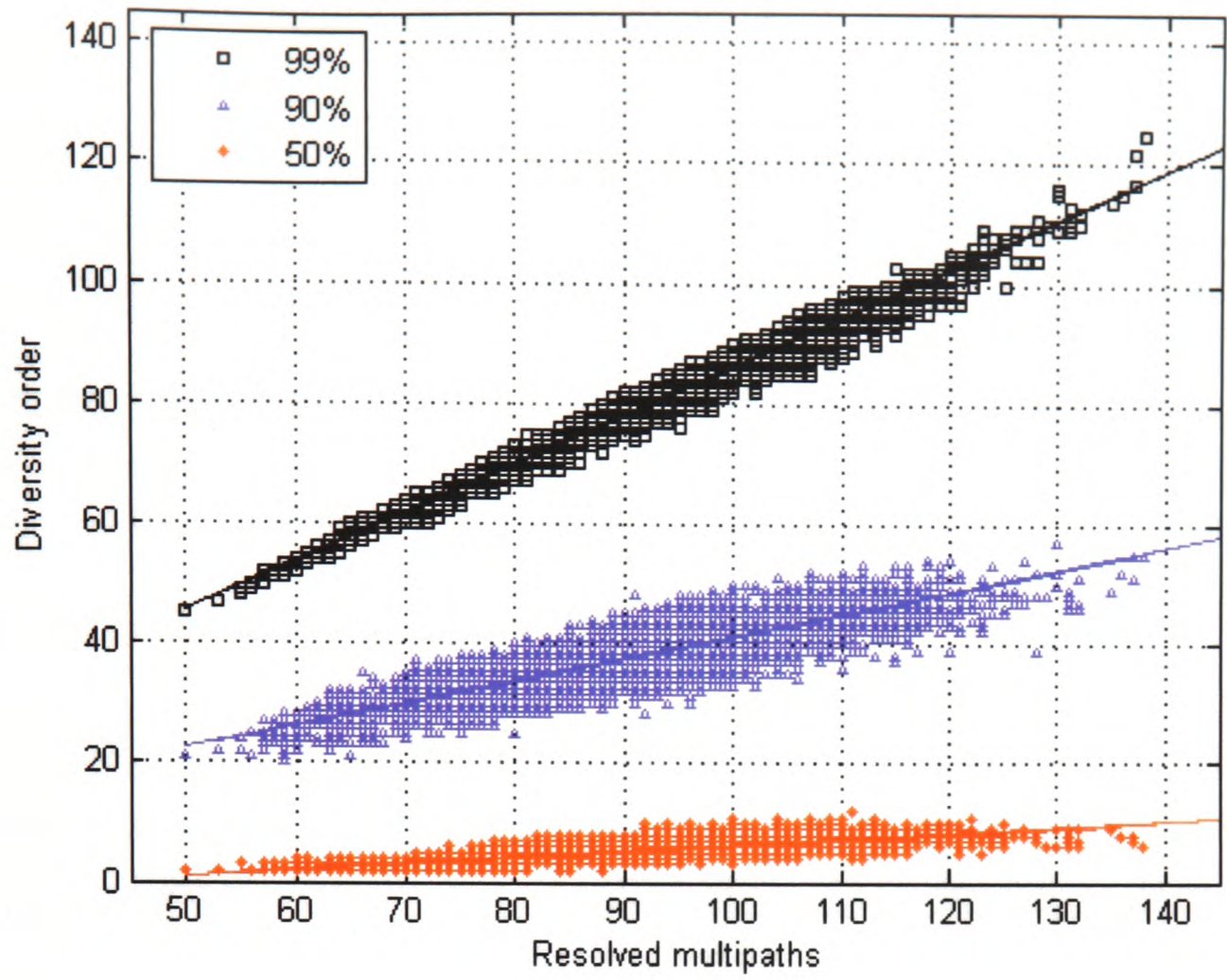
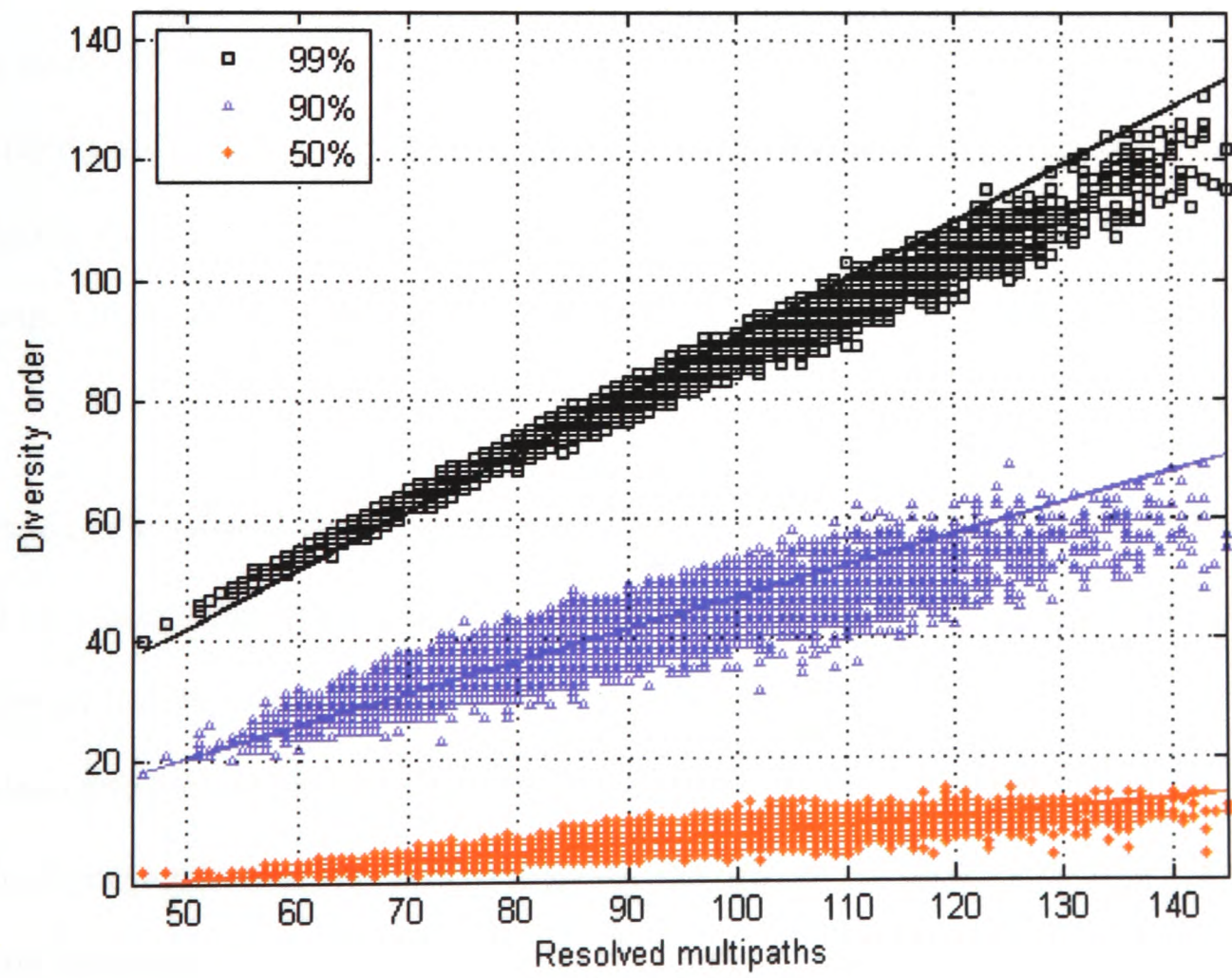


Figure 5.6 H-S/MRC rake output power, normalised across the measurement ensemble and diversity orders L , as a function of number of resolved multipaths at full bandwidth in (a) LOS and (b) NLOS



(a)



(b)

Figure 5.7 Rake diversity order required to capture a given percentage of the incident power at the full bandwidth $B=7.5$ GHz in (a) LOS and (b) NLOS channels

Table 5.2 Rake performance–complexity tradeoff determinants

Parameter	Regression	Operating Conditions
Diversity order / no. of paths	0.45 fingers/path	For 90% power
Diversity order / bandwidth	5 fingers/GHz	For 90% power
Diversity gain / no. of paths	-0.07 dB/path	For $L=3$
Diversity gain / bandwidth	0.8 dB/GHz	For $B>2$ GHz

5.8. Error Rate Improvement and Capacity Enhancement

SNR enhancement at the detector as a result of rake reception significantly improves system performance. As the bit-error rate (BER) is a monotonically decreasing function of SNR for any modulation format, rake reception improves the BER by overcoming fading and increasing the mean receiver SNR. Also, as the joint probability of blockage of all multipaths combined by the rake receiver is many orders lower than that with a single-multipath detector, the system outage probability reduces drastically. In addition, the various channel parameters affecting the diversity gain of the rake as discussed earlier in this section also impact its information theoretic performance. The diversity gain has been shown to decrease with an increase in N , and as the SNR ρ is linked to the system capacity C through the well-known Shannon expression [136]

$$C = B \log_2 (1 + \rho), \tag{5.13}$$

it can be directly inferred that an increase in N lowers the mutual information capacity of the UWB channel even with perfect rake combining. This confirms and extends the observations in [137], where it is proposed that the mutual information, I , is inversely proportional to N for a spread-spectrum signal in a wideband fading channel assuming uncorrelated, uniform-power multipaths.

Introducing a rake also improves the spectral efficiency, C/B . It follows directly from (5.13) that a 3 dB gain increases the spectral efficiency by 1 b/s/Hz, at high operating SNR. Thus for a full-band pulsed UWB system, a sufficiently high order H-S/MRC rake receiver, with an additional 6 dB SNR over an SC rake, can increase the raw channel capacity of a full-band UWB system by up to 15 Gbps.

5.9. Summary

The SNR improvement from the use of optimal and sub-optimal rake receivers has been analysed for measured UWB channels. The incremental rake improvement is found to fall rapidly with increase in the diversity order, confirming earlier theoretical predictions based on simple channel models. The diversity gain of unordered, or partial, combining asymptotically approaches that of selective rake combining, but the performance difference is large with realistic complexity especially in NLOS channels. The rake diversity gain increases monotonically with increase in bandwidth but the incremental gain diminishes. To capture 90% of the incident multipath-dispersed power, approximately 5 fingers/GHz are required. The rake output deteriorates as the number of paths rises, but the performance degradation factor falls with diversity order. The required diversity order increases linearly with the number of paths. Combining 45% of the paths yields 90% of the incident power. Information theoretic analysis shows that an optimal rake receiver can enhance the mutual information capacity of a short-range UWB system by up to 15 Gbps.

These results relate rake output and required number of fingers with signal bandwidth and number of incident multipaths, and provide readily usable information for system design. A quantitative answer to the ultimate rake receiver design question – how many fingers – is thus provided. The next chapter analyses UWB signal distortion due to antennas.

6. ANTENNA DISTORTION

6.1. Introduction

Traditional antenna characterisation is centred around fundamental antenna parameters related to the input characteristics and radiation patterns [71]. Antenna design techniques focus on impedance matching, low VSWR and group delay, and high radiation efficiency [138]. While this usually suffices for narrowband antennas, several issues arise in the case of UWB antennas that merit deeper analysis. In the context of antenna modelling and characterisation, the wide bandwidth of UWB signals necessitates some enhancements to traditional techniques. Antenna efficiency measurements, for instance, require modified Wheeler Cap methods [139].

In the wideband regime, the end-to-end signal distortion introduced by the transmitter and receiver antennas is effectively a filtering operation such that the antenna response is convolved with the signal waveform. The physical size and shape of the antenna also affects its time-frequency response. For example, a log-periodic antenna disperses a UWB signal as the low frequency components are launched earlier than the high frequency components when the connection to the transmission line is at the broad end [140]. In such case, the phase centre of the antenna translates with frequency resulting in temporal distortion [141]. This form of distortion can be described in terms of the group delay through time-domain transient response analysis [142, 143].

Apart from these, the frequency dependence of the radiated power as defined by the Friis transmission equation [144] for freespace propagation also causes waveform distortion due to the non-uniform power flux density across the UWB band. After transmission and reception through the antennas, the UWB signal power spectrum slope is determined by the antenna combination [72]. This can be interpreted as a differentiation or integration operation in the time domain [145], introducing phase shifts and waveform distortion in the signal [146], and is thus highly detrimental to the performance of both single-band and multi-band UWB communications links [147]. The characterisation of frequency dependence of various antenna parameters is thus of paramount importance for UWB system design.

Besides spectral and temporal dispersion, angle is another domain where the behaviour of an antenna can have a significant impact. This is particularly important for indoor communications systems such as wireless personal area networks, where dense multipath propagation gives rise to a wide range of angles of arrival [98, 148]. In the absence of the theoretical isotropic radiator, a broadcast transmitter using an omni-directional antenna is expected to radiate the signal in all directions within the azimuthal plane with symmetric gain and group delay such that the wavefronts have near-uniform intensity and planar shapes. In practice, however, this condition is sometimes violated by UWB antennas, causing look-angle dispersion [141].

The performance of antennas operating in the FCC UWB band is investigated in this chapter. The sensitivity of the radiation pattern of a simple dipole antenna to frequency is discussed. The dependence of radiated power jointly on frequency and direction is probed experimentally using discone [43, 149] and Vivaldi [150-153] antennas that represent the general cases of omni-directional and directional UWB antennas respectively. While the details of the dispersive behaviour will vary from antenna to antenna, the results illustrate the phenomenon and highlight its consequences in the broad sense. In addition to quantifying this behaviour in terms of classical antenna parameters, effective angular bandwidth is introduced as a measure of antenna dispersion. UWB signal distortion is evaluated and the effect on the communications link is highlighted. A technique for removing the link aberration using direction-of-arrival information is proposed. Some ideas for the mitigation and exploitation of antenna effects using a cognitive radio are presented.

6.2. Dipole Radiation Pattern Variation with Frequency

The dipole is inherently a narrowband antenna but is representative of many other, more complex antenna shapes [71]. Also, its field equations are analytically tractable and well understood. On the basis of its compact, simple and economical design, the dipole has been suggested for use in short-range UWB transmission [154]. For a vertically oriented dipole, the elevation plane far field radiation pattern can be expressed as a function of the angle of elevation θ and frequency f , i.e.[155]

$$F(\theta, f) = \frac{\cos\left(\frac{kl}{2}\right) - \cos\left(\frac{kl \sin \theta}{2}\right)}{\cos \theta} \quad (6.1)$$

where l is its length, $k=2\pi/\lambda$ is the wavenumber and λ is the wavelength. For a 0.1 m long dipole, the power radiation pattern varies significantly within the UWB band (3.1–10.6 GHz). This is shown in Figure 6.1, where the normalised elevation plane radiation patterns at four equidistant frequencies within that band are plotted. An increase in operating frequency leads to higher directivity and an increase in the number of lobes. The directions of the lobes also change with frequency, increasing in elevation. This observation points to the important problem of frequency-dependent antenna radiation patterns. The next sections probe the existence of similar phenomena in UWB antennas.

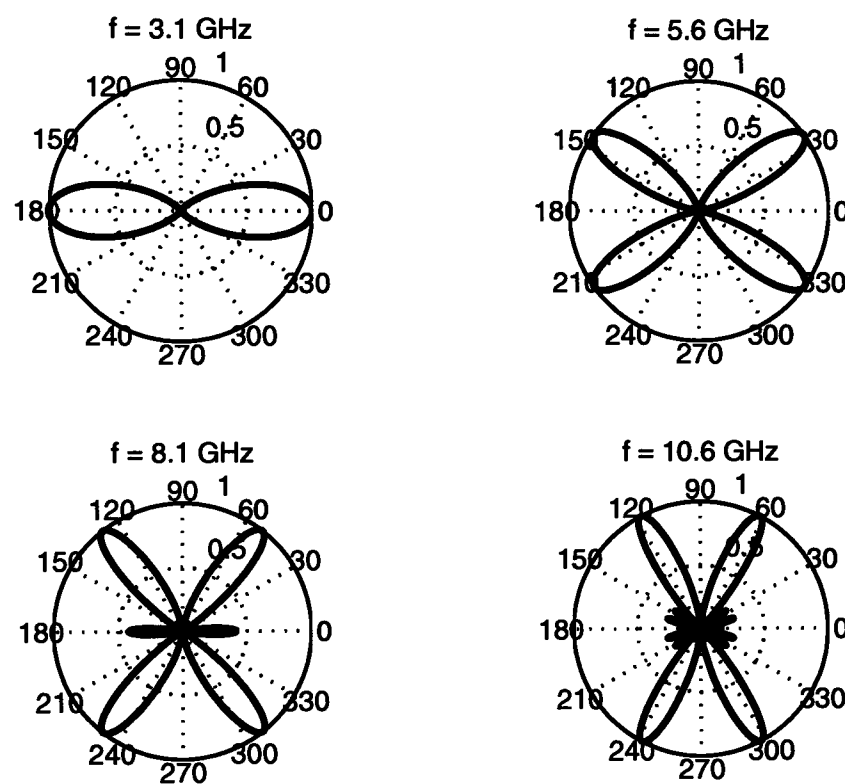


Figure 6.1 Normalised elevation plane field radiation patterns of a vertically polarised dipole antenna at various frequencies, with the circular axis signifying the elevation plane angle θ

6.3. Analysis Procedure

Many types of UWB antennas have been proposed in literature, and designs such as the biconical, bowtie, d-dot, log-spiral, Vivaldi, TEM horn and ridged circular horn are among the most popular [156-162]. In this work, the discone and Vivaldi are chosen for the experimental analysis. The discone

is a popular choice of antenna for omni-directional, linearly polarised systems. With a conveniently manufactured design and bandwidth exceeding a decade [149, 163], it is commonly used in UWB channel measurements [32, 164]. The radiation pattern of the discone is similar to electrically small dipole, biconical, and bowtie antennas. The balanced antipodal Vivaldi antenna [150-152, 159] is an elliptically tapered slotline structure with an instantaneous bandwidth greater than a decade, low cross-polarisation and high gain [153].

A pair of identical Vivaldi antennas was manufactured using a photolithographic process, while the discones were made by machining. Each discone antenna, made from aluminium, consisted of a hollow cone and a large circular ground plane, with its geometry described by the cone length $l=0.065$ m, cone half-angle $\psi=45^\circ$, and ground plane diameter $d=0.2$ m, as shown in Figure 6.2. The balanced antipodal Vivaldi antenna, also shown in the figure, had dimensions of $d=0.04$ m and $l=0.1$ m. It had a triplanar form with two layers of dielectric sandwiched between three layers of duroid substrate.

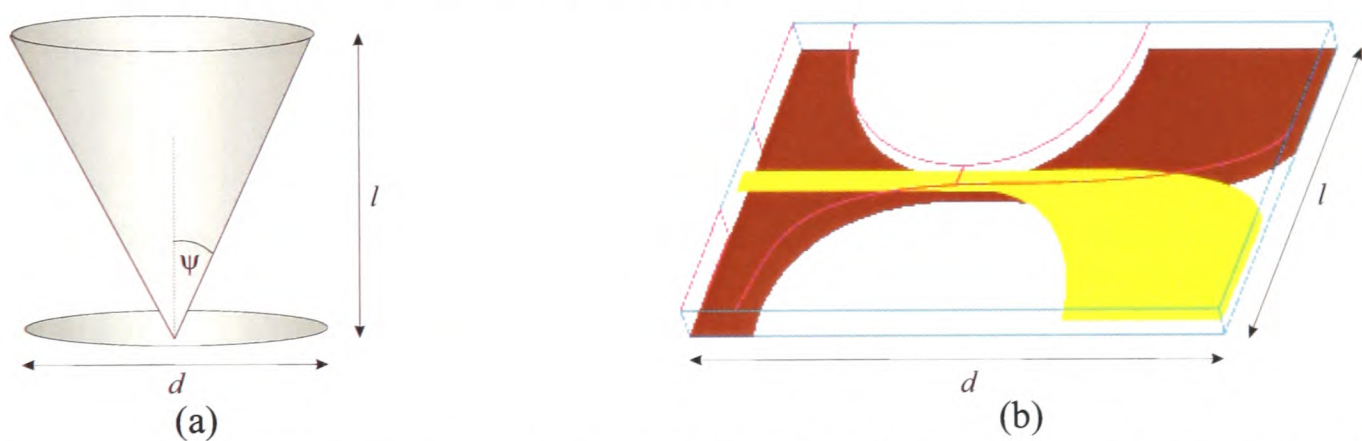


Figure 6.2 Construction of UWB antennas: (a) a discone antenna with a large circular ground plane; and (b) a balanced antipodal Vivaldi antenna

Both pairs of antennas were extensively characterised in an anechoic chamber. A vector network analyser (VNA) was used to measure the $S_{11}(f)$ parameter. The VNA was calibrated prior to the measurement to remove the attenuation, delay and phase distortion of the cables and connectors. $S_{11}(f)$ is used to derive the reflection coefficient $\Gamma(f)$, which yields the antenna return loss in logarithmic scale often used to characterise the radiation performance of an antenna over a given frequency band. Figure 6.3 shows the reflection loss of the four antennas. For the discone antennas, the reflection loss lies well below -10 dB for the entire UWB band, i.e. $f_l=3.1$ to $f_h=10.6$ GHz. This is also true for the Vivaldi except for a small range of frequencies near 9 GHz, due to which the radiation efficiency of these

Vivaldi antennas will be higher in the lower portion of the UWB band. Furthermore, the corresponding antennas in each pair have similar $\Gamma(f)$, confirming that they are identical.

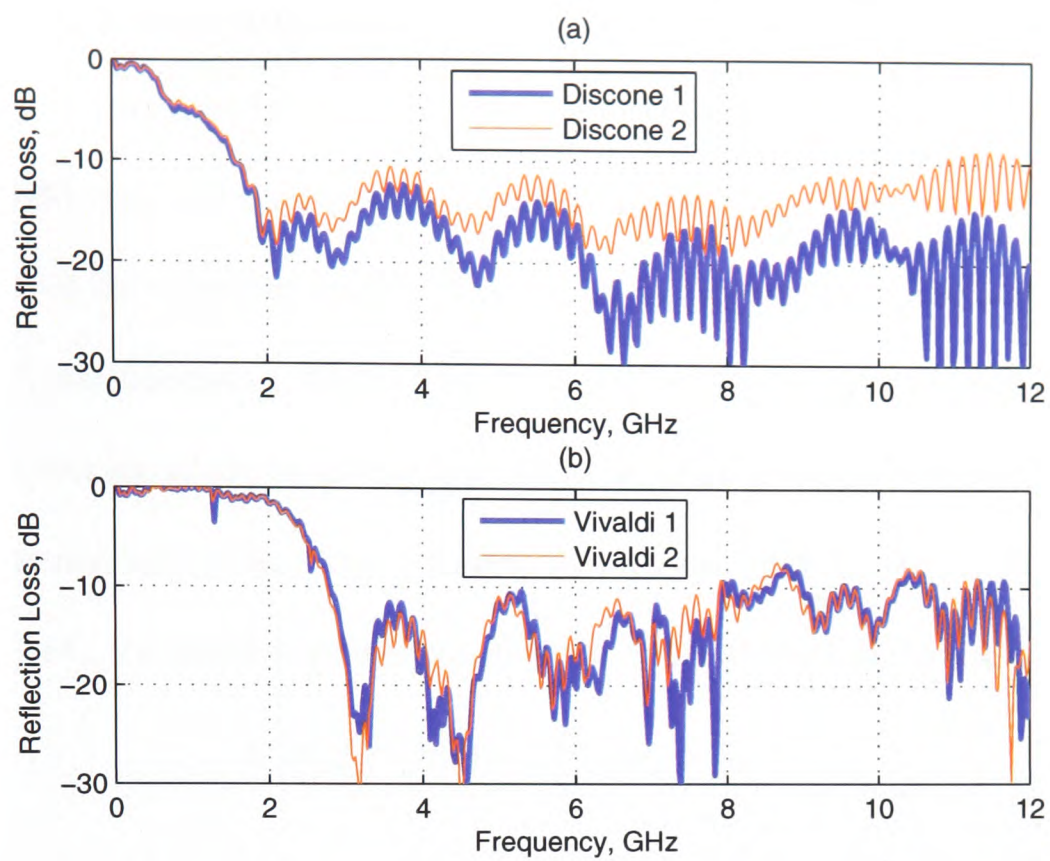


Figure 6.3 Measured reflection loss of the (a) discone and (b) Vivaldi antennas

The power radiation patterns were measured in the anechoic chamber using the setup shown in the plan view in Figure 6.4.

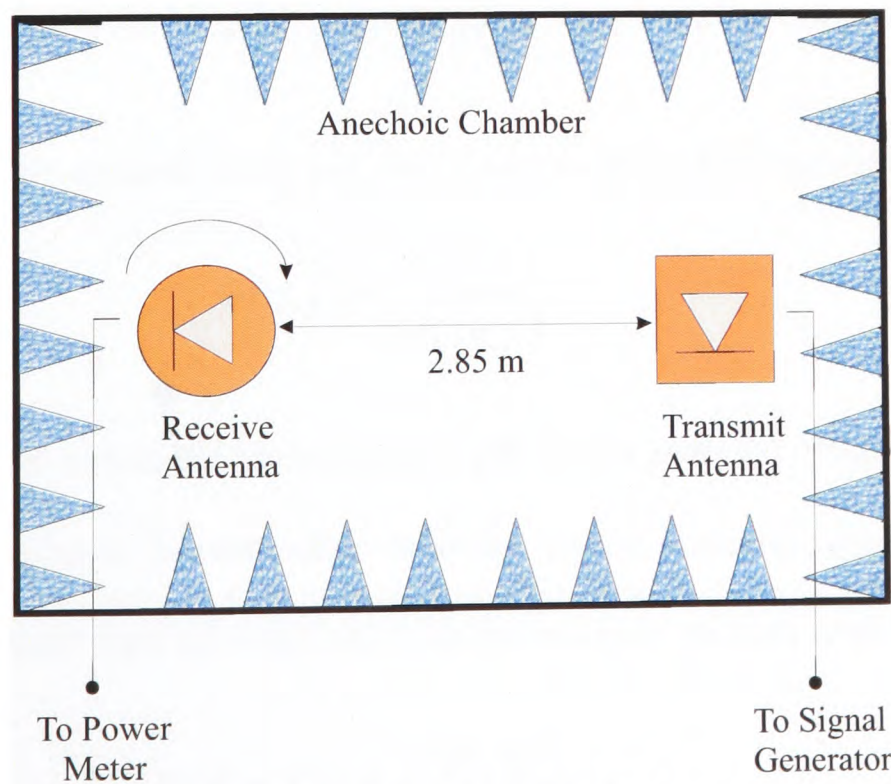


Figure 6.4 Antenna radiation pattern measurement in an anechoic chamber

A set of $n_f=16$ discrete equidistant frequencies within the UWB band were measured with a continuous-wave signal generator and power metre, providing a frequency resolution of $\Delta f=500$ MHz. Both antennas were placed on electromagnetic absorbent blocks to prevent any ground reflections, and were separated by a distance of $r=2.85$ m. The transmitted power was $P_t=15$ dBm, and a low-noise amplifier with 30 dB gain was connected to the receiving antenna. The transmitting antenna was fixed at one position while the receiving antenna was placed on a computer-controlled turntable and rotated in the elevation, θ , and azimuth, φ , planes from 0° to 360° with angular resolution $\Delta\theta=\Delta\varphi=1^\circ$.

Consider the rotation of the receiving antenna by an angle α_r while the transmitter is stationary at $\alpha_t=0^\circ$, with the antenna separation r kept constant. According to the Friis equation for continuous-wave free-space links [144], the received power is a function of the product of the antenna gains,

$$P_r(\alpha, f) = G_t(\alpha_t, f) G_r(\alpha_r, f), \quad (6.2)$$

where $G_t(\alpha_t, f)$ and $G_r(\alpha_r, f)$ are the gains of the transmitting and receiving antennas respectively, λ is the wavelength corresponding to f , and the transmitted power and free-space loss terms have been neglected. With one stationary antenna, the received power in our measurement can be represented by

$$P_r(\alpha, f) = G(\alpha, f) G(0, f), \quad (6.3)$$

where $G_t(\alpha, f)=G_r(\alpha, f)=G(\alpha, f)$ for the identical antennas. It is straightforward to see that

$$G(0, f) = \sqrt{P_r(0, f)}, \quad (6.4)$$

so that the gain of a single antenna-under-test can be derived from the received power as

$$G(\alpha, f) = \frac{P_r(\alpha, f)}{\sqrt{P_r(0, f)}}. \quad (6.5)$$

This procedure is used to obtain the multi-frequency radiation patterns, $P(\theta, \varphi, f)$, of the antenna in the elevation and azimuth planes. Normalisation with the highest received signal intensity removes the effect of the radiator-sensor separation and provides the radiation patterns used subsequently, i.e.,

$$P_n(\theta, \varphi, f) = \frac{P(\theta, \varphi, f)}{\sup_{\theta, \varphi} \{P(\theta, \varphi, f)\}}. \quad (6.6)$$

6.4. UWB Antenna Radiation Patterns

To develop an insight into the radiation characteristics of UWB antennas, the power radiation patterns of the discone and Vivaldi antennas are studied in this section. The discone is linearly polarised, circularly symmetric and omni-directional in the azimuthal plane. The far-field elevation pattern of a thin discone resembles that of a monopole, but for large values of the cone half-angle ψ , the presence and number of any sidelobes depend on ψ . The electrical length of the cone, $k_r=2\pi/l$, determines the cutoff frequency and the elevation plane beam pattern. The ground plane introduces a slant to the elevation pattern depending on the electrical diameter $k_d=2\pi/d$ and frequency f .

The Vivaldi antenna is a stripline-to-freespace transformer with a very broadband input match. The radiation pattern, however, varies considerably with frequency as the phase centre shifts within the structure. The advantage of the design is that maximum gain is maintained in the boresight direction. The lower cutoff frequency of the Vivaldi is determined by the flare angle, such that the aperture is half a wavelength wide at that frequency.

Figure 6.5 shows the normalised, co-polar power radiation patterns of the discone and Vivaldi antennas in the elevation and azimuth planes across the UWB band. In the figure, the discone elevation boresight is nominally at $\pm 90^\circ$, for the Vivaldi it is at 0° . Keeping in view the range of the radiated power, the images are thresholded to a lower level of -15 dB.

From Figure 6.5(a), the elevation-plane radiation pattern of the discone evolves with frequency. There is very low power when the two discone antennas are perpendicular to each other, as expected. Towards the lower end of the band, a large, well-defined mainlobe is observed close to the equatorial direction. Considerable radiation is also present at a wide range of angles due to the flow of current to the edges of the ground plane, since d is approximately twice the wavelength at the lower frequency range. With increasing frequency, the boresight pattern contracts and translates, while the backplane propagation vanishes. The most noticeable changes occur close to the 6 GHz region, beyond which the beam diverges and the mainlobe continues to narrow. At wavelengths shorter than l , the antenna starts behaving like a multiple-wavelength monopole instead of a quarter-wave monopole, resulting in narrower beams and an increase in the number of sidelobes and nulls [71]. The sidelobes are produced

by the reflection of the wave from the discontinuity at the far end of the finite cone as explained in [165], where the relationship of antenna dimensions with beamwidth, beam tilt and sidelobe level is discussed for a finite, large-angle conical antenna without a ground plane.

The discone’s radiation pattern in Figure 6.5(b) is uniform with the elevation angle as expected for an omni-directional antenna but there is considerable variation with f due to the non-uniform return loss and antenna efficiency. The Vivaldi antenna’s elevation pattern in Figure 6.5(c) and azimuth pattern in Figure 6.5 (d) both show strong boresight gain. At the lower end of the spectrum there is appreciable backplane propagation, especially in the case of the elevation pattern which is nearly omni-directional up to the centre frequency. The desired angular focussing is achieved only at higher frequencies. The elevation mainlobe narrows with increasing f and the sidelobes vanish. The azimuth pattern has more ripples than the elevation pattern, indicating greater leakage through sidelobes, which decay at lower frequencies for most angles.

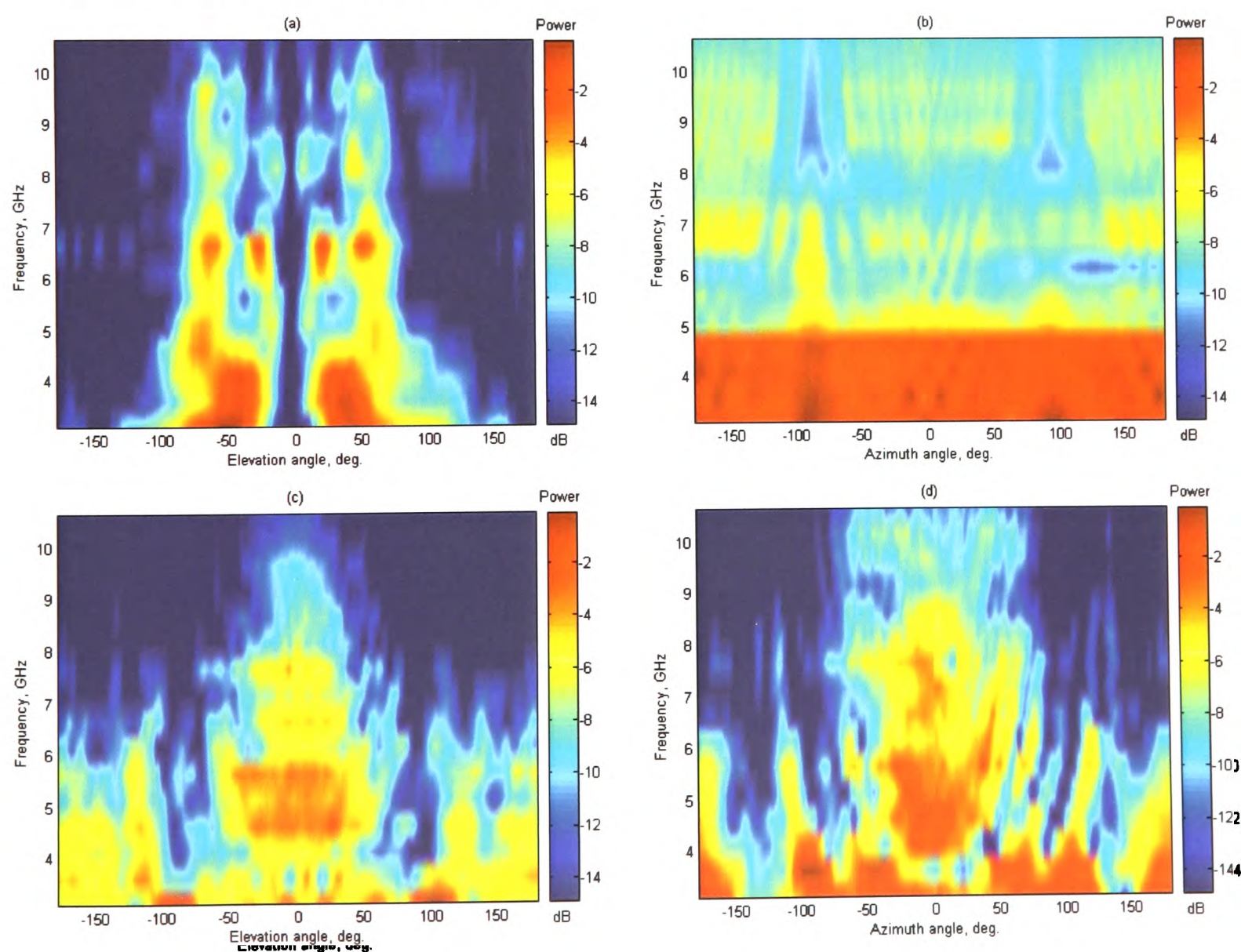


Figure 6.5 Evolution of UWB antenna radiation patterns with frequency: (a) discone elevation pattern; (b) discone azimuth pattern; (c) Vivaldi elevation pattern; and (d) Vivaldi azimuth pattern

Thus the elevation radiation pattern of the discone antenna is strongly dependent on frequency. The antenna's frequency properties at one look-angle are starkly different from those at another. Alternatively, the frequency components are radiated non-uniformly in different directions. This is illustrated further in Figure 6.6 where the mainlobe tilt and splitting with frequency is clearly visible.

In a typical indoor wireless application, a terminal using the discone antenna in a vertically polarised configuration held by a user sitting down would find a different frequency response, and thus performance, than while standing. Also, it is common to place the transmitting and receiving antennas in the horizontal plane. In such a situation, a signal with only a subset of the frequencies would be launched in the antenna plane, while a wider range of frequencies might be available at a steeper angle. All of these effects would cause waveform distortion and power loss in single-band UWB systems, while in multi-band UWB, the efficiency of the sub-bands will vary substantially with the look-angle.

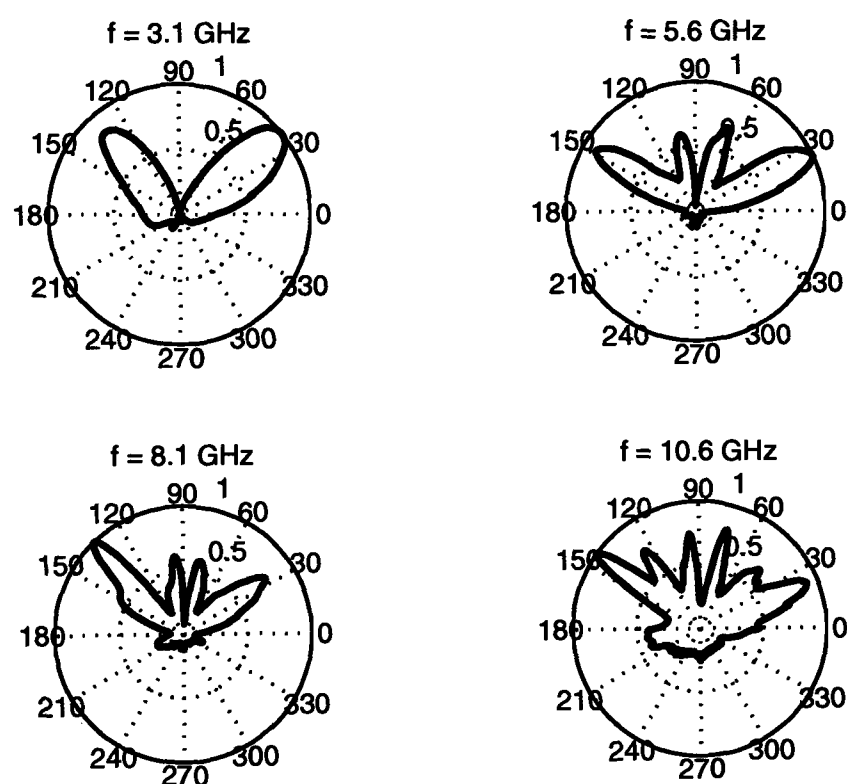


Figure 6.6 Elevation plane field radiation pattern of a discone antenna at various frequencies

6.5. Radiation Properties

The antenna beamwidth and directivity can be used to quantify and consolidate the angular-spectral variation of UWB antennas as observed in the previous section. The mainlobe beamwidth can be evaluated in terms of the beam solid angle

$$\Omega_A(f) = \int_0^{2\pi} \int_0^\pi P_n(\theta, \varphi, f) d\Omega = \int_0^{2\pi} \int_0^\pi P_n(\theta, \varphi, f) \sin\theta d\theta d\varphi. \quad (6.7)$$

Figure 6.7(a) shows $\Omega_A(f)$ for the discone and Vivaldi. Both have a large beam solid angle at low frequencies, lying in the region of 2–3 sr, which falls exponentially with frequency and in a nearly identical fashion for the two antennas reaching 0.1 sr at f_h . Even though the beam solid angle of the Vivaldi remains smaller, the discone also shows a small Ω_A for large f , behaving like a directional antenna. This is confirmed by the directivity,

$$D(f) = \frac{4\pi}{\Omega_A(f)} \quad (6.8)$$

which, for the discone, starts at 3 dBi at f_l and increases to 23 dBi, as seen in Figure 6.7(b).

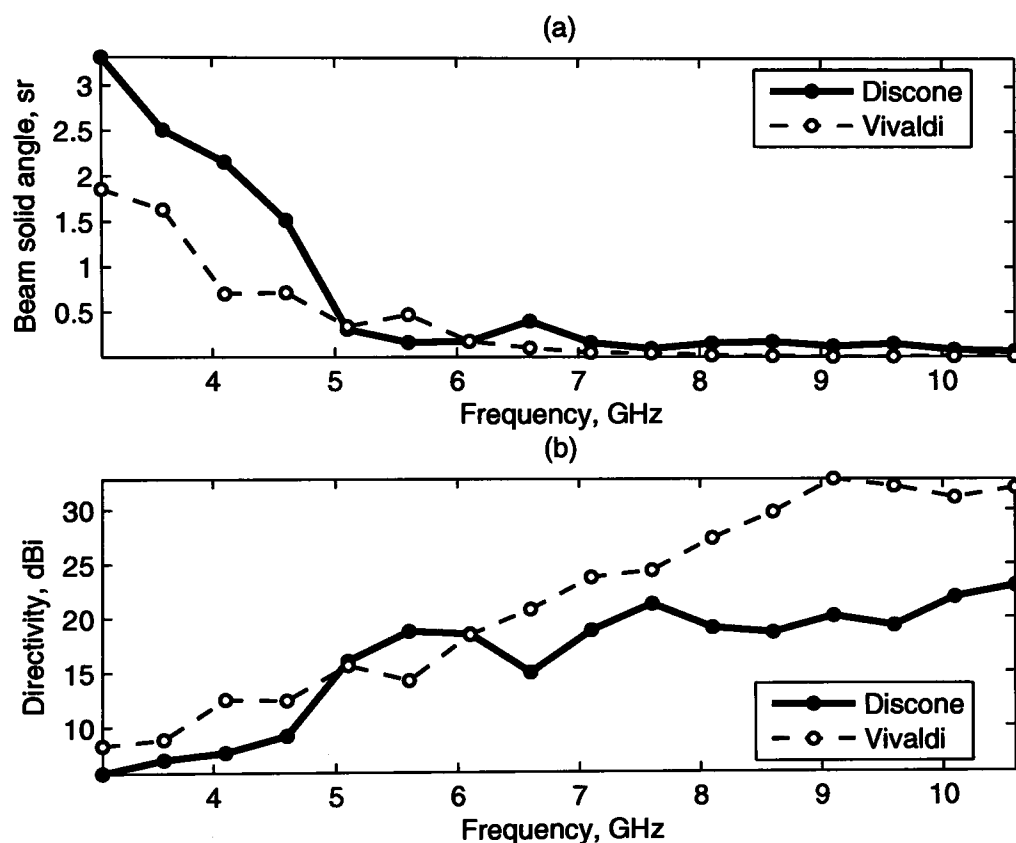


Figure 6.7 Variation of (a) beam solid angle, and (b) directivity of UWB antennas with frequency

Interestingly, near the mid-band region the discone is even more directive than the Vivaldi. The increased directivity of the discone in the upper band has the positive consequence of extending the link as long as the transmitting and receiving antennas are oriented along the main beam. However, this variation in directivity can also have a significant impact on the transmitted power. The system designer needs to ascertain that the transmitted signal power at any frequency and angle does not

exceed the EIRP regulatory limits as a result of variable directivity. The above results confirm that the omni-directional UWB antennas analysed here do not have a constant gain over the UWB band.

6.6. Signal Waveform Distortion

The effect of antennas on UWB system performance can be quantified by defining a suitable figure of merit for the quality of the communications link. We define angular bandwidth, $B_n(\theta, \varphi)$, as the n dB bandwidth at a given angular location, where n is a suitable threshold commonly taken as 3 or 10 compared to the peak of the radiation pattern. This is the effective bandwidth perceived by a terminal at that location and is a direct measure of angular-spectral dispersion. Figure 6.8 shows the elevation plane angular bandwidth $B_n(\theta, \varphi)|_{\varphi=\text{constant}}$ for the discone and Vivaldi, with the full system bandwidth $B_{\text{max}}=7.5$ GHz also marked for reference. Most noticeably, there is only a small window of θ for a discone antenna to approach the full $B_{10}(\theta, \varphi)$, close to $\theta=50^\circ$. This is significantly displaced from the equatorial plane usually deemed ideal for such omni-directional antennas, at which $B_{10}(\theta, \varphi)$ is less than 2 GHz. The Vivaldi antenna retains a third of its maximum $B_{10}(\theta, \varphi)$ at all values of θ , which amounts to partial spectral leakage in undesired directions.

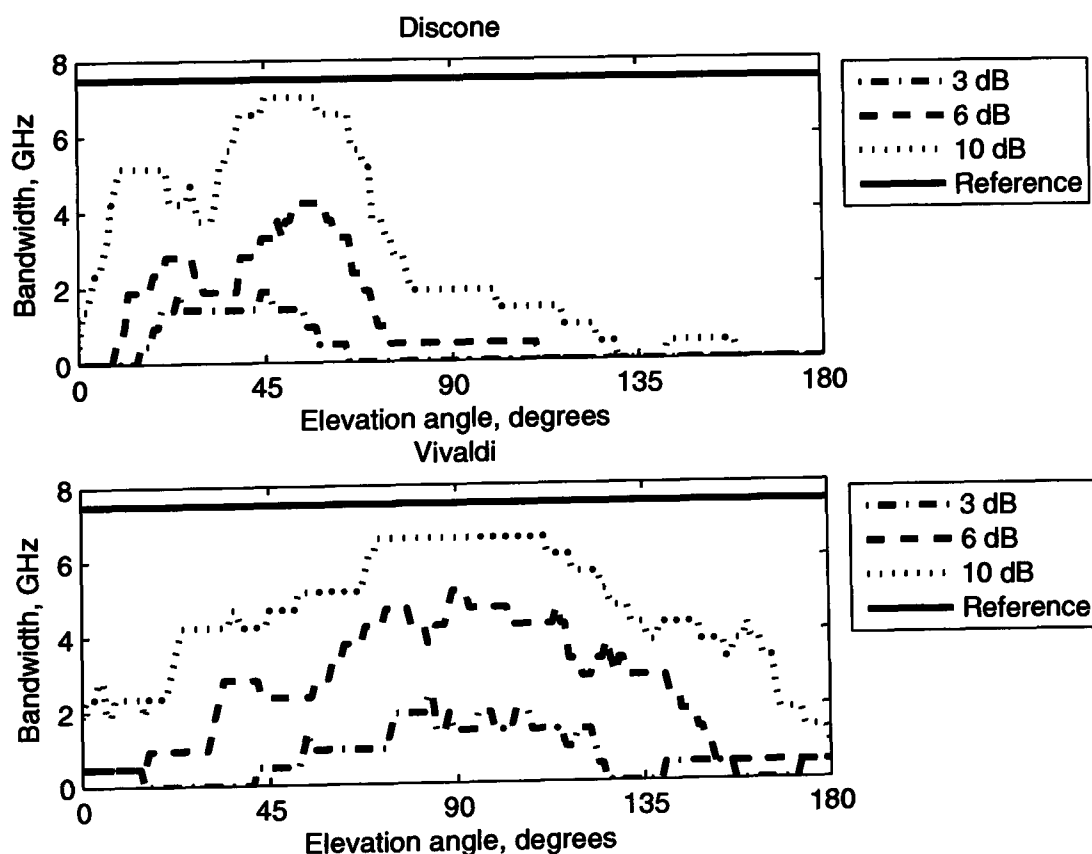


Figure 6.8 Effective angular bandwidth of discone and Vivaldi antennas in the elevation plane. The reference line indicates the full 7.5 GHz bandwidth of a UWB signal occupying the FCC UWB band

Figure 6.9 illustrates the measured antenna transfer functions

$$|H(\theta, \varphi, f)| = \sqrt{P_n(\theta, \varphi, f)} \quad (6.9)$$

at $\theta = \{0^\circ, 30^\circ, 60^\circ, 90^\circ\}$. The least attenuation and relative uniformity in the transfer function is observed at $\theta = 60^\circ$, which agrees with the conclusions drawn from Figure 6.8, and therefore the least spectral distortion would be caused to the signal radiated at that angle.

The edge reflections in a finite antenna cause radiation of multiple transients whose separation varies with the look angle [166]. This can be overcome with the use of tapered resistive [167] or capacitive [168] loading. Edge reflection analysis, however, is not included in this treatment.

The response of the antenna to signal $s(t)$ is given by

$$R(\theta, \varphi, f) = S(f)H(\theta, \varphi, f) \quad (6.10)$$

where $S(f)$ is the Fourier transform of $s(t)$ and $R(f)$ is the spectrum of the distorted signal. The effect of antenna dispersion on a UWB signal is illustrated with the help of a Gaussian monopulse [169],

$$s(t) = S_0 \frac{t}{\tau} e^{-\frac{t^2}{2\tau^2}} \quad (6.11)$$

where τ is the time decay constant and S_0 the amplitude.

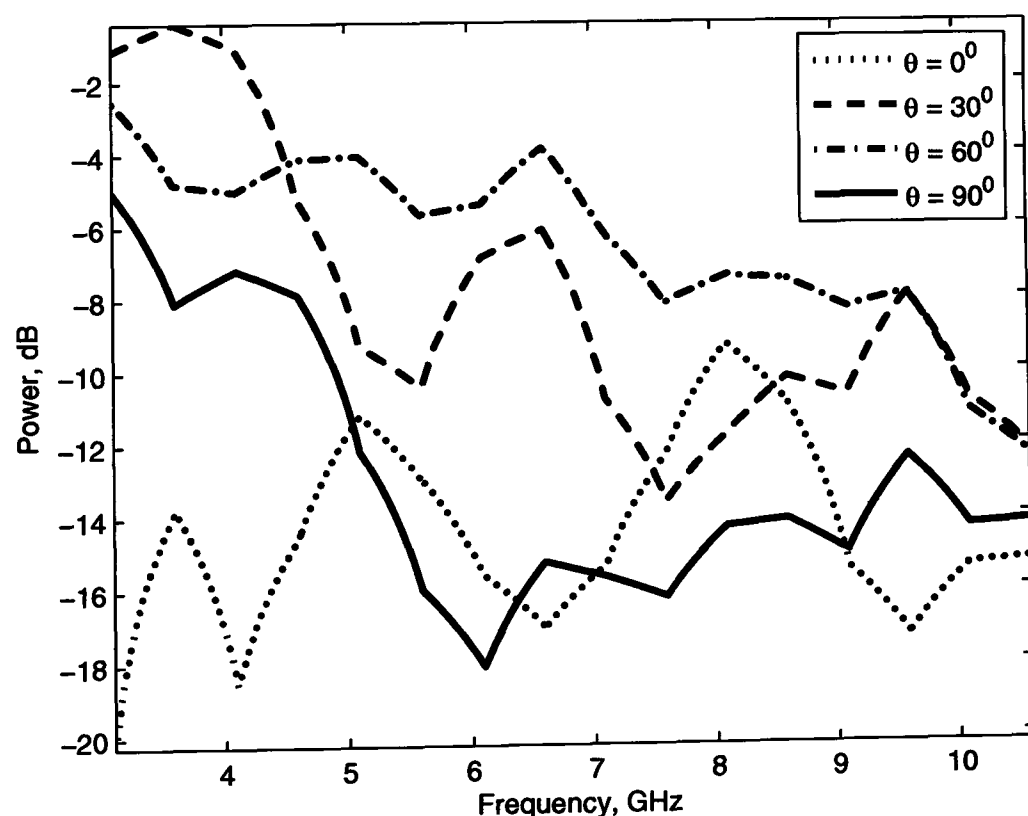


Figure 6.9 Frequency transfer functions of the discone antenna at various angles of elevation

The output waveform is given by

$$r(\theta, \varphi, t) = \int_{-\infty}^{\infty} s(t - \xi) h(\theta, \varphi, t) d\xi \quad (6.12)$$

where $h(\theta, \varphi, t)$ is the directional impulse response of the antenna and ξ is the integration variable.

Figure 6.10(a) shows the original monocycle $s(t)$ with a 3 dB bandwidth spanning the UWB band (3.1–10.6 GHz), while Figure 6.10(b) through (d) show the distorted waveforms $r(\theta, \varphi, t)$ for various θ . It is seen that a discone antenna alters the pulse shape depending on the elevation, and the waveform at $\theta=60^\circ$ has greater similarity to the transmitted (template) waveform than those at 30° and 90° , due to a flatter antenna response and wider effective angular bandwidth at 60° as discussed earlier.

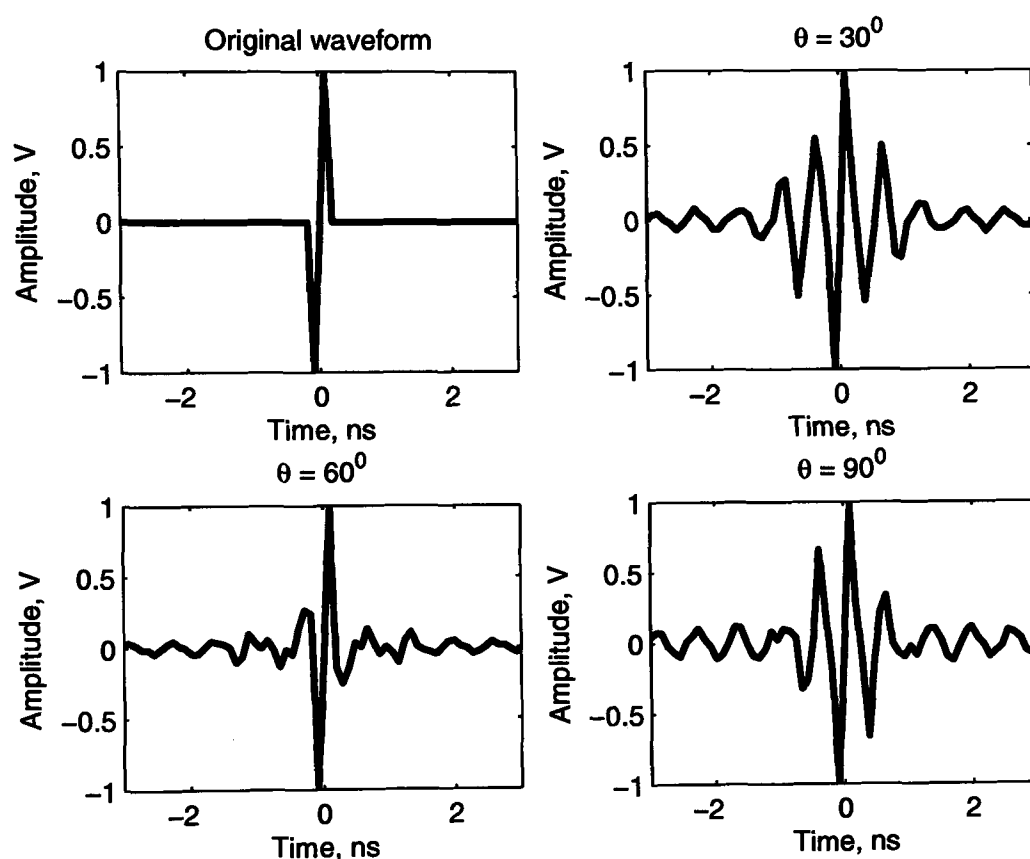


Figure 6.10 Waveform distortion by a discone antenna at various elevations. The amplitude is normalised

The angular-spectral effects of other antennas are also expected to be similar. Other UWB waveforms will also be distorted, and both pulsed and multi-band systems would be affected. This includes the orthogonal frequency division multiplexing (OFDM) [170] or direct-sequence (DS) [171] modulated signals used in multi-carrier UWB schemes [21]. In a communications link, the signal radiated by a UWB transmitter is filtered by the antenna based on its direction-of-departure. Multipath components of this antenna-filtered signal, after transmission, may undergo a combination of

propagation mechanisms altering the frequency content [14]. In the indoor UWB channel, a large number of multipaths reach the receiver with a wide range of angles of arrival [98]. The antenna transfer function again performs angular-spectral filtering, causing the pulse shape of each multipath arrival to be individually altered depending on its direction of arrival [17, 172], and consequently presenting a highly distorted composite signal to the subsequent receiver components. Antenna distortion can thus degrade the signal-to-noise ratio (SNR) at the receiver considerably. In UWB signals that already operate at low SNR, this can drastically increase the outage probability and symbol error rate, reduce the capacity, and limit the coverage range. The mitigation of spectral and directional antenna distortion can thus boost the link performance considerably.

6.7. Normalisation of Antenna-Induced Link Aberration

An antenna-independent channel model can be used to evaluate system performance in the channel using any given antenna, or to compare the performance of different antennas. Most UWB channel characterisation studies, however, largely ignore the effect of the antenna, making no mention of its contribution to the measured channel characteristics. This is misleading, as it has been shown that channel statistics such as the rms delay spread depend on antennas [173]. If the antenna is implicitly assumed to be a part of the channel, the model thus derived is valid only when the same antenna is used. As measurements are usually performed using large, high-performance antennas while actual systems require compact, economic designs, this discrepancy can lead to flawed system design. Removing antenna effects is thus critical for accurate channel modelling.

This section proposes a simple technique to extract antenna-independent UWB channel information through normalisation of the receiving antenna's frequency and directional characteristics. The effectiveness of the method is demonstrated experimentally for both vertically and horizontally polarised antenna orientations. Since the antenna dispersion is related to angle, the performance of the compensation technique is best evaluated by converting the channel profiles received over the synthetic aperture grid into the corresponding angular power spectra. Alternatively, the normalisation procedure can also be applied by first transforming the antenna radiation pattern into the Cartesian

coordinate system corresponding to the measured grid data. The former approach is followed here as it provides a visual representation of the multipath arrivals and the removal of the antenna's angular distortion. The multipath direction-of-arrival (DoA) information for each frequency component is obtained through the Fourier-based procedure described in Chapter 3. Super-resolution algorithms like SAGE, MUSIC or ESPRIT can also be used [174-176], but the proposed method is independent of the DoA determination technique. To perform the antenna normalisation, the angular-spectral antenna response is deconvolved from the channel response. In the frequency domain,

$$H_n(\theta, \varphi, f) = \frac{H(\theta, \varphi, f)}{A_a(\theta, \varphi, f)}. \quad (6.13)$$

where $A_a(\theta, \varphi, f)$ is the field pattern of the receiving antenna and the $H(\theta, \varphi, f)$ is the measured channel data. Since the radiation pattern of an omni-directional antenna does not vary with φ , (6.13) reduces to

$$H_n(\theta, \varphi, f) = \frac{H(\theta, \varphi, f)}{A_a(\theta, f)}. \quad (6.14)$$

The rest of this section provides some typical results in the form of pre- and post-normalisation angular power spectra for LOS and NLOS propagation environments with discone antennas. In each case, the wideband angular power spectrum is calculated from $H_n(\theta, \varphi, f)$ as

$$P_n(\theta, \varphi) = \int_{f_l}^{f_h} H_n^2(\theta, \varphi, f) df. \quad (15)$$

The transmitter is located at $(\theta, \varphi) = (\pi/2, -\pi/2)$, given the measurement configuration and definition of axes for the synthetic aperture. Also, each image is normalised to the highest intensity level for comparison, which is then referred to as 0 dB, and a -20 dB threshold is applied.

Figure 6.11 shows the pre- and post-normalisation angular power spectra for vertically polarised transmit and receive antennas. The strongest multipath in the LOS pre-normalisation image, that arrives from $\theta = 45^\circ$ indicating a reflection from the ceiling, appears to be stronger than even the direct path, which is counter-intuitive and is due to the discone antenna's high gain in that direction. Some other clusters are also visible but contain comparatively lower power. After normalisation, the direct path becomes dominant and the 45° is de-emphasised, while a cluster of scatterers can now be

observed in the horizontal direction, which corresponds to the actual propagation conditions. From the figure, it can be observed that greater angular resolution is also achieved by normalisation.

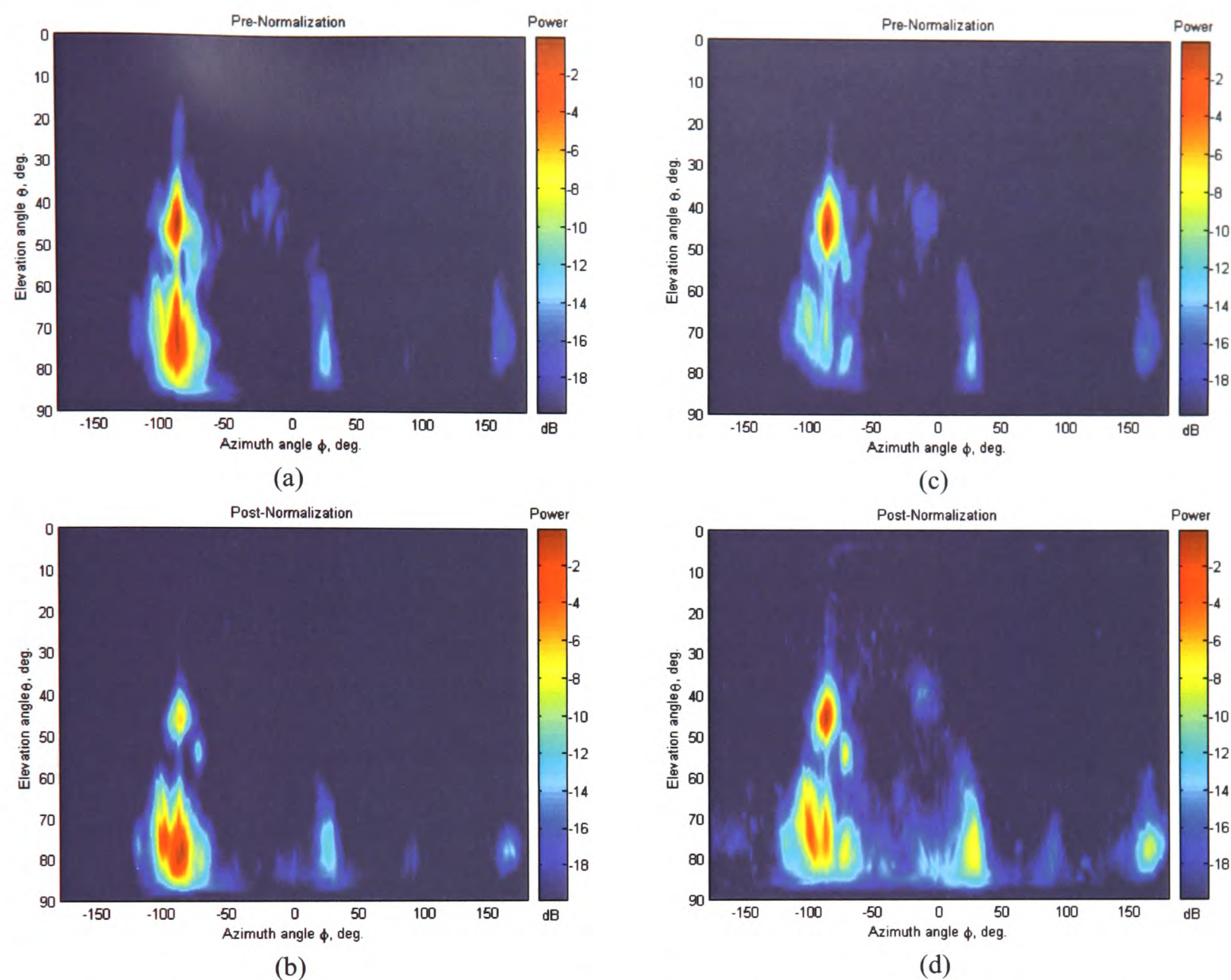


Figure 6.11 Angular profile of the received power (a) before and (b) after antenna normalisation, in the LOS channel with vertically polarised transmit and receive antennas. Similarly, (c) and (d) show the pre- and post-normalisation angular power spectra for the NLOS channel

The NLOS images also present a similar situation, with the effect of normalisation quite obvious. As the ceiling reflection is de-emphasised, the signal propagating in the horizontal plane becomes noticeable and many clusters are observed, which is expected in the NLOS scenario [16]. Some power also arrives from the direction of the transmitter, which is due to diffraction or reflection from the nearby objects.

Similarly, Figure 6.12 shows the pre- and post-normalisation angular power spectra for horizontally polarised antennas.

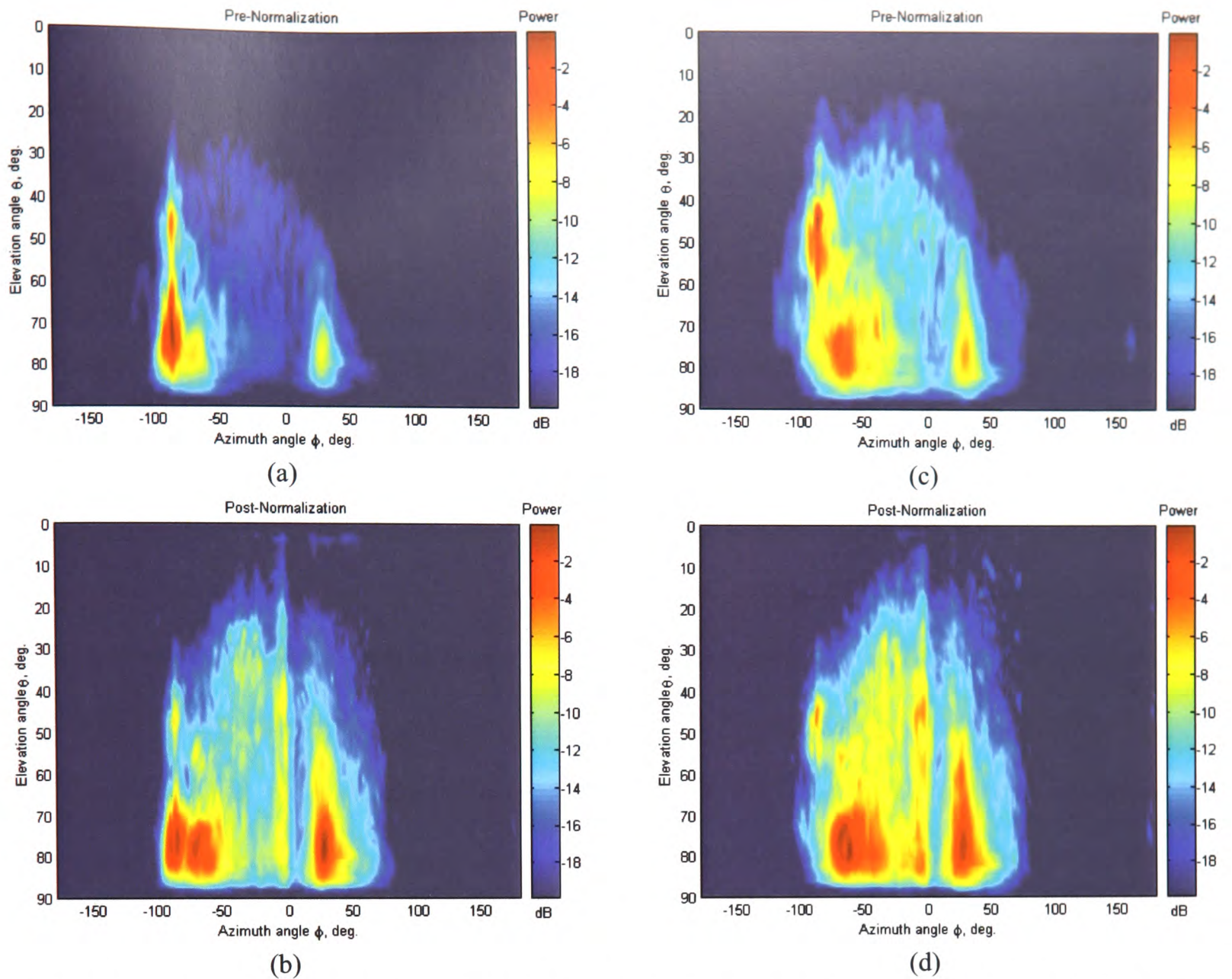


Figure 6.12 Angular profile of the received power with horizontal antennas (a) before and (b) after antenna normalisation, in the LOS channel. Also shown are the corresponding NLOS spectra in (c) and (d)

Prior to normalisation, the horizontal LOS spectrum differs remarkably from the NLOS spectrum. The post-normalisation LOS and NLOS angular spectra, however, are identical except for the presence of the direct component at $\phi=90^\circ$. This observation is intuitively correct, as the only difference between the LOS and NLOS propagation environments is the presence of the direct path. The strong scatterers in the azimuthal plane are emphasised by normalisation, as with the vertical polarisation. The differences between the vertical and horizontal post-normalisation channel profiles are due to the non-uniform response of scatterers to signals incident with varying polarisations.

This analysis confirms the validity of the technique for both vertical and horizontal electric field polarisations and antenna orientations.

6.8. Mitigation and Exploitation

Some of the dispersive effects of antennas and the likelihood of substantial degradation in system performance as a result of antenna artifacts have been highlighted above. Interestingly, some of these very characteristics specific to UWB antennas can also be used in performance enhancement techniques. The radiation of certain frequency components only in certain directions opens new avenues for UWB communications in terms of multiplexing, coexistence, and robustness.

Cognitive radio, a recent concept whose prime motivation is improved system reliability and spectrum utilization, is a possible means for exploiting this antenna behaviour [65]. With its roots in software-defined radio, it incorporates awareness of the radio environment through a feedback mechanism, which amounts to the estimation of interference temperature and the detection of spectrum holes. Based on this information, spectral and power control is applied dynamically and adaptively.

On the one hand, the cognitive radio architecture provides the flexibility to enable operation only at the part of the spectrum that is amenable to error-free, reliable communications. This is conceptually a feedback system in which the double-ended antenna effects are taken into account during transmission and consequent signal loss is prevented through appropriate pre-distortion filtering. It is obvious that both ends of the communications link require the adequate degree of adaptability to allow waveform shaping at the transmitter and template formation at the receiver respectively. A special case of this technique is soft-spectrum adaptation, proposed originally to comply with regulatory spectral masks in a flexible manner [177]. Thus the instantaneous dispersive behaviour of the antennas is equalized by an adaptive signalling scheme and the distortion is mitigated.

On the other hand, a more exciting utility of a cognitive radio in this scenario is its ability to transform the distortive antenna behaviour into a form of angular-spectral multiplexing in a multiuser environment. This technique would be more advantageous in a low spread multipath channel such as the line-of-sight short-range indoor channel designated CM1 in the IEEE 802.15.3a WPAN channel model specification [40]. A relatively small number of multipath arrivals with a low angular spread follow from this requirement. In such a situation, the multipath copies of the signal reaching a particular receiver are characterised by the corresponding set of antenna angular-spectral transfer

functions, a set that is small due to the condition imposed earlier. This antenna dispersion, a function of the small range of angles, is thus associated with the receiver location relative to the transmitter, and also with the geometry of the propagation environment. In other words, a distortion signature can be associated with each transmitter-receiver pair. It directly follows that such a system could be able to exploit the dispersion to implement a waveform discrimination strategy for a multiuser scenario, with a level of orthogonality provided by the waveforms altered during propagation through the antenna pair.

There can be other techniques for mitigating and exploiting antenna dispersion. A direction-aware radio receiver, though not incorporating the cognitive radio philosophy, can use the correlation between angular and spectral antenna dispersion to compensate for it. With DoA information, such a correlation receiver can optimise demodulation by choosing the appropriate template from its database.

6.9. Summary

Substantial variation in the radiation properties of UWB antennas with frequency has been demonstrated. Measurement results show that the beam pattern at one portion of the frequency band can be quite dissimilar from that at another. This variation can be quantified by expressing the beam solid angle and directivity as functions of frequency. Both omni-directional and directional antennas typically exhibit higher directivity in a frequency range that increases almost four-fold with respect to the lowest frequency in the UWB band. The effective bandwidth of a system is found to be highly sensitive to the angular disposition of the antenna at each end of the link and severely limited in some directions. Careful orientation of antennas is thus required to achieve full UWB operation for a range of antenna designs. With two vertically polarised UWB antennas operating in the same plane, the 10 dB bandwidth can be as low as 2 GHz when the joint angular-spectral dispersion is considered. Signal waveform distortion with look-angle is observed as a consequence, severely degrading the performance of a UWB receiver. The relationship between angular and spectral dispersion can be exploited in direction finding and localisation. A cognitive radio can use this phenomenon to perform adaptive waveform shaping and enable a directionally- and spectrally-discriminated multiuser system. A method has been presented for the normalisation of receiver antenna distortion in array systems

based on direction-of-arrival information. As demonstrated experimentally for a variety of channels, the proposed technique can provide antenna-independent channel measurements and can also be used in receivers to improve signal detection.

Having characterised the behaviour of a single wideband antenna under UWB signalling, the next chapter assesses the performance of multiple-antenna systems.

7. MULTIPLE-ANTENNA (MIMO) TECHNIQUES

7.1. Introduction

Antenna diversity techniques have been used extensively in narrowband systems to mitigate temporal and spatial fading [80, 178-181]. Similar in principle to other forms of diversity that utilise multipath, frequency or temporal redundancy, antenna diversity relies on independently fading signals at the multiple receiving antennas to reduce the probability of outage and improve the quality of service. Beside spatial antenna diversity, polarisation diversity is a more recently proposed alternative that allows more compact arrays with colocated antennas [83, 91, 111, 182-184]. Further extensions to antenna diversity techniques have demonstrated the advantage of arrays at both the transmitter and receiver for substantially increasing the spectral efficiency in fading channels [185-189]. Termed as multiple-input multiple-output (MIMO), this system architecture provides more efficient spectrum usage through spatial multiplexing and holds promise for future wireless systems with ever-increasing data-rate requirements. MIMO systems transmit simultaneous independent streams of information over the eigenmodes of a randomly fading matrix channel, exploiting the additional degrees of freedom to achieve higher data-rates than possible otherwise [190]. Diversity and spatial multiplexing in MIMO systems are closely related and one can be sacrificed for the other depending on the requirements, as shown for single-user and multiple-access channels in [191, 192] and [193] respectively. For MIMO systems with asymmetric diversity orders, selection between the excess antennas can provide diversity gain beside the multiplexing gain from the symmetric antenna subset [194, 195].

Most of the reported work on antenna diversity has focussed on narrowband systems, with the objective of improving signal reliability by mitigating small-scale fading [196]. Since the UWB channel does not suffer from deep power fades [20], the motivation for antenna diversity is instead derived from the regulatory control on power emission levels that in turn adversely affects the range, data rate and robustness of the link. As a severe penalty in throughput and quality of service is experienced beyond short distances, high data-rate UWB communications systems are severely range-restricted [197]. The increase in SNR offered by antenna diversity can help extend the range, overcome

fading, and enable the use of higher order modulation for higher data-rates. A 5 dB diversity gain can, for instance, compensate for the transmission loss of a wall, enabling operation in a through-wall propagation scenario [16]. The shallow small-scale fading in the UWB channel can be largely overcome so that a temporally and spatially uniform signal level is maintained.

While UWB already provides much higher data-rates than competing wireless technologies, the use of MIMO spatial multiplexing can extend the lead much further. A small number of antenna elements on each end can increase the system capacity manifold, providing the ultimate in high-speed short-range communications. Thus a MIMO-equipped UWB system has the potential to support applications such as simultaneous wireless transmission of several high-definition multimedia streams or multi-gigabit radio links. As the antenna separation required for orthogonal MIMO signalling depends on frequency, UWB systems allow compact placement of antenna elements and smaller device sizes. Antenna diversity is thus more feasible with UWB than with lower frequency narrowband systems. Furthermore, the frequency diversity of the UWB channel can yield much higher outage capacity compared to narrowband channels [198]. These features of UWB make the use of MIMO techniques attractive, and motivate the study undertaken in this chapter.

Impelled by the theoretical predictions of MIMO performance gain on canonical channels, experimental investigations of narrowband MIMO channels were undertaken in [198, 199], while [200, 201] studied the effect of physical considerations for frequency-selective channels with a semi-analytical treatment. Limited work on MIMO performance in the UWB arena has been reported, however. Error rate reduction in UWB channels with space-time coded MIMO systems was analysed in [202, 203] and with a rake-aided MIMO system using pulse-based UWB signals in [204]. The performance of binary antipodal signalling was investigated in [205], while [206] studied time-hopping and direct-sequence UWB with MIMO. A highly simplified channel model was adopted for UWB to determine the MIMO capacity in [207].

Recent work in information theory has focused on the interplay between bandwidth, SNR and MIMO capacity. In view of the low SNR per degree of freedom in UWB channels, the terms ‘wideband capacity’ and ‘low SNR capacity’ are sometimes considered synonymous [208]. It has been shown that the channel capacity saturates at large bandwidth with optimal signals but vanishes with

spread-spectrum signals. A comprehensive treatment of spectral efficiency in the wideband regime was presented in [209], generalising the term to refer to all scenarios where the information rate per receive dimension is small, such as when several receive antennas are present. Earlier, it was shown in [210] that ‘peaky’ signals, localised in time or frequency, are optimal for achieving capacity in wideband multipath channels, while for ‘white-like’ signals the mutual information is inversely related to bandwidth [137]. The latter was confirmed by [211], which showed that signals uniform over both time and frequency can perform well with UWB systems only in quasi-static – and not in time-varying – channels. A similar decline in capacity was also noticed for uncorrelated scattering fading channels in [212]. Extending these wideband capacity results to multiple antennas, the analysis in [213] showed that linear MIMO capacity scaling with antennas is conditional on the quadratic growth of the number of resolved multipaths. It also argued that the capacity varies directly with the SNR at large bandwidth when the channel is known at the receiver, failing which it was found to be quadratic in SNR in [214].

This study is inspired by the promise of UWB and MIMO technologies for high data-rate and reliable communications, and seeks to exploit their synergetic performance gains. The performance of MIMO systems is investigated in the UWB channel under the assumption of perfect receiver-side channel information. For three-branch spatial and polar antenna arrays, the measured MIMO sub-channel fading correlation, branch power ratios and power gains are analysed. The consequent increase in system robustness and information rate is evaluated.

7.2. Analysis Procedure

A measurement of the MIMO channel over the FCC-allocated UWB frequency range was conducted using the procedure discussed in Chapter 2. Both LOS and NLOS measurements were undertaken, with the blockage for the latter created by placing a 1.25 m x 0.75 m x 0.75 m aluminium box 0.5 m away from the transmitter. MIMO systems with two types of antenna array configurations were measured, namely spatial and polar, with up to three diversity orders on each end so that the number of antennas assumed a maximum value of $n_T=n_R=3$. Figure 7.1 shows the arrangement of antenna elements for the construction of spatial and polar arrays.

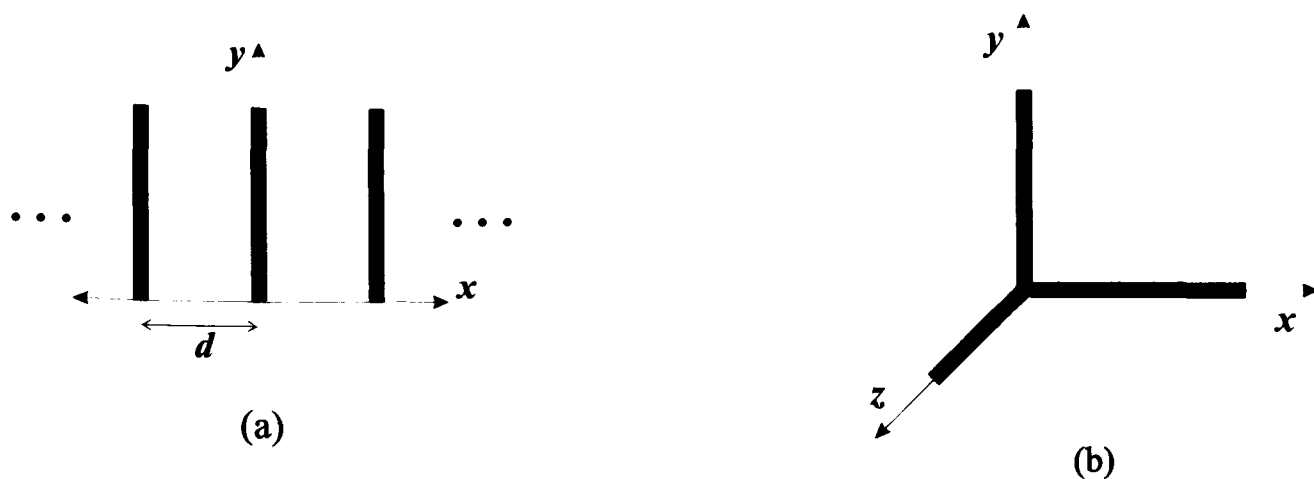


Figure 7.1 (a) Spatial and (b) polar MIMO antenna array configurations. The antenna separation d for spatial arrays should exceed one half wavelength. Polar antennas are dimension-limited but colocated

While alternate configurations exist for antenna placement in spatial MIMO systems [215], a virtual uniform linear array was used in this study owing to its popularity and simplicity. The spatial array elemental spacing d was 0.06 m, ensuring that the condition

$$d \geq \frac{\lambda_l}{2}, \quad (7.1)$$

was met, where $\lambda_l=0.097$ m is the largest wavelength in the measurement corresponding to the lowest frequency, $f_l=3.1$ GHz. Thus the spatial antenna elements, denoted by S_1 , S_2 and S_3 , were separated by over half a wavelength for all of the constituent frequency components. The transmit antenna arrays were synthesised by means of sequential measurements assuming stationarity of the indoor channel. Similarly, the tri-element receiver array was synthesised by extracting channel transfer functions from grid positions d apart. The reason for non-concurrent MIMO sub-channel measurements was to remove antenna coupling, which was considered peripheral to the main focus of the study, as commercially designed MIMO antennas would be expected to enjoy low coupling under the operating conditions.

Polar MIMO measurements were conducted by placing each of the three orthogonally polarised antennas sequentially at the same location to emulate a colocated tri-polar array. The rest of the measurement procedure was similar to that for spatial MIMO. The use of three orthogonally polarised antennas along the Cartesian axes provided a basis spanning the entire polarisation signal space, and the extraction of all three electric field components was thus possible at each measurement location. In the rest of this paper, the polar antenna elements will be represented by V , H_1 and H_2 , corresponding to

the vertical and the two horizontal orientations, of which the last is oriented along the line joining the transmitting and receiving antennas.

Thus $n_d=30$ receiver positions were scanned on the grid in each direction for each of the combinations of spatial, polar, LOS and NLOS measurements, yielding $n_d^2=900$ MIMO channel transfer function matrices in each case.

7.3. MIMO Channel Formulation

For each measurement point (x,y) on the grid, the complex MIMO transfer function

$$G_{x,y}^{(m,n)}(f) = \sum_{k=1}^{n_f} A_{x,y,k} e^{j\psi_{x,y,k}} \delta(f - k\Delta f), \quad (7.2)$$

is obtained, where m and n signify the transmitting and receiving antennas respectively, such that $m,n \in \{S_1, S_2, S_3\}$ or $\{V, H_1, H_2\}$ for spatial or polar arrays respectively. Power normalisation across the antennas and frequencies is then performed in order to remove the pathloss and divide the power between the antennas [199], to obtain

$$H_{x,y}^{(m,n)}(f) = G_{x,y}^{(m,n)}(f) \left(\frac{1}{n_T n_R n_f} \sum_{i=1}^{n_T} \sum_{j=1}^{n_R} \sum_{k=1}^{n_f} |G_{x,y}^{(m,n)}(f)|^2 \right)^{\frac{1}{2}}. \quad (7.3)$$

In the subsequent analysis, a comparison of the performance of 1x1 SISO, 1x2 and 1x3 SIMO, and 2x2 and 3x3 MIMO systems is made. Thus in eq. (7.3), n_T and n_R are defined accordingly in each case to apply the unity average power constraint for fair comparison.

For each measurement point, the wideband MIMO channel matrix, $\mathbf{H}_{x,y,f} \in \mathbb{C}^{n_R \times n_T \times n_f}$, is formulated by collating the narrowband MIMO channel matrices at all frequency components. Thus for the point (x,y) ,

$$\mathbf{H}_{x,y,f} = \begin{bmatrix} H_{x,y}^{(1,1)}(f) & \dots & H_{x,y}^{(n_T,1)}(f) \\ \vdots & \ddots & \vdots \\ H_{x,y}^{(1,n_R)}(f) & \dots & H_{x,y}^{(n_T,n_R)}(f) \end{bmatrix}, \quad f = \{1, \dots, n_f\}, \quad (7.4)$$

represents the wideband MIMO channel with n_f frequency components. It is worth emphasising that this expression is different from the conventional MIMO channel representation for narrowband channels with frequency-flat fading, in which the channel coefficients are scalar quantities without frequency dependence.

For a deterministic matrix channel, the effective degrees of freedom can be evaluated from its rank. Random matrices arising from multipath fading, however, have full rank in general, which is therefore not a reliable measure of the degrees of freedom. The sub-channel correlation, the covariance matrix eigenvalue distributions and branch power ratios can be used instead to derive the actual number of independent channels supported by a MIMO system and their gains, as analysed next.

7.4. Sub-Channel Correlation

Sub-channel fading independence is a well-established criterion for the evaluation of MIMO performance [216, 217]. A correlation coefficient of up to 0.5 causes little loss in capacity [218], which is therefore commonly used as the correlation threshold.

The frequency-domain power correlation coefficients are computed as

$$r_{\mathbf{u}\mathbf{v}} = \frac{E[\mathbf{u} \cdot \mathbf{v}] - E[\mathbf{u}]E[\mathbf{v}]}{\sqrt{E[\mathbf{u}^2 - (E[\mathbf{u}])^2]E[\mathbf{v}^2 - (E[\mathbf{v}])^2]}}, \quad (7.5)$$

where ‘.’ denotes element-by element vector multiplication, $E[.]$ denotes expectation, while $\mathbf{u}$ and $\mathbf{v}$ denote two sub-channel power spectra or the squared transfer functions. The power transfer function $\mathbf{u}$ can be written as

$$\mathbf{u} = \left| H_{x,y}^{(m,n)}(f) \right|^2, \quad (7.6)$$

and $\mathbf{v}$ is defined similarly.

The transmit correlation is defined as the correlation between the signals transmitted by two antennas and received by a single antenna. As an example, for the transmit correlation between the

signals transmitted by the two horizontally polarised antennas and received by the vertically polarised antenna, we have,

$$\mathbf{u} = \left| H_{x,y}^{(H_1,V)}(f) \right|^2 \quad \text{and} \quad \mathbf{v} = \left| H_{x,y}^{(H_2,V)}(f) \right|^2 \quad (7.7)$$

for a given (x,y) . Similarly, the receive correlation is a measure of the statistical independence between the two subchannels formed by one transmitting antenna and two receiving antennas.

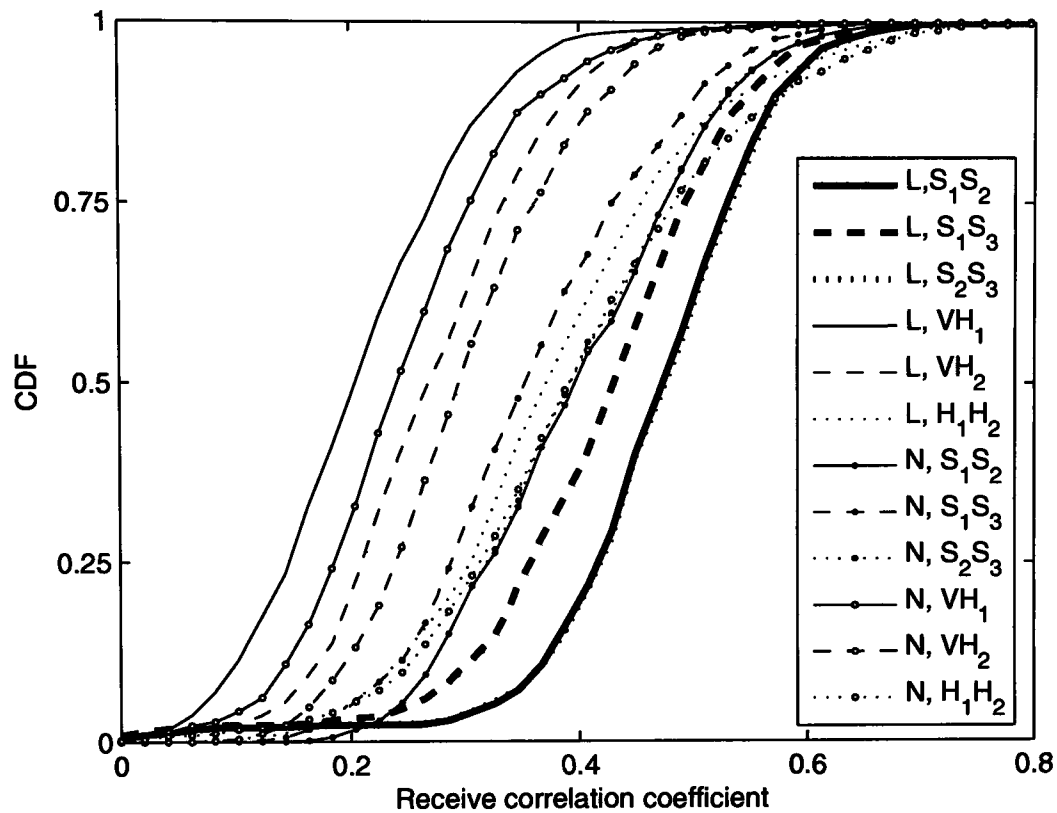


Figure 7.2 Empirical CDFs of MIMO sub-channel receive correlation coefficient for spatial and polar arrays in LOS and NLOS scenarios

Figure 7.2 shows the statistical distributions of the receive correlations computed over the measurement ensemble of size n_d^2 . With the spatial arrays, the signals received by adjacent antennas are more correlated than those received by the extreme end antennas. The reason lies in the well-known fact that the signal correlation falls as a Bessel function with antenna separation [219]. Polar arrays also experience different correlation between the antenna pairs, with the highest correlation coefficient $r_{H_1H_2}$ obtained with the two horizontal antennas. Also, r_{VH_1} is consistently lower than r_{VH_2} , suggesting that the two horizontally polarised antennas would provide unequal robustness to fading and capacity gains.

In the LOS channel, the spatial array correlation has a lower mean but higher variance than in the NLOS channel, while that of the polar arrays is largely unaffected. The correlation coefficients are mostly lower than 0.5, with higher values obtained only in a small number of cases. They are almost always lower than 0.7.

This experimental analysis shows that the UWB MIMO sub-channels generally undergo independent fading, and multiple-antenna systems can provide significant capacity and diversity gain.

7.5. Power Gain Analysis

The eigen decomposition of the MIMO channel's covariance matrix provides information about the relative strengths of the independent transmission modes supported by a multiple-antenna system [220]. The number of significant eigenvalues determines the effective degrees of freedom.

For each measurement point, the MIMO covariance matrix, which is a Wishart matrix, is evaluated from the MIMO channel matrix as

$$\mathbf{R}_{x,y,f} = \mathbf{H}_{x,y,f} \mathbf{H}_{x,y,f}^H, \quad (7.8)$$

with $(.)^H$ representing Hermitian transposition. Each $\mathbf{R}_{x,y,f}$ can be decomposed into its n_R eigenvalues.

The eigen decomposition is given by

$$\mathbf{R}_{x,y,f} = \mathbf{Q}_{x,y,f} \mathbf{\Lambda}_{x,y,f} \mathbf{Q}_{x,y,f}^H, \quad (7.9)$$

where $\mathbf{Q}_{x,y,f}$ is an $n_R \times n_R$ unitary matrix and $\mathbf{\Lambda}$ is a diagonal matrix containing the eigenvalues λ_i , $i=\{1,2,\dots,n_R\}$, of $\mathbf{R}_{x,y,f}$.

Nominally, all n_R eigenvalues are significant, indicating finite power gain in each sub-channel and thus the existence of multiple transmission paths that a MIMO system can exploit. As $\mathbf{H}_{x,y,f}$ is a random matrix, λ_i is also random and its statistical properties can be analysed.

Figure 7.3 shows the empirical distributions of the eigenvalues for the three-branch spatial and polar MIMO channels. In each case, three significant eigenvalues are observed, indicating that the power gains of the three sub-channels are sufficient for MIMO diversity and spatial multiplexing. They also turn out to be distinct, suggesting power imbalance across the branches. Thus uniform power

allocation across the antennas, used in the unavailability of channel state information at the transmitter (CSI-T), would perform worse than waterfilling [137, 221] over these channels. With CSI and waterfilling, singular value decomposition (SVD) provides the unitary matrices containing the optimal weights for the receiving and transmitting arrays. These can be used to maximise the spatial multiplexing gain through optimal power allocation to the antennas. Similarly, SVD-generated weights can be used to maximise the diversity gain for transmit diversity using waterfilling and for receive diversity using maximal-ratio combining (MRC). In terms of spatial multiplexing, the availability of CSI-T and its exploitation with waterfilling performs better than the $\min(n_T, n_R)$ capacity scaling factor that is ideally obtained otherwise.

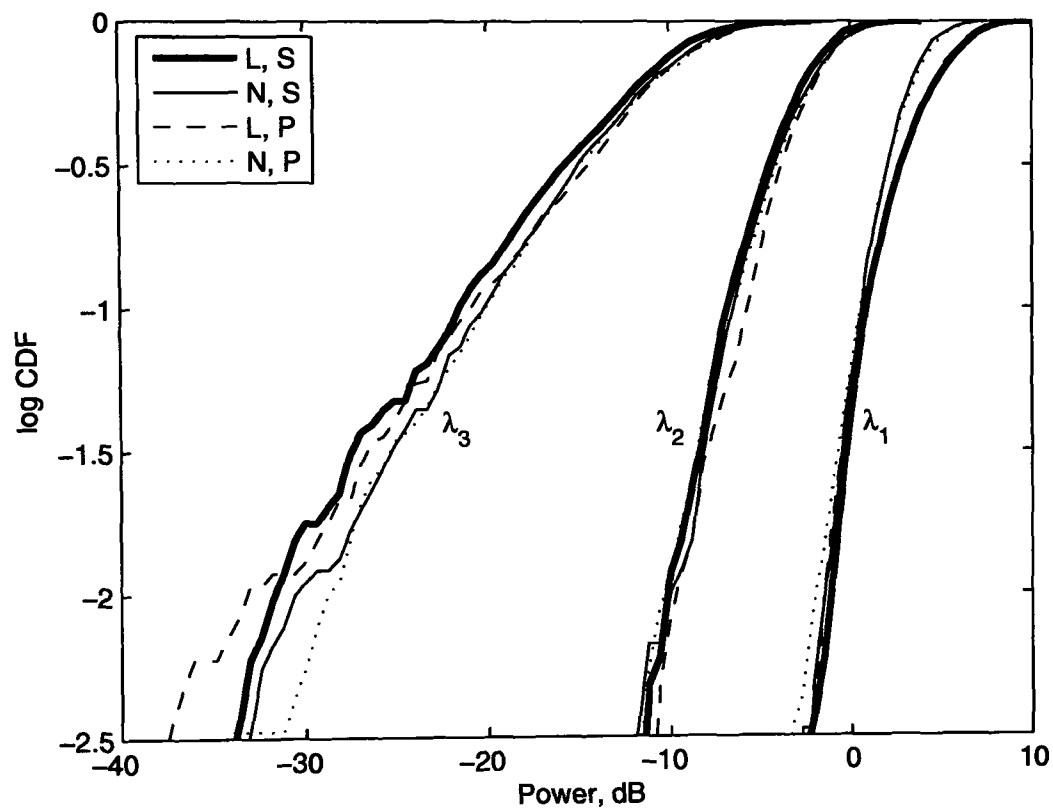


Figure 7.3 Empirical CDFs of the covariance matrix eigenvalues for the tri-element MIMO systems

7.6. Branch Power Ratios

The performance of a diversity system is optimal when the mean SNRs of the signals incident at the receiver antennas are uniform [91]. The branch power ratio, BPR, provides a comparison of the signal power levels at the diversity branches. Uniform BPR is especially important for equal gain combining (EGC) schemes [80], where the excess noise contribution from a low signal power branch would incur an SNR penalty on the combined signal.

The BPR of the n^{th} diversity branch at the receiver is calculated as [183]

$$\text{BPR}_n = 10 \log_{10} \frac{P_n}{P_{\text{total}}}, \quad n = \{1, \dots, n_R\}, \quad (7.10)$$

where P_n is the received power level in the n^{th} branch and

$$P_{\text{total}} = \sum_{n=1}^{n_R} P_n, \quad (7.11)$$

is the total power received by the n_R -element array. It follows from (7.10) that the ideal BPR for each branch in a tri-antenna receive diversity system is -4.8 dB.

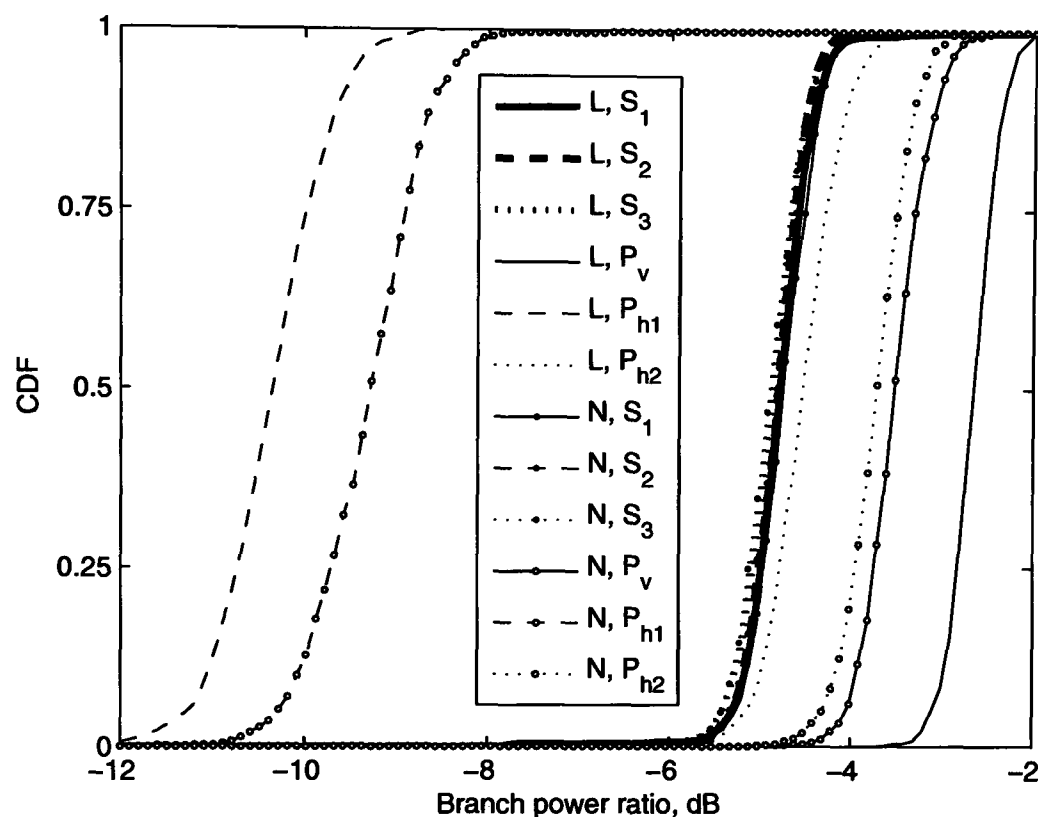


Figure 7.4 Empirical CDFs of branch power ratio for three-branch spatial and polar SIMO arrays

Figure 7.4 shows the BPR of each of the spatial and polar antennas when they are taken in 1x3 combinations. The transmitter in each case is a vertically polarised antenna at array position 1, i.e. S_1 or V . It can be observed that each of the branches in the spatial array receives almost identical power levels, with a median value of about -4.8 dB. There is, however, a large power difference in the polar arrays especially in the LOS channel. In both LOS and NLOS, the BPR at H_1 is the lowest, indicating that very little power is coupled to that polarisation when a vertically polarised signal is transmitted. This can be related plausibly to wave propagation and scattering in an indoor environment, as

reflections from objects with faces oblique to the incident wave cause a polarisation rotation of the wave, suggesting the observed behaviour when the geometry of the measurement environment, shown in Figure 2.1, is considered. This information predicts that three-branch spatial diversity would perform well in all channel conditions and provide symmetric gain irrespective of the antenna positions, but the performance of dual-polar diversity would be highly sensitive to antenna orientation.

The next section provides a quantitative analysis of MIMO performance in terms of diversity and spatial multiplexing gain. It is assumed that there is no feedback path and thus only receiver-side channel state information (CSI-R) is available.

7.7. Antenna Diversity Gain

With appropriate combining, an increase in the number of diversity antennas can improve the received signal quality and reduce error rates. Antenna diversity is classically implemented in single-input multiple-output (SIMO) configurations. Multiple-input single-output (MISO) architectures exploiting maximal ratio transmission show similar diversity gain but only in the presence of CSI-T [222]. Continuing with the assumption of only CSI-R, this section evaluates the $1 \times n_R$ SIMO diversity gain with spatial and polar arrays using MRC. Defined as the difference between the input and output SNRs of the diversity system, the diversity gain is highest when the multipath effect is significant, the sub-channel signals are uncorrelated and the branch power ratios are equivalent. The existence of these conditions in indoor UWB channels has been verified in the preceding discussion, and a considerable diversity gain is expected consequently, as analysed below.

The output SNR of the SIMO antenna diversity system with MRC is given by [80]

$$\gamma_{SIMO} = \sum_{n=1}^{n_R} \gamma_n, \quad (7.12)$$

where γ_n is the SNRs of the n^{th} branch. From this expression, the SIMO system's SNR is higher than that of any individual diversity branch, and increases by a factor of n_R ideally.

The empirical cumulative distribution functions of SIMO diversity gain in UWB channels are shown in Fig. 6. It can be observed that a 2.2–3 dB SNR gain is achieved with $n_R=2$, depending on the

array and channel type. Three-branch spatial diversity provides a total SNR gain of 4 dB. A comparison of the performance of spatial and polar arrays with similar diversity orders shows that the former consistently provides a much higher gain. The difference is most pronounced in LOS channels with $n_R=3$, where the output SNRs differ by 2 dB. In NLOS channels with $n_R=2$, however, spatial and polar diversity perform equally well. Also, spatial diversity performance is not affected by LOS conditions, but the same is not true for polar diversity.

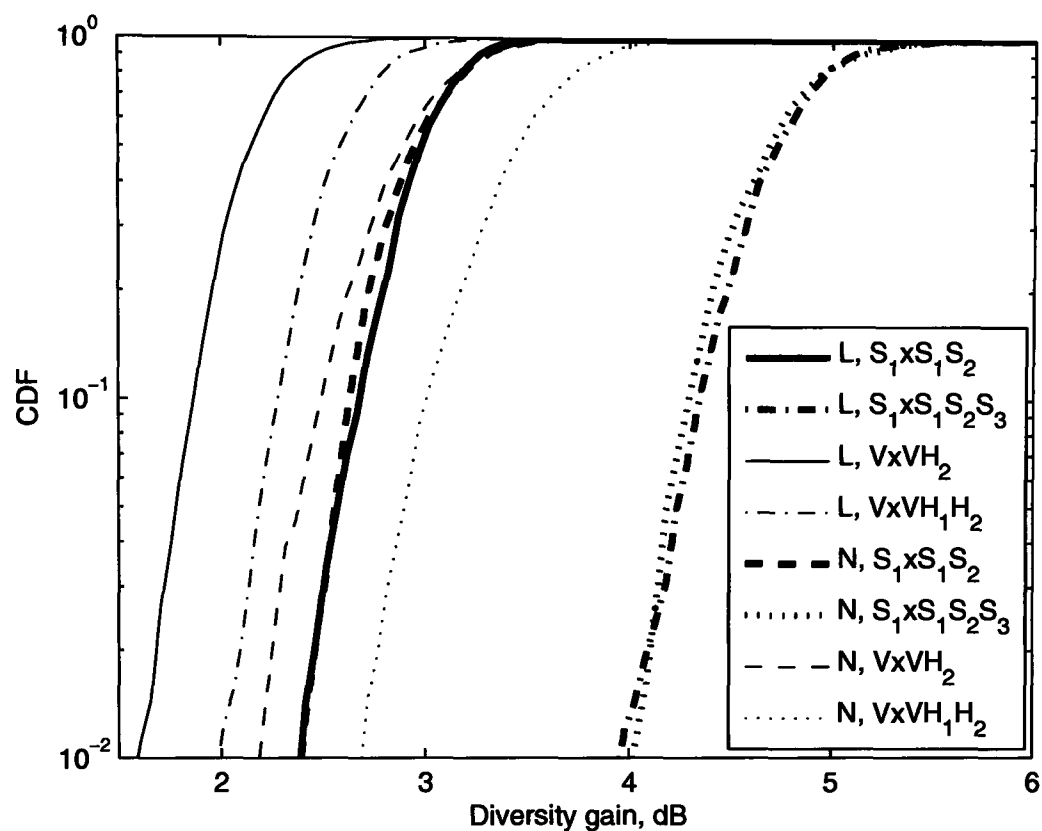


Figure 7.5 Empirical CDFs of diversity gain from spatial and polar SIMO (1x2 and 1x3) arrays

The above values of diversity gain have been calculated at 1% outage probability, or 99% reliability, which is a practical and popular measure. The median is also sometimes considered, as it indicates the gain achieved in half of the channel instantiations. The outage and median diversity gains are listed in Table 7.1. The comparative observations made above for the outage diversity gain also hold, in general, for the median diversity gain. The table shows that the SNR increases by up to 4.7 dB in half of the cases. Also, the median and outage diversity gain differ by less than 1 dB, demonstrating low variance in the diversity gain. Thus with three-branch diversity at the receiver, the SNR can be improved by up to 4 dB at the outage level, which is significant compared to the 5–6 dB fades in the UWB channel [20], and can thus be used effectively to overcome channel fading.

Table 7.1 Diversity gain with the SIMO antenna configuration

Antenna	Antenna Combination		Median, dB		At 1% Outage, dB	
Array Type	Transmitter	Receiver	LOS	NLOS	LOS	NLOS
Spatial	S ₁	S ₁ S ₂	3.0	3.0	2.4	2.4
	S ₁	S ₁ S ₂ S ₃	4.7	4.7	4.0	4.0
Polar	V	VH ₂	2.2	2.9	1.6	2.2
	V	VH ₁ H ₂	2.6	3.5	2.0	2.7

7.8. Spectral Efficiency Gain

The MIMO channel capacity quantifies the increase in the information rate achievable with spatial multiplexing, and is a measure of the highest data-rate that can be supported over the channel with acceptably low error rates. Fashioned after the Shannon capacity [136], the MIMO capacity of a flat-fading matrix channel, $\mathbf{H}$, is the mutual information maximised over the transmission alphabet probability. The capacity in bps/Hz of a flat-fading MIMO channel without CSI-T is given by [185, 186]

$$C_{MIMO} = \log_2 \det \left(\mathbf{I}_{n_R} + \frac{\rho}{n_T} \mathbf{H} \mathbf{H}^H \right) = \sum_{i=1}^{n_R} \log_2 \left(1 + \frac{\rho}{n_T} \lambda_i \right), \quad (7.13)$$

where a full-rank covariance matrix has been assumed and

$$\rho = \frac{E_s}{N_0} \quad (7.14)$$

is the average SNR at the receiver antennas with E_s representing the total symbol energy and N_0 the noise power spectral density. A special case is SIMO, for which (7.13) reduces to

$$C_{SIMO} = \log_2 \left(1 + \rho \sum_{i=1}^{n_R} |h_i|^2 \right), \quad (7.15)$$

where the h_i represent the elements of the SIMO vector channel. Eq. (7.15) shows that SIMO systems provide a logarithmic increase in capacity with n_R . MISO offers similar capacity enhancement under

CSI-T [222]. Multiple antennas at both ends can, however, establish multiple parallel transmission modes and scale the capacity by a factor of $\min(n_T, n_R)$, providing the sub-channels are independent.

The capacity of a frequency-selective channel is derived as the expectation of the capacities of its narrowband components. Thus the wideband capacity expression becomes

$$C_{MIMO} = \frac{1}{n_f} \sum_{f=1}^{n_f} \log_2 \det \left(\mathbf{I}_{n_R} + \frac{\rho}{n_T} \mathbf{H}_{x,y,f} \mathbf{H}_{x,y,f}^H \right). \quad (7.16)$$

This expression is used to calculate the MIMO capacity of the UWB channel in this section.

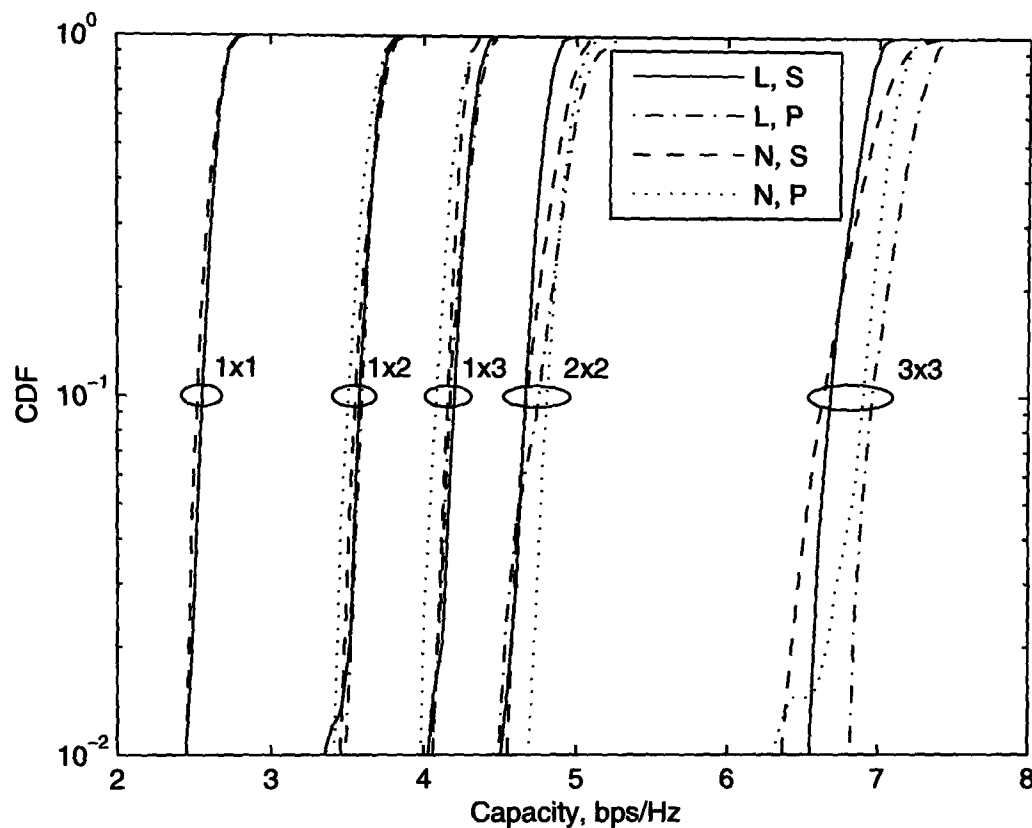


Figure 7.6 Empirical CDFs of MIMO capacity in UWB channels with spatial and polar arrays

Figure 7.6 shows the CDFs of MIMO capacity obtained from the measured channels assuming CSI-R and 10 dB average SNR at the antennas. The capacities obtained with SISO (1x1), SIMO (1x2 and 1x3) and MIMO (2x2 and 3x3) array configurations are shown. From the figure, spatial and polar arrays of the same diversity order provide similar capacities in both LOS and NLOS channels. The incremental SIMO capacity gain evidently decreases with n_R . The MIMO capacity, when $n_T=n_R$, however experiences a greater increase with diversity order. This is seen more clearly in Table 7.2, in which the ergodic, or median, capacity and the 1% outage capacity for each of the cases under consideration are listed. With tri-element arrays at 10 dB SNR, the capacity can be up to 6.85 bps/Hz.

While the capacity does not scale exactly linearly with the number of antennas, it is nevertheless orders of magnitude higher than that achieved with SIMO. The sublinear increase in capacity can be attributed to nonzero correlation and unequal branch powers, which are the limiting factors in the performance of a MIMO system. These results verify experimentally that the MIMO capacity of UWB channels increases almost linearly with the number of antennas for at least up to three orders of spatial multiplexing.

Table 7.2 MIMO capacity at 10 dB SNR without CSI-T

Antenna	Antenna Combination		Ergodic Capacity, bps/Hz		1% Outage Capacity, bps/Hz	
Array Type	Transmitter	Receiver	LOS	NLOS	LOS	NLOS
Spatial	S ₁	S ₁	2.68	2.66	2.47	2.47
	S ₁	S ₁ S ₂	3.70	3.72	3.37	3.48
	S ₁	S ₁ S ₂ S ₃	4.33	4.35	4.05	4.09
	S ₁ S ₂	S ₁ S ₂	4.81	4.92	4.54	4.58
	S ₁ S ₂ S ₃	S ₁ S ₂ S ₃	6.91	7.24	6.58	6.85
Polar	V	V	2.68	2.66	2.47	2.47
	V	VH ₂	3.70	3.66	3.53	3.43
	V	VH ₁ H ₂	4.27	4.23	4.09	4.00
	VH ₂	VH ₂	5.00	4.97	4.52	4.72
	VH ₁ H ₂	VH ₁ H ₂	6.98	7.10	6.40	6.31

In the infinite bandwidth limit, the mutual information capacity approaches the ergodic capacity with appropriate signaling [213]. Thus, while there is little improvement in the ergodic capacity due to the frequency diversity of a frequency-selective channel, the more practical metric, viz the outage capacity, increases significantly. In other words, the capacity of the random wideband channel asymptotically becomes deterministic as the increase in bandwidth tightens the capacity bounds. This

was shown experimentally for wideband channels in [198]. From our statistical analysis, the standard deviation of MIMO capacity obtained is up to 0.2 bps/Hz, increasing with the diversity order. Also, spatial MIMO capacity enjoys a slightly lower standard deviation. These observations confirm that the above-mentioned results for wideband channels relating the bandwidth to the variance of capacity can be extended to UWB channels. This result underlines the advantage of MIMO spatial multiplexing with UWB systems and establishes that the combination of the two technologies provides the ultimate performance in short-range wireless communications.

7.9. Summary

MIMO performance in indoor UWB channels has been characterised experimentally for spatial and polar antenna arrays with up to three elements. It has been shown that MIMO signals are decorrelated sufficiently due to scattering, paving the way for considerable performance gain. The branch power ratios are identical for spatial arrays but markedly different for polar arrays, suggesting that the performance of dual-polar antenna arrays is highly sensitive to antenna orientation. Spatial SIMO diversity with two or three receiver antennas increases the SNR by 2.4 or 4 dB respectively, and is insensitive to the presence of LOS. Polar diversity provides a much reduced diversity gain especially in LOS channels. In NLOS channels it yields an outage diversity gain of 2.2 and 2.7 dB with dual- and tri-element arrays respectively. With both spatial and polar arrays, the capacity scales almost linearly with the number of antennas at each end in symmetric MIMO configurations. Thus the 1% outage capacity with 3x3 arrays is as high as 6.85 bps/Hz at 10 dB SNR. The standard deviation of SIMO diversity gain is only about 1 dB, while the MIMO capacity varies by up to 0.2 bps/Hz. The variance is small due to the inherent frequency diversity and low fading in the UWB channel. This work has established that UWB systems with MIMO configurations hold the key to robust, multi-gigabit wireless links for short-range indoor networks.

The next chapter presents a framework for the optimal end-to-end design of a UWB communications system using the channel and subsystem information gathered so far.

8. OPTIMAL SYSTEM DESIGN

8.1. Introduction

The preceding chapters have presented a comprehensive analysis of the UWB channel properties using experimental techniques. The performance of various system components over the FCC-allocated UWB frequency range has also been characterised. This chapter presents a framework for the design of an optimised UWB communications system with the aim of demonstrating the value of the prior analysis. As enhanced performance often comes with increased system complexity, practical constraints are taken into account to arrive at the optimal tradeoffs for the design of high-performance yet physically realisable and commercially viable systems.

8.2. A Simple Behavioural Model

In order to keep the analysis meaningful without including unnecessary detail, a simple UWB system model is considered. A pulsed transmission scheme is assumed, with the signals well localised in time and spanning the FCC UWB band (3.1–10.6 GHz). This can be achieved, for instance, with bipolar derivative-of-Gaussian pulses, but the exact pulse shape and modulation is unimportant for this discussion. A rake receiver of diversity order L using hybrid-selection/maximal-ratio combining (H-S/MRC) is assumed (see Chapter 5). Each of the rake branches consists of a correlator that correlates the incoming signal with a locally generated template waveform, using only the amplitude parts of both signals. The decision device is a slicer that produces an estimate of the transmitted data bit.

Multiple-antenna arrays consisting of omni-directional, linearly polarised elements are considered at both transmitter and receiver in order to include an assessment of MIMO performance, with the associated pre- and post-processing (see Chapter 7). The availability of channel state information is assumed only at the receiver, enabling receiver antenna diversity and spatial multiplexing without the requirement of channel estimation for the feedback path or transmitter-side non-uniform power

allocation. Diversity combining is performed using MRC, and the performance is evaluated at 1% outage, or equivalently, 99% availability. This system model is illustrated in Figure 8.1.

By far the most likely operating scenario for high data-rate radio communications using UWB technology is in wireless personal area network (WPAN) configurations at a short range within an indoor environment, and is therefore used in this analysis. For the sake of simplicity, it is assumed that the system is operating in a single-user environment and there is no multi-user interference. The analysis is for a single pulse, and therefore time-hopping or other forms of pulse repetition coding are also not considered.

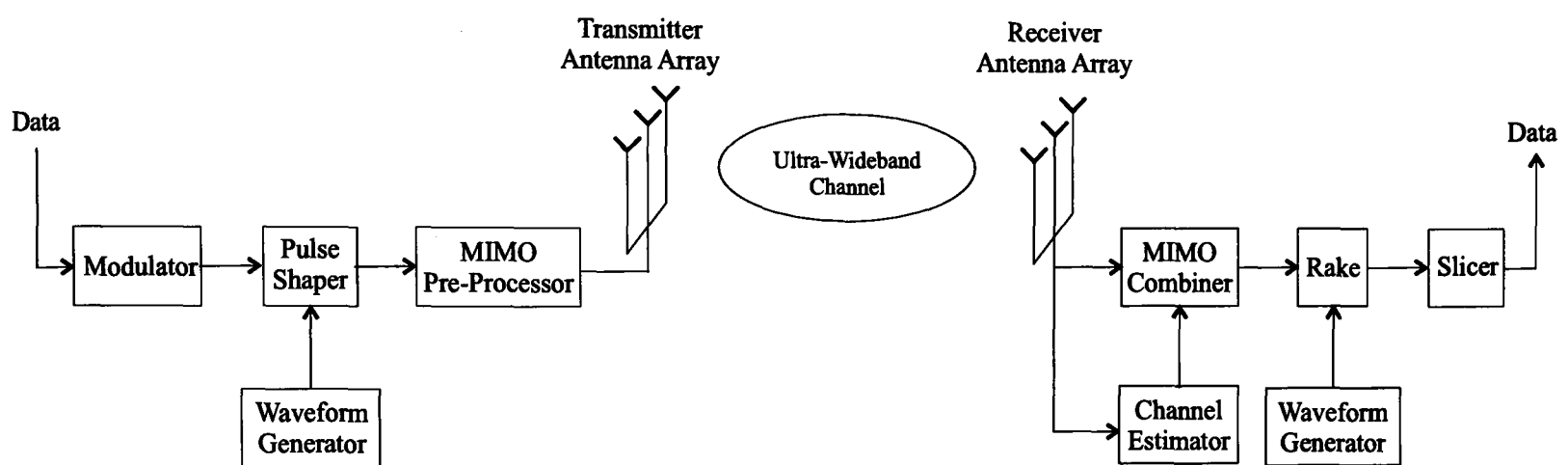


Figure 8.1 An example of a simple UWB communications system

8.3. Operating Scenarios

In order to demonstrate the impact of the operating conditions on system performance, a number of distinct scenarios are chosen that reflect some of the most frequently encountered situations. In this analysis, the key factors distinguishing these configurations are identified as the bandwidth of the transmitted signal, its electromagnetic polarisation and the presence or absence of an unobstructed propagation path (the line-of-sight or LOS path) between the transmitter and the receiver. The following three representative scenarios are considered.

1. With LOS present, the signal is vertically polarised and instantaneously spans the full UWB band.
2. The vertically polarised signal instantaneously spans 1 GHz, and LOS is present. In order to disregard any characteristics specific to certain sub-bands that are peripheral to the present

discussion, the average behaviour across the band is considered by assuming that the signal covers the full UWB band in a given time interval, such as with a frequency-hopping spread spectrum implementation.

3. The LOS is blocked, and the horizontally polarised signal spans the full band instantaneously. In the light of the previous chapters, the horizontally polarised NLOS case presents a harsh propagation scenario in terms of signal dispersion and attenuation. It is, however, of considerable practical importance given that user movement and changes in the device orientation can often lead to unintended horizontally polarised signal transmission even in an otherwise static environment. The NLOS scenario can also frequently arise in the indoor wireless environment due, for instance, to signal blockage by human movement.

8.4. Channel Impairments

A multitude of channel impairments affect signal propagation and reception fidelity under the operating conditions listed above. The small-scale fading is particularly benign in UWB systems due to the frequency diversity and multipath resolution provided by the large bandwidth, and will therefore not be included in this analysis. Fast fading is practically non-existent in indoor wideband channels, and will also be ignored. Some of the other channel effects, as characterised in the earlier part of this thesis, will be discussed in the following treatment. This should not, however, be treated as a comprehensive and all-encompassing assessment of channel distortion, as it is only an illustrative example. Some of the channel impairments most relevant to this discussion are the following.

1. Multipath propagation determines the frequency-selectivity of the channel, apart from the material composition of the propagation environment. The number of resolved multipaths, N , can be well approximated with a discrete normal random variable. The average number of resolved multipaths is $\bar{N}=80$ at the full bandwidth, and falls linearly as the bandwidth is decreased.
2. Frequency-dependent pathloss (FDP) leads to the distortion of the propagating waveform. The dispersion index, ζ , is a measure of the FDP and can be modelled as a continuous normal random

variable. According to its characterisation in the preceding chapters, ζ does not vary with bandwidth but depends on the signal polarisation and the presence of LOS.

3. The presence of a large number of scatterers in the indoor channel with various orientations and inclinations results in the depolarisation of the transmitted signal. This effect can be quantified in terms of the cross-polar discrimination (XPD), which is well approximated as a normal random variable. Significant variations in the XPD are observed when the LOS path is blocked or the transmitted signal polarisation is altered.

These are some of the primary channel impairments whose significance and severity in the UWB channel have been established previously in this work. Their effect on the various signal types will be analysed and system configurations optimal to combat them will be devised.

8.5. Transmitter and Receiver Sub-Systems

The design techniques that can combat the above-mentioned channel impairments effectively have been comprehensively analysed earlier in this thesis. Specifically, the following system components are used in this analysis to mitigate the channel effects.

1. A rake receiver captures and combines the multipath power, improving the SNR. The performance of a rake receiver is dependent on its diversity order, bandwidth and number of paths (see Chapter 5). Under consideration here will be sub-optimal, L -finger H-S/MRC rake receivers, with large L representing high complexity.
2. The template waveform used by the correlator is generated locally at the receiver. With Gaussian signalling, the use of templates matched to the channel distortion can improve the correlator output and partially overcome the FDP. This can be based on the antennas, as those are the major source of FDP. A further refinement can include template adaptation based on channel estimation, but with additional cost in terms of complexity (see Chapter 3).
3. Multiple diversity antennas along orthogonal polarisations can recombine the signal depolarised by the channel. A receiver diversity array with a $1 \times n_R$ configuration is considered in this analysis, where n_R represents the number of receiving antennas (see Chapter 7).

8.6. Performance–Complexity Optimisation

An evaluation of the highest achievable performance with the lowest system complexity constitutes the optimal system design in a given channel. The figure of merit for system performance considered here is the signal-to-noise ratio (SNR). As the bit error rate (BER) of a given signalling scheme is a monotonic function of SNR, the latter can be easily used to derive information about the expected error rates. The mutual information capacity of the system is also a function of the SNR, and is thus covered in this analysis. The system complexity, on the other hand, is limited by practical constraints related to fabrication and deployment. As a precise mathematical definition of system complexity true for all possible systems is not simple, the present treatment limits the scope of the problem to the subset of components listed in the previous section.

Mathematically, this situation can be formulated as a multivariate optimisation problem that can be solved with maximum *a posteriori* (MAP) estimation techniques for the maximum SNR gain and minimum complexity [223]. Alternatively, the problem can be treated through evolutionary algorithms such as the cross-entropy or particle swarm optimisation methods [224]. In the interest of keeping the present discussion intuitive and relating it to physical insight, however, a heuristic approach is used here, and the evaluation of the SNR gain *vis à vis* complexity, as undertaken in the previous chapters, is employed as the performance improvement measure. The fractional performance curves for the sub-systems are used in the determination of the optimal system components in relation with the channel parameters for each of the operating scenarios under consideration.

8.7. Quantitative Results

Table 8.1 lists the three operating scenarios of Section 8.3 along with some measures of the distortion suffered by the signal during propagation. As previously mentioned, the bandwidth, transmit polarisation, and presence or absence of LOS serve as the independent variables. These are used to determine the number of incident and resolved multipaths, the dispersion index, and the cross-polar discrimination in the light of the statistical analysis carried out earlier in this thesis. The mean channel

behaviour, and therefore the average value for each quantity, is considered. The optimal configurations for the sub-systems are also shown, along with the SNR gain provided by them. These figures are derived directly from the characterisation of measured UWB channels and the performance of the sub-systems in those channels, undertaken in the previous chapters.

Table 8.1 Signal impairments in the UWB channel, and the system components to mitigate them with the resulting gain in SNR under some illustrative operating scenarios

Channel and System Components	Scenario 1	Scenario 2	Scenario 3
Bandwidth, GHz	7.5	1	7.5
Transmit polarisation	V	V	H
Propagation channel	LOS	LOS	NLOS
Number of resolved multipaths N	80	11	80
Dispersion index ξ	2.04	2.04	2.39
VH cross-polar discrimination, dB	4.50	4.50	2.31
Correlation template matching	Antenna	Antenna	Channel
Rake receiver diversity order L	30	4	30
Polar antenna diversity order n_R	2	2	3
Correlation gain, dB	3.3	3.3	2.9
Rake diversity gain, dB	7	5	20
Polar antenna diversity gain, dB	1.6	1.6	2.7

As shown in the table, a large number of the total incident multipaths are resolved with the full-band signals of scenarios 1 and 3, and a high diversity order rake is required as a consequence.

The rake yields a large SNR gain in the NLOS channel benchmarked against the first arriving path. This gain is close to its asymptotic value but achieved by combining only a fraction of the resolved multipaths. In scenario 2, the rake diversity gain is reduced as the bandwidth is lowered, and the optimal diversity order turns out to be quite low. The deployment of a high order rake incurs a cost

in terms of complexity but the gain is not proportionate, and there is thus little value in going beyond the figures suggested by Table 8.1.

The dispersion index in scenarios 1 and 2 is approximately equal to 2, which is the dispersion caused by the omni-directional antenna pair. Matching the template waveform with the expected antenna-dispersed waveform can thus increase the SNR by 3.3 dB in the light of the findings in Chapter 3. In terms of complexity, this does not incur a heavy burden on the system, as the template estimation is static. In scenario 3, however, the mean dispersion index is significantly higher than that caused by the antenna. Synthesising the template using this channel's dispersion index improves the SNR by 2.9 dB, which is higher than the 2.5 dB that would be achieved with the antenna-matched template. Additional complexity, however, is introduced by adaptive template generation using the channel estimate at the receiver.

The deployment of two orthogonal, linearly polarised diversity antennas at the receiver provides an additional 1.6 dB in scenarios 1 and 2. As adding the third antenna element increases the unit size and cost, in addition to increasing the signal processing overhead, it is not desirable in these scenarios. In scenario 3, however, the signal suffers greater depolarisation due to horizontal NLOS propagation, as evident from the lower value of the XPD, necessitating a full, tripolar antenna array that provides a gain of 2.7 dB.

The above comparison of the operating scenarios and the differences between the sub-systems that optimise the performance–complexity tradeoff underlines the importance of an holistic system design approach that takes into account the effect of the channel attributes on the performance of individual components.

8.8. Summary

A framework for the design of UWB communications systems optimised to channel conditions has been presented. Typical operating scenarios have illustrated the significance of channel and sub-system characterisation in predicting the performance of the system as a whole. Modular system analysis has been used to present a methodology for the design of a system optimised jointly in

performance and complexity. The strategy presented here can be extended to incorporate other signalling formats and the corresponding system components in order to arrive at the optimal design for a robust, reliable and feasible system.

9. CONCLUSION AND FUTURE WORK

This work has presented a comprehensive assessment of the UWB channel with respect to its frequency-selective properties using experimental techniques in order to build the basis for the design of a wireless communications system optimised for high performance and low complexity. The analysis is based on extensive frequency-domain measurements conducted in typical indoor propagation environments in the FCC UWB band (3.1–10.6 GHz).

From the experimental investigation, the indoor channel coherence bandwidth has been found to be 30 MHz, which is much smaller than the system bandwidth, 7.5 GHz, imparting a highly frequency-selective character to the channel. This gives rise to signal dispersion, whose statistics are useful in the prediction of fading, inter-symbol interference and maximum achievable data-rates, and have thus been characterised in this project. It has been shown that a full-band UWB system can typically resolve 80–100 multipaths, but the multipath degrees of freedom have been found to be as low as 20, providing an estimate for the required order of a sub-optimal multipath diversity receiver. The variation of the channel's signal propagation characteristics across the band has been investigated by evaluating the sub-band pathloss and delay spread, and shown to be significantly non-uniform. The comparison of multi-frequency angular power spectra has provided evidence of frequency-selective scattering in the indoor environment. The signal distortion caused by frequency-dependent pathloss and the consequent decline in system performance has been quantified. A computationally efficient, deterministic channel prediction technique has been presented which can be used to obtain a first approximation of the response of a propagation environment to arbitrary UWB signals without physical measurement.

Comprehensive statistical characterisation of the channel's polarisation behaviour has been undertaken using an elliptical basis and the Mueller-Stokes formalism. An analysis of the polarisation characteristics of the channel has shown that the UWB cross-polar discrimination (XPD) behaves in a similar manner to narrowband channels, with comparable mean XPD but a much greater spread. The investigation of the XPD's temporal and spectral evolution, however, shows significant fluctuation across these quantities, which is unique to UWB channels.

The comparison of the performance of rake receivers with a variety of configurations has demonstrated that nearly all of the incident power can be captured from less than half of the resolved multipaths when path selection is employed. As the signal bandwidth increases, the required diversity order scales with a factor of 5 fingers/GHz. Thus the interplay between the rake diversity order, bandwidth, number of incident and resolved multipaths, and the rake diversity gain has been examined, establishing the effect of each on the performance of the rake receiver.

Antenna radiation patterns are shown to vary with frequency, with the consequence that only a fraction of the signal spectrum is radiated in a given direction. Considerable signal waveform distortion is reported, that leads to reduced and often unpredictable receiver performance. A technique for removing the antenna-induced link aberration has been presented that uses the multipath direction-of-arrival information, and is useful for producing antenna-independent channel measurements.

Uncorrelated MIMO sub-channel fading has been observed with spatial or polarisation arrays in UWB channels. It has been found that the use of only two- or three-level antenna diversity can yield an outage SNR gain of up to 2.4 or 4 dB respectively, depending on the antenna arrangement in the array. An almost linear increase in system capacity has also been observed, with the added advantage that the inherent frequency diversity renders the capacity asymptotically deterministic. The results have shown that MIMO arrays can be effective in enhancing the operating range and signal quality of UWB systems. With high scalability and feasible device sizes, spatial multiplexing technique can be used with UWB signalling to design multi-gigabit wireless links, removing the fundamental limitation of throughput from radio communications systems.

A framework for the design of communications systems optimised jointly in performance and complexity has been presented that takes into account the channel conditions. Depending on the preferred transmitter and receiver architectures, this can be formulated as a multivariate optimisation problem and solved for the optimum level of multipath or antenna diversity, modulation, etc. Thus a high performance yet resource-efficient wireless communications system can be designed based directly on the channel and sub-system characterisation undertaken in this thesis.

A natural progression of this work would lead to the design of a highly adaptive and environment-aware system extending the cognitive radio architecture. This should include an investigation of the

optimal data modulation and coding formats. The interference of UWB systems with other electronic and communications equipment remains a contentious topic, and is of great interest in the context of coexistence of future wireless communications systems. While this work has focussed on single-user links, a detailed study of the efficient design of a UWB cellular architecture is also required.

The ever-growing demand for higher capacity will require the use of complex signal processing schemes for capacity enhancement. This thesis has shown the performance advantage of techniques such as multiple-antenna signalling on UWB channels. It is expected that MIMO processing will be computationally feasible for commercial units in the future with the rapid pace of advancement in device integration technology. Such systems will require compact antenna arrays for practicable unit sizes. The design of compact antennas and a comparative study of various array configurations will be valuable in that direction. This work has also highlighted the distortion caused by antennas, and designs that mitigate such effects will improve the link budget and signal quality of UWB systems.

The variability in sub-band characteristics as demonstrated in this work also makes it worthwhile to analyse the system behaviour at frequencies outside the FCC band, particularly on the higher end of the spectrum, and explore the suitability of those bands for UWB signalling. With advances in ultra-fast electronics, practical implementations of electronic circuits for wide bandwidth, millimetre-wave communications are becoming more and more feasible, and this work will therefore be of much value.

The coverage range of low-power UWB signals remains a limiting factor, and the performance significantly deteriorates beyond a few metres in accordance with the well-known range-throughput tradeoff. Design strategies to combat this constraint will be very useful. This work has presented schemes for improvement in the quality of service and range based primarily on multipath and antenna diversity. Other approaches can include intelligent network architectures, such as resilient ad-hoc networks or hybrid radio-and-fibre configurations, that would enable connectivity at longer ranges than are currently considered possible. The latter is especially interesting because of its potential to bring together two very high capacity communications techniques, namely UWB and fibre optics, with a range of novel applications such as medium-haul distribution networks and office-wide high data-rate connectivity.

The high resolution, material penetration capability and harmless nature of low-power UWB signals offer the potential to revolutionise the world of medical diagnostics through tissue imaging. It appears plausible that UWB can alter the image acquisition process fundamentally from the established realms of nuclear magnetic resonance imaging (MRI) and computer tomography, with such practical advantages as vastly reduced operating time and cost, and greater comfort for the patient. However, this is still unexplored territory and will require fundamental inter-disciplinary research involving studies of the UWB–tissue interaction, the underlying physics of the raw image acquisition, and image reconstruction enhancement algorithms.

There are thus an infinitude of perceivable applications of UWB, many of which are expected to unfold in the days to come. It is expected that this work will provide the foundation for many of those future research endeavours.

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