Space-time channel modeling for emerging wireless techniques : a system-dependent approach

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Space-Time Channel Modeling for Emerging Wireless Techniques: A System-Dependent Approach

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfilment of the requirement for the degree of
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Statement of Originality

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

05/04/2006
Date

Chen Yi Fan

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ABSTRACT

In the past decade, the wireless industry continues to revolutionize and develop new technologies. Emerging wireless systems like ultrawideband (UWB) radio and smart antennas have defied channel characterization by the paradigm of traditional approaches. This is due to the fact that the technologies either occupy a large fractional bandwidth or utilize a large array to form narrow beams, which will significantly influence the definition of the channel. The need is now greater than ever for more accurate space-time channel models with emphasis laid on the question how to integrate the system dependency into the parametric model.

In the first part of the thesis, we present a comprehensive analysis on the geometric channel models from a system-dependent viewpoint. First, a new perspective of channel characterization has been developed under the rubric of “visible channel” and “invisible channel”, which sheds some light on the system-dependent nature of multipath channels. Subsequently, a unified skeleton for geometry-based channel models, which aims at providing the underlying connection between the canonical model and the physical propagation process, has been proposed. Following from the general discussion, we probe into the multipath propagation in street-dominated environments. A parabolic scattering model that complements the widely used circular scattering model for urban macrocells has been introduced to reproduce the waveguiding effects observed in street canyons. The generic approach is also applied to explore the impact of direction-sensitive antennas on scatterer distributions in an urban macrocellular environment. Finally, by extending the unified framework to incorporate the frequency-dispersive effect of

wideband channels, a space-time model for UWB communication systems in indoor environments has been proposed. Each of the abovementioned contributions will prove valuable to high-capacity system design and performance evaluation.

The second part of the thesis ties together the main theme of the system-dependent channel-modeling philosophy, and investigates the effect of system-dependency on emerging wireless technologies. First, we present techniques to estimate the diversity (DIV) and degrees-of-freedom (DOF) gains in multiple-input-multiple-output (MIMO) channels. The new paradigm of DIV (DOF) density is used to provide a more accurate bound of the two measures. The other important aspect that we have looked into is UWB source localization in indoor clustered environments. A broadband generalized Capon estimator has been proposed for frequency-dependent multipath channels. The newly invented particle-swarm-optimization (PSO) algorithm is applied to perform the multi-dimensional search of the model parameters. To the best of our knowledge, this is the first application of the PSO algorithm in the field of signal processing research.

LIST OF ABBREVIATIONS

1G	first generation
2G	second generation
2.5G	enhanced second generation
3G	third generation
4G	fourth generation
1D	one dimensional
2D	two dimensional
3D	three dimensional
AC	angular constriction
ACA	array-constrained angular
ADPS	azimuth-delay power spectrum
AFDGM	azimuth-frequency domain geometric microcell
AoA	angle-of-arrival
AoD	angle-of-departure
APAP	average power azimuthal profile
APD	angular power distribution
AS	angle spread
BER	bit error rate
BS	base station
BW	bandwidth
CDMA	code-division multiple access
CPSO	corporative particle swarm optimization

CSM	circular scattering model
DGA	dynamic global area
DIV	diversity
DLA	dynamic local area
DOF	degrees-of-freedom
DRT	deterministic ray tracing
DS	Doppler spectrum
DUS	doubly underspread
EPAS	effective power azimuthal spectrum
EPL	effective path length
ESM	elliptical scattering model
FACF	frequency-aperture correlation function
FCC	Federal Communications Commission
FuTURE	future technologies for universal radio environment
GA	genetic algorithm
GBSBM	geometrically based single bounce macrocell
GCM	geometric channel model
GDP	geometry-dependent parameter
GSDM	Gaussian scatterer density model
HIPERLAN	high performance local area network
HPBW	half power beamwidth
i.i.d.	identically and independently distributed
ISI	intersymbol interference
ISP	intermediate scattering process

IVC	invisible channel
KP	keyhole population
LoS	line-of-sight
MB-OFDM	multiband orthogonal frequency division multiplexing
MBS	mobile broadband system
MCMB	multiple-cluster-multiple-bounce
MIMO	multiple-input-multiple-output
MPAP	mean power azimuthal profile
MPC	multipath component
MS	mobile station
NFLF	Nakagami fading loss factor
NLoS	non-line-of-sight
NSDIM	normalized space-dependent intensity measure
OFDM	orthogonal frequency division multiplexing
PAS	power azimuthal spectrum
pdf	probability density function
PDF	probability distribution function
PDS	power delay spectrum
PSC	parabolic scattering cluster
PSD	power spectrum density
PSM	parabolic scattering model
PSO	particle swarm optimization
QoS	quality of service
QSB	quasi-single-bounce

QWSSUS	quasi wide sense stationary uncorrelated scattering
Rx	receiver
SCSB	single-cluster-single-bounce
SGBS	semi-geometrically based statistical
SGBSM-AP	semi-geometrically based statistical macrocell model inclusion of antenna pattern
SISO	single-input-single-output
SNR	signal-to-noise ratio
SRC	stochastic ray tracing
STDC	spatial-temporal dynamic channel
SV	Saleh-Valenzuela
TA	transit area
ToA	time-of-arrival
Tx	transmitter
ULA	uniform linear array
UMTS	universal mobile telecommunications system
UWB	ultrawideband
VC	visible channel
WBAN	wireless body area networks
WCAN	wireless chip area networks
WLAN	wireless local area networks
WPAN	wireless personal area networks
WSS	wide sense stationary
WSSUS	wide sense stationary uncorrelated scattering

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CHAPTER 1 INTRODUCTION

1.1 Background

In 1901, Marconi received the Morse code letter “S” in Newfoundland transmitted by Fleming from England [DES36]. With this achievement, the era of long-distance wireless communication had begun. In the past century, human beings have witnessed the remarkable technology innovations in worldwide mobile cellular services, from the first generation (1G) analogue systems to the second-generation (2G) all-digital signal transmission. Currently, the third generation (3G) mobile systems are developed in the Universal Mobile Telecommunications System (UMTS) framework [CMO99, OP98, PMK00], which is expected to provide both symmetric circuit-switched services for real-time applications such as voice and video, and increasing asymmetric packet-switched services for non-real-time applications such as mobile internet access. However, there are two limitations with 3G: difficulty of extension to very high data rates such as 100Mbps/s with code-division multiple access (CDMA) due to excessive interference between services; and difficulty of providing a full range of multirate services with various quality of service (QoS) and performance requirement [YAN05]. Therefore, the fourth generation (4G) systems including the IEEE 802.11 standard at around 2.4, 5.3 and 5.8 GHz [OP98], the High Performance Local Area Network (HIPERLAN) standard at around 5.2, 5.6, and 17.2 GHz [VP00], and Mobile Broadband System (MBS) which is planned to operate at around 40 and 60 GHz frequency ranges are currently being investigated [COR97]. Advanced wideband techniques will be employed in future-generation mobile radio

networks. The next generation radio systems are intelligent and adaptable to the radio propagation environment by using advanced radio interface techniques, e.g., the Rake receiver in UMTS [OP98], Orthogonal Frequency Division Multiplexing (OFDM) [VP00] in HIPERLAN and MIMO-OFDM (multiple-input-multiple-output-OFDM) in China's Future Technologies for Universal Radio Environment (FuTURE) project [GLS05].

From the above retrospect, a more detailed exploration and characterization of the propagation are of great importance when designing radio systems. The performance of a system is highly sensitive to the characteristics of the radio propagation channel. If an accurate channel model is available, it is possible to implement particular system architecture and design appropriate algorithms that achieve a high performance by exploiting the properties of the channel. The type of channel model that is desired depends decisively on the type of system that is considered. For instance, different channel models are needed for narrowband, broadband, single antenna, and multiple antenna systems. Early channel modeling work concentrates on non-directional radio interface. This led to statistical models of the fading of the received signal [AUL79, CLA68, JAK74]. With the advent of antenna array systems, the spatial domain is being explored in more and more detail [MFG++99, KM96, SJJ++00, ECS++98, SBR02, YO02]. To make optimum use of all features of the spatial domain requires much more comprehensive understanding about the nature of wave-propagation process. The geometry-based modeling that traces individual propagation paths along the propagation environment in a semi-stochastic way turns out to be the only candidate to account for all the information offered by high-resolution spatial-temporal measurement techniques [STE01]. In the following

section, we will discuss 4 basic challenges in the emerging wireless scenarios which envisage the imperative need for improved space-time channel models.

1.2 Basic Challenges in the Emerging Wireless Scenarios

1.2.1 Higher Data Rates in a Shrinking Coverage Area

Wireless networks shrink their coverage to support an ever-increasing information flow. The traditional 1G analogue, 2G digital, and enhanced 2G (2.5G) systems support data transmission at low bit rates. In the past few years, various 3G standards like WCDMA, cdma2000, and TD-SCDMA [LXC++05] are being developed, which are optimized for transmission of packet data at high bit rates (up to 2Mbits/s) for a large geographical coverage (100 m to 1000 m). Even higher bit rates for limited coverage areas (10 m to 100 m) can be achieved with wireless local area networks (WLANs), operating at 11 Mbits/s. The wireless personal area networks (WPANs) operate at 100 Mbits/s for 10 m and the wireless body area networks (WBANs) at 600 Mbits/s for 1 m [VS02]. In recent years, researchers even looked into the wireless chip area networks (WCANs) for chip-scale wireless systems, which was able to operate at 3 Gbits/s with a low bit error rate ($\text{BER} < 10^{-3}$) over the entire chip of size $20 \times 20 \text{ mm}^2$ [ZHA04].

Since future wireless applications will operate in virtually every conceivable environment, there will be exceedingly dissimilar varieties of channels that demand much research effort on channel characterization. In fact, many of these new environments will challenge the paradigms of conventional channel models.

1.2.2 Adaptive Antennas

Adaptive arrays and other types of smart antenna techniques can lead to the following improvements in mobile communication systems [SWS++04]:

- The beamforming gain provided by the adaptive array increases the received power in both uplink and downlink, which effectively increases the coverage area of the base station. The boost in received power can also be utilized to reduce the power consumption for transmission, which enables a longer battery life at the mobile terminal.
- The angular-selective transmission and reception results in a reduction of the cochannel interference and the multipath delay spread. Many multi-user communication systems have operated below their designed capacity due to the interference-constrained operating environment. Smart antenna technology is a cheap and effective solution for mitigating the problem of cochannel interference. Furthermore, a smaller delay spread decreases the intersymbol interference and thereby eliminates the need of applying temporal equalization at the receiver.
- Multiple receive antennas help to introduce spatial diversity and thereby improve the link quality dramatically.

Obviously, employment of directional antennas at the receiver or transmitter drastically changes the channel characteristics. Therefore, channel models that were once applied to omnidirectional antennas have to be revisited and improved to account for the new spatio-temporal distortion of the channel by the antenna directivity.

1.2.3 Ultrawideband (UWB) Radio Technology

UWB technology is defined by the Federal Communications Commission (FCC) as any wireless transmission scheme that occupies a fractional bandwidth of greater than 20%, or more than 500 MHz of absolute bandwidth [RFS++04, SM04, DG04]. There are almost infinite ways in which to utilize UWB in applications, which are distributed among three categories [SM04]:

- Data communication systems – From an information-theoretic point-of-view, data rates of the order of Gbits/s are realizable over short-range WLANs in the unlicensed regime. Moreover, the signals appear to be very low-level background noise to an unintended narrowband receiver. Out of many varieties of UWB, two flavors have been developed – impulse-based UWB which uses short pulses to transmit data and Multi-band OFDM (MB-OFDM) UWB, which makes use of spread spectrum techniques to spread the information across the available frequency band.
- Position location and tracking – UWB localizer is an excellent candidate for indoor geolocation, which can be placed strategically in several wireless signposts along a trail to mark the route [LEE02, LS02]. Moreover, UWB tracker allows the precise determination of location and the tracking of moving objects within an indoor space to an accuracy of several centimeters. It can also work in complex unstructured environments with heavy traffic.
- Vehicular radar systems – Operation of vehicular radar in the frequency range 22 to 29 GHz is permitted under the UWB standardization. It can be used to avoid collision and enable a faster response to road conditions.

Apparently, channel models developed for the conventional narrowband (in RF engineering) systems are inadequate to characterize wireless systems where the signal fractional bandwidth are orders-of-magnitude greater.

1.2.4 MIMO Systems

One of the most promising future trends in wireless communications is known to be MIMO. MIMO system uses dual arrays at both ends of the radio link, which helps to achieve huge capacity and signal diversity gains by spatial multiplexing and space-time coding techniques, respectively. With MIMO it is possible to establish in parallel multiple data pipes between transmit and receive sites. Thus, the achievable capacity is a multiple of the capacity of a single-input-single-output (SISO) system. The separation of the data streams happens in the spatial domain, which is possible since there are usually many propagation paths between transmitter and receiver. The antenna array allows one to distinguish among these propagation paths at the same time and within the same frequency band. The emerging MIMO systems bring in a huge philosophical quake in how wireless communications systems are designed. Most important are the following three points:

- The multipath phenomenon is deemed not as a nuisance, but rather as a useful source to provide higher capacity and diversity gains.
- Space diversity at both ends of the radio link provides the basis for a dramatic improvement in channel capacity or spectral efficiency.
- Increasing channel capacity with MIMO is achieved by increasing computational complexity while maintaining the primary communication resources like power and bandwidth.

Early MIMO models [TEL99, FG98] assume that the fading is identically and independently distributed across all antenna pairs and show that the number of orthogonal communication modes is the minimum of the number of transmit and receive antennas. Furthermore, the maximum diversity gain is the multiplication of the number of transmit and receive antennas. However, this performance improvement is too optimistic in real-life environments. Hence, a more realistic spatio-temporal channel model is in order.

1.3 Major Contribution of the Thesis

Following from the discussion in Section 1.2, this dissertation aims to provide solutions to the abovementioned problems in a unified framework. The primary contribution of this thesis is two-fold. The first part, including Chapter 3 to Chapter 6, describes the space-time channel models for emerging wireless techniques from a system-dependent perspective. Contributions include

- proposal of a theoretical skeleton for geometric channel models, which provides the underlying connection between the canonical model and the physical wave-propagation process [CD03b, CD04a]. The generic analysis is applicable to a wide range of wireless data transmission, with various bandwidths, carrier frequencies, antenna beam patterns, or environments of operation.
- development of channel models for street-dominated propagation environments, which would be a common mobile service area for future-generation wireless applications [CD05a].
- development of channel models when direction-sensitive antennas are employed at two ends of the radio link [CD04d]

- development of space-time models for UWB communication systems in indoor environments [CD04b, CD04c]

Each of these contributions should prove valuable to high-capacity system design and performance evaluation.

The second part, including Chapter 7 and Chapter 8, discusses the influence of system-dependency of multipath channels on emerging wireless systems.

Contributions include

- formulation of degrees-of-freedom (DOF) and diversity (DIV) for a system-dependent MIMO channel [CD05b, CD05f]. We will take a microscopic view of the multiplexing and diversity gains to provide better estimate of the two measures.
- development of a broadband generalized Capon estimator for frequency-dependent UWB channels [CD05g]. The newly invented particle swarm optimization (PSO) algorithm is applied to search for the model parameters. To the best of our knowledge, this is the first application of the PSO algorithm in the field of signal processing research.

1.4 Roadmap of the Thesis

Chapter 2 starts with an introduction of existing channel models, emphasizing the system and correlation functions of wide sense stationary uncorrelated scattering (WSSUS) and quasi-WSSUS (QWSSUS) channels. We then classify the channel models into several categories and survey some of the state-of-the-art space-time models. In particular, the existing geometry-based channel models are critically

reviewed. The final section of Chapter 2 states our viewpoint of the system-independent and system-dependent modeling concepts.

Chapter 3 outlines the semi-geometrically based statistical (SGBS) model, which defines the distribution of principal scatterers contributing to the last reradiation of multipath components to the receiver. A set of heuristic rules is then proposed to provide the underlying connection between the canonical model and the propagation process which it represents. The generic analytical tool is then used to revisit several widely used geometric channel models for narrowband systems with omnidirectional antennas employed at both ends of the radio link. The last part of the chapter deals with the geometric model in a discrete Manhattan world which is applicable to signal propagation in city-street grids.

Chapter 4 proposes a parabolic scattering model (PSM) to complement the widely used circular scattering model (CSM) for street-guided mobile environments. Starting from the SGBS model, we then discuss the assumptions made in the PSM and provide the expressions for certain important channel parameters. The recently introduced angular constriction is used to provide a quantitative measure on the constriction of received signals in the azimuth domain. The model results are then compared with the measurement data available for street-dominated environments.

Chapter 5 provides a systematic investigation of the effect of antenna directivity on the spatial pattern of scatterers in a macrocell. Through the application of the SGBS model and the heuristic rules, a quantitative analysis on the geometric distribution of scatterers when antennas with different beam patterns are used at one end of the link has been provided. The usefulness of the proposed model is illustrated by applying it to study the channel characteristics in several numerical examples.

Based on the generic analysis, an azimuth-frequency domain geometric microcell (AFDGM) model for UWB transmission is established in **Chapter 6**. Theoretical derivations of power azimuthal spectrum and power delay spectrum are presented. It is shown that the channel properties derived using the proposed model exhibit a standard deviation of ray arrivals in the azimuth domain and a rate of decay in the delay domain that are comparable to the empirical results reported in the literature.

Chapter 7 presents the DIV and DOF calculation for system-dependent wideband directional channels. Starting with a degenerate single-cluster-single-bounce channel and perfect receiving conditions, the analysis is extended to more general scattering and receiver-operating scenarios. The utility of the proposed methodology is demonstrated by presenting various numerical examples to provide insight into the impact of signal power spectrum on the DIV and DOF estimates.

Chapter 8 extends the simple yet accurate generalized-Capon-estimator for narrowband stationary signals [HSG04] to UWB scenarios based on a realistic signal/channel model in the angular-frequency domain. The particle swarm optimizer [KES01], which is a robust stochastic evolutionary computation technique and is new to the signal processing community, is then implemented to perform the multidimensional search for the parametric setting of the underlying channel model.

Chapter 9 highlights the key issues of this thesis and proposes some relevant future work.

Notations: The following notations will be used throughout this dissertation. We will use boldface upper case letters for matrices and lower case letters for vectors. $\mathbf{I}$ is the identity matrix. For a set S , $|S|$ denotes its Lebesgue measure. $[\cdot]^*$, $[\cdot]^T$ and $[\cdot]^H$

denote complex conjugate, transpose and Hermitian operators, respectively. $E\{\cdot\}$ is the expectation operator.

CHAPTER 2 THE MOBILE RADIO CHANNEL

MODELING – LITERATURE REVIEW AND SYSTEM-INDEPENDENT AND SYSTEM-DEPENDENT VIEWPOINTS

2.1 Introduction

Fig. 2.1. A definition of channels in generic wireless systems

Multipath channel modeling is the most important and fundamental amongst all the research activities on wireless systems that have taken place in the past years. The term “channel” takes on different meanings in the scientific literature depending on the relevant technical issues [VA04]. For example, the physical propagation paths that result from multipath environment are called the (electromagnetic) *propagation channel*. The signal received by a loaded antenna system is referred to as the output from the *signal channel*, which comprises the functional blocks of the propagation

channel, the effects of the interaction of the antenna and the field, and the antenna itself. Throughout the stages of downconversion, demodulation and decoding (baseband processing), different types of channels can be defined like *baseband radio channel*, *digital channel*, *raw data channel*, etc [VA04]. A pictorial illustration of the channel definitions is shown in Fig. 2.1. In this thesis we will consider the antenna and its associated signal processing and front-end circuitry as “system”, which can effectively alter the antenna interaction with the free space waves.

The wireless channel is the primary contributor to many of the problems and limitations that overwhelm wireless systems. One obvious example is multipath propagation which is a major characteristic of mobile radio channels. It is caused by reflection, diffraction and scattering from terrain features and buildings, which leads to distortion in analogue communication systems and affects the performance of digital systems by reducing the signal-to-noise and signal-to-interference ratios. A physical understanding and consequent mathematical modeling of the channel is very important because it facilitates a better prediction of system performance and provides the mechanism to evaluate methods for mitigating deleterious effects caused by the harsh environment. However, owing to the appearance of newer wireless networks which shrink their coverage and require broader bandwidth and larger aperture size to support high data rate transmission as described in Chapter 1, new “mind's eye” of the wireless channel becomes mandatory. (Arguably, any modeling activities are simply “inventions of the human mind” as described by the quotation from Papoulis' now classic book [PP02]:

“...if we agree, as we must, that scientific theories are not discoveries of the laws of nature but rather inventions of the human mind.”)

2.2 Existing Channel Models – Scanning the Literature

Channel modeling activities usually involve two types of propagation prediction corresponding to the large-scale path loss information and small-scale fading statistics. The path loss is critical for wireless network coverage planning and any other applications relevant to the received signal strength. The small-scale parameters provide information on local field variations and thereby help to improve receiver design and combat the multipath fading. In the following, we will focus on the small-scale modeling theories. The interested readers can read review papers [KCW93, BHM++94, HAS93, MOL91, IY02, ECS++98, YO02, MFG++99, SBR02] and books [KEN68, VA04, BLA00, BA02, BER00, DUR03, RAP02, PAR00, SAU99] to complement the brief review in this section.

2.2.1 System and Correlation Functions

Conventionally, time-variant non-directional channels can be described by one of the four system functions first presented by Bello in a complete and rigorous formulation [BEL63]. The most popular system function in context with mobile radio communications is the time-variant impulse response $h(t, \tau)$. The other system functions, including the time-variant transfer function $T(t, f)$, the Doppler-variant impulse response also called spread function $s(\nu, \tau)$, and the Doppler-variant transfer function $H(\nu, f)$, can be derived by Fourier transforms with respect to either the time t or the delay τ . The cyclic Fourier relationship is shown in Fig. 2.2(a).

The four system functions can only be fully described by the multidimensional probability density function (pdf), which is too complex to use in practice. By

Fig. 2.2. (a) System and (b) correlation functions for time-variant non-directional channels; (c) system and (d) correlation functions for time-variant directional channels

assuming the wide sense stationary uncorrelated scattering (WSSUS) random process which is equivalent to wide sense stationary (WSS) in both time t and frequency f , the analysis is greatly simplified. Due to this stationarity the dependence of the correlation functions on four variables in the general case reduces to a dependence on two variables, i.e., the time difference Δt and the frequency difference Δf . Therefore we end up with

$$R_h(t_1, t_2, \tau_1, \tau_2) = E\{h(t_1, \tau_1)h^*(t_2, \tau_2)\} \quad (2.1.1)$$

$$\stackrel{\text{WSSUS}}{\equiv} E\{h(t_1, \tau_1)h^*(t_1 + \Delta t, \tau_2)\} = P_h(\Delta t, \tau)\delta(\tau_2 - \tau_1)$$

$$R_s(\nu_1, \nu_2, \tau_1, \tau_2) = E\{s(\nu_1, \tau_1)s^*(\nu_2, \tau_2)\} \stackrel{\text{WSSUS}}{\equiv} C_s(\nu, \tau)\delta(\nu_2 - \nu_1)\delta(\tau_2 - \tau_1) \quad (2.1.2)$$

$$\begin{aligned}
 R_T(t_1, t_2, f_1, f_2) &= E\{T(t_1, f_1)T^*(t_2, f_2)\} \\
 &\stackrel{\text{WSSUS}}{=} E\{T(t_1, f_1)T^*(t_1 + \Delta t, f_1 + \Delta f)\} = R_T(\Delta t, \Delta f)
 \end{aligned}
 \tag{2.1.3}$$

$$\begin{aligned}
 R_H(\nu_1, \nu_2, f_1, f_2) &= E\{H(\nu_1, f_1)H^*(\nu_2, f_2)\} \\
 &\stackrel{\text{WSSUS}}{=} E\{H(\nu_1, f_1)H^*(\nu_2, f_1 + \Delta f)\} = P_H(\nu, \Delta f)\delta(\nu_2 - \nu_1)
 \end{aligned}
 \tag{2.1.4}$$

The corresponding cyclic Fourier relationship is illustrated in Fig. 2.2(b).

The non-directional system and correlation functions have been applied successfully until quite recently, where the advent of antenna array systems for both interference cancellation, capacity improvement, and position location has motivated the exploration of the spatial domain of the wireless channel [ECS++98, YO02, VA04, MFG++99, SBR02]. A formal extension of Bello's system functions to directional time-variant channels is usually done by introducing an additional dependence on the direction φ of the incoming or outgoing multipath component (MPC), i.e.,

$$h(t, \tau) \rightarrow h(t, \tau, \varphi)$$

By extending to the 3-dimensional propagation and taking into account of both transmit and receive sites, a 6-dimensional time-variant impulse response is resulted:

$$h(t, \tau) \rightarrow h(t, \tau, \varphi_T, \varphi_R, \theta_T, \theta_R)$$

where the subscripts T and R denote the transmitter (Tx) and receiver (Rx), respectively. In this thesis, we consider the 2-dimensional (2D) scenario where only the azimuth at both ends of the radio link is taken into account. Similar with the non-directional case, other system functions can be obtained by Fourier transforms with respect to the time t , the delay τ , or the angle φ . In analogy to the Fourier relationships between time t and Doppler shift ν as well as between frequency f and

delay τ , a Fourier transform with respect to $\sin\phi$ leads to system functions in the one-dimensional aperture domain a [KAT02, VAU00, FLE00]. To simplify the expression, Herdin considered the non-linear but monotonic transform between the aperture a and the angle-of-arrival (AoA) or angle-of-departure (AoD), ϕ [HER04]. In total, 8 possible system functions are defined:

- $q(v, \tau, a)$: Doppler-aperture-variant impulse response
- $H(v, f, a)$: Doppler-aperture-variant transfer function
- $s(v, \tau, \phi)$: Doppler-variant angle-resolved impulse response
- $G(v, f, \phi)$: Doppler-variant angle-resolved transfer function
- $g(t, \tau, a)$: Time-aperture-variant impulse response
- $T(t, f, a)$: Time-aperture-variant transfer function
- $h(t, \tau, \phi)$: Time-variant angle-resolved impulse response
- $Q(t, f, \phi)$: Time-variant angle-resolved transfer function

The corresponding 8 correlation functions can be calculated by ensemble average in a similar manner with the equations from (2.1.1) to (2.1.4). A pictorial illustration is shown in Fig. 2.2(c) and 2.2(d). Note that in such a case a WSSUS directional channel shows wide sense stationary with respect to t, f , and a .

In the real environment, the WSSUS condition is not strictly fulfilled since the channel has to remain unchanged over infinite time, frequency, and aperture intervals. Usually a quasi-WSSUS (QWSSUS) assumption is imposed if the channel is approximately stationary within a specific time, frequency and aperture observation window. A more general characterization for non-WSSUS non-directional channels was first introduced by Matz [MAT03a, MAT03b] and

subsequently extended to the spatial domain by Herdin [HER04]. Instead of the WSSUS scattering function $C_s(v, \tau, \varphi)$, a local scattering function is defined

$$C_s(v, \tau, \varphi, f, t, a) \triangleq \iiint R_T(\Delta t, \Delta f, \Delta a, f, t, a) e^{-j2\pi(-v\Delta f + \tau\Delta f + \sin\varphi\Delta a/\lambda)} d\Delta f d\Delta t d\Delta a \quad (2.2)$$

This definition is consistent with the directional-WSSUS case, for which $C_s(v, \tau, \varphi, f, t, a)$ reduces to $C_s(v, \tau, \varphi)$. Under the assumption of doubly underspread (DUS) channels (refer to [MAT03b] for a formal definition of DUS channels), $C_s(v, \tau, \varphi, f, t, a)$ represents the mean power of those signal components that have a Doppler/delay/angle shift of (v, τ, φ) at frequency f , time t , and aperture a . Subsequently, a novel channel correlation function is formulated as

$$R_T(\Delta v, \Delta \tau, \Delta \varphi, \Delta f, \Delta t, \Delta a) \triangleq \iiint R_T(\Delta t, \Delta f, \Delta a, f, t, a) e^{-j2\pi(-f\Delta \tau + \tau\Delta f + \Delta v + a/\lambda \sin\varphi\Delta \varphi)} df dt d\varphi \quad (2.3)$$

which is the correlation function in all 6 dimensions. More details on non-WSSUS wireless channels can be found in [MAT03a, MAT03b]. In this thesis, we will only focus on the WSSUS channel and leave the more general scenarios for future investigation.

We can now grossly characterize channels by first extending the measures introduced by Lerner [LER59, KEN68] to the space domain: Doppler spread σ_v , delay spread σ_τ , angle spread σ_φ , and total spread σ . Bearing in the mind that the exact definition is relatively unimportant, we adopt the following formulations by extending the definitions in [KEN68]

$$\sigma_v = \left[\int s_v^2(v) dv \right]^{-1} \quad (2.4.1)$$

$$\sigma_\tau = \left[\int s_\tau^2(\tau) d\tau \right]^{-1} \quad (2.4.2)$$

$$\sigma_{\varphi} = \left[\int s_{\varphi}^2(\varphi) d\varphi \right]^{-1} \quad (2.4.3)^1$$

and

$$\sigma = \left[\iiint s^2(\nu, \tau, \varphi) d\nu d\tau d\varphi \right]^{-1} \quad (2.4.4)$$

where

$$s_{\nu}(\nu) = \iint s(\nu, \tau, \varphi) d\tau d\varphi$$

$$s_{\tau}(\tau) = \iint s(\nu, \tau, \varphi) d\nu d\varphi$$

$$s_{\varphi}(\varphi) = \iint s(\nu, \tau, \varphi) d\nu d\tau$$

and $s(\nu, \tau, \varphi)$ is the Doppler-variant angle-resolved impulse response. The quantity σ_{ν} provides a rough measure of the frequency interval over which the Doppler shifts of the scatterers are spread, whereas σ_{τ} provides a measure of the interval over which their path delays are spread. Finally, σ_{φ} quantifies to which extent multipaths concentrate about a single incoming direction. Also note that the four-parameter description $(\sigma_{\nu}, \sigma_{\tau}, \sigma_{\varphi}, \sigma)$ tends to become useless if the scattering function consists of several disjoint pieces. Moreover these definitions may not reflect some crucial differences that can exist between different scattering functions even when the function is composed of a single piece.

Next, the signal duration T_s , bandwidth W and angular concentration γ could be defined by consistently extending the expressions in [KEN68] to the space domain

$$T_s = \left[\int |u(t)|^4 dt \right]^{-1} \quad (2.5.1)$$

¹ Strictly speaking, the integral kernel should be with respect to $\sin\varphi$ rather than φ . However, we will assume that the difference caused by the change of argument is relatively insignificant.

Table 2.1. Waveform characteristics for doubly dispersive channels [KEN68]

$\sigma_v T_s$	$W \sigma_r$	σ_v / W	σ_r / T_s	Distorted	Dispersed in		Remarks
					Time	Frequency	
$\ll 1$	$\ll 1$	$\ll 1$	$\ll 1$	No	No	No	Implies $\sigma_v, \sigma_r \ll 1$
$\ll 1$	$\gg 1$	$\ll 1$	$\ll 1$	Yes	No	No	Implies $\sigma_v, \sigma_r \ll 1$, and $T_s W \gg 1$
$\ll 1$	$\gg 1$	$\ll 1$	$\gg 1$	Yes	Yes	No	
$\gg 1$	$\ll 1$	$\ll 1$	$\ll 1$	Yes	No	No	Implies $\sigma_v, \sigma_r \ll 1$, and $T_s W \gg 1$
$\gg 1$	$\gg 1$	$\ll 1$	$\ll 1$	Yes	No	No	Implies $T_s W \gg 1$
$\gg 1$	$\gg 1$	$\ll 1$	$\gg 1$	Yes	Yes	No	Implies $\sigma_v, \sigma_r \gg 1$, and $T_s W \gg 1$
$\gg 1$	$\ll 1$	$\gg 1$	$\ll 1$	Yes	No	Yes	
$\gg 1$	$\gg 1$	$\gg 1$	$\ll 1$	Yes	No	Yes	Implies $\sigma_v, \sigma_r \gg 1$, and $T_s W \gg 1$
$\gg 1$	$\gg 1$	$\gg 1$	$\gg 1$	Yes	Yes	Yes	Implies $T_s W \gg 1$

$$W = \left[\int |U(f)|^4 df \right]^{-1} \tag{2.5.2}$$

and

$$\gamma = \left[\int |u(\varphi)|^4 d\varphi \right]^{-1} \tag{2.5.3}$$

where $U(f)$ is a Fourier transform of the signal $u(t)$, and

$$u(t) = \int u(t, \varphi) d\varphi$$

$$u(\varphi) = \int u(t, \varphi) dt$$

For segmented waveforms (2.5.1) to (2.5.3) tend to reflect time (frequency, direction) over which the waveform is nonzero rather than the time (frequency, direction) from the beginning to the end of the waveform. In almost all realistic situations, none of the frequency dispersion σ_v , the time dispersion σ_r , and the direction dispersion σ_φ vanishes. However, the system may behave as though one or all of them are zero. That is, the effects of one or all of these dispersions are indiscernible in the received waveform, which depends upon both the nature of the transmitted waveform and the channel scattering function. Table 2.1 gives a gross summary on the classification of

Doubly dispersive channel by only considering the time and frequency domains [KEN68]. It can be easily extended to include the space domain by taking into consideration the σ_φ and γ measures.

Following from the above discussions, many commonly used small-scale channel models such as Rayleigh fading envelope model, Clarke/Jakes model, two-ray model are simply a first approximation to the various situations as listed in Table 2.1. The underlying thought is that it might be instructive to distinguish the real-life channel from the canonical channel, where the former refers to a complete picture of the propagation environment whereas the latter refers to a simplification of the former by ignoring the imperceptible details caused by the limited resolving ability of the systems.

2.2.2 Model Classification

The azimuth-delay power spectrum (ADPS) $P_h(\tau, \varphi)$ [COR01] and frequency-aperture correlation function (FACF) $R_T(\Delta f, \Delta a)$ are mostly studied in context with directional WSSUS channels. The two functions can be interpreted as a kind of temporally averaged delay-angle cross power spectrum density (PSD), $P_h(\Delta t, \tau, \varphi)$, and time-frequency-aperture correlation function, $R_T(\Delta t, \Delta f, \Delta a)$, respectively (cf. Fig. 2.2(d)). This kind of description contains sufficient information for study of space-time channels, which are of special interest in most of the existing systems. These functions may be explicitly derived either by physically based or non-physically based models.

Physically based modeling is one of the mostly applied approaches for predicting radio channels used in the deployment tools. It treats the electromagnetic waves as

traveling along localized ray paths. The paths themselves and the associated fields can be traced by sequentially accounting for the local features of the medium as they are encountered by the rays. Physically based models can be either completely deterministic such as the site-specific ray tracing techniques [PL95] or stochastic such as the geometrically based statistical models (e.g., [COR01]). The non-physically based models rely on the channel statistical information using non-physical parameters (e.g., [SJJ++00]), which are generally easy to simulate and provide accurate channel characterization for the situation under which they were identified.

2.2.3 Physically Based Channel Models

2.2.3.1 Deterministic Ray Tracing Models

One of the main approaches is the deterministic ray tracing (DRT) [SR94, LB98, KGW++99, IY02]. It assumes that the locations and reflection properties of all the important objects inhibiting propagation are accurately known. Then all the possible paths connecting the Tx and the Rx are found through the specified complicated environment. The basic method is the ray-launching algorithm as illustrated in Fig. 2.3(a) [SR94, LCL89, CJ97, YWK98, SM97, DPR97, CPD++98]. First, numerous test rays are launched from the Tx. The algorithm then searches through all the rays to find out which pass closest to the Rx location. Subsequently, both large-scale and small-scale channel information can be calculated from the traced paths using geometrical optics. The implementation of ray-launching schemes has three fundamental steps: 1) launch test rays, which can be ray tubes or ray cones [SR94, YWK98], 2) find rays received by the receiving antenna by applying the reception

Fig. 2.3. Pictorial illustration of (a) ray-launching algorithm and (b) image method

test [SR94, DPR97], and 3) obtain channel parameters from these paths. By using spatial discrimination techniques such as bounding volume hierarchies, the calculation time exhibits a logarithmic dependence on the number of surfaces in the database [DPR97].

A second approach is the image method, which is a simple yet accurate method for determining the ray paths between the transmitting and receiving antennas. The method of images uses a concept similar to image theory to place artificial sources in the environment that model reflections from the smooth planes in an environmental database (Fig. 2.3(b)) [TT96, FPY96, LM94, ANM00, AN00]. The image method is accurate but suffers from the inefficiency when the number of reflective surfaces is large. Every additional wall doubles the number of images available to the Rx. For realistic implementations, image-based ray tracing is usually accelerated by purging the branches of a visibility graph corresponding to extra sources that do not illuminate other reflective surfaces [LM94, ANM00, AN00].

In recent years, extensive research work has been carried out to reduce the computational load of the ray-tracing algorithm. The acceleration is achieved by

reducing the number of objects on which actual ray-object intersection will perform (e.g., the angular Z-buffer [CPD++98] and the ray-path search algorithms [AFR++00] for the ray-launching method; “illumination zone” concepts proposed in [LM94, AN00, ANM00] for the image method). Another way to improve the efficiency is to accelerate the intersection test (e.g., the 2-D/two-and-one-half dimensional method proposed by Rizk *et al.* [RWG97] and the vertical plane launch method in [LB98]). A comprehensive overview of the state-of-the-art DRT models is given in [IY02].

2.2.3.2 Geometric Channel Models

Essentially, the site-specific floor-plan database and the information about the building materials in a DRT model can be regarded as a particular realization of the stochastic environment, which denotes an ensemble of typical propagation scenarios dictated by the morphological characteristics of sites like Generalized Typical Urban, Generalized Bad Urban, and Generalized Rural Area for outdoor macrocell; Generalized Street Canyon Line-of-Sight (LoS), Generalized Street Non-Line-of-Sight (NLoS), and Generalized Street Crossing for outdoor microcell; Generalized Small Office LoS, Generalized small office NLoS, and Generalized Factory/Hall for indoor picocell [COR01, STE01]. An alternative approach is the stochastic ray tracing (SRC) and represents a semi-deterministic modeling of the environment, which provides a universal characterization of the stochastic environment.

One of the mostly used SRC models is the geometric channel model (GCM). GCM assumes certain distribution of scatterers around the two ends of the radio connection and then derives the azimuth-delay power spectrum by applying the

fundamental laws of specular reflection, diffraction, and scattering (see Fig. 2.4). Various geometric models have been proposed with each corresponding to a specific environment type. Circular scattering model (CSM) [FMB98, JAN02, PRR02, MHS02, MOL04] has been suggested for macrocell type of environments where the base station (BS) antenna array is elevated above rooftop level of the surrounding buildings. The scatterer distribution can take various forms of linear [PRR02], one-sided Gaussian [MOL04, CD04a], hyperbolic [MHS02], and Rayleigh [JAN02]. On the other hand, elliptical scattering model (ESM) [ER99, LI95] is suitable for micro- or pico-cell type of environments, in which the BS antenna is placed at a relatively low height and the geometry does not support the notion that all significant scatterers are clustered around one end of the link. In this case a reciprocal spatial distribution would be an appropriate assumption. While all of the above models are 2D in nature, a quasi 3D model has been developed in [NA98], in which the scatterers are distributed on ellipsoids and only the intersection between the ellipsoid and the ground is being used for analysis. In addition, the parameters of the model could be adjusted using measured data as illustrated in [NA98]. Most of the existing GCMs take into account only the local scattering cluster [PRR02, NA94, JAN02], which is always located around the mobile unit with few available models defining the shape and distribution of far clusters [COR01, 3GPP/LUC, 3GPP/IMS]. In street-dominated environments where the BS antenna is elevated above rooftop level and the mobile station (MS) is located in the street canyon. In such a case, almost all the long-delayed paths travel along the directions of the streets where the MS is located [KRB00, FRB97, COX73, KLV++03, LKT++02]. Such a waveguiding effect cannot be properly reproduced using the CSM. Kuchar *et al.* [KRB00] proposed two

Fig. 2.4. Geometric channel-modeling method

superimposed scatterer distributions in street canyons: a circular disk for the short-delayed components and a rectangular distribution for long-delayed waves. However, the shape and scatterer density of the scattering areas have not been clearly specified in [KRB00].

Generally, the aforementioned GCMs have two main drawbacks. First, only a single specular reflection is accounted for, and none of the scattering, diffraction and multiple-bouncing is considered. This hypothesis may not be realistic as has been found in many measurement campaigns [KRB00, FRB97, KLV++03]. Multiple scattering can be taken into account by either assuming an equivalent single-bounce model or by placing a scattering object at the last reradiation and approximating the preceding scattering as a stochastic process [CD04a, CD05a]. Second, the geometry of scattering areas in the existing GCMs is obtained on a rather undefined plane and is thus questionable. In the analysis of a GCM, one subject is often overlooked by researchers – the actual constitution of a realistic GCM. The axioms corresponding to the *a priori* properties of the channel that is modeled using geometric approach must represent realistic wave propagation. GCM is one type of geometric model, which simulates a process or phenomenon that has a geometric analog [MOT97]. In GCM, a

shape (spatial pattern of scatterers) is defined by a set of logical relationships satisfying a set of axioms (wave propagation obeying diffraction, reflection and scattering laws), which are in turn interpreted as true statements about the model. Subsequently, one can safely infer the process behavior from the model behavior if and only if the model closely corresponds to the process. However, the shape and scatterer density of the scattering areas in the existing GCMs have been defined with a certain level of ambiguity, with neither rigorous mathematical statement nor sound physical reasoning available to support the validity of the various models. A first attempt to circumvent these two problems has been presented recently [CD04a, CD05a].

In recent years, there has been much research effort on the investigation of direction-sensitive smart antenna deployment in broadband systems [MFG++99, ECS++98, EGG99]. It is conceptually well known that the directivity of transmit and receive antennas has a significant influence not only on receive power but also on angular spread of the multipath propagation channel. Narrow adaptive pencil beams also help to overcome the equalization problem. Hence, knowledge of the effects of antenna directivity on the channel characteristics is crucial in improving system performance. When GCMs are used to determine the channel properties, the antenna effect is accounted for by simply multiplying the angle-resolved impulse response with the beam pattern [PRR02, AJ93, FMB98]. However, due to the finite sensitivity of the measurement device, the effect of directional antenna is actually twofold [CD04d]. Firstly it alters the spatial distribution of scatterers by providing a different sample space for the random field, and secondly, it collects the signal components from the AoAs by weighted combination. As such, it will be more appropriate to

present an antenna-dependent channel model by integrating the antenna effect into the model geometry [CD04d, ZN02].

The GCM can also be applied to study the channel characteristics of an Ultrawideband (UWB) system [CD04c]. UWB signal is one that occupies a large fractional bandwidth (its transmission bandwidth divided by its center frequency) over 20%, or more than 500 MHz of absolute bandwidth. The existing indoor channel models mainly focus on systems with bandwidths much narrower than those for UWB [HAS93, BMS89, RST91, HP92]. Only in the last few years have UWB measurement campaigns been performed (e.g., [CWM02, GJR++04]). When an UWB pulse reflects off an object or penetrates through materials in the process of multipath propagation, the effects are frequency sensitive and therefore the waveform is distorted in some way. Now the receiving waveforms associated with a propagation pathway are dependent on that path. Each path will have its own impulse response or frequency transfer characteristic [QIU02, QIU96]. This is the essential difference between a narrowband and UWB multipath channel. On one hand, for UWB signals, various diffractions and reflections pertain to different frequency-dependent behaviors. For example, consider a UWB system operating at a frequency of 5 GHz with a bandwidth (BW) of 2.5 GHz [PCR++02]. In this frequency band, the difference in scattering properties is significant for a sphere/cylinder of radius 0.1 m [SVA01]. On the other hand, study of propagation through different materials also capitalizes the statement of spectrum dependency. For example, cloth partition shows higher loss due to support elements inside the partitions as reported in [MUQ03]. Attenuation constants for the door, wood, and structure wood sample are also presented in [MUQ03], which sheds more light on the effects of various materials on

the filtering of UWB pulses. To the best of our knowledge, the first attempt to incorporate the frequency-dependent feature of rays to the GCM is given in [CD04c].

2.2.3.3 Other Stochastic Ray Tracing Models

A range of models in which different levels of tradeoff between the complexity of the building profile and the accuracy of the prediction have been developed. An extreme case would be the percolation model by Franceschetti *et al.* [FMP99, MPF99], which proposes to model the environment as a homogeneous area with uniformly distributed obstacles of given density. They consider a random walk formulation of the problem of optical ray penetration inside the medium as illustrated in Fig. 2.5(a). This model has been refined to include signal attenuation and non-Manhattan-type grid (obstacle) pattern in [FBS04, FRA04], which is applicable to microcellular environments with a large number of small scattering objects (see Fig. 2.5(b)). This school of thought is to develop statistical models characterized by the fewest number of parameters that can still give some global information of the field properties. Along the similar line of thought, another SRT model is proposed by Hassan-Ali and Pahlavan which requires three site-specific parameters to generate the channel: mean free distance, transmission coefficient, and reflection coefficient [HP02]. A more comprehensive framework ought to be the one by Ullmo and Baranger [UB99], where the scattering by walls and local clutter is considered through either a relaxation-time approximation in a Boltzmann equation or by using a diffusion equation. They include partial information about the building (e.g., its geometry and the typical wall separation) and treat the unknown aspects using a probabilistic model for the scattering.

Fig. 2.5. Illustration of (a) a ray impinging on a random lattice and (b) a random-walk photon propagating in a cluttered environment

2.2.4 Non-Physically Based Channel Models

One of the mostly used non-physically based models ought to be the Saleh-Valenzuela (SV) model [SV87, MFP03, IEEE802/02/490]. The time-only impulse response is given by

$$h(t) = \sum_{l=0}^{\infty} \sum_{k=0}^{\infty} \alpha_{kl} e^{j\phi_{kl}} \delta(t - T_l - \tau_{kl}) \quad (2.6)$$

where the sum over l corresponds to the scattering clusters and the sum over k corresponds to the rays within a cluster. T_l and τ_{kl} are the cluster and ray delays. The expression of (2.6) is derived based on the clustering phenomenon observed in the experimental data [SV87]. The cluster amplitude α_{kl} is Rayleigh distributed with

Fig. 2.6. Illustration of the clustered channel impulse response in the delay and azimuth domains

$$E\{\alpha_{kl}^2\} = E\{\alpha_{00}^2\} \exp(-T_l/\Gamma) \exp(-\tau_{kl}/\gamma) \quad (2.7)$$

where the shape parameters Γ and γ are the inter- and intra-cluster decay constants, respectively. By including the statistics of AoA and AoD, the SV model has been extended to the space domain in [SJJ++00]. The combined spatio-temporal channel impulse response is taken to be

$$h(t, \varphi) = \sum_{l=0}^{\infty} \sum_{k=0}^{\infty} \beta_{kl} e^{j\phi_{kl}} \delta(t - T_l - \tau_{kl}) \delta(\varphi - \Phi_l - \phi_{kl}) \quad (2.8)$$

where Φ_l and ϕ_{kl} are the cluster and the ray angles. A pictorial illustration of the clustering phenomenon in the channel impulse response is shown in Fig. 2.6. The non-physically based channel models assume certain distributions of power azimuthal spectrum (PAS) and power delay spectrum (PDS). The most common

shapes suggested for the PAS include uniform [ECS++98], Gaussian [BD91, ECS++98, MAR98], Laplacian [POO04, PMF00], and von Mises [ABK02] depending on the propagation environments. On the other hand, the PDS is usually assumed to be an exponentially decaying function. Other non-physically based models provide for the performance study of various cellular systems by deriving the joint distribution of path gain, delay spread, angle spread, and shadow fading [GEY++97, APM02].

2.3 System-Independent and System-Dependent Channel Modeling Viewpoints

2.3.1 General Remarks

For a system-dependent characterization of mobile radio channels, the so-called double-directional channel concept is proposed [SMB01, ZHR++04]. In order to separate the influence of Tx/Rx sites and the propagation channel itself, it is a versatile practice to distinguish between the *radio channel* and the *propagation channel*. The former is described by the non-directional channel response whereas the latter by the double-directional channel response excluding both the Tx and Rx antennas (Fig. 2.7(a)). In between is the so-called single-directional channel with the introduction of directionality at the Rx site.

With arbitrary inclusion of system-specific information such as antenna beam patterns and operating-frequency, the system-independent approach features an improved scatterer identification which offers the following advantages [SMB01]

- insertion of any desired system configuration (e.g., antenna patterns)

Fig. 2.7. Illustration of (a) system-independent and (b) system-dependent channel modeling concepts

- evaluation of path dispersion and dissimilarities [SÖH++02]
- a generic channel model for multiple-input-multiple-output (MIMO) systems that enables the signal-processing optimization process at both ends of the radio link.

In general, the double-directional channel models separate three functional blocks in the radio transmission chain:

- Transmitting block, which operates at a specific frequency band and distributes the signal into the desired AoD. These AoDs can be selected from the group of input directions provided by the double-directional channel measurements.
- Propagation channel, which includes all the resolvable propagation paths between the transmit and receive sites. Each path is fully characterized by its impulse response including the excess delay, the proper complex amplitude, and the associated AoA and AoD. In general, all the parameters will also depend on the absolute time, and the set of MPCs may also vary with time.
- Receiving block, which collects the signal from the various AoAs by weighted combination and filters the signal by a bandpass filter.

The perspective of system-dependent channel characterization is completely different from the abovementioned philosophy. Referring to Fig. 2.7(b), the transmitter behaves like a bulb which probes the environment by illuminating the lights. The receiver behaves like the eyes of an observer who is trying to obtain any useful information from the transmitter by receiving the lights. Depending on the *brightness* of the source and the *eyesight*, only certain environmental details are discernible to the eyes. Therefore, the observer will interpret the environment as what he has seen, which also conveys the meaningful information to him. In a similar manner, in system-dependent channel characterization, we distinguish between the visible channel (VC) and the invisible channel (IVC). The former represents a pool of meaningful information, from which the useful signal can be extracted for further processing given the specified system sensitivity and the radio channel structure.

Surrounding the VC, there is a large invisible region that extends to infinity. Apparently, the boundary line of the VC is dependent on the received signal properties such as signal strength, which in turn is together determined by the system properties and the multipath compositions. Also note that it is usually unfeasible to define the boundary line of the VC in a deterministic manner. A more realistic approach would be defining the VC boundary probabilistically.

In general, a thorough description of the propagation environment is almost impossible due to the sophisticated channel structure. There are in effect an infinite number of paths that could be enumerated in an ideal double-directional channel, which should be system-independent according to its original definition. However, certain system-dependency has been introduced in the actual implementation of double-directional channel models by extracting the “effective multipath” [STE01]. A path is considered to be effective only if it carries a significant percentage of the total power and it is separable in the angle and delay domains, both of which are determined by the system sensitivity and the useful information that is required for a certain application (e.g., MIMO capacity computation, position location, etc). From this viewpoint, channel shall not be defined before the properties of the transmit and receive systems have been specified. A study case will be presented in Appendix A to further exemplify how the system sensitivity may affect the definition of the channel.

2.3.2 Influence of System-Dependency of Multipath Channels on Some of the Emerging Wireless Systems

2.3.2.1 Degrees-of-freedom and Diversity of MIMO Channels

In recent years, considerable attention has been drawn to MIMO communication techniques due to the prospect of significant improvement of system performance [PNG03, LS03]. Relevant literature has discussed the performance improvement from two different directions. One set of techniques called as space-time coding, improves the link quality by utilizing the diversity gains offered by dual antenna arrays [TSC98, ALA98]. Whereas the other scheme called as spatial multiplexing, utilizes the parallel spatial sub-channels by exploiting channel information at the receiver [FG98]. Tradeoff between the two approaches has been analyzed from information-theoretic [ZT03, TVZ04, GH02] and system performance [CGE03] point-of-view recently. In the existing literature, diversity (DIV) [TSC98] and degrees-of-freedom (DOF) [ZT03, TVZ04, GH02] have been widely used as the simple measures of performance gains obtained from using MIMO systems. DIV is commonly defined as the negative exponent of signal-to-noise-ratio (SNR) in the probability of error expression and DOF is referred to as the coefficient of $\log \text{SNR}$ occurring in the expression for capacity, both in the high-SNR region. To shed a light on the nature of the two measures, DIV can be thought of as the redundancy of the transmitted signal in a particular communication system whereas DOF is considered as the number of independent spatial-temporal channels available for communication.

Nonetheless, little work has been reported incorporating the theoretical framework with real-life channel conditions to obtain a mathematical model that allows system designers to develop a deeper understanding on these two schemes. In the past, a few models have been proposed to relate the spatial DOF to the array geometry and the scattering field patterns [SAY02, RC98, GBG++02, BGP02, BUR03]. However, in most of the models the propagation channel is characterized by

a set of system-independent multipath/cluster parameters. This separation hypothesis seems implausible in real environments such as the measurement campaigns performed by Poon *et al.* [PH03] and Beach *et al.* [BAK04]. The empirical results in the former showed that the number of clusters and their subtended angular intervals decreased with increasing operating frequency. This observation was later incorporated into the mathematical model developed in [PBT05] for DOF calculation. In [BAK04], it was shown that significant spatial decorrelation could occur in urban deployments for carrier separations as small as 5 MHz in frequency division duplex networks. In this regard, it would be more appropriate to introduce the paradigm of DIV and DOF densities [CD05b, CD05f], both of which facilitate a microscopic examination of the two measures, thereby capturing the system-dependent nature of multipath channels. Subsequently, the DIV and DOF are computed as the volume enclosed by the surfaces jointly determined by channel and signal boundaries. More details can be found in Chapter 7.

2.3.2.2 UWB Source Localization

The emerging UWB technology demands much research effort to achieve improved operational capabilities and cost-effective system solutions for broad commercial and military applications [SM04]. Theoretical study in the area of signal modeling and signal processing is thus essential for the development and performance analysis of practical UWB systems. Despite that a large volume of literature has been devoted to handle problems like source localization and sidelobe reduction for the narrowband case, fundamental differences exist between the

conventional and UWB array signal processing techniques [HUS02, GCB02, CRA00].

Recent UWB indoor channel measurements showed that physical paths were clustered around transmit/receive directions and each cluster corresponded to a major path to the receiver [CRA00, POO04, MFP03, IEEE802/04/0112]. Arrivals within clusters are likely the result of closely associated objects that are part of a very similar group of paths to the receiver (e.g., reflection from walls, floors and ceilings, scattering from furniture and displays, diffraction from doorway openings, transmission through soft partitions [IEEE802/04/0112, MUQ03]). Furthermore, each cluster, corresponding to an individual scattering center, will have its own impulse response or frequency transfer characteristics depending on the scattering mechanism [QIU02, QL99, QIU96, QIU04a, QIU04b, PH03, POO04] as mentioned in Section 2.2.3.2. To sum up, the UWB system structure features three distinct characteristics: presence of large angle spreads for multiple incoherently distributed scattering clusters, wideband array signals, and frequency-dependent dispersive effects. All of them have greatly complicated the estimation process and have not been jointly considered in the existing literature. In this regard, new methods for estimating the angular information of UWB sources and frequency-dependence of the underlying propagation modes are on demand.

2.4 Conclusions

This chapter has been designed to be a general introduction to the concepts and techniques of radio channel modeling. Starting with a thorough description of the system and correlation functions, the existing wireless channel models have been

classified into two categories: physically based and non-physically based. Next, some of the long-established and emerging channel models have been reviewed with a special emphasis on geometry-based modeling approaches. We have also presented our views of system-independent and system-dependent channel modeling concepts. In the subsequent chapters, more details will be provided on the modeling aspects and impact of system-dependent channel characterization.

CHAPTER 3 AN UNIFIED ANALYSIS FOR GEOMETRIC CHANNEL MODELING METHODS

3.1 Introduction

This chapter provides a unified analysis to circumvent the two problems of existing geometric channel models (GCMs) as discussed in Chapter 2: first, only a single specular reflection is taken into account, and neither diffuse scattering nor multiple reradiation is considered; second, the geometry of scattering areas in the existing GCMs is obtained on a rather undefined plane and is thus subject to debate. The semi-geometrically based statistical (SGBS) model is first presented. It is parameterized in such a way that only the distribution of scatterers contributing to the last reradiation is emulated while the preceding multiple-bounce is modeled as a stochastic process, which has a lognormal shadowing with Nakagami fading [STU01, SA00]. Such a modeling approach has two distinct merits. Firstly, it reproduces the basic stochastic processes of the received signal vector over space and time (e.g., power azimuthal spectrum (PAS), power delay spectrum (PDS), azimuthal and temporal dispersions) and is useful in studying the higher-order moments about the origin of the stationary processes (e.g., variation of power received in the azimuth and delay domains). Secondly, it paves the way for the postulation of the suppositions to be introduced in this chapter. An attempt is then made to unveil the understated assumptions made in arbitrary GCMs involving the physical propagation process, which is achieved by the introduction of three parameters: the effective path

length (EPL), the geometry-dependent parameter (GDP), and the normalized space-dependent intensity measure (NSDIM). The EPL is related to the large-scale path loss suffered by multipath components (MPCs) via the scatterers whereas the GDP bears relation to the spatial patterns of scattering points. With these three parameters, two heuristic rules are subsequently proposed from the wave-propagation viewpoint to establish a relationship between the geometry and the physical channel. The rules also imply the system-dependent channel modeling in its presentation. To demonstrate the accuracy and applicability of these rules, the prior assumptions made in several widely used GCMs are further exploited, thereby revisiting the validity of these models.

The organization of the chapter is as follows. Section 3.2 presents the SGBS model and describes the employed parameters. This section also gives the expressions of EPL, GDP and NSDIM with an explanation of their physical significance. Subsequently, the two heuristic rules are presented. Section 3.3 then proceeds to deal with the accuracy of a few widely used GCMs. Section 3.4 extends the generic model to discrete Manhattan world, which is applicable to wave propagation in city-street grids. Finally, the conclusion is presented in Section 3.5.

3.2 The Semi-Geometrically Based Statistical Model

3.2.1 General Remarks

The wireless channel is typically a multipath channel due to the natural and man-made objects, which are situated between the base station (BS) and mobile station (MS). At the receiver (Rx), plane waves arriving from different directions are

combined vectorially to produce a composite received signal. If only one discrete scatterer is present, an expression can be easily found for the signal at the Rx by using fundamental wave propagation and scattering laws. However, if many objects are present, things become much more complicated since the interaction between the different objects must be accounted for. Consequently a very complex model would result if multiple bounces were to be accurately modeled. By placing a scattering object at the last reradiation and approximating the preceding scattering as a stochastic process, some of the properties of multiple bounces can be retained while providing for a much simpler model.

When this approach is applied to a GCM, a compound random point field would first be defined¹ [CI80, BD91]. This field is determined by scatterer positions $S_1, S_2, S_3, \dots, S_k$ with a real valued scatterer response process $p_k(t)$ corresponding to each position. The points are samples from a spatial point process, which will be described as a nonhomogeneous Poisson process in the subsequent discussion. The $p_k(t)$'s represent the instantaneous power contributed by individual scatterers at the Rx, which are statistically independent and wide sense stationary. We will model this function as a shadowed Nakagami fading process, since the Nakagami distribution is known to provide a closer match to some experimental data than the Rayleigh, Ricean, or lognormal distribution [STU01]. Beyond its empirical justification, the Nakagami distribution is often used for the following reasons. First, it can model fading conditions that are either more or less severe than Rayleigh fading. Second, the Ricean distribution can be closely approximated by using the following relation

¹ It has been assumed that the number of scattering objects in the environment is large enough to form an ensemble of the point process. Furthermore, the finite-dimensional distribution of the points satisfies the symmetry and consistency conditions as stated in Kolmogorov theorem [CL67].

between the Rice factor K and the Nakagami shape factor m [STU01].

$$K = \frac{\sqrt{m^2 - m}}{m - \sqrt{m^2 - m}} \quad m > 1 \quad (3.1)$$

Though scatterers are buildings, cars and other pertinent structures in real life situations rather than delta functions, they will be indistinguishable from spatial delta functions if their dimensions are small relative to the spatial resolution of the system, as pointed out by Braun and Dersch [BD91]. Moreover, the reradiation from a scatterer is considered as narrowband in this chapter, i.e., the difference in the scattering properties within the frequency band is negligible due to the small bandwidth of the system as compared to the carrier frequency (usually much less than 1% such as the HIPERLAN/2 system). Nonetheless, recent advancement in wireless communications has generated a renewed interest in ultrawideband (UWB) systems. As defined by the Federal Communication Commission, such systems operate either with a fractional bandwidth greater than 20% or a –10 dB bandwidth that is beyond 500 MHz [FCC02]. In such cases, when a wave reflects off an object or penetrates through a material during multipath propagation, the effects are frequency sensitive, thus resulting in the waveform being filtered to some extent. The pulse shape associated with a propagation path is dependent on that path, with each path having a unique impulse response or frequency transfer characteristic [QIU02]. Hence the distribution of scatterers would have to be defined in a three-dimensional space, with two dimensions representing the physical space and one dimension representing the frequency domain [CD03a, CD04c]. In this chapter, focus would be placed solely on the narrowband scenario, with spatial pattern of the point process considered to be indifferent across the whole spectrum. Thus the effect of frequency variations will

Fig. 3.1. Pictorial representation of scattering from a scatterer S

not be taken into account at first. More details on UWB channel modeling will be given in Chapter 6.

3.2.2 Model Description

Based on the abovementioned methodology, the following SGBS model is proposed by taking into account multiple bounces. Fig. 3.1 shows the multipath propagation from the transmitter (Tx) to the Rx². In the sequel, it is assumed that the signals received at the Rx are plane waves arriving from the horizon. In addition, the following terms are introduced:

- S The scatterer contributing to the final reradiation of the incoming MPCs to the Rx.
- $r_R(r_T)$: The distance between the Rx (Tx) and the scatterer S .

² We will consider an uplink transmission in all the subsequent discussion.

$\hat{S}_1(\hat{S}_2)$: Local objects around S , scattering from which results in local subpaths to S .

$\check{R}_1(\check{R}_2)$: Objects around the Tx which cause the MPCs to undergo multiple reflections and/or diffractions.

β : Scattering coefficient of S .

$G_R(\varphi_R)$ ($G_T(\varphi_T)$): The directive gain of the antenna at the Rx (Tx) which satisfies the normalization function $\int_{-\pi}^{+\pi} G(\varphi) d\varphi = 2\pi$ for two-dimensional (2D) propagation.

It is further assumed that the Tx-to- S path has a shadowed Nakagami fading. Subsequently, the normalized area-mean power received at S has the general form

$$p_{a,T} = G_T(\varphi_T) r_T^{-n} = [G'_T(\varphi_T) r_T]^{-n} \quad (3.2)$$

with n being the path loss exponent. Assuming shadowing to be superimposed on (3.2), the local-mean power is lognormally distributed about the area mean, viz.,

$$f_{p_{l,T}}(p_{l,T}) = \frac{1}{p_{l,T} \sigma_T \sqrt{2\pi} \ln 10 / 10} \exp \left\{ -\frac{(10 \log_{10} p_{l,T} - p_{a,T}(\text{dBm}))^2}{2\sigma_T^2} \right\} \quad (3.3)$$

where σ_T is the standard deviation of the shadowing, expressed in natural units. Multipath reception at the scatterer causes Nakagami fading; with the squared-envelope having a Gamma-log-normal density function

$$f_{p_{s,T}}(p_{s,T}) = \int_0^\infty \frac{m^m \underline{p}_{s,T}^{m-1}}{\underline{p}_{s,T}^m \Gamma(m)} \exp \left(-\frac{m \underline{p}_{s,T}}{\underline{p}_{s,T}} \right) \frac{\exp \left[-\frac{(10 \log_{10} \underline{p}_{s,T} - p_{a,T}(\text{dBm}))^2}{2\sigma_T^2} \right]}{\underline{p}_{s,T} \sigma_T \sqrt{2\pi} \ln 10 / 10} d(\underline{p}_{s,T}) \quad (3.4)$$

For the special case when the multipath is Rayleigh-distributed ($m = 1$), (3.4) reduces to a composite exponential/lognormal probability density function (pdf).

As shown in [STU01], the composite pdf can be approximated by a lognormal

distribution with new mean and variation

$$P_{a,T}(\text{new}) = P_{a,T} 10^{10[\psi(m) - \ln(m)]/\ln 10} \quad (3.5)$$

$$\sigma_T^2(\text{new}) = (\ln 10/10)^{-2} \zeta(2, m) + \sigma_T^2 \quad (3.6)$$

where $\psi(\cdot)$ is the Euler *psi* function, and $\zeta(\cdot)$ is the Riemann's *zeta* function.

Let $p_{a,T}(\text{new}) = G_T(\phi_T)[r_T(\text{new})]^{-\eta}$, by applying (3.2) and (3.5) to the equation we obtain

$$r_T(\text{new}) = \eta r_T \quad (3.7)$$

where $\eta = 10^{-10[\psi(m) - \ln(m)]/(n \ln 10)}$ is defined as the Nakagami fading loss factor (NFLF).

It is observed that the effect of Nakagami fading in (3.4) would decrease the mean ($\eta \geq 1$) and increase the variance. However, such an effect reduces as the shape factor m increases [STU01]. In the situation when m approaches to infinity, η approaches to 1.

Apparently, the received power reradiated by the scatterer S at the Rx is also lognormally distributed with mean and standard deviation

$$P = [G'_R(\phi_R) G'_T(\phi_T) (r_R * \eta r_T)]^{-\eta} \beta \quad (3.8)$$

$$\sigma = \sigma_T(\text{new}) \quad (3.9)$$

with $G'_R(\phi_R) = [G_R(\phi_R)]^{-1/\eta}$. The operator $*$ represents either the product or the sum of the modulus of its arguments, depending on whether scattering or specular reflection dominates. Equation (3.8) reveals that the multipath structure from Tx to S has been changed by the Nakagami fading. As lognormals tend to remain lognormal under addition and subtraction [FEN60], the locally averaged total path gain at a given location thus tends to be lognormal as commonly observed in many field trials [STU01].

It is worth mentioning here that unlike conventional geometrically based models and ray-tracing techniques, the constellation of scatterer distribution in the SGBS model is in general nonreciprocal except in the single-bounce situation. Nonetheless, the reciprocity is satisfied if the channel is a quasi-single-bounce (QSB) one. That is, S belongs to a group of strong scatterers having line-of-sight (LoS) connection to both Tx and Rx. Other reradiators, such as $\hat{S}$ and $\check{R}$, only cause the Tx-S-Rx path to undergo power fading without changing the average information of time-of-arrival (ToA) and angle-of-arrival (AoA) associated with that path. We only consider this type of channel in the subsequent discussion. In particular, for waveguided environments such as indoor hallways or street canyons, S may be considered as a hypothetical scatterer which effectively keeps the ToA and AoA information in the model with some omission of details of the propagation process (see Fig. 3.2).

3.2.3 The Underlying Connection between the Spatial Pattern of Scatterers and Actual Wave Propagation Process

Fig. 3.2. Hypothetical scatterers in a waveguided environment

Assuming that the spatial position of a scattering point is denoted by the vector ζ with components $(\underline{r}, l)$ related to an artificial coordinate in a 2D Euclidean space $\mathbb{R}$, where

$$\underline{r} = \mathcal{H}(G_R(\varphi_R), G_T(\varphi_T), r_R, r_T)$$

is a continuous and differentiable function with the modulus of the arguments defined in Section 3.2.2 (or refer to Fig. 3.3). $\underline{r}$ is named as the geometry-dependent parameter. l is a point at the oriented curve $\underline{C}$: $\underline{r}(l) = \underline{r}$ with l_s being an arbitrarily defined starting point and the positive sense on $\underline{C}$ being clockwise as illustrated by the arrow in Fig. 3.3. In our formulation of the GCM, the function $\mathcal{H}$ is defined such that the nonhomogeneous Poisson process could be constructed using the intensity measure at any point ζ . Firstly let $\rho_1, \rho_2, \dots, \rho_m$ be the intensities of many identically and independently distributed homogeneous Poisson processes³ on discrete curves $\underline{C}_1: \underline{r}(l) = \underline{r}_1, \underline{C}_2: \underline{r}(l) = \underline{r}_2, \dots, \underline{C}_m: \underline{r}(l) = \underline{r}_m$. Since $\underline{r}$ is a continuous function, the number of points on a fixed line $\underline{C}$ will be zero with probability one. Thus the lines should be fattened up to strips

$$\{(\underline{r}, l) : \underline{r}_{m-1} < \underline{r} < \underline{r}_m, l \in \underline{C}_{m-1}\}$$

as shown in Fig. 3.4. The points in different strips (non-overlapping differential regions) are independent, which is essential for the uncorrelated scattering properties assumed in many channel models. A more formal realization of this process can be described below following from [CI80]. Let the intensity measure $\rho(\zeta)$ be a positive function of ζ with dimension $[\text{area}]^{-1}$. Let δ_ζ be a small neighborhood of ζ of area

³ It is assumed that the intensity measure of the Poisson process defined here only depends on $\underline{r}$. Nevertheless, this conjecture holds an intuitive appeal (all the factors affecting the distribution of scatterers have been embedded in the arguments of $\underline{r}$) and has been implicitly assumed in many existing GCMs.

Fig. 3.3. Artificial coordinate (l, l)

Fig. 3.4. Realization of the spatial nonhomogeneous Poisson process

$|\zeta|$ (Lebesgue measure) with all the points of the process occurring outside δ_ζ denote by $\varepsilon_{\delta_\zeta}$. If $N(\mathbb{A})$ denote the number of points in an area $\mathbb{A}$, it is required that

$$\text{pr}\{N(\delta_\zeta) = 0 | \varepsilon_{\delta_\zeta}\} = 1 - \rho(\zeta)|\delta_\zeta| + o(|\delta_\zeta|)$$

$$\text{pr}\{N(\delta_\zeta) = 1 | \varepsilon_{\delta_\zeta}\} = \rho(\zeta)|\delta_\zeta| + o(|\delta_\zeta|)$$

$$\text{pr}\{N(\delta_\zeta) \geq 2 | \varepsilon_{\delta_\zeta}\} = o(|\delta_\zeta|)$$

where $\text{pr}\{\cdot\}$ denotes the probability. Since the intensity measure ρ depends only on r , it is rewritten as $\rho(r)$ in the subsequent discussion.

When the total number of points in the bounded region $\mathbb{A}$ is normalized to 1, $\rho(\underline{r})$ actually represents the variation of scatterer density $f_{E,l}(\underline{r}, l)$ at any position on the curve $\underline{C}$, which is uniform along the curve due to the fundamental characteristics of a homogeneous Poisson process on $\underline{C}$. Such a parameter is thus defined as the *normalized space dependent intensity measure*. Apparently the distribution of $\underline{r}$ can be obtained using the line integral of $f_{E,l}(\underline{r}, l)$ over $\underline{C}$, viz.,

$$f_{\underline{r}}(\underline{r}) = \int_{\underline{C}} f_{E,l}(\underline{r}, l) dl = f_{E,l}(\underline{r}, l) \cdot \ell_{\underline{C}} \quad (3.10)$$

where $\ell_{\underline{C}}$ denotes the length of the curve $\underline{C}$.

Next, the wave propagation in the wireless link is considered. The EPL is defined as

$$\bar{\underline{r}} = G'_R(\varphi_R) G'_T(\varphi_T) (r_R * \eta r_T) \quad (3.11)$$

As shown in (3.8), the expectation of received power at the Rx from any scattering point is inversely proportional to $\bar{\underline{r}}^n$ ($p(\bar{\underline{r}}) \propto \bar{\underline{r}}^{-n}$). Therefore, the EPL is a measure of the mean path loss suffered by MPCs whereas the NSDIM is relevant to the geometry of canonical models.

• The Two Heuristic Rules

The following two heuristic rules governing the relationship between the geometric distribution of scatterers ($f_{\underline{r}}(\underline{r})$ and $f_{E,l}(\underline{r}, l)$) and the power distribution $p(\bar{\underline{r}})$ are proposed. A pictorial representation of the two rules is shown in Fig. 3.5.

1. *Bijjective mapping rule*: Let $\underline{A}$ denote the set of all curves in the region B with each curve representing a collection of scatterers with the same value of $\underline{r}$ and let

$\hat{\underline{A}}$ denote the set of all curves with each representing a collection of scatterers in the same region, which contribute to a constant average-received-power at the Rx. The mapping $\mathcal{F}: \underline{A} \rightarrow \hat{\underline{A}}$ defines a one-to-one correspondence between $\underline{A}$ and $\hat{\underline{A}}$ through the function $\hat{\underline{r}} = \mathcal{G}(\underline{r})$.

2. *Tapering-off NSDIM rule:* If $\mathcal{G}$ defines a monotonically increasing function, the NSDIM gradually tapers off as $\underline{r}$ increases where the maximum NSDIM corresponds to the direct path between the Rx and the Tx in the radio link. This rule implies that, as the average-received-power at the Rx due to reradiation from a scatterer at a certain location decreases, the intensity measure at that location decreases as well.

A useful analogy can be used to establish the relationship between the two heuristic rules and the principles governing the equipotential lines in electromagnetic field theory. In the geometric channel modeling problems, the set $\underline{A}$ in the first rule resembles the visualization of an electrical field and the set $\hat{\underline{A}}$ is similar to the collection of equipotential lines. The electrical potential is constant on each equipotential line as depicted in Fig. 3.6. (*In a similar manner, scatterers are uniformly distributed on each curve $\underline{C}$ in the GCMs.*) These curves will not intersect each other (*bijective mapping rule*). Assuming the potential at infinite distance to be zero, the further the lines are to the charges, the smaller the absolute value of the potential (*tapering-off NSDIM rule*).

The two heuristic rules have provided a venue to exploit the understated assumptions made in an arbitrary GCM concerning the power distribution in the

Fig. 3.5. Pictorial representation of the two heuristic rules

Fig. 3.6. Equipotential lines due to a pair of configuration charges

wireless link. Nevertheless, it has to be noted that they are only a first approximation in a spatially regular scattering environment, where the waves arriving at the Rx stem from the same propagation mechanism and/or scattering process [LKT++02], thereby specifying a unique nonhomogeneous Poisson process.

- **Geometric Classification**

For the case of a more complex wireless channel where the propagation of signals could be governed by various mechanisms, the distribution of scatterers can be classified into a few geometric clusters as distinguished by their respective spatial point processes [CD03b, CD05a]. Following from the two heuristic rules, a definition for geometric classification can be proposed as follows

Definition 1 (Geometric classification): Let $\mathbb{S}$ be the sample space of the spatial point process and $\mathcal{S}_i \subseteq \mathbb{S}$, a subset of the space, such that the GDP $\underline{r}$ can be parameterized using the same $\mathcal{H}_i$ for any element ε in the subset. Furthermore, let $\mathcal{P}$ represents a partition of $\mathbb{S}$ into $\mathcal{S}_i$, then the smallest partition $\mathcal{P}^*$ equals the set of geometric clusters.

By Definition 1 each cluster is a geometric abstraction of certain radio environment and the shape of each cluster is defined by a set of logical relationships (i.e., the functions $\mathcal{G}$ and $\mathcal{H}$) corresponding to the *a priori* properties of wave-propagation processes in the channel. The identifiable clusters correspond to different physical propagation routes and mechanisms, such as street-guided propagation with multiple reflections from the street canyon walls, or scattering from high-rise objects over the rooftop. Note that the current classification need not imply separable topography, which is different from the methodology used in [FMB98] and similar with the one in [LKT++02]. The channel impulse responses due to various clusters may be overlapping in the azimuth-delay plane.

3.3 Revisiting Some Mostly Used Geometric Channel Models

3.3.1 Macrocellular Environments

In what follows it is assumed that omnidirectional antennas are used at both ends of the connection, i.e., $G_R(\varphi_R) = G_T(\varphi_T) = 1$. The EPL can thus be simplified as

$$\hat{r} = r_R * \eta r_T \quad (3.12)$$

The circular scattering model (CSM) is shown in Fig. 3.7(a), where the scatterer density is assumed to be uniform in azimuth around the Tx (which is the MS for the uplink transmission). Such models have been suggested for macrocell type of environments with moderate values of angular spread and large values of delay spread. It has been assumed that in macrocellular environments, there will be no signal scattering from locations near the BS since the antenna heights are relatively large. Furthermore, it is also assumed that the scatterers surrounding the MS are about the same height or higher than the mobile.

The curve representing an aggregate of scatterers with the same value of $\hat{r}$ is a circle defined by $\underline{C}$: $\underline{r}(l) = r_T$. Applying the bijective mapping rule, the EPL $\hat{r}$ is a bijection of r_T , which is valid when the following relationship is satisfied

$$\hat{r} = r_R * \eta r_T \equiv \mathcal{G}(r_T) \quad \forall r_T \quad (3.13)$$

Equation (3.13) implies

$$r_R + \eta r_T \approx D + \eta r_T \quad \text{if the symbol } * \text{ designates sum} \quad (3.14)$$

$$r_R \cdot \eta r_T \approx \chi r_T \quad \text{if the symbol } * \text{ designates product} \quad (3.15)$$

χ is a constant in expression (3.15). In this case, the EPL $\hat{r} = D + \eta r_T$ or χr_T is a monotonically increasing function with respect to r_T , thereby showing that the necessitated condition of the tapering-off NSDIM rule is fulfilled. Apparently, the CSM implies that the majority of the scattering elements are densely clustered around

Fig. 3.7. (a) Circular scattering model and (b) elliptical scattering model

the Tx (MS) and the width of the scattering area is much less than the distance between the Tx and the Rx. These two statements are generally valid in a dense urban macrocell [ECS++98].

Three distributions of r_T which are currently available in the literature will be considered in this section. It should be noted that only the scatterers contributing to the last reradiation of MPCs are emulated in the models in order for them to be analyzed under the proposed framework.

- A linear distribution of r_T

$$f_{r_T}(r_T) = \begin{cases} \frac{2r_T}{R^2} & r_T \leq R \\ 0 & \text{elsewhere} \end{cases} \quad (3.16)$$

This pdf leads to the geometrically based single bounce macrocell (GBSBM) model [PRR02], where the scatterers are uniformly distributed within a circle with radius R . Employing (3.10), the NSDIM can be obtained as

$$f_{r_T,l}(r_T, l) = \frac{f_{r_T}(r_T)}{2\pi r_T} = \frac{1}{\pi R^2} \quad (3.17)$$

which is a constant and does not abide by the tapering-off NSDIM rule. In fact, (3.17) implies that as the received power due to rebounding from a scatterer at a certain location decreases, the scatterer density at that location remains constant. Apparently, this assumption of the GBSBM model does not accurately describe the observed phenomena (average power of signal decreases systematically with increasing distance) and requires further investigation.

- A normal distribution of the x - and the y - coordinate measured from the position of the MS leads to a Rayleigh distribution of r_T .

$$f_{r_T}(r_T) = \frac{r_T}{\sigma_R^2} \exp\left[-\frac{r_T^2}{2\sigma_R^2}\right] \quad (3.18)$$

where σ_R is the standard deviation of the local scatterer density. This pdf leads to the Gaussian scatterer density model (GSDM) [FMB98, JAN02, 3GPP/LUC], where the spatial distribution of the scatterers is approximated by the bivariate Gaussian pdf. Similarly, the NSDIM is given by

$$f_{r_T,l}(r_T, l) = \frac{f_{r_T}(r_T)}{2\pi r_T} = \frac{1}{2\pi\sigma_R^2} \exp\left[-\frac{r_T^2}{2\sigma_R^2}\right] \quad (3.19)$$

It is apparent from (3.19) that the NSDIM decreases monotonically as the EPL

increases, which abides by the second rule. Therefore, implementing this model to resemble the physical wave-propagation process actually implies that as the average received power due to reradiation from a scatterer at a certain location decreases, the intensity measure at that location decreases as well, which holds an intuitive appeal.

- A one-sided Gaussian distribution is represented using

$$f_{r_T}(r_T) = \frac{2}{\sqrt{2\pi}\sigma_r} \exp\left(-\frac{r_T^2}{2\sigma_r^2}\right) \quad (3.20)$$

with σ_r being the standard deviation of the scatterer density. This distribution leads to the generic multiple-input-multiple-output (MIMO) channel model [MOL04]. Subsequently, the NSDIM follows from (3.10) as

$$f_{r_T,l}(r_T, l) = \frac{f_{r_T}(r_T)}{2\pi r_T} = \frac{2}{(2\pi)^{3/2} \sigma_r r_T} \exp\left[-\frac{r_T^2}{2\sigma_r^2}\right] \quad (3.21)$$

This model also abides by the tapering-off NSDIM rule. Further comparison of (3.19) and (3.21) reveals that as r_T increases, the mean number of relevant scatterers in the one-sided Gaussian distribution model decreases more significantly as compared to the preceding model (by comparing the slopes of the curves expressed in (3.19) and (3.21)). Though an exact mapping of the average signal power to the NSDIM is difficult to justify, the two heuristic rules nonetheless reveal a close relationship between the spatial pattern and the physical propagation process.

3.3.2 Microcellular or Picocellular Environments

The elliptical scattering models (ESM) [ER99, LI95] (Fig. 3.7(b)) are proposed for the picocell or microcell where antenna heights are relatively low and the Tx and the Rx are the foci of the elliptical area. With low antennas, the Rx will receive

MPCs from scatterers distributed around the Tx as well as itself. All angles in the azimuth plane are involved in this case, and the maximum delay is determined by the size of the ellipse. In the physical sense, it is possible to ignore components with larger delays since signals with longer delays have relatively low power as compared to those with shorter delays, due to the longer path length traveled.

The curve representing a collection of scatterers with the same value of $\underline{r}$ is an ellipse defined by $\underline{C}$: $\underline{r}(l) = r_T + r_R$. Applying the bijective mapping rule, the EPL $\underline{\hat{r}}$ should have a one-to-one correspondence with $\underline{r}$. This statement is justified when the following relationship is satisfied

$$\underline{\hat{r}} = r_R * \eta r_T \equiv \mathcal{G}(r_R + r_T) \quad (3.22)$$

which implies

$$\eta = 1 \text{ and the symbol } * \text{ designates sum} \quad (3.23)$$

In such a case, the EPL consistently increases with $\underline{r}$. The expression in (3.23) implies that for the ESMs, specular reflection is the dominant mechanism of interaction between the electromagnetic waves and the surfaces of scattering objects. Other mechanisms, such as diffraction and diffuse scattering can be ignored. Moreover, the transmitted signal has mild short-term fading around the MS. Such observations are generally valid in indoor propagation [HAS93].

In the Liberti's model [ER99, LI95], the scatterers are uniformly distributed within an ellipse with semi-major axis a and semi-minor axis b as illustrated in Fig. 3.7(b). The pdf of $\underline{r} = r_R + r_T$ can be approximated as

$$f_{\underline{r}}(\underline{r}) = \frac{1}{2b} \frac{2(\underline{r}/2a)^2 - (D/2a)^2}{\sqrt{(\underline{r}/2a)^2 - (D/2a)^2}}, \quad \frac{D}{2a} \leq \frac{\underline{r}}{2a} \leq 1 \quad (3.24)$$

The NSDIM following from (3.10) is

Table 3.1. Summary of implication of the two rules

Geometric models	Implication of the bijective mapping rule
CSM [FMB98, JAN02, PRR02, MHS02, MOL04]	The scattering elements are densely clustered around the Tx and the width of the scattering area is much less the distance of the direct path between Tx and Rx.
ESM [ER99, LI95]	Specular reflection is the principal mechanism; mild short-term fading around the MS.
Tapering-off NSDIM rule	
GBSBM [PRR02]	Not abide
GSDM [FMB98, JAN02]	Abide
MIMO model [MOL04]	Abide
Liberti's model [ER99, LI95]	Abide

$$f_{e,l}(\underline{r},l)=\frac{f_e(\underline{r})}{4aH(e,\pi/2)} \tag{3.25}$$

where

$$H\left(e,\frac{\pi}{2}\right)=\frac{\pi}{2}\left[1-\left(\frac{1}{2}\right)^2e^2-\left(\frac{1\cdot3}{2\cdot4}\right)^2\frac{e^4}{3}-\dots\right] \& e=D/2a \tag{3.26}$$

It can be shown from (3.25) that the pdf has a maximum and an integral singularity at $\underline{r} = D$, where the peak corresponds to the LoS path. The GBSBEM model thus abides by the tapering-off NSDIM rule and is a good approximation to the physical reality from a wave-propagation viewpoint.

A brief summary of this section is listed in Table 3.1. It has been demonstrated that both the circular and elliptical scattering models can be investigated under the proposed analytical skeleton. The validity of the models has also been reexamined using the two rules proposed. The analysis clearly points out that it may be necessary to reconsider the understated assumptions in some of the widely used GCMs.

3.4 Further Remarks – Geometric Model in a Discrete Manhattan World

3.4.1 Background

Signal propagation along streets surrounded by buildings, which are considerably taller than the height of both Tx and Rx antennas, is very different from propagation in conventional cells. The geometry of city-street blocks determines the statistical properties of multipaths arriving at the receive antenna. The widely used GCMs discussed in the previous sections are usually not applicable in these environments. In order to extend the unified analysis presented in the previous sections to the street-guided propagation environment, the taxicab norm or Manhattan distance has to be used to compute the EPL between two points due to the grid structure of the overlay profile. Furthermore, since scatterers are not continuously distributed in the space due to the structure of street grids, the two heuristic rules must be reformulated for them to be applied in a discrete Manhattan world. Note that this analysis would not hold in the case when either Tx or Rx antenna is above the height of the buildings, where other mechanisms will dominate the process; or when either Tx or Rx antenna is close to the rooftop, where diffraction over the rooftops becomes significant.

3.4.2 Model Descriptions

The rectilinear orientation of the streets is shown in Fig. 3.8(a). A microcell site $Tx(x_T, y_T)$ is located in one street called the main street, and the subscriber $Rx(x_R, y_R)$ is positioned on the perpendicular street (cross street). For the sake of clarity, the width of the streets and other parallel and perpendicular streets are omitted from the

figure. The propagation mechanism of radio signals is as follows. The transmit energy flows along the main street, and some of it couples into the cross streets due to multipath reflections. In turn, some of the energy flowing along the cross streets couples into the main streets. Based on this mechanism, the falloff with distance of received signal power at the Rx decreases as $\hat{r}^{-n}$, with $\hat{r}$ being the EPL between the Tx and Rx via the scatterer S in city-street grids and n the path loss exponent. Apparently, instead of Euclidean distance, the taxicab norm or Manhattan metric is applied to define $\hat{r}$, i.e.,

$$\hat{r} = r_T * \eta r_R = (|x_T - x_S| + |y_T - y_S|) * \eta (|x_R - x_S| + |y_R - y_S|) \quad (3.27)$$

when omnidirectional antennas are used at both ends of the radio link. If specular reflection is the primary mechanism for radio wave propagation and the NFLF $\eta = 1$, (3.27) reduces to

$$\hat{r} = |x_T - x_S| + |y_T - y_S| + |x_R - x_S| + |y_R - y_S| \quad (3.28)$$

Fig. 3.8. (a) Rectilinear orientation of the street and (b) curves representing the collection of scatterers with constant $\hat{r}$

We further define the bijection in the bijective mapping rule as $\underline{r} = \mathcal{G}^{-1}(\hat{\underline{r}}) = \tilde{\underline{r}}$.

The locus of scatterers giving rise to a constant $\underline{r}$ can thus be expressed as

$$\underline{r} = \begin{cases} -2x_S + 2y_S + (x_T - y_T) + (x_R - y_R) & \textcircled{1} \\ 2y_S + (x_T - y_T) - (x_R + y_R) & \textcircled{2} \\ 2x_S + 2y_S - (x_T + y_T) - (x_R + y_R) & \textcircled{3} \\ -2x_S + (x_T + y_T) + (x_R - y_R) & \textcircled{4} \\ (x_T + y_T) - (x_R + y_R) & \textcircled{5} \\ 2x_S - (x_T - y_T) - (x_R + y_R) & \textcircled{6} \\ -2x_S - 2y_S + (x_T + y_T) + (x_R + y_R) & \textcircled{7} \\ -2y_S + (x_T + y_T) - (x_R - y_R) & \textcircled{8} \\ 2x_S - 2y_S - (x_T - y_T) - (x_R - y_R) & \textcircled{9} \end{cases} \quad (3.29)$$

where the 9 regions in the city map are illustrated in Fig. 3.8(a). From (3.29), the curves in the scattering region with each representing a collection of scatterers with the same value of $\underline{r}$ are octagonal and shown in Fig. 3.8(b). Apparently, scatterers causing echoes with identical delays are situated on each octagon as well.

Subsequently, the tapped-delay-line model is implemented to reveal the discrete nature of city-street blocks [CD03c]. Each tap is characterized by the probability distribution function (PDF) of $\underline{r}_k^D$, $\forall k \in (1, 2, \dots, K)$, where $\underline{r}_k^D$ is the discrete GDP of the k^{th} tap and K is the total number of taps.

3.5 Conclusions

In this chapter, we have introduced an SGBS model for multipath propagation environments. This model defines the distribution of principal scatterers contributing to the last reradiation of MPCs to the Rx. The composite Nakagami lognormal pdf

was used to model the preceding multiple-reflection process. Three parameters, namely, the EPL, GDP, and NSDIM, were introduced to better understand the relationship between the geometry of the canonical models and the actual propagation process. Subsequently, the methodology was used to revisit several widely used GCMs for narrowband stationary environments with omnidirectional antennas employed at both ends of the communication link. It was shown that both the circular and elliptical scattering models could be inspected using the proposed analytical tool. The investigation also revealed that the understated assumptions in some of the widely used GCMs were questionable. Through extension of the unified analysis to discrete Manhattan space, an octagonal tapped delay line model was implemented to describe the distribution of scatterers in a typical rectilinear street environment.

CHAPTER 4 PARABOLIC DISTRIBUTION OF SCATTERERS IN STREET-DOMINATED ENVIRONMENTS

4.1 Introduction

In the final sections of Chapter 3, we consider the propagation scenario for rectilinear city-street grids when both the transmit (Tx) and receive (Rx) antenna heights are below the rooftop level of the surrounding buildings. In this chapter, we will discuss another type of street-dominated environment as depicted in Fig. 4.1, for which the base station (BS) antenna is elevated above rooftop level and the mobile station (MS) is located in the street canyon. In such a case, almost all the long-delayed paths travel along the directions of the streets where the MS is located [KRB00, FRB97, COX73, KLV++03, LKT++02]. Such a waveguiding effect cannot be properly reproduced using the conventional circular scattering model (CSM). The main object of this chapter is thus to provide a more accurate model for street-guided mobile environments. Following from the general analysis in Chapter 3, Section 4.2 presents the parabolic scattering model (PSM) to simulate the guidance of streets in dense urban areas. Expressions are also provided for useful channel parameters like the power azimuthal spectrum (PAS), angular constriction (AC), variance of received power in the azimuth domain, and Doppler spectrum (DS). Subsequently, Section 4.3 gives a comparison between the theoretical and experimental results so as to verify the validity of the proposed model. Finally, a conclusion is presented in Section 4.4.

Fig. 4.1. A typical street-dominated propagation environment

4.2 Parabolic Scattering Cluster for Longer-delayed Waves in Street-dominated Environments

4.2.1 General Remarks

For short-delayed waves present in street-dominated environments, the local objects in the vicinity of the MS like a cluster of plant leaves, walking pedestrians or nearby building structures dominate the propagation [KRB00]. In such cases, diffuse scattering becomes the principal mechanism of interaction between the electromagnetic field and the scatterer surfaces. Moreover, the majority of the

Fig. 4.2. (a) The parabolic scattering cluster and (b) the upper bound of the integrand r for the PAS at the MS

scatterers are clustered around the MS and the width of the local scattering area is much less than the BS-to-MS distance D ($D \gg r_M$). Subsequently, the CSM presented in Section 3.3.1 would be suitable in such a case, which also corresponds to the circular model for the small-delay components in [KRB00].

Conversely, propagation via street canyons would dominate for large excess delays. The energy is now concentrated in azimuth sectors aligned with the street directions. In such a scenario, the Nakagami fading loss factor (NFLF) η in (3.11) would be a direction-sensitive variable $\eta(\theta)$, since it can be noted that the closer the scatterers are to the street-direction, the less severe the Nakagami fading is due to fewer multiple bounces. Moreover, the waveguiding in street canyons occurs primarily via reflections from buildings and hence the operator $*$ would read sum. A parabolic scattering cluster (PSC) as shown in Fig. 4.2(a) would fit into the aforesaid situation and serves to be very convenient for the analysis of the channel characteristics.

A parabolic line can be parameterized as (see also Fig. 4.2(a))

$$D + \hat{p} = r_B \cos \varphi_B + r_M = \underline{r} \quad (4.1)$$

which can be rewritten as

$$r_B + r_M / \cos \varphi_B = \underline{r} / \cos \varphi_B \approx \underline{r} \text{ for small } \varphi_B \quad (4.2)$$

where $\hat{p}$ is half of the latus rectum of the parabolic line. Comparing (3.11) and (4.2) yields $\eta(\varphi_B) = 1/\cos \varphi_B$ for omnidirectional antennas. Also note that η increases as φ_B increases.

Additionally, the following basic assumptions are made in order to fully develop the model.

- To simplify the analysis, we assume that there is no line-of-sight (LoS) connection between the BS and MS. This hypothesis is valid as long as the coupling of signal power into streets is via reflections from buildings/diffractions around street corners, rather than dominant reflectors rising above the average rooftop level acting as pseudo signal sources. Nonetheless, the effect of major

reflector can be easily included by adding a strong wave into the angular-delay resolved impulse response.

- The scattering area is divided into a “clear zone” and a “scattering zone”, as depicted in Fig. 4.2(a). It is assumed that only the elements in the scattering zone would effectively redirect the incoming signals to the receiver in order to ensure that the scatterer distribution does not extend to infinity. The width of the clear zone is denoted by a , which is usually much greater than the street width so as to provide for the validity of small-angle approximations in (4.2). The scatterers are positioned along a family of parabolic lines $y^2 = -2\hat{p}(x-D-\hat{p}/2) = -2(\underline{r}-D)(x-D/2-\underline{r}/2)$ and the MS is the common locus point of these parabolas.
- Following from the tapering off NSDIM rule, the scatterer density should gradually taper off with the geometry-dependent parameter (GDP) $\underline{r}$, as remote scatterers generally yield less contribution compared to close-in scatterers. Herein a general family of pdfs which encompass the one-sided Gaussian, Laplacian, and uniform pdfs is proposed to provide for various possible distributions of $\underline{r}$.

$$f_{\underline{r}}(\underline{r}) = \frac{\exp\left[-(|\underline{r}-\underline{r}_{\min}|/b)^{2/c}/2\right]}{b2^{c/2}\Gamma(1+c/2)} \quad \underline{r} \geq \underline{r}_{\min} \quad (4.3)$$

where the coefficient of kurtosis $\kappa = \Gamma(5c/2)\Gamma(c/2) / [\Gamma(3c/2)]^2 - 3$. The minimum value of GDP is $\underline{r}_{\min}$, Γ is the gamma function, the scale parameter $b > 0$ and the shape parameter $c > 0$. For $b = c = 1$, the generic pdf corresponds to a standard normal distribution, while for $b = 1/2$ and $c = 2$ the pdf is Laplacian. As c tends to zero, the pdf tends toward a rectangular distribution. The kurtosis κ is a statistical measure to describe the tail weight for a distribution. A useful physical interpretation for κ is as follows. If the density of scattering points decreases

steeply as r increases, in such cases, the pdfs would have tails which decline more rapidly thus yielding smaller values of kurtosis. However if the dropping of the scatterer density is gentle, the pdf decays slowly, thus resulting in a larger kurtosis.

The subsequent analysis would focus on the signal reception in the azimuth domain at the MS, which has greater relevance to the majority of available measurement data collected in street-dominated environments. Nevertheless, the characteristics of received signals at the BS could be derived in a similar way by assuming a quasi-single-bounce (QSB) channel.

4.2.2 PAS, Angular Constriction, and Variance of Received Power in the Azimuth Domain at the MS

From Fig. 4.2(a), the circumference of the parabola $y^2 = -2\bar{p}(x-D-\bar{p}/2)$ is

$$\ell_C = L_{MN} = \bar{p} \left[\sqrt{\frac{2k}{\bar{p}} \left(1 + \frac{2k}{\bar{p}} \right)} + \ln \left(\sqrt{\frac{2k}{\bar{p}}} + \sqrt{1 + \frac{2k}{\bar{p}}} \right) \right] \quad (4.4)$$

where $k = D - a + \bar{p}/2$. Thus the pdf of angle-of-arrival (AoA) at the MS conditioned on r can be expressed as

$$f_{\varphi_M|r}(\varphi_M|r) = \frac{\sqrt{r_M^2 + (dr_M/d\varphi_M)^2}}{\ell_C} \quad (4.5)$$

where r_M is obtained from the elementary geometry as

$$r_M = \frac{r - D}{1 - \cos \varphi_M} \quad (4.6)$$

Following from the analysis in Section 3.2.2, the instantaneous power, p_s , received at the MS due to reradiation from a scatterer S would be lognormally

distributed due to the reciprocity of the QSB channel. Therefore, the joint pdf of p_s and φ_M , conditioned on the mean power p is written as

$$f_{p_s, \varphi_M | p}(p_s, \varphi_M | p) = f_{p_s | p}(p_s | p) f_{\varphi_M | p}(\varphi_M | p) \quad (4.7)$$

where p is related to the GDP $\underline{r}$ by the function $p = \beta \underline{r}^{-n}$ assuming a mirror mapping in the bijective mapping rule (i.e., $\underline{r} = \bar{\underline{r}}$). β denotes the average attenuation loss.

Applying the transformation of variables to (4.7), the joint density $f_{p_s, \varphi_M | \underline{r}}(p_s, \varphi_M | \underline{r})$ can be expressed as

$$f_{p_s, \varphi_M | \underline{r}}(p_s, \varphi_M | \underline{r}) = f_{p_s, \varphi_M | p}(p_s, \varphi_M | p) \Big|_{p=\beta \underline{r}^{-n}} \quad (4.8)$$

After some manipulations, the PAS at the MS can be obtained as

$$\begin{aligned} f_{\varphi}(\varphi_M) &= \Sigma_{\varphi} E \{ E \{ p_s, \varphi_M | \underline{r} \} \} \\ &= \Sigma_{\varphi} \iint_{\Xi} p_s f_{p_s, \varphi_M | \underline{r}}(p_s, \varphi_M | \underline{r}) f_{\underline{r}}(\underline{r}) d\underline{r} dp_s \\ &= \Sigma_{\varphi} \int_D^{\underline{r}_{\max}} f_{\underline{r}}(\underline{r}) f_{\varphi_M | \underline{r}}(\varphi_M | \underline{r}) d\underline{r} \int p_s f_{p_s | \underline{r}}(p_s | \underline{r}) dp_s \\ &= \Sigma_{\varphi}' \int_D^{\underline{r}_{\max}} \frac{\exp \left[-(|\underline{r} - \underline{r}_{\min}|/b)^{2/c} / 2 \right]}{b^{2c/2} \Gamma(1+c/2)} \frac{\sqrt{r_M^2 + (dr_M/d\varphi_M)^2} \beta \underline{r}^{-n}}{\ell_{\underline{r}}} d\underline{r} \end{aligned} \quad (4.9)$$

where Σ_{φ} and Σ_{φ}' are normalization factors to ensure that $f_{\varphi}(\varphi_M)$ is a pdf, and Ξ is the integration region. The upper bound of the integrand $\underline{r}$ in (4.9) is calculated as

$$\underline{r}_{\max} = \begin{cases} +\infty & \text{when } 90^\circ \leq \varphi_M \leq 270^\circ \\ D + \hat{p}_{\max} = D + \frac{(D-a)(1-\cos \varphi_M)}{\cos \varphi_M} & \text{otherwise} \end{cases} \quad (4.10)$$

with the geometric relationship illustrated in Fig. 4.2(b). The closed-form expression for (4.9) is intractable and thus has to be solved numerically. Note that the PAS obtained in (4.9) is independent of the lognormal fading statistics around the MS, which implies that the spatial pattern of the random field has retained all the necessary information required to characterize the first-moment statistics of the

received power in the azimuth domain.

For the street-guided propagation class, waves with almost equal amount of power arrive at the mobile unit from two opposite directions of the street canyons, producing no clear bias in the AoA. Therefore the value of the angular spread provides less insight into the power distribution in the azimuth domain. To measure how multipaths concentrate about two azimuthal directions, the angular constriction introduced by Durgin *et al.* [DR00] is applied, which is defined as

$$\gamma = \frac{|F_0 F_2 - F_1^2|}{F_0^2 - |F_1|^2} \quad (4.11)$$

where F_n is the n^{th} complex Fourier coefficient of the angular distribution of multipath power, $f_\varphi(\varphi_M)$, and is given by

$$F_n = \int_0^{2\pi} f_\varphi(\varphi_M) \exp(jn\varphi_M) d\varphi_M \quad (4.12)$$

The measure of the angular constriction remains invariant under changes in transmitted power or any series of rotational or reflective transformations of $f_\varphi(\varphi_M)$ [DR00]. The possible values of γ range from 0 to 1, with 0 denoting no clear bias in the two AoA directions and 1 denoting the extreme case of having exactly two ray arrivals coming from two different directions.

In order to study the aggregate effect of multifold scattering on signal reception, it would be useful to obtain the higher-order statistics of the stationary process of received power. The variance of the received power along the azimuth angle φ_M is derived as

$$\begin{aligned}
 V_p(\varphi_M) &= E\{p_s^2(\varphi_M)\} - \{E\{p_s(\varphi_M)\}\}^2 \\
 &= \iint_{\mathcal{D}} p_s^2 f_{p_s, \varphi_M | \underline{r}}(p_s, \varphi_M | \underline{r}) f_{\underline{r}}(\underline{r}) d\underline{r} dp_s - [f_{\varphi}(\varphi_M)]^2 / \Sigma_{\varphi}^2 \\
 &= \int_D^{\underline{r}_{\max}} f_{\underline{r}}(\underline{r}) f_{\varphi_M | \underline{r}}(\varphi_M | \underline{r}) d\underline{r} \int p_s^2 f_{p_s | \underline{r}}(p_s | \underline{r}) dp_s - [f_{\varphi}(\varphi_M)]^2 / \Sigma_{\varphi}^2 \quad (4.13) \\
 &= \int_D^{\underline{r}_{\max}} f_{\underline{r}}(\underline{r}) f_{\varphi_M | \underline{r}}(\varphi_M | \underline{r}) \beta \underline{r}^{-n} \exp(2\sigma_p^2) d\underline{r} - \frac{[f_{\varphi}(\varphi_M)]^2}{\Sigma_{\varphi}^2}
 \end{aligned}$$

The physical significance of V_p is that it describes the power fluctuations in a local area along the direction φ_M .

4.2.3 Doppler Spectrum

The received signal at the MS undergoes Doppler spread due to the motion of the mobile unit. Signals arriving from scatterers directly ahead of the vehicle will be shifted to a higher frequency whereas those directly behind will be shifted negatively. The Doppler spectrum is given by $S(\omega) = \chi f_{\omega}(\omega)$, where χ is the average power, and $f_{\omega}(\omega)$ is the pdf of the distribution of Doppler frequency. It is shown in [PRR02] that

$$f_{\omega}(\omega) = \frac{f_{\varphi}(\theta_v + |\cos^{-1}(\omega/\omega_m)|)}{\omega_m \sqrt{1 - (\omega/\omega_m)^2}} + \frac{f_{\varphi}(\theta_v - |\cos^{-1}(\omega/\omega_m)|)}{\omega_m \sqrt{1 - (\omega/\omega_m)^2}} \quad |\omega| \leq \omega_m \quad (4.14)$$

where ω_m is the maximum Doppler shift and can be represented by $\omega_m = v/\lambda$ with v being the traveling speed of the vehicle and λ the wavelength. θ_v is the angle between the direction of motion of the MS and the direct path from the BS to MS (see Fig. 4.2(a)). The expression of $f_{\varphi}(\cdot)$ is given by (4.9).

4.3 Numerical Results and Comparisons with Measurements

In street-dominated environments, waves with long excess-delay travel along the street, both from the front and back [KRB00, FRB97, COX73]. In the latter case, longer echoes arriving from the back arise from distant extended objects blocking off the end of the street. This situation is depicted in Fig. 4.1, which is defined as *Street_Classic* in [KRB00] and can be equivalently represented in the geometric model by introducing a secondary resource (BS2) at the back of the mobile terminal (refer to the simplified representation in Fig. 4.1). In this section, we present results for this type of environment. The main purpose of this section is twofold. First, we would look into the effects of different street widths on the channel properties, where the width is generally related to the standard deviation of r , σ_r . To establish the validity of the model and the expressions provided, we would further compare the PAS to the experimental data available in [KRB00]. Kuchar *et al.* [KRB00] performed measurements of angular power distribution at the MS in downtown Paris at 890 MHz. The average rooftop level was about 25-35 m and the BS antenna was placed about 47 m above ground level. The transmitter was a 10-dBi gain omnidirectional antenna while the receiver was a quarter wavelength monopole mounted on a conducting shield and placed at a height of 2 m. RX27 was a representative location for *Street_Classic* with no LoS conditions present during the measurements. At RX27 the receiver was located in *Rue de Rivoli*, a 27-m-wide street, about 700 m away from the transmitter. The PAS for large delays with an excess delay $\tau > 0.4 \mu\text{s}$ at RX27 was presented in [KRB00] and will be compared with the theoretical results from the model. Second, the RF Doppler spectrum obtained using the PSM would be compared with the empirical results in [COX73]. Cox [COX73] conducted wideband measurements in a heavily built-up urban mobile

Table 4.1. Parameters used in the numerical examples

Parameters	Numerical Values
BS1-to-MS distance, D_1	700m
BS2-to-MS distance, D_2	900m
Width of the clear zone for BS1, a_1	200m
Width of the clear zone for BS2, a_2	700m
Path loss exponent, n	2
Attenuation loss of the scattering area, β	1
Direction of the vehicle motion for the DS in Fig. 4.6(a)	10°
Direction of the vehicle motion for the DS in Fig. 4.6(b)	0°

radio environment at 910 MHz in New York City. The statistics covered vehicle traveling-distances on the order of 30 m along streets. The BS antenna was located on top of a 120-m high building, which was tall relative to the streets though not so when compared to the surroundings. The receiver was installed in a vehicle about 2½ meters high with an omnidirectional quarter wavelength vertical antenna. The two RF Doppler spectra were obtained at the narrow Pine Street, where the tall buildings on both sides formed a deep canyon. The Doppler shift at the excess delay of 0 μ s was caused by the energy, which was either forward scattered or reflected from buildings ahead of the vehicle, and the Doppler shift at 8 μ s was mainly contributed by the energy reflected from Brooklyn traveling straight down the street from behind the vehicle. The former could be reproduced using the PSM with the signal source BS1 and the latter with the pseudo signal source BS2 as illustrated in the simplified representation in Fig. 4.1. In all the numerical examples, the Tx-to-Rx distance is 700 m for BS1 and 900 m for BS2, which correspond to the figures provided in [KRB00, FRB97, COX73]. Similarly, the carrier frequency is $\omega = 900$ MHz corresponding to

the frequency of channel sounders used in [KRB00, FRB97, COX73]. Lastly the pdf of r is Gaussian in all the test cases. The parameters used in the examples are summarized in Table 4.1.

From Fig. 4.3(a) it can be shown that as σ_r decreases from 25 m to 5 m to represent a decrease in the street width, the angular range of ray arrivals becomes more confined around the street direction, which can also be intuitively explained by considering that the narrower the street, the better the waveguiding effect. This is also in agreement with the observation derived from the outdoor measurement performed in downtown Paris [FRB97], in which the echoes arriving at the receiver located in the narrow street *Rue du Chien Vert* were much narrower in their angular spread as compared to the azimuthal distribution of the incoming paths in the wider street *Boulevard Beaumarchais*. Deep nulls are predicted when the azimuth angle approaches to the directions of the street canyon (0° , 180°) because the non-line-of-sight (NLoS) propagation in street-guided environments is assumed, which prohibits the occurrence of echoes from the direct path. Similar behavior has been observed in measurements [KRB00, FRB97] (see the encircled data points in Fig. 4.4). Fig. 4.3(b) quantifies the effects depicted in Fig. 4.3(a) by using the angular constriction γ . γ decreases from 0.5386 to 0.1507 as σ_r increases from 5 m to 25 m, which signifies that the angular distribution of power becomes less constricted around the two opposite directions.

Fig. 4.4 shows a comparison of the PAS with the measured data provided in [KRB00], in which the authors conjectured that the vehicle was not parked exactly in parallel with the street (90° , 270°). Hence, an attempt was made to fit the theoretical results with the measurements by rotating the PAS and varying the value of σ_r . By

inspection, the theoretical PAS gives the best fit with the measured data when $\sigma_r = 5$ m. It is worth noting that the street-width of *Rue de Rivoli* is 27 m, which is in good agreement with the value of $6\sigma_r$ (for Gaussian distribution of $\underline{r}$, more than 99% of the scatterers are confined within the range of $D \pm 3\sigma_r$).

The variances of received power in the azimuth domain for two different standard deviations of lognormal shadowing around the MS are plotted in Fig. 4.5. The variance is greater for the echoes coming from the rear of the vehicle (encircled portions in Fig. 4.5) due to an increase in multifold bounces, subsequently leading to more severe power fluctuations. Again a deep null is predicted as the AoA approaches to 180° , which may be explained by the same hypothesis used for the dips in Fig. 4.3(a). Furthermore, Fig. 4.5 shows that as σ_p increases by 0.5 unit, the variance also increases by 30 dB. Hence, the shadowing effect can be observed to have a substantial impact on the power fluctuations in a local area for the azimuth domain.

Finally a comparison was made between the RF Doppler spectrum obtained using (4.14) with the experimental results of [COX73] as shown in Fig. 4.6(a) and 4.6(b). The general multiple-peak nature of the spectrum was due to discontinuities in measuring vehicle velocity, which resulted from irregularities in the street surface, as claimed in [COX73]. In Fig. 4.6(a), an assumption that the direction of the vehicle motion was not exactly aligned with the street canyon ($\theta_v = 10^\circ$) is made in order to best fit the encircled portion. It suffices to conclude that the agreement of the PSM with measured data is much better compared with the classical Jakes spectrum, which is obtained by assuming a uniform distribution of angular power at the receiver. Note that similar “narrow” Doppler spectra were also observed in [ABK02, ZKV++03]

due to the narrow angular ranges in the azimuth plane. The superiority of the PSM over the CSM in modeling street-dominated environments is further verified in Fig. 4.6(b) for which most of the energy arrives from behind the vehicle and a negative Doppler shift is observed to dominate the spectrum.

In summary, it can be concluded that the PSM for longer-delayed waves would serve to be a more useful solution in the simulation studies of street-guided environments as compared to the typical CSM.

4.4 Conclusions

We have proposed a PSM to complement the widely used CSM for street-dominated environments. Starting from the SGBS model, discussions were brought up on the assumptions made in the PSM and expressions were provided for certain important channel properties. The recently introduced AC was used to provide a quantitative measure on the constriction of received signals in the azimuth domain. The model results were then compared with the measurement data available for street-dominated environments, which include the PAS and DS. Finally it was concluded that the PSM for longer-delayed waves would serve to be a more useful solution in the simulation studies of street-guided environments as compared to the typical CSM.

Fig. 4.3. (a) PAS and (b) AC for the PSM with different values of σ_r

Fig. 4.4. PAS for the PSM with $\sigma_r = 5$ m compared with measurements conducted in downtown Paris [KRB00]

Fig. 4.5. Variance of received power in the azimuth domain for the PSM with $\sigma_r = 5$ m

Fig. 4.6. RF Doppler spectrum due to (a) BS1 (actual signal source) and (b) BS2 (pseudo signal source) using the PSM compared with measurements conducted in New York City [COX73]

CHAPTER 5 EFFECT OF ANTENNA DIRECTIVITY ON ANGULAR POWER DISTRIBUTION AT THE MOBILE IN URBAN MACROCELLULAR ENVIRONMENTS

5.1 Introduction

In this chapter we provide a systematic way to analyze the effect of antenna directivity on the spatial pattern of scattering elements. Starting with the unified analysis presented in Chapter 3, Section 5.2 then proposes the versatile von Mises function to model the antenna power pattern because it includes and/or approximates many important functions like uniform, impulse, cardioids, Gaussian and wrapped Gaussian [MAR72], and thus is flexible for quantitative analysis. We also model the probability density function (pdf) of the geometry-dependent parameter (GDP) r as an exponential power distribution for a wide variety of scatterer clustering situations. Section 5.3 derives the important channel parameters including the angular power distribution (APD), effective power azimuthal spectrum (EPAS), Doppler spectrum, and multipath shape factors [DR00] based on the proposed model. Subsequently, Section 5.4 presents various numerical examples. Finally some concluding remarks are given in Section 5.5.

5.2 Modeling the Antenna Power Gain Pattern and Distribution of Geometry Dependent Parameter

Fig. 5.1. Von Mises antenna beam patterns

We propose the versatile von Mises function to represent the directive gain of antennas at the transmitter (Tx) or the receiver (Rx). After proper normalization, the von Mises function reads

$$G(\varphi) = \frac{\exp[\mu \cos(\varphi - \varphi_p)]}{I_0(\mu)} \quad (5.1)$$

where $I_0(\mu)$ is the zero-order modified Bessel function, φ_p accounts for the direction of maximum directive gain, and μ controls the width of the antenna beam. The pattern defined here is single smooth-lobed over the whole angular region and any sidelobe structure is assumed to have a negligible effect¹. Fig. 5.1 shows $G(\varphi)$ in linear coordinates for different values of μ and $\varphi_p = 0$. For $\mu = 0$, the antenna is isotropic; while $\mu \rightarrow \infty$ yields $G(\varphi) = \delta(\varphi)$ where $\delta(\cdot)$ is the delta function representing an extremely narrow beam. A useful geometrical explanation of μ is that for large μ , the inflexion points of $G(\varphi)$ are approximately equal to $\pm 1/\sqrt{\mu}$.

¹ The von Mises function assumption may not be accurate, but it captures the important feature of every antenna power pattern, namely a controllable null-to-null beamwidth. In fact, as to be made clear in the following discussion, the general methodology of analyzing antenna-directivity effect on scatterer distribution applies to any antenna patterns.

Therefore, for a unidirectional antenna beam pattern, the null-to-null beamwidth is roughly equal to $2/\sqrt{\mu}$. Also, when μ is large the function resembles a Gaussian pdf with mean φ_p and a standard deviation $1/\sqrt{\mu}$, which is similar to the circular function in $\cos^\mu \varphi$ form proposed in the context of mobile communications by Lee [LEE73].

Making use of (3.11) and (5.1), the effective path length (EPL) when directional antennas are employed at both ends of the radio link can be obtained as

$$\underline{r} = \left[I_0(\mu_T) I_0(\mu_R) \right]^{\frac{1}{n}} \exp \left[-\frac{\mu_T \cos(\varphi_T) + \mu_R \cos(\varphi_R)}{n} \right] (r_R * \eta r_T) \quad (5.2)$$

We further define the bijection in the bijective mapping rule as a mirror mapping

$$\underline{r} = \mathcal{G}^{-1}(\hat{\underline{r}}) = \hat{\underline{r}} \quad (5.3)$$

Equation (5.3) defines the geometry of the scattering area in terms of the parametric function of $\underline{r}$.

Based on the tapering NSDIM rule, $f_{\underline{r}}(\underline{r}, l)$ should decrease monotonically as $\underline{r}$ increases. As such, the exponential power distribution proposed in Section 4.2.1 is used here to represent the distribution of $\underline{r}$, which is expressed in a different form as

$$f_{\underline{r}}(\underline{r}) = \frac{c_1(\alpha)}{\sigma} \exp \left[-c_2(\alpha) \left| \frac{\underline{r} - \underline{r}_{\min}}{\sigma} \right|^{\frac{2}{1+\alpha}} \right] \quad \underline{r} \geq \underline{r}_{\min} \quad (5.4)$$

with

$$c_1(\alpha) = \frac{\left\{ \Gamma[3(1+\alpha)/2] \right\}^{1/2}}{(1+\alpha) \left\{ \Gamma[(1+\alpha)/2] \right\}^{3/2}} \quad \text{and} \quad c_2(\alpha) = \left\{ \frac{\Gamma[3(1+\alpha)/2]}{\Gamma[(1+\alpha)/2]} \right\}^{\frac{1}{1+\alpha}}$$

where the coefficient of kurtosis $\kappa = \Gamma[5(\alpha+1)/2] \cdot \Gamma[(\alpha+1)/2] / \{ \Gamma[3(\alpha+1)/2] \}^2 - 3$. The minimum value of $\underline{r}$ is $\underline{r}_{\min}$. Γ is the gamma function, the scale parameter $\alpha > -1$ and the variance of the distribution is σ^2 . For $\alpha = 0$, the pdf corresponds to a normal

Fig. 5.2. Some members of the exponential power distribution family, $\sigma = 100$ m

distribution, while for $\alpha = 1$ the pdf is Laplacian. As α tends to -1, the pdf tends toward a rectangular distribution. Some members of the family are presented in Fig. 5.2. As a final remark of this section, we should point out that the assumptions of von Mises antenna beam pattern and exponential power distribution of $\underline{r}$ are an oversimplification in the model, but they are both mathematically convenient.

5.3 The SGBS Macrocell Model Inclusion of Antenna Pattern (SGBSM-AP)

5.3.1 General Remarks

In macrocellular environments, antenna heights are relatively large. There will be no signal scattering from locations near the base station (BS). If diffuse scattering is the principal mechanism of signal propagation and the scattering objects are densely

clustered around the mobile station (MS), in such a case, $r_R^* \eta r_T \approx \eta D r_T$ for uplink transmission where D is the direct path between the BS and MS. This results in a circular scattering model when omnidirectional antennas are used at both ends of the radio link, which has been investigated in Section 3.3.1. In the following discussion, the subscripts M and B will be used to replace T and R. Subsequently, we assume that a pencil-beam antenna is used at the BS to illuminate the scattering region around the MS. In such a case, μ_B is large and the von Mises function resembles the truncated Gaussian function. The antenna pattern can thus be rewritten as

$$G_B(\varphi_B) \approx \Sigma_G \exp\left(-\frac{\mu_B \varphi_B^2}{2}\right) \quad |\varphi_B| \leq \pi \quad (5.5)$$

with Σ_G being a normalization factor. It is mathematically convenient to write the expression of $\underline{r}$ in terms of the polar coordinates (r_M, φ_M) . Based on the elementary geometric relationship shown in Fig. 3.3, φ_B can be expressed as follows by making use of the small angle approximation

$$\varphi_B \approx \tan(\varphi_B) = \frac{r_M \sin \varphi_M}{D - r_M \cos \varphi_M} \approx \frac{r_M \sin \varphi_M}{D} \quad (5.6)$$

Substituting (5.5) and (5.6) into (5.2) and (5.3), $\underline{r}$ can be rearranged as follows

$$\underline{r} = \left(\frac{I_0(\mu_M)}{\Sigma_G} \right)^{\frac{1}{n}} \exp\left(-\frac{\mu_M \cos \varphi_M - \mu_B r_M^2 \sin^2 \varphi_M / 2D^2}{n}\right) (\eta D r_M) \quad (5.7)$$

Equation (5.7) defines the shape of the scattering area in an urban macrocell. As a first inspection, (5.7) may be written in the form of $\underline{r} \approx [I_0(\mu_M)]^{1/n} \cdot \eta D r_M \cdot \exp[-\mu_M \cos(\varphi_M)/n]$ by ignoring the impact of directional antenna at the BS. When $\underline{r}$ remains unchanged, as φ_M increases from 0° to 180° , r_M decreases. Apparently, the curve representing an aggregate of scatterers with the same $\underline{r}$ is *neither a circle nor*

part of a circle obtained by the deletion from the constellation of any scatterers not contained within the antenna beam sector. Rather, the shape is warped and it has a peak value in the MS-to-BS direction. From a physical propagation viewpoint, the stochastic distribution of scatterers is altered due to the uneven distribution/collection of signal energy at the MS/Bx end.

A useful analogy exists between the abovementioned phenomenon and the rules governing the electric field intensity in electromagnetic theory. Consider an initially uniform electric field sketched in Fig. 5.3(a) (*in a similar manner, consider the spatial pattern of scatterers sketched in Fig. 5.3(b) when direction-insensitive sensors are used at two ends of the radio link*). After the conducting sphere is introduced into the electric field (*direction-sensitive sensors are employed*), a redistribution of charges will take place in such a way that the electric field intensity within the sphere is zero and outside the sphere the field lines will intersect with the surface normally as shown in Fig. 5.3(c) (*reshuffling of spatial scattering points takes place in order to satisfy the two heuristic rules proposed from physical wave-propagation viewpoint*). Therefore, the field lines are not simply disunited by the sphere surface, but curved and twisted in order to satisfy the boundary condition of the electrostatic problem (*sample space of scattering points are not obtained by simply deleting the constellation of points not contained within the antenna beam sector. Instead, the sample space is defined by equation (5.7) as pictured in Fig. 5.3(d)*).

5.3.2 Directional Antenna Used at the MS and Omnidirectional Antenna at the BS

Fig. 5.3. An analogy to illustrate the behavior of the antenna directivity altering the spatial pattern of scatterers

Mobile terminal antennas used next to the head in speaking position will be directional because part of the radiation facing the head will be reflected and partly attenuated [KP02]. Nevertheless, due to the constraint of mobile size, the beamwidth of the directional antenna is usually large. The expression of GDP $\underline{r}$ in (5.7) can be simplified as

$$\underline{r} = \left(I_0(\mu_M) \right)^{\frac{1}{n}} \exp\left(-\frac{\mu_M \cos \varphi_M}{n} \right) \cdot \eta D r_M \quad (5.8)$$

Solving (5.8) for r_M gives the subsequent close-form formula

$$r_M = \frac{[I_0(\mu_M)]^{\frac{1}{n}} \cdot r}{\eta D} \exp\left(\frac{\mu_M \cos \varphi_M}{n}\right) \quad (5.9)$$

The circumference of the curve $\underline{C}$ parameterized by $\underline{r}$ is given as

$\ell_C = \int_{-\pi}^{\pi} \sqrt{r_M^2 + [r'_M(\varphi_M)]^2} d\varphi_M$. It is then straightforward to show that the pdf of angle-of-departure (AoD) at the MS conditioned on $\underline{r}$ can be written as

$$f_{\varphi_M|\underline{r}}(\varphi_M|\underline{r}) = \sqrt{r_M^2 + [r'_M(\varphi_M)]^2} / \ell_C \quad (5.10)$$

5.3.2.1 Angular Power Distribution at the MS

Following from the assumption made in Section 3.2.2, the instantaneous power at the MS due to reradiation from scatterer S is lognormally distributed. Therefore, the joint pdf of the instantaneous power, p_s , and AoD at the MS, φ_M , conditioned on the area mean power p , can be expressed as

$$f_{p_s, \varphi_M|p}(p_s, \varphi_M|p) = f_{p_s|p}(p_s|p) f_{\varphi_M|p}(\varphi_M|p) \quad (5.11)$$

where $p = \beta(\underline{r})^{-n}$ with β being the average attenuation loss of the scattering objects.

Applying the transformation of variables to (5.11), we can obtain

$$f_{p_s, \varphi_M|\underline{r}}(p_s, \varphi_M|\underline{r}) = f_{p_s, \varphi_M|p}(p_s, \varphi_M|p) \Big|_{p=\beta \underline{r}^{-n}} \quad (5.12)$$

We thus obtain the APD at the MS after some straightforward manipulation as

$$\begin{aligned} f_{\varphi}(\varphi_M) &= \sum_{\varphi} E\{p_s(\varphi_M)\} = \sum_{\varphi} E\{E\{p_s(\varphi_M)|\underline{r}\}\} \\ &= \sum_{\varphi} \int_{\underline{r}} p_s f_{p_s, \varphi_M|\underline{r}}(p_s, \varphi_M|\underline{r}) f_{\underline{r}}(\underline{r}) d\underline{r} dp_s \\ &= \sum_{\varphi} \int_{\underline{r}} f_{\underline{r}}(\underline{r}) f_{\varphi_M|\underline{r}}(\varphi_M|\underline{r}) \left[\int p_s f_{p_s|\underline{r}}(p_s|\underline{r}) dp_s \right] d\underline{r} \\ &= \sum_{\varphi} \int_{\underline{r}_{\min}}^{\infty} \frac{c_1(\alpha)}{\sigma} \exp\left[-c_2(\alpha) \left| \frac{\underline{r} - \underline{r}_{\min}}{\sigma} \right|^{\frac{2}{1+\alpha}}\right] \frac{\sqrt{r_M^2 + [r'_M(\varphi_M)]^2} \beta \underline{r}^{-n}}{\ell_C} d\underline{r} \end{aligned} \quad (5.13)$$

where Σ_φ and Σ_φ' are normalization factors to ensure that $f_\varphi(\varphi_M)$ is a pdf, Ξ is the integration region and r_M is given in (5.9).

5.3.2.2 Effective Power Azimuthal Spectrum (EPAS)

The EPAS is referred to as the effective angular distribution of ray arrivals after the antenna effects are deconvolved from the measured APD. This metric can be expressed as

$$\tilde{f}_\varphi(\varphi_M) = k \frac{f_\varphi(\varphi_M)}{G_M(\varphi_M)} \quad (5.14)$$

where k is a normalization constant to ensure that $\tilde{f}_\varphi(\varphi_M)$ is a pdf. More details on the EPAS will be given in Chapter 7. It is worth emphasizing here that the environment response $\tilde{f}_\varphi(\varphi_M)$ shall not be defined separately from the antenna pattern as commonly practiced. The underlying statement is that it may cause systematic prediction errors if others want to use the same set of data to study the overall channel response when different directional antennas are to be used, even for analysis in the same physical environment, because the sample space of scatterers, or the channel connecting the Tx and Rx has been altered.

5.3.2.3 Doppler Spectrum at the MS

Following from the discussion in Section 4.2.3, the Doppler spectrum is given by $S(\omega) = \chi f_\omega(\omega)$, where χ is the average power, and

$$f_\omega(\omega) = \frac{f_\varphi(\theta_v + |\cos^{-1}(\omega/\omega_m)|) + f_\varphi(\theta_v - |\cos^{-1}(\omega/\omega_m)|)}{\omega_m \sqrt{1 - (\omega/\omega_m)^2}} \quad |\omega| \leq \omega_m \quad (5.15)$$

where ω_m is the maximum Doppler shift, θ_v is the angle between the direction of motion of the MS and the direct path from the BS to MS. The expression of $f_w(\cdot)$ is given by (5.13). It is worth mentioning here that the antenna effect has been taken into account in (5.13).

5.3.2.4 Multipath Shape Factors and Forbidden Zones of Correlation

In the future, more than one antenna in a mobile terminal could be a reality [KVP00]. Multiple-input-multiple-output (MIMO) schemes require different antennas to experience unique radio channels [PNG03]. Therefore, knowledge of the correlation is crucial to calculate the performance of a mobile terminal including an antenna system. The shape factors as discussed in Section 4.2.2 can also be applied to relate the azimuth angle spectrum and antenna gains to the properties of spatial selectivity. The angular spread and angular constriction are described as [DR00, DUR03]

$$\Lambda = \sqrt{1 - \frac{|F_1|^2}{F_0^2}} \quad (5.16)$$

$$\gamma = \frac{|F_0 F_2 - F_1^2|}{F_0^2 - |F_1|^2} \quad (5.17)$$

where F_n is given by

$$F_n = \int_0^{2\pi} f_\varphi(\varphi_M) \exp(jn\varphi_M) d\varphi_M \quad (5.18)$$

Multipath shape factor theory allows the system designers to study correlation properties as a function of azimuthal orientation, φ , in space instead of one-dimensional position [DUR03]:

Fig. 5.4. (a) The geometry of element separation and (b) the forbidden zone of correlation determined by the multipath shape factors and the correlation criterion

$$-\ln \mathcal{R} = 23\Lambda^2 \left(1 + \gamma \cos[2(\varphi - \varphi_{\max})]\right) \left(\frac{\Delta x}{\lambda}\right)^2 \quad (5.19)$$

where $\mathcal{R}$ is the maximum acceptable amount of correlation between two coplanar antenna elements, $\varphi_{\max}$ is the direction of maximum fading, and Δx is a small spatial

separation (see also Fig. 5.4(a)). Equation (5.19) defines an elliptical forbidden zone of correlation as shown in Fig. 5.4(b). For the space inside the ellipse, the envelope correlation is higher than $\mathcal{R}$; outside the ellipse, the envelope correlation is lower than $\mathcal{R}$. Therefore, the size and shape of the ellipse is a useful tool for spacing antenna elements.

5.3.3 Directional Antenna Used at the BS and Omnidirectional Antenna at the MS

In this case, (5.7) may be further simplified as

$$\begin{aligned} \underline{r} &= \Sigma_G^{-1/n} \exp\left(\frac{\mu_B r_M^2 \sin^2 \varphi_M}{2nD^2}\right) \cdot \eta D r_M \\ &\approx \eta D \Sigma_G^{-1/n} r_M \left(1 + \frac{\mu_B r_M^2 \sin^2 \varphi_M}{2nD^2}\right) \\ &= a_1 r_M^3 \sin^2 \varphi_M + a_2 r_M \end{aligned} \quad (5.20)$$

which is obtained by ignoring the higher-order terms of the Maclaurin series of the exponential function for $r_M \ll D$. $a_1 = \eta D (\Sigma_G)^{-1/n} \mu_B / (2nD^2)$ and $a_2 = \eta D (\Sigma_G)^{-1/n}$.

Solving the third-order equation for r_M gives the following close-form formula

$$r_M = \sqrt[3]{-\frac{p_1}{2} + \sqrt{\left(\frac{p_1}{2}\right)^2 + \left(\frac{p_2}{3}\right)^3}} + \sqrt[3]{-\frac{p_1}{2} - \sqrt{\left(\frac{p_1}{2}\right)^2 + \left(\frac{p_2}{3}\right)^3}} \quad (5.21)$$

where $p_1 = -\underline{r} / (a_1 \sin^2 \varphi_M)$ and $p_2 = a_2 / (a_1 \sin^2 \varphi_M)$. Following the similar procedures in the previous section, the APD and other critical parameters can be readily obtained.

As a final remark, though the analysis presented here only focuses on the angular domain, the usefulness of the generic approach is certainly not limited to this domain.

Owing to the ray-tracing nature of the QSB models, behaviors of the mobile radio channel in other domains can be easily acquired.

5.4 Numerical Examples

The main purpose of this section is twofold. We first look into the effect of antenna directivity on the channel characteristics when direction-sensitive antenna is used at either the BS or MS (*case study 1*). The comparison is made under the same channel conditions except for the antenna power gains. The distribution of $\underline{r}$ is assumed to be Gaussian. Second, we consider the influence of different distributions of $\underline{r}$ (*case study 2*), which reproduce different conditions of scatterer clustering around the MS. In both the two scenarios, the distance of separation between the BS and MS is 1 km. The path loss exponent $n = 2$ and the attenuation loss of the local scattering area $\beta = 1$. The carrier frequency $f = 1800$ MHz. Furthermore, the Nakagami fading loss factor (NFLF) $\eta = 2.4088$ corresponding to a Nakagami fading shape parameter $m = 3$, which is typical in urban macrocellular environments [STU01].

5.4.1 Case Study 1

We compare four different values of μ for directive antennas at the BS and MS, respectively. The following two sets of data are used, $\mu_M = [0.5 \ 2 \ 5 \ 10]$ and $\mu_B = [20 \ 80 \ 200 \ 640]$. The diagram of the antenna beam patterns for different μ is illustrated in Fig. 5.1. Note that due to the size constraint, it is more difficult to form a narrow beam at the MS. The shape parameter of the exponential power distribution $\alpha = 0$

(Gaussian distribution) and the standard deviation of scatterer distribution is 100 m (for the scaled GDP $\underline{r}' = \underline{r} / \eta D$).

Fig. 5.5 illustrates the scatterer constellation when directional antenna is used at either the mobile terminal or base station, respectively. The MS is located at the origin of the polar plot and $\varphi_M = 0$ corresponds to the direction of LoS path between the MS and BS. As can be seen from Fig. 5.5, the geometry deviates from a conventional circular scattering model. When directional antenna is used at the MS (Fig. 5.5(a)), scatterers are densely clustered near the LoS path and become sparser as φ_M increases from 0° to 180° (or in a symmetric manner, as φ_M reduces from 360° to 180°). On the other hand, Fig. 5.5(b) demonstrates that when the directional antenna is employed at the BS, the spatial configuration is similar with an eye diagram. Note that the symmetry with respect to y -axis is due to the small-angle approximation applied in equation (5.6). Still another trend may be observed from Fig. 5.5. When r_M is small, the distribution is approximately a circle. As r_M increases, the curve is deformed more severely, indicating that the effect of antenna directivity becomes more remarkable.

The non-uniform distribution in azimuth at the MS has also been verified in the empirical measurements reported by Knudsen *et al.* [KP02], for which an outdoor BS and an indoor mobile are assumed for system operation. At the MS the major part of the signal energy comes through openings in the building such as windows. Since typically, only one of the sides in a room has windows, and it is seldom that a room has windows on all sides, Knudsen *et al.* have proposed a new statistical spherical outdoor to indoor power spectrum model to calculate the directional performance of MS with a single or multiple antennas, which consists of a major scattering area in

one direction and more uniformly distributed minor scatterers in the other directions. It is worth mentioning here that the “opening” in the building in effect acts as the extended aperture of a virtual directional antenna used at the MS. Thus, our model is also applicable to the scenario of outdoor to indoor propagation.

The APD at the MS when directional antenna is used at the mobile terminal is illustrated in Fig. 5.6(a). As may be predicted from Fig. 5.5(a), the APD should have a higher value for ray arrivals closer to the direct path when directional antenna is used at the MS, which has been verified in Fig. 5.6(a). However, for a large value of μ_M , the model results predict that the angular response does not peak at $\varphi_M = 0^\circ$ or 360° due to the severely curved scatterer constellation. It is also evident from Fig. 5.6(b) that the angular dispersion of the EPAS decreases as the antenna becomes more directive, which confirms with the statement we have made earlier in the abstract – the effect of directional antennas is twofold. First, it provides a constricted sample space for the random field and hence reshuffles the spatial distribution of scatterers. Second, it distributes the signal components into AoD and/or collects them from AoA by weighted combination.

The Doppler spectra for $\theta_v = 0^\circ$ are presented in Fig. 5.7. As μ_M increases, the “horned” Doppler spectra are changed into the “narrow” Doppler spectra (the names are quoted from [ZKV03]), implying a channel with slower temporal fluctuations and a larger coherence time. Thus, the effects of multipath interference that prevails in wireless communication systems can be considerably mitigated by using a directional antenna at the mobile unit. Note that similar conclusions are drawn in [AJ93, ND02, CLA68] for a simplified model constellation and antenna beam pattern.

The trend between multipath shape factors and μ_M is depicted in Fig. 5.8(a). As the antenna at the MS becomes more directive, the angular spread decreases whereas the angular constriction increases. Subsequently, a channel with a smaller Λ and a larger γ will likely require more than two elements. A design with two elements in the worst-case scenario (i.e., the decorrelation distance is the major axis of the forbidden zone ellipse) would require about 0.1 m of separation for $\mathcal{R} = 0.6$ and $\mu_M = 10$ as depicted in Fig. 5.8(b) – impractical for a mobile phone but feasible for some other user terminals such as laptops. Another important trend observed from Fig. 5.8 is that as μ_M increases, the eccentricity approaches to 1 and the corresponding forbidden zone stretches along its major axis. In such a case, a design with two elements in the average-case scenario (i.e., the decorrelation distance is an intermediate value between the major and minor axes of the forbidden zone) would create some inadequate, high-correlation links in the neighborhood.

The APD at the MS when directional antenna is used at the BS is shown in Fig. 5.9. As may be predicted from Fig. 5.5(b), multipath tends to cause an 180° bifurcation in the arriving propagation. Thus, the APD of the environment response indicates a double-sectored multipath model as depicted in Fig. 5.9. Similar with Fig. 5.6(a), two deep nulls at $\varphi_M = 0^\circ$ and $\varphi_M = 180^\circ$ are predicted for a large value of μ_B due to the curving of the scatterer distribution. The double-sector azimuth spectrum always has an angular spread of $\Lambda = 1$. It can be seen from Fig. 5.10 that the shapes of Doppler spectra are close to the “horned” Doppler spectra. Different antenna directivities just change the relative amplitudes of the spectra. Therefore, level-crossing-rate in the fading envelope would be less sensitive to the BS antenna beam patterns.

The relationship between the angular constriction and μ_B is depicted in Fig. 5.11(a). As the antenna at the BS becomes more directive, the angular constriction increases. For the case of $\mu_B = 640$, we get an angular constriction of $\gamma = 0.23$. A design with two elements in the worst-case scenario would require less than 3 cm of separation (Fig. 5.11(b)) – a rather reasonable amount for a mobile unit. Moreover, the eccentricity of the forbidden zone is small, resulting in a desirable shape for spacing antenna elements using the average-case-scenario design.

5.4.2 Case Study 2

In this case, we set the antenna parameters $\mu_M = 2$ when the directional antenna is implemented at the MS and $\mu_B = 80$ (very narrow beam) when the directional antenna is at the BS. However, we compare four different values of the shape parameters α ($\alpha = [-0.99 \ 0 \ 1 \ 5]$) with the same standard deviation of scatterer distribution.

As shown in Fig. 5.12(a), when directional antenna is at the MS, the change in APD at the MS is unobservable as the distance measure of scatterer distribution from gaussianity changes. However, in the case when directional antenna is at the BS, Fig. 5.12(b) shows that the angular constriction will slightly decrease if the distribution of r has a heavier tail weight (or equivalently the number of relevant scatterers drops in a slower rate). In general, the effect of nongaussianity on the channel characteristics is insignificant in either case when the directional antenna is used at the MS or BS.

5.5 Conclusions

A general investigation of the effect of antenna directivity on the spatial pattern of scattering points has been presented in this chapter. The main thrust is to achieve

an analytically tractable channel model, incorporating the feature of antenna and channel properties. It has been made clear that the scatterer distribution for directional antennas is not just a subset of the distribution for omnidirectional antennas. Armed with the generic model presented in Chapter 3, we have provided a quantitative analysis on the geometric distribution of scatterers when antennas with different beam patterns are used at either end of the radio link. The versatile von Mises function including a variety of nonisotropic scenarios is proposed to parameterize the beam shapes and the exponential power distribution encompassing various classes of densities is used to model the distribution of r . Specifically, we have presented the model in urban macrocell environments. The utility of the proposed model has been illustrated by applying it to study the channel characteristics in several numerical examples. It has been further demonstrated that implementing the directional antenna at the MS is more effective than implementing it at the BS for mitigation of multipath interference. Nevertheless, the former introduces a much larger forbidden zone of correlation as compared to the latter, which is undesirable when the MIMO systems are to be implemented to improve the link quality or increase the system capacity.

Fig. 5.5. Spatial pattern of scattering points when (a) directional antenna with $\mu_M = 2$ is used at the MS and omnidirectional antenna at the BS and (b) directional antenna with $\mu_B = 20$ is used at the BS and omnidirectional antenna at the MS

Fig. 5.6. (a) APD and (b) EPAS at the MS when directional antennas with different μ_M are used at the MS

Fig. 5.7. Normalized Doppler spectrum when directional antennas with different μ_M are used at the MS

Fig. 5.8. (a) The trend between multipath shape factors and μ_M and (b) the diagram of the corresponding forbidden zones ($\mathcal{R} = 0.6$)

Fig. 5.9. (a) APD at the MS when directional antennas with different μ_B are used at the BS

Fig. 5.10. Normalized Doppler spectrum when directional antennas with different μ_B are used at the BS

Fig. 5.11. (a) The trend between multipath shape factors and μ_B and (b) the diagram of the corresponding forbidden zones ($\mathcal{R} = 0.6$)

Fig. 5.12. APD at the MS for different scatterer clustering situations when (a) directional antenna with $\mu_M = 2$ is used at the MS and (b) directional antenna with $\mu_B = 80$ used at the BS. For the sake of clarity, the curves for $\alpha = 0, 1$, and 5 in (a) have been shifted upward by $0.1, 0.2$, and 0.3 units, respectively.

CHAPTER 6 GEOMETRIC CHANNEL MODEL FOR UWB SIGNAL PROPAGATION

6.1 Introduction

This chapter aims to provide a robust and reliable geometric channel model (GCM) incorporating frequency-dependent features of rays. As mentioned in Chapter 2, the pulse shape for the ultrawideband (UWB) signal associated with a propagation path is frequency-dependent. Hence the distribution of scatterers will be defined in a hypothetical three-dimensional (3D) space, where two dimensions are for the azimuth domain and one for frequency domain. Three heuristic rules will be proposed to provide the link between the canonical model and the physical environment it represents, which are an extension of the analysis presented in Chapter 3 to UWB signal propagation in an indoor environment.

This chapter is structured as follows. Section 6.2 presents the wideband semi-geometrically based statistical (SGBS) model which is an extension of the narrowband SGBS model described in Chapter 3. The three heuristic rules are also introduced in this section. Section 6.3 presents the azimuth-frequency domain geometric microcell (AFDGM) model and exploits the effect of different frequency components on the scatterer density based on the analysis in Section 6.2. It also gives the general formulation of the channel properties such as power azimuthal spectrum (PAS) and power delay spectrum (PDS). Section 6.4 studies the application of the AFDGM model in an indoor microcell environment for UWB wireless propagation channel, and finally Section 6.5 presents a number of conclusions.

6.2 Extension of the Unified Analysis to Wideband Channels

6.2.1 The Wideband SGBS Model

The following wideband SGBS model is implemented by taking into account the frequency-selective effects. Fig. 6.1 shows the multipath propagation from the transmitter (Tx) to the receiver (Rx). The following terms are further introduced:

S : The scatterer contributing to the final reradiation of the incoming multipath components (MPCs) to the Rx.

$r_R(r_T)$: The distance between the Rx (Tx) and the scatterer S .

$\hat{S}_1(\hat{S}_2)$: Local objects around S , scattering from which results in local subpaths.

$\tilde{R}_1(\tilde{R}_2)$: Objects around the Tx, which cause the MPCs to undergo multiple reflections and/or diffractions.

$\beta(\omega)$: Scattering coefficient of S for the frequency component ω .

It is assumed that the Tx-to-Rx path has a shadowed Nakagami fading. Subsequently,

Fig. 6.1. Pictorial representation of scattering from a scatterer S

the normalized area mean power received at S due to the continuous-wave component has the general form

$$p_{a,T} = |V(\omega)|^2 (\omega/\omega_0)^{-n_w} r_T^{-n} \quad (6.1)$$

where $|V(\omega)|^2$ is the power spectrum of the signal pulse, n is the path loss exponent, ω the frequency, ω_0 the reference frequency, and n_w characterizes the frequency dependency of the path loss.

Assuming shadowing to be superimposed on (6.1), the local mean power is lognormally distributed about the mean, viz.,

$$f_{p_{l,T}}(p_{l,T}) = \frac{1}{p_{l,T} \sigma_T \sqrt{2\pi \ln 10/10}} \exp \left\{ -\frac{(10 \log_{10} p_{l,T} - p_{a,T}(\text{dBm}))^2}{2\sigma_T^2} \right\} \quad (6.2)$$

where σ_T is the standard deviation of the shadowing expressed in dB. Multipath reception at the scatterer causes Nakagami fading, with the squared envelope having a Gamma-log-normal density function. As shown in Section 3.2.2, the composite probability density function (pdf) can be approximated by a lognormal distribution with mean and standard deviation

$$p_{a,T}(\text{new}) = p_{a,T} 10^{10[\psi(m) - \ln(m)]/\ln 10} \quad (6.3)$$

$$\sigma_T^2(\text{new}) = (\ln 10/10)^{-2} \zeta(2, m) + \sigma_T^2 \quad (6.4)$$

where $\psi(\cdot)$ is the Euler *psi* function, $\zeta(\cdot)$ is the Riemann's *zeta* function and m is the shape parameter of the Gamma distribution.

Let $p_{a,T}(\text{new}) = |V(\omega)|^2 (\omega/\omega_0)^{-n_w} [r_T(\text{new})]^{-n}$. Applying (6.1) and (6.3) to the equation yields

$$r_T(\text{new}) = \eta r_T \quad (6.5)$$

with $\eta = 10^{-10[\psi(m) - \ln(m)]/(n \ln 10)}$ being the Nakagami fading loss factor (NFLF).

Apparently, the received power reradiated by the scatterer S at the Rx is also lognormally distributed with mean and standard deviation

$$p = \left| V(\omega) \right|^2 (r_R * \eta r_T)^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega) \quad (6.6)$$

$$\sigma = \sigma_T (\text{new}) \quad (6.7)$$

The operator $*$ represents either product or sum depending on whether scattering or specular reflection dominates.

6.2.2 The Three Heuristic Rules

Originally, the GCMs have been applied to narrowband modeling. Therefore, the spatial patterns of randomly distributed scattering elements are solely defined in the physical space. Analogous to the fact that statistical modeling of the time-variant transfer function $T(t, \omega)$ can be regarded as a consistent extension of $T(t)$ to wideband channels, in the geometric modeling, the extension is realized by defining the scattering point as a vector ξ in a *virtual* space Ξ with the coordinates $(\underline{r}, l, \omega)$. $\underline{r} = \mathcal{H}(r_R, r_T)$ is the geometry-dependent parameter (GDP) as defined in Section 3.2.3 assuming that omnidirectional sensors are employed at both ends of the link. ω is the frequency and l is a point at the curve $\underline{C}$: $\underline{r}(l) = \underline{r}$ with l_s being an arbitrarily defined starting point and the positive sense on $\underline{C}$ being clockwise as illustrated in Fig. 3.3. The scattering object is now represented by a hypothetical azimuth-frequency point S (see Fig. 6.2). In our formulation of GCM, the function $\mathcal{H}$ is defined such that the scatterer density $f_{\underline{r}, l|\omega}(\underline{r}, l|\omega)$ is uniform on each oriented curve $\underline{C}$ for any ω , following which $\mathcal{H}$ determines the geometry of the canonical model. Apparently the conditional probability density function (pdf) $f_{\underline{r}|\omega}(\underline{r}|\omega)$ can be obtained

Fig. 6.2. Virtual coordinates (r, l, ω)

using the line integral of $f_{r,l|\omega}(r, l|\omega)$ over the oriented curve $\underline{C}$. viz.,

$$f_{r|\omega}(r|\omega) = \int_{\underline{C}} f_{r,l|\omega}(r, l|\omega) dl = \ell_{\underline{C}} \cdot f_{r,l|\omega}(r, l|\omega) \quad (6.8)$$

where $\ell_{\underline{C}}$ denotes the length of the curve $\underline{C}$.

Similar with the analysis presented in Section 3.2.3, we define the effective path length (EPL) as

$$\hat{r} = r_R * \eta r_T \quad (6.9)$$

with η being the corresponding NFLF in the targeted scattering cluster. Thus, the expectation of received power at the Rx from any scatterer S is inversely proportional to $\hat{r}^n$.

The following heuristic rules are thus imposed to show the relationship between the geometric distribution of scatterers $f_{\xi}(r, l, \omega)$ and the power distribution $p(\hat{r})$. These rules are an extension of the suppositions proposed in Section 3.2.3. A pictorial representation of the rules is shown in Fig. 6.3.

1. *Weak bijective mapping rule (Rule 1)*: For any frequency component ω within

the transmission bandwidth $\Delta\omega$, let $\underline{A}$ denote the set of all curves in the region Ξ_ω (Ξ_ω is the collapse of the virtual space Ξ into the 2D hyperplane at ω), with each curve representing a collection of scatterers with the same value of $\underline{r}$ and let $\hat{\underline{A}}$ denote the set of all curves with each representing a collection of scatterers with the same value of $\hat{\underline{r}}$. The mapping $\mathcal{F}: \underline{A} \rightarrow \hat{\underline{A}}$ defines a bijective function between $\underline{A}$ and $\hat{\underline{A}}$ in terms of $\hat{\underline{r}} = \mathcal{G}_\omega(\underline{r}) \quad \forall \omega$.

2. *Tapering-off conditional NSDIM rule (Rule 2)*: For any frequency component ω , if $\mathcal{G}_\omega$ defines a monotonically increasing function, the conditional pdf $f_{\underline{r},l|\omega}(\underline{r},l|\omega)$ gradually tapers off as $\underline{r}$ increases and the local maximum conditional pdf corresponds to the Tx-to-Rx path.
3. *Strong bijective mapping rule (Rule 3)*: $f_\xi(\underline{r}_1, l_1, \omega_1) = f_\xi(\underline{r}_2, l_2, \omega_2)$ when $p(\hat{\underline{r}}_1)|_{\omega=\omega_1} = p(\hat{\underline{r}}_2)|_{\omega=\omega_2}$, $p(\hat{\underline{r}}_i)|_{\omega=\omega_i}$ ($i = 1, 2$) is the average power received at the receiver via the point $(\underline{r}_i, l_i, \omega_i)$.

The two bijective mapping rules assert a one-to-one correspondence between the equi-scatterer-density curves and the equi-power curves. Rule 1 declares the relationship for a single frequency component, whereas Rule 3 is for the cross-spectrum components. The tapering-off conditional NSDIM rule actually states that as the average-received-power at the Rx due to reradiation from a scatterer at a certain location decreases, the scatterer density at that location decreases as well.

Fig. 6.3. The three heuristic rules

6.3 The AFDGM Model – A General Formulation of the Model Structure

6.3.1 General Remarks

Consider the spatial pattern of the scattering points for each frequency component ω . This can intuitively be interpreted as a slice of the overall distribution of scatterers in the higher-dimensional space Ξ into a hyperplane Ξ_{ω} . When Rule 3 is satisfied, we can establish the correspondence between the two cross-spectrum scattering positions $\xi_1(\underline{r}_1, l_1, \omega_1)$ and $\xi_2(\underline{r}_2, l_2, \omega_2)$ with the same scatterer density (i.e., $f_{\xi}(\underline{r}_1, l_1, \omega_1) = f_{\xi}(\underline{r}_2, l_2, \omega_2)$) as

$$|V(\omega_1)|^2 [\mathcal{G}_{\omega_1}(\underline{r}_1)]^{-n} \omega_1^{-n_s} \beta(\omega_1) = |V(\omega_2)|^2 [\mathcal{G}_{\omega_2}(\underline{r}_2)]^{-n} \omega_2^{-n_s} \beta(\omega_2) \quad (6.10)$$

Subsequently, if the conditional pdf of scatterer density for ξ_1 can be expressed as $f_{\underline{r},l|\omega_1}(\underline{r}_1, l_1 | \omega_1)$, the conditional pdf for ξ_2 is obtained as

$$f_{\underline{r},l|\omega_2}(\underline{r}_2, l_2 | \omega_2) = \frac{f_{\underline{r},l|\omega_1}(\underline{r}_1, l_1 | \omega_1) f_{\omega}(\omega_1)}{f_{\omega}(\omega_2)} \quad (6.11)$$

If $\underline{r} = \mathcal{G}_{\omega}(\underline{r}) = \underline{r} \quad \forall \omega$ defines a mirror-mapping function, the relationship between $\underline{r}_1$ and $\underline{r}_2$ is given by

$$\underline{r}_1 = \underline{r}_2 \left[\left(\frac{\omega_2}{\omega_1} \right)^{-n_s} \frac{\beta(\omega_2) |V(\omega_2)|^2}{\beta(\omega_1) |V(\omega_1)|^2} \right]^{\frac{1}{n}} \quad (6.12)$$

At this point, some interesting conclusions can already be drawn from (6.12). The most important implication is that the propagation channel is not defined as a separate functional block from the Tx and Rx blocks. (Note that $|V(\omega)|^2$ is embedded in the expression of (6.12), which implies that the signal properties affect the distribution of the scatterers.) Rather, it has to be treated as a nontrivial solution under the initial conditions specified by the Tx and Rx properties and the physical environments. Secondly, the geometry (e.g., circle, ellipse, or parabola) of the curves on which each scatterer has the same value of $\underline{r}$ and the statistical functions (e.g., Gaussian, exponential or linear distribution) used to describe the distribution of $\underline{r}$ could be different across various frequency components.

6.3.2 Power Azimuthal Spectrum and Power Delay Spectrum

We first consider the channel properties for a zero-bandwidth channel at the frequency bin ω . With reference to Section 3.2.2, the instantaneous power received at the Rx due to reradiation from a scatterer S (Fig. 3.1) would be lognormally

distributed. Therefore, the joint pdf of the instantaneous power p_s and AoA φ_R conditioned on the area mean power p and ω can be expressed as

$$f_{p_s, \varphi_R | p, \omega}(p_s, \varphi_R | p, \omega) = f_{p_s | p, \omega}(p_s | p, \omega) f_{\varphi_R | p, \omega}(\varphi_R | p, \omega) \quad (6.13)$$

p is related to the GDP $\underline{r}$ by $p = |V(\omega)|^2 \underline{r}^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega)$ with $\beta(\omega)$ being the average attenuation loss of the scattering object for a particular frequency bin. ω_0 is the reference frequency. Applying the transformation of variables to (6.13), the joint density $f_{p_s, \varphi_R | \underline{r}, \omega}(p_s, \varphi_R | \underline{r}, \omega)$ is found using

$$f_{p_s, \varphi_R | \underline{r}, \omega}(p_s, \varphi_R | \underline{r}, \omega) = f_{p_s, \varphi_R | p, \omega}(p_s, \varphi_R | p, \omega) \Big|_{p=|V(\omega)|^2 \underline{r}^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega)} \quad (6.14)$$

The PAS at the Rx conditioned at ω is thus obtained as

$$\begin{aligned} f_{\varphi_R | \omega}(\varphi_R | \omega) &= \Sigma_{\varphi}(\omega) \iint_{\Xi_{\omega}} p_s f_{p_s, \varphi_R | \underline{r}, \omega}(p_s, \varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) d\underline{r} dp_s \\ &= \Sigma_{\varphi}(\omega) \int_{\Xi_{\omega}} f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) d\underline{r} \int p_s f_{p_s | \underline{r}, \omega}(p_s | \underline{r}, \omega) dp_s \\ &= \Sigma_{\varphi}(\omega) \int_{\Xi_{\omega}} f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) |V(\omega)|^2 \underline{r}^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega) \exp(-\sigma_p^2/2) d\underline{r} \\ &= \Sigma'_{\varphi}(\omega) \int_{\Xi_{\omega}} f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) \underline{r}^{-n} d\underline{r} \end{aligned} \quad (6.15)$$

where σ_p is the standard deviation of the lognormal fading around the Tx and Rx, and is assumed to be frequency-independent. $\Sigma_{\varphi}(\omega)$ and $\Sigma'_{\varphi}(\omega)$ are normalization constants.

The set Ξ_{ω} denotes the integration domain for $\underline{r}$, which is the collapse of Ξ into the two-dimensional hyperplane at ω . The overall PAS is calculated as

$$f_{\varphi_R}(\varphi_R) = \Lambda_{\varphi} \int_{\omega_L}^{\omega_H} |V(\omega)|^2 f_{\varphi_R | \omega}(\varphi_R | \omega) f_{\omega}(\omega) d\omega \quad (6.16)$$

where $f_{\omega}(\omega)$ is the marginal pdf of scatterer distribution in the frequency domain, and Λ_{φ} a normalization factor. ω_L and ω_H are the lower and upper 3 dB points of the

signal spectrum, respectively.

To evaluate the PDS, we first find the joint pdf of p_s and the path delay τ conditioned on $\underline{r}$ and ω using

$$f_{p_s, \tau | \underline{r}, \omega}(p_s, \tau | \underline{r}, \omega) = \frac{f_{p_s, \varphi_R | \underline{r}, \omega}(p_s, \varphi_R | \underline{r}, \omega)}{|d\tau/d\varphi_R|} \bigg|_{\varphi_R = h(\tau, \underline{r}, \omega)} \quad (6.17)$$

By averaging over the random variable $\underline{r}$, the conditional PDS is shown to be

$$\begin{aligned} f_{\tau | \omega}(\tau | \omega) &= \Sigma_{\tau}(\omega) \iint_{\mathbb{E}_s} p_s f_{p_s, \tau | \underline{r}, \omega}(p_s, \tau | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) d\underline{r} dp_s \\ &= \Sigma_{\tau}(\omega) \int_{\mathbb{E}_s} \left| \frac{d\varphi_R}{d\tau} \right| f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) d\underline{r} \int p_s f_{p_s | \underline{r}, \omega}(p_s | \underline{r}, \omega) dp_s \\ &= \Sigma_{\tau}(\omega) \int_{\mathbb{E}_s} \left| \frac{d\varphi_R}{d\tau} \right| f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) \underline{r}^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega) |V(\omega)|^2 \exp(-\sigma_p^2/2) d\underline{r} \\ &= \Sigma'_{\tau}(\omega) \int_{\mathbb{E}_s} \left| \frac{d\varphi_R}{d\tau} \right| f_{\varphi_R | \underline{r}, \omega}(\varphi_R | \underline{r}, \omega) f_{\underline{r} | \omega}(\underline{r} | \omega) \underline{r}^{-n} d\underline{r} \end{aligned} \quad (6.18)$$

with $\Sigma_{\tau}(\omega)$ and $\Sigma'_{\tau}(\omega)$ being normalization constants.

The overall PDS is thus calculated as

$$f_{\tau}(\tau) = \Lambda_{\tau} \int_{\omega_L}^{\omega_H} |V(\omega)|^2 f_{\tau | \omega}(\tau | \omega) f_{\omega}(\omega) d\omega \quad (6.19)$$

with Λ_{τ} being a normalization constant.

Interestingly, as shown in (6.16) and (6.19) both the PAS and PDS are independent of the first-moment fading statistics around the Rx and Tx, which means that the geometry of the canonical model preserves all the necessary information to characterize the power distribution of channel output in the azimuth and delay domains. As a final remark of this section, wide sense stationary uncorrelated scattering (WSSUS) channel conditions are required in the proposed model to relate a channel's frequency correlation function to its average power delay profile through a Fourier transform [BUL02].

6.4 Modeling the Indoor Propagation Environments for UWB Transmission

6.4.1 Model Description

In indoor propagation where antenna heights are relatively low, the Rx will receive MPCs from scatterers distributed around the Tx as well as itself. All angles in the azimuth plane are involved. In this type of environment, specular reflection is the major mechanism of the interaction between the electromagnetic waves and the surfaces of scatterers. Other mechanisms such as diffuse scattering can be ignored. Moreover, the transmitted signal has mild short-term fading around the Tx. In such cases, the NFLF $\eta = 1$ and the operator $*$ refers to sum. From (6.9), it is evident that the curve representing a collection of scatterers with constant distribution of $\underline{r}$ is an ellipse defined by $\underline{C}: \underline{r} = \hat{\underline{r}} = r_R + r_T = 2a$ as depicted in Fig. 6.4. The Rx and the Tx are the foci of the elliptical area respectively. This reduces to an elliptical scattering model (ESM) [ECS++98].

We further make the following assumptions in order to fully develop the model.

- 1) The frequency dependence of the scattering coefficient is taken into account by assuming the following relationship

$$\beta(\omega) = \omega^\mu \quad \mu \leq 0 \quad (6.20)$$

which can be employed to explain most practical scattering phenomenon occurring on buildings, windows, and furniture as claimed in [QIU96]. For example, a ray coming from the wall edge has the frequency dependence of $\beta(\omega) = \omega^{-1}$, and a ray coming from the diffraction at the desk corner has $\beta(\omega) = \omega^{-2}$ [QIU96].

Fig. 6.4. The elliptical scattering model

Fig. 6.5. Gaussian monocycles and their corresponding normalized frequency spectrum. Dash dot lines: $\tau_m = 0.2$ ns, dash lines: $\tau_m = 0.8$ ns, solid lines: $\tau_m = 1.5$ ns

- 2) The signal at the input of the transmit antenna uses a class of waveforms known as Gaussian waveforms [WS02], according to

$$v_0(t) = A_0 \exp \left[-2\pi \left(\frac{t - t_d}{\tau_m} \right)^2 \right] \quad (6.21)$$

where t_d represents the location of the pulse center in time and τ_m is a parameter

which determines the temporal width of the pulse. Under this assumption, the ideal transmitted pulse shape can be modeled as the derivative of (6.21)

$$v_1(t) = A_1(t - t_d) \exp \left[-2\pi \left(\frac{t - t_d}{\tau_m} \right)^2 \right] \quad (6.22)$$

The signal defined by (6.22) is called as the Gaussian Monocycle. By representing the $v_1(t)$ as linear combination of Hermite functions $v_1(t) = \sum a_k H_k$, the magnitude of the frequency spectra of the Gaussian monocycle is given by

$$V_1(\omega) = \sqrt{2\pi} \sum_k (-1)^k a_k H_k(\omega) = \frac{A_1 \omega \tau_m^3}{2\sqrt{2}} \exp \left(-\frac{\pi \omega^2 \tau_m^2}{2} \right) \quad (6.23)$$

Subsequently the 3-dB cutoff points ω_L and ω_H can be shown to be the solutions of the nonlinear equation $\omega^2 \tau_m^2 \exp(-\pi \omega^2 \tau_m^2) = 1/(4\pi e)$. Fig. 6.5 illustrates the normalized Gaussian Monocycles, and their frequency-domain representations for $\tau_m = [0.2 \ 0.8 \ 1.5]$ ns.

- 3) Following from the strong bijective mapping rule, there is a bijective mapping between the multidimensional pdf $f_{\xi}(\underline{r}, l, \omega)$ and $p(\underline{\hat{r}})|_{\omega}$. The function $f_{\xi}(\underline{r}, l, \omega) = \mathbb{Z}(p(\underline{\hat{r}})|_{\omega})$ is a strong function of the architecture of the building itself and secondary reflections off furniture or other objects, which captures the main features of the propagation channel and gives an indication about the robustness of the radio channel against slight variations of any other input parameters (e.g., temporal width and/or pulse shape). Furthermore, $\mathbb{Z}$ has to be a monotonically increasing function following from the second heuristic rule. Subsequently we extend the linear function in [CD04b] and consider the following versatile variable-exponent function which has previously been used by Blaunstein *et al.* to

Fig. 6.6. The normalized $Z(p)$ function for various parameters n_p

model the height profile of the buildings inside the city layer [BKC++01]

$$\begin{aligned} f_{\varepsilon}(\underline{r}, l, \omega) &= Z\left(p(\underline{r})\big|_{\omega}\right) = Z(p) \\ &= k \left[1 - \hbar(p_1 - p) - \hbar(p - p_1) \hbar(p_2 - p) \left(\frac{p_2 - p}{p_2 - p_1} \right)^{n_p} \right] \end{aligned} \quad (6.24)$$

where $p(\underline{r})\big|_{\omega} = p(\underline{r})\big|_{\omega} = |V(\omega)|^2 \underline{r}^{-n} (\omega/\omega_0)^{-n_s} \beta(\omega)$. k is a normalization constant. $\hbar(x)$ is the Heaviside step function. $\hbar(x)$ is equal to 1 if $x > 0$ and it is equal to 0 if $x < 0$. p_1 and p_2 are the lower and upper bounds, respectively. The former determines the minimum power below which the probability of finding an effective scatterer is zero, and the latter is determined by the maximum power contributed by the scatterers in all multipaths. A useful physical interpretation of p_1 is that it can be regarded as the noise floor of receiving equipment. When the power of impinging rays is below p_1 , they become undetectable and can be effectively treated as “nonexistent”. On the other hand, p_2 is the power contributed by the strongest MPC (usually from the direct path). The graph of this

function versus the power is presented in Fig. 6.6. For $n_p \ll 1$, $\mathbb{Z}(p)$ describes the case when scatterers contributing power lower than p_2 exist very rarely, i.e., the arriving signals from most of the scatterers are at the level of maximal power, p_2 . The case when the distribution of scatterers becomes more uncorrelated with p is given by $n_p \gg 1$. For n_p close to zero, most scatterers have approximately the same power which is equal to p_2 . For $n_p = \infty$, we have the case of scatterers uniformly distributed in the range p_1 to p_2 .

The following three conditional pdfs can thus be computed as

$$f_\omega(\omega) = \int_D^{\infty} \ell_{\underline{c}}(\underline{r}) f_{\xi}(\underline{r}, l, \omega) d\underline{r} = \int_D^{\infty} 2\underline{r} H\left(e, \frac{\pi}{2}\right) \mathbb{Z}(p) d\underline{r} \quad (6.25)$$

$$f_{\underline{r}|\omega}(\underline{r}|\omega) = \ell_{\underline{c}}(\underline{r}) \cdot f_{\xi, l|\omega}(\underline{r}, l|\omega) = \ell_{\underline{c}}(\underline{r}) \cdot \frac{f_{\xi}(\underline{r}, l, \omega)}{f_\omega(\omega)} = 2\underline{r} H\left(e, \frac{\pi}{2}\right) \cdot \frac{\mathbb{Z}(p)}{f_\omega(\omega)} \quad (6.26)$$

$$f_{\varphi_R|\underline{r}, \omega}(\varphi_R|\underline{r}, \omega) = \frac{\sqrt{r_R^2 + (dr_R/d\varphi_R)^2}}{\ell_{\underline{c}}(\underline{r})} = \frac{(\underline{r}^2 - D^2) \sqrt{\underline{r}^2 - 2\underline{r}D \cos \varphi_R + D^2}}{4\underline{r} H\left(e, \frac{\pi}{2}\right) (\underline{r} - D \cos \varphi_R)^2} \quad (6.27)$$

$$\text{with } H\left(e, \frac{\pi}{2}\right) = \frac{\pi}{2} \left[1 - \left(\frac{1}{2}\right)^2 e^2 - \left(\frac{1 \cdot 3}{2 \cdot 4}\right)^2 \frac{e^4}{3} - \dots \right] \text{ and } e = D/\underline{r}.$$

The overall PAS can be obtained by substituting (6.25)-(6.27) into (6.15) and (6.16). In a similar manner, by noting that $f_{\tau|\underline{r}, \omega}(\tau|\underline{r}, \omega) = \delta(\tau - \underline{r}/v_c)$ where τ is the path delay, v_c is the speed of the electromagnetic waves, and $\delta(\cdot)$ the delta function, the overall PDS can be computed by substituting (6.25)-(6.27) into (6.18) and (6.19).

As a final remark of this section, the abovementioned assumptions are an oversimplification in the model, but they are mathematically expedient. The implementation of the suppositions to represent the real-world situation would have to be justified by further measurements.

6.4.2 Numerical Examples and Comparisons with Measurements

This section presents examples for characteristics of the UWB channel as generated by the proposed model for indoor micro- and picocell types of environments. The main purpose of this section is threefold. In Test Case 1 (TC1), the dependency of PAS and PDS on the temporal width of the pulse is studied. Three different values of τ_m are used in the numerical calculation. In Test Case 2 (TC2), we investigate the effect of different Tx-to-Rx distances, D , on the PAS and PDS. Finally we look into the effect of different values of n_p on the channel properties in the third study case (TC3). The parameters used in the test cases are summarized in Table 6.1, which are comparable to the typical signal properties and channel conditions for UWB system [SV87, CSW02, SJJ++00].

Consider first the plots obtained in Fig. 6.7 for Test Case 1. It can be observed from the model results that the influence of different widths of the Gaussian Monocycles on the PAS and PDS is *insignificant* for the range of τ_m specified in TC1. In regards to Fig. 6.7(a), several points are of note. The PAS best fits the Laplacian

Table 6.1. Parameters used in the test cases

Symbol	Numerical values (TC1)	Numerical values (TC2)	Numerical values (TC3)
τ_m	[0.2 0.8 1.5] ns	0.8 ns	0.8 ns
ω_0	1 GHz	1 GHz	1 GHz
n_ω	2.5	2.5	2.5
μ	-1.5	-1.5	-1.5
n	4	4	4
n_p	1	1	[0.01 1 10]
p_1	0	0	0
p_2	Direct-path power	Direct-path power	Direct-path power
D	25 m	[10 25 40] m	25 m

distribution with a standard deviation σ_θ of 42° . This figure is in good agreement with the empirical result, 37° , reported in [CSW02] and 25.5° in [SJJ++00]. However, our model envisages a heavier tail against a Laplacian distribution (for $|\phi_R| \geq 100^\circ$). Fig. 6.7(b) shows that the PDS is fairly close to an exponential distribution with a standard deviation σ_τ of 18 ns. This result compares to values of 28.6 ns in [SJJ++00] and 20 ns in [SV87] for the intra-cluster rate of decay, or the rate at which the energy in the signal arrivals within a cluster falls off as a function of the delay, since we only consider the single-cluster environment here. A non-linear (on a log scale) fit to the data may also be attempted in order to obtain a model which weights the strong arrivals earlier in time more heavily [CSW02]. In fact, the result in Fig. 6.8(a) shows that as the Tx-to-Rx distance increases, the decay of the energy can be better fit into an exponential distribution.

The results for TC2 are plotted in Fig. 6.8. It is made apparent from the plots that as D increases, the angular spread decreases and the delay spread increases. The latter is supported by the published data in [GEY++97] for outdoor environments and the references therein. The PAS plots for several values of D are shown in Fig. 6.8(a) in order to demonstrate the consistency of the observation that the PAS can be best fit by a Laplacian distribution. It is also of interest to note that the Laplacian function would reflect the tail weight characteristics of the PAS better for a larger Tx-to-Rx distance. The consistency of the inspection that the PDS can be approximated by an exponential distribution is illustrated in Fig. 6.8(b).

Fig. 6.9 shows the PAS and PDS for TC3, respectively. Both the angle spread (AS) and delay spread (DS) increase as n_p increases. This shows that when the scatterers are more uniformly distributed in the range p_1 to p_2 , the power of ray

arrivals is more widely spread in the azimuth and delay domains, which certainly holds an intuitive appeal. This also illustrates that there is a *positive* cross correlation between the AS and DS, if each pairs of AS and DS could be considered as the medians for different indoor scenarios (room-to-room, within-the-room, in hallways, etc.) and signal propagation through walls made of various building materials (e.g., dry wall, wall board, structure wood, glass sheet [MUQ03]). The same phenomenon has been observed in the outdoor measurement campaign [APM02]. Furthermore, the results also indicate that the mechanisms leading to azimuthal and temporal dispersion are related and the potential space and frequency diversity gains for UWB systems are highly correlated.

6.5 Conclusions

In this chapter, we have introduced an AFDGM model for UWB transmission. This model defines the distribution of principal scatterers in a hypothetical multidimensional azimuth-frequency space. Three heuristic rules that are an extension of the work presented in Chapter 3, have been introduced to exploit the underlying assumptions made by any geometric model concerning the actual wave propagation process in an UWB channel. It has been shown that channel should not be defined as a separate functional block from the Tx and Rx blocks. In fact, the signal properties affect the scatterer distribution in the azimuth-frequency domain. We have also derived the important channel parameters including the PAS and PDS. The generic AFDGM model was then applied to model an UWB channel. Theoretical derivations of PAS and PDS have shown that the pulse duration does not have a significant influence on these two parameters for practical sounding pulses. It was

also shown that the channel properties derived using the proposed model exhibited a standard deviation of ray arrivals in the azimuth domain and a rate of decay in the delay domain that were comparable to the empirical results reported in other works.

(a)

(b)

Fig. 6.7. (a) PAS at the Rx (Curves for $\tau_m = 0.8$ ns and $\tau_m = 1.5$ ns have been shifted up by 0.1 and 0.2 units for the sake of clarity), and (b) PDS in Test Case 1 (Curves for $\tau_m = 0.8$ ns and $\tau_m = 1.5$ ns have been shifted up by 5 and 10 units for the sake of clarity)

Fig. 6.8. (a) PAS at the Rx and (b) PDS in Test Case 2

Fig. 6.9. (a) PAS at the Rx and (b) PDS in Test Case 3 (The three best-fit exponential curves have been shifted down for the sake of clarity).

CHAPTER 7 DIVERSITY AND DEGREES-OF-FREEDOM: A SYSTEM-DEPENDENT CHANNEL MODELING PERSPECTIVE

7.1 Introduction

In this chapter, we obtain more fundamental bounds to the available diversity (DIV) and degrees-of-freedom (DOF) of wideband directional channels from a system-dependent channel modeling perspective. In urban and indoor wireless channels, scatterers are not distributed uniformly throughout the whole coverage area, but rather occur in clusters [COR01]. The scattering of each cluster in physical environments will be characterized by a set of aperture-frequency-dependent parameters including the subtended angular and delay intervals, cluster radius and keyhole radius, where the first two serve to characterize the cluster boundaries and the last two reflect its internal radio propagation conditions.

This chapter is organized as follows. Section 7.2 presents the modeling methodology for the DIV and DOF computation, starting with a degenerate single-cluster-single-bounce (SCSB) channel and perfect receiving conditions [CD05b]. Section 7.3 then extends the approach in Section 7.2 to more general scattering and receiver-operating scenarios. Section 7.4 shows the utility of the proposed methodology by presenting various numerical examples to provide insight into the impact of signal power spectrum on the DIV and DOF estimates. Finally, we conclude in Section 7.5.

7.2 DIV and DOF in A Single-Cluster-Single-Bounce Channel

7.2.1 A General Description of the Multipath Structure

In this section, we consider a two-dimensional single-bounce environment with scattering objects spanning over a single solid angle Ω_T as depicted in Fig. 7.1, where Ω_T (Ω_R) is obtained by first projecting the scattering cluster onto the unit circle enclosing the transmit (receive) array and then projecting the arc onto the arrays. This domain will be called as the array-constrained angular (ACA) domain in the following discussion. For instance, in the ideal Clarke's scattering field the channel solid angle in the ACA domain is 2.

Without loss of generality we represent the scattering cluster as a circular area, $\mathcal{C}$, with radius R (see Fig. 7.2(a)), where there are total N scatterers, each of which is randomly located in this area independent of other scatterers. We further define a keyhole¹, $\mathcal{K}$, as the circular sub-area with radius r within which the orthogonality of the received waves at any two points, $\mathbf{a}_{C1}$ and $\mathbf{a}_{C2}$, is below a pre-specified critical value ς regardless of the monochromatic transmitting sources. Stated alternatively,

$$\int_{\mathcal{A}_T} G^H(\mathbf{a}_T, \mathbf{a}_{C2}, \lambda) G(\mathbf{a}_T, \mathbf{a}_{C1}, \lambda) d\mathbf{a}_T \leq \varsigma, \quad \forall \mathcal{A}_T$$

where $G(\mathbf{a}_T, \mathbf{a}_C, \lambda)$ is the Green's function, or the possible waves at position $\mathbf{a}_C$ resulting from a point source at position $\mathbf{a}_T$ within any transmit space $\mathcal{A}_T$. In this work we consider continuous one-dimensional (1D) arrays which are composed of an infinite number of antennas separated by infinitesimal distances. This eliminates the

¹ Note that the "keyhole" defined here is a bit different from the original definition [CFG++02]. We refer to keyhole as any scenario that causes a significant reduction of the rank of the channel transfer matrix $\mathbf{H}$ as compared to the idealized infinite-rank case.

Fig. 7.1. Illustration of a single-cluster environment

Fig. 7.2. Multipath structure of a scattering cluster in the azimuth-delay domain

need to specify *a priori* the number of antennas and their relative positions on the arrays.

The channel response from the transmitter to the receiver is

$$\mathbf{H}(\mathbf{\kappa}_T, \mathbf{\kappa}_R) = \sum_{n=1}^N \alpha_n \delta(\mathbf{\kappa}_T - \mathbf{\kappa}_{T,n}) \delta(\mathbf{\kappa}_R - \mathbf{\kappa}_{R,n}) \quad (7.1)$$

where α_n is the attenuation and polarization between the transmit direction $\mathbf{\kappa}_T$ and receive direction $\mathbf{\kappa}_R$. $\mathbf{H}(\cdot, \cdot)$ can be well approximated as

$$\mathbf{H}(\mathbf{\kappa}_T, \mathbf{\kappa}_R) \approx \sum_{i=1}^M \alpha'_i \sum_{j \in P_{K,i}} \delta(\mathbf{\kappa}_T - \mathbf{\kappa}_{T,j}) \delta(\mathbf{\kappa}_R - \mathbf{\kappa}_{R,j}) \quad (7.2)$$

where α'_i is the average attenuation of the i th keyhole and $P_{K,i}$ is the set of propagation paths in the i th keyhole.

The keyhole radius describes the correlation scale of multipath components (MPCs) within a scattering cluster. Subsequently, we introduce another two parameters to characterize the cluster boundaries, namely, the path delay spread T and the subtended angular interval Ω_T (Ω_R). Then, to ensure the well-conditionedness of $\mathbf{H}(\cdot, \cdot)$, the channel response satisfies

$$\mathbf{H}(\mathbf{\kappa}_T, \mathbf{\kappa}_R) \neq 0 \text{ if and only if } (\mathbf{\kappa}_T, \mathbf{\kappa}_R) \in \Omega_T \times \Omega_R$$

We further assume that it is possible to identify a set of joint azimuth-delay taps as depicted in Fig. 7.2(a) [YLC++04]². Each tap point (scattering center) is enclosed by its corresponding keyhole $\mathcal{K}$ and all the $\mathcal{K}$'s will cover the whole cluster with no overlap and gap. The point spectrum of $\mathbf{H}(\cdot, \cdot)$ ranges between 1 and infinite as r varies between R and 0. For example, a horizontal roof-edge-diffraction brings the rank of $\mathbf{H}(\cdot, \cdot)$ to unity because all MPCs go through a single point and hence $R/r \rightarrow 1$.

² It has been found in [YLC++04] that the average delay-tap angle spread (AS) is just few degrees less than the AS of that cluster. Therefore, the reference system in Fig. 7.2(a) that describes the location and size of each tap can be converted to the Cartesian coordinate in Fig. 7.2(b).

Another example of degenerate channel is when the scatterers have smooth, well-reflecting surfaces such as pure glass front.

7.2.2 Diversity Gain Calculation

7.2.2.1 General Formulation

The number of independent diversity paths associated with a channel is in general infinite. However, only a finite number of them are large enough to be regarded as significant. Moreover, in many problems the significant eigenvalues are of the same variance [KEN68]. In such cases, it is appropriate to define the number of dominant eigenvalues as the diversity estimate associated with a system.

For wideband directional channels, each signal point can be treated as a vector field $\rho(a_T, f)$ on the signal space $\mathcal{S} = \mathcal{A}_T \times \mathcal{F}$, where a_T and f correspond to the transmit aperture position and frequency, respectively. For brevity, it is assumed in this work that $a_T \in \mathcal{A}_T$ which is a linear continuous array. $\mathcal{F}$ is the frequency space spanned by the signal spectrum. We start the discussion with a fully scatterered environment (i.e., an infinitesimal keyhole structure) and assume that the receive space $\mathcal{A}_R$ is an infinite-long linear aperture which is able to form an arbitrarily narrow beam. The delay-ACA-domain expression of the complex received process $z(t, \mathcal{G}_R)$ can thus be written as

$$z(t, \mathcal{G}_R) = \frac{1}{\sqrt{2E}} \sum_j \sum_k a_{j,k} u(t - \tau_j, \mathcal{G}_T \mapsto \mathcal{G}_{R,j,k}) \exp(-i2\pi\tau_j) \quad (7.3)$$

where E is the average received energy. The two solid angles $\mathcal{G}_T$ and $\mathcal{G}_R$, representing angle-of-departure (AoD) and angle-of-arrival (AoA) in the ACA domain, are viewed from the transmitter (Tx) and receiver (Rx), respectively. τ_j is the propagation delay

caused by the j th scattering strip (i.e., the intersection of an ellipse and the scattering cluster) as illustrated in Fig. 7.1. $a_{j,k}$ is an identically and independently distributed (i.i.d.) random variable and it is Rayleigh-distributed when the received process is Gaussian. $\mathcal{G}_T \mapsto \mathcal{G}_{R,j,k}$ denotes the mapping of waveforms from $\mathcal{A}_T$ to $\mathcal{A}_R$ in the ACA domain by the k th subdivision of the j th scattering strip. The channel output is just a summation of delayed, attenuated and scattered replicas of the transmitted signal $u(\cdot, \cdot)$.

We then invoke the approximation that a signal of bandwidth W cannot change its value substantially in a time interval of duration W^{-1} [KEN68], which means

$$u\left(t - \tau_j\right) \approx u\left(t - \frac{n}{W}\right) \quad \text{provided that } \left|\tau_j - \frac{n}{W}\right| \leq \frac{1}{2W}, \text{ where } n = 0, 1, 2, \dots \quad (7.4)$$

Casting (7.4) onto linear arrays, we can obtain the following analogy in the ACA domain. A waveform of aperture extent L/λ_0 , when Fourier-transformed to the ACA domain, cannot have an appreciable change in an interval of $(L/\lambda_0)^{-1}$, where λ_0 is the center frequency of a narrowband signal. That is,

$$\begin{aligned} u\left(\mathcal{G}_T \mapsto \mathcal{G}_{R,j,k}\right) &\equiv u\left(\mathcal{G}_T \mapsto \mathcal{G}_{T,j,k}\right) \approx u\left(\mathcal{G}_T \mapsto \frac{m}{L/\lambda_0}\right) \\ &\text{provided that } \left|\mathcal{G}_{T,j,k} - \frac{m}{L/\lambda_0}\right| \leq \frac{1}{2L/\lambda_0}, \text{ where } m = 0, 1, 2, \dots \end{aligned} \quad (7.5)$$

for an SCSB channel.

The next step is to divide the range of summation in (7.3) into the set of intervals

$$\left|\tau_j - \frac{n}{W}\right| \leq \frac{1}{2W}, \quad \left|\mathcal{G}_{T,j,k} - \frac{m}{L/\lambda_0}\right| \leq \frac{1}{2L/\lambda_0}, \quad (n, m) = 0, 1, 2, \dots$$

to obtain

$$z(t, \mathcal{G}_R) \approx \frac{1}{\sqrt{2E}} \sum_n \sum_m \sum_{\Xi_n} \sum_{\Xi_m} a_{j,k} u\left(t - \frac{n}{W}, \mathcal{G}_T \mapsto \frac{m}{L/\lambda_0}\right) \exp(-i2\pi\tau_j) \quad (7.6)$$

where Ξ_n and Ξ_m denote the sets of those j and k for which (7.4) and (7.5) are satisfied. By defining

$$b_{n,m} \approx \frac{1}{\sqrt{2E}} \sum_{\Xi_n} \sum_{\Xi_m} a_{j,k} \exp(-i2\pi\tau_j) \quad (7.7)$$

we obtain

$$z(t, \mathcal{G}_R) \approx \sum_n \sum_m b_{n,m} u\left(t - \frac{n}{W}, \mathcal{G}_T \mapsto \frac{m}{L/\lambda_0}\right) \quad (7.8)$$

The complex variables $b_{n,m}$ that multiply these functions are statistically independent because the values of different $b_{n,m}$ are determined by different scatterers. If the channel can be characterized by a set of cluster boundaries: path delay spread T and angle spread Ω_T , it would be a simple task to show that the function z is a combination of $W \times T \times (L/\lambda_0) \times \Omega_T$ known functions, which indeed gives the upper bound for diversity estimate [KEN68], provided that the following three conditions are all fulfilled:

- 1) The wireless transmission scheme occupies a small fractional bandwidth W/f_0 such that L/λ_0 is a good approximation to the normalized array length across the entire frequency band.
- 2) The signal space $\mathcal{S}$ is of rectangular shape.
- 3) T and Ω_T are fixed parameters independent of the signal power spectrum density (PSD) functions.

Nevertheless, the deployment of ultrawideband (UWB) technology as a candidate physical layer mechanism for IEEE 802.15 Wireless Personal Area Network

[RFS++04], which utilizes large bandwidths to achieve increasingly fine resolution of multipath arrivals leading to reduced fading per resolved path, has nullified the first condition. Furthermore, the setting of Condition 2 restricts the signaling strategy to a particular scenario, which may not be feasible for the diversity/capacity gain analysis for some emerging techniques in multiple access networks. One such example is the recently proposed space-time linearly precoded orthogonal frequency-division multiple access [SLG02]. As a final point, in real-life channels, both T and Ω_T would exhibit array-length and frequency dependencies as to be discussed below.

7.2.2.2 System-dependent Channel Characterization

We will first characterize the relationship for either Ω_T or T versus the array length L . Consider the situation depicted in Fig. 7.3(a), where a large number of multipaths emanating from a finite region are impinging on a continuous linear array $\mathcal{A}$ of length L placed symmetrically along the a -axis, which has a space factor

$$F_{\mathcal{A}}(\theta) = \int_{-L/2}^{L/2} I(a) e^{j[k a \sin \theta + \varphi(a)]} da \quad (7.9)$$

where $k = 2\pi/\lambda$ is the free space wavenumber. $I(a)$ and $\varphi(a)$ represent, respectively, the amplitude and phase distributions along the source. Some commonly used line-source distributions are uniform, triangular, cosine for amplitude and linear, quadratic, and cubic for phase [BAL97]. For brevity, we assume a normalized uniform current distribution of the form $I(a) = I_0/L$ and $\varphi(a) = 0$. In such a case, (7.9) reduces to

$$F_{\mathcal{A}}(\theta) = I_0 \sin(kL \sin \theta/2) / (kL \sin \theta/2) \quad (7.10)$$

We further assume that the true distribution of angles has a Laplacian density:

$$f_{\text{tr}}(\theta) = C_1 \exp(-|\theta - \theta_p|/s) \quad \theta \in (-\pi, \pi] \quad (7.11)$$

$\theta_p \in (-\pi, \pi]$ accounts for the mean AoA (Fig. 7.3(a)). The variance of the distribution is s^2 , and C_1 is a normalization factor to ensure that f_{tr} is a pdf.

At each θ , the received signal will be a weighted sum of many incident rays along that direction. The sum will vary randomly due to the random amplitude and phase of each incident ray. Assuming that the magnitude of the received complex envelope $|H(\theta)|$ has a Nakagami fading, $|H(\theta)|^2$ would be a Gamma variate. The expectation of $|H(\theta)|^2$ is

$$E\{|H(\theta)|^2\} = f_{\text{tr}}(\theta) [F_{\mathcal{A}}(\theta)]^2 \quad (7.12)$$

Fig. 7.3. (a) Multiple rays impinging on a continuous linear array (b) normalized EPAS viewed from $\mathcal{A}$ for different array lengths. The system dynamic range is set

to $\zeta = 10$ dB.

We can interpret (7.12) as representing the *mean* power azimuthal profile (MPAP). Since the usual way to eliminate the effect of instrumental noise is setting to zero the values of MPCs whose power lies below the noise threshold [ROS99], the effective MPAP viewed from $\mathcal{A}$ is given by

$$\begin{aligned} p_{\text{eff}}(\theta) &= \int_{\mathcal{N}} p \cdot f_{p|\theta}(p|\theta) dp \\ &= \int_{\mathcal{N}} p \cdot \left\{ \frac{m}{E\{|H(\theta)|^2\}} \right\}^m \frac{p^{m-1}}{\Gamma(m)} \exp\left\{-\frac{mp}{E\{|H(\theta)|^2\}}\right\} dp \end{aligned} \quad (7.13)$$

where $f_{p|\theta}$ is the probability of received power conditioned on angle, m is the Nakagami shape parameter, and $\Gamma(\cdot)$ is the Gamma function. $\mathcal{N}$ denotes the noise threshold which is quantified through the system dynamic range ζ , defined as the number of dB below the maximum average received power. It is of special interest to find the effective power azimuthal spectrum (EPAS), $f_{\text{eff}}(\theta)$, referring to the actual distribution of angles after the antenna effects are deconvolved from the measured data. The EPAS is derived from (7.13) as

$$f_{\text{eff}}(\theta) = C_2 \cdot \frac{p_{\text{eff}}(\theta)}{[F_{\mathcal{A}}(\theta)]^2} \quad (7.14)$$

with C_2 being a normalization factor to ensure that f_{eff} is a pdf. It is worth mentioning that as $\mathcal{N}$ approaches to 0, $f_{\text{eff}}(\theta)$ approaches to $f_{\text{tr}}(\theta)$ as could be expected.

Fig. 7.3(b) illustrates the plots of EPAS for $m = 1$ (Rayleigh distribution), $\zeta = 10$ dB, and $s = 10^\circ$. The maximum values of all the curves have been normalized to 0 dB. Notably, the 10-dB angular spread, Θ , of the EPAS decreases from around 15° to 5° as the array length increases from 0.5λ to 7.5λ . This is due to the fact that the distribution of angles at a small array is spread over a wider angular range and thus

the AoA statistics would be less distorted by truncation of the tails that are below the noise cutoff. Fig. 7.4(a) captures the relationship for Θ versus L/λ for different noise thresholds, which can be fitted by an exponential regression curve, leading to

$$\Theta(L/\lambda) = \beta^{(\Theta, L)} e^{\gamma^{(\Theta, L)} L/\lambda} \quad (\text{degree}) \quad (7.15)$$

In general, a larger $\beta^{(\Theta, L)}$ and $\gamma^{(\Theta, L)}$ are observed for a larger value of ζ , leading to a wider angular spread.

Next, we consider the relationship for T versus L . It is well known that the use of directive antennas at the remote terminal is an effective way to reduce the multipath delay spread [RH92, MMI96]. For example, Manabe *et al.* conducted propagation measurements at 60 GHz in a conference room within a typical modern office building. A monopole antenna with an omnidirectional radiation pattern was used as the transmit antenna, and three directive antennas with wide (a scalar feed horn), medium (a pyramidal standard horn), and narrow (a scalar lens horn) beamwidths were used to receive the signals. It was demonstrated that the delay spread decreased from 13.59 ns to 1.05 ns as the antenna 3-dB beamwidth decreased from 60° to 5° (equivalently, the array length would increase from approximately λ to 11.46λ). A ray-tracing tool was used to simulate the abovementioned measurement campaign. It was observed that the estimated delay spread decreased from 10.91 ns to 0.79 ns as the beamwidth of the receive antenna decreased, which was consistent with the empirical observation. For brevity, we propose to model the relationship as follows

$$T(L/\lambda) = \beta^{(T, L)} e^{\gamma^{(T, L)} L/\lambda} \quad (\text{ns}) \quad (7.16)$$

Equation (7.16) is shown to have a fairly good agreement with the results reported in [MMI96] as illustrated in Fig. 7.4(b) for the limited number of data points.

Fig. 7.4. (a) AS and (b) DS [MMI96] for different array lengths and their exponential regression curves; (c) AS and (d) DS in the townhouse environment [PH03] and their exponential regression curves

We may further quantify the relationship for either Θ or T versus the frequency f . In [PH03], Poon and Ho reported UWB measurements conducted in a typical townhouse. The frequency was swept from 2 to 8 GHz with a resolution of 3.75 MHz. 11 data sets (6 in non-line-of-sight (NLoS) and 5 in line-of-sight (LoS)) were collected in a two-floor townhouse. Each data set comprised a set of 20 receive antenna positions lined up linearly and oriented either toward or perpendicular to the

transmit array. The resulting plot of average Θ versus f , shown in Fig. 7.4(c), can be well fitted in the exponential form

$$\Theta(f) = \beta^{(\Theta, f)} e^{\gamma^{(\Theta, f)} f} \quad (\text{degree}) \quad (7.17)$$

In a similar manner, the average T versus f can also be fitted by an exponential curve following from the measurement data reported in [PH03], viz.,

$$T(f) = \beta^{(T, f)} e^{\gamma^{(T, f)} f} \quad (\text{ns}) \quad (7.18)$$

This observation has been illustrated in Fig. 7.4(d). The combination of (7.15)–(7.18) forms a useful model for estimating DIV in a system-dependent wideband directional channel. Although the models are derived from the situation where the finite-length array is employed at the Rx site, they would be applicable to a transmit finite-length array due to the theorem of reciprocity.

7.2.2.3 DIV Computation

To better reveal the system-dependent nature of fading channels, we examine the DIV on a microscopic scale. First, we represent each signal point as a vector $\mu(y, f)$ on the signal space $\hat{\mathcal{S}}$. The y -domain can be interpreted using Fig. 7.5(a), which illustrates a general line source distribution, $\mathcal{D}$, in the aperture domain. $\mathcal{D}$ can be considered as the combination of many differential arrays each having a uniform distribution:

$$\mathcal{D} \equiv \sum_i 2|a_r(y_i)| \cdot \delta y_i \quad (7.19)$$

for any symmetric $\mathcal{D}$ peaked at $a = 0$. Let $\delta\mu$ be the area of a small neighborhood centered on $\mu(y, f)$. δy and δf are the differential intervals in the respective domains.

As illustrated in Fig. 7.5(b), the DIV on $\delta\mu$ can be calculated as the volume of each box scaled by $\delta y/y(0)$, which may be expressed as

$$\delta \text{DIV} = \left\{ \delta f \times [T(2a_T(y), f)] \times \frac{2a_T(y)}{\lambda} \times [\Omega_T(2a_T(y), f)] \right\} \times \frac{\delta y}{y(0)} \quad (7.20)$$

Fig. 7.5. (a) A non-uniform line-source distribution $\mathcal{D}$ where $y(a_T) = [I(a_T)]^2$ and (b) a microscopic view of DIV

The total number of DIV is computed as

$$\text{DIV} = \sum \delta \text{DIV} = \iint_{\mathcal{S}} \frac{2a_r(y) f \Omega_r(2a_r(y), f) T(2a_r(y), f)}{y(0) v_c} dy df \quad (7.21)$$

where v_c is the speed of electromagnetic waves. The integral kernel in (7.21) is referred to as the DIV density.

If the multipath channel is ill-conditioned caused by propagation effects like diffraction or waveguiding leading to a reduction of the rank of $\mathbf{H}(\cdot, \cdot)$, the size of $\mathcal{K}$ could be comparable to the cluster size. In such a case, the keyhole population (KP) may not be sufficient to support the propagation of all diversity paths. Since the propagation mechanisms that cause channel rank-deficiency is usually sensitive to the system-operating conditions [CFG++02], we can similarly take a microscopic view of the KP on $\delta\mu$ as

$$\delta \text{KP} = \left[\frac{R(2a_r(y), f)}{r(2a_r(y), f)} \right]^2 \frac{\delta y \cdot \delta f}{y(0) \cdot W} \quad (7.22)$$

Summation of (7.22) yields

$$\text{KP} = \sum \delta \text{KP} = \iint_{\mathcal{S}} \left[\frac{R(2a_r(y), f)}{r(2a_r(y), f)} \right]^2 \frac{1}{Wy(0)} dy df \quad (7.23)$$

with the integrand being referred to as the KP density. Subsequently, the overall diversity estimate is given by

$$\overline{\text{DIV}} = \min \{ \text{DIV}, \text{KP} \} \quad (7.24)$$

7.2.3 Degrees-of-freedom Calculation

DOF is essentially the number of pairs of functions (communication modes), with each pair consisting of one function for the transmit space and one for the receive

space. The different communication modes are mutually orthogonal and define the two sets of functions with the strongest possible couplings between the spaces. We first consider the DOF in the time domain. In waveform channels, if a transmitted signal is first frequency-limited to $[-1/2W, 1/2W]$ and then approximately time-limited to $[-\tilde{T}/2, \tilde{T}/2]$, the dimension of the subspace satisfying these two physical constraints will give the number of significant singular values [SP61]:

$$W\tilde{T} + \chi \ln(W\tilde{T}) + o(\ln(W\tilde{T})) \quad (7.25)$$

for $W\tilde{T} \gg 1$ and χ is a constant. Theoretically, multipath propagation channel increases the signal duration from $1/W$ to the delay spread T . However, the signal bandwidth is also decreased from W to $1/T$ after being convolved with the channel impulse response. Consequently, no extra DOF can be achieved in the time domain. Stated other way, it is impossible to utilize different delay taps as parallel data pipes since all the taps will be excited once the signal is transmitted. Therefore, *Nature* forces the entire signal space to be utilized only for DIV gains.

The next step is to solve for the spatial DOF based on a heuristic approach which is similar to the rationale presented in [MIL00]. We first consider the monochromatic wave communication between a well-conditioned cluster $\mathcal{C}$ and the Tx array $\mathcal{A}_T$. A key step in calculating the DOF lies in understanding the degree to which the waves in the scattering cluster being generated from different points in $\mathcal{A}_T$ are orthogonal to one another. Consider two source points, $\mathbf{a}_{T1}$ and $\mathbf{a}_{T2}$, in $\mathcal{A}_T$ as shown in Fig. 7.6. Apparently if the points are close, the waves generated by them in $\mathcal{C}$ will be almost identical and will not be orthogonal. As the points move further apart, a rapidly varying phase results, which causes the waves to undergo alternating constructive and

Fig. 7.6. Illustration of the source points $\mathbf{a}_{T1}$ and $\mathbf{a}_{T2}$ in the transmit array and the scattering cluster $\mathcal{C}$. Also shown is an array segment Δl near $\mathbf{a}_{T1}$. Other sources in Δl are assumed to produce waves in $\mathcal{C}$ that are largely similar to the waves from the source point $\mathbf{a}_{T1}$, whereas sources outside Δl are assumed to produce waves in $\mathcal{C}$ that are substantially orthogonal to those from $\mathbf{a}_{T1}$.

destructive interference. An assumption made here is that the best possible choice of relative phase between the two sources is used to eliminate this particular interference effect. With this assumption, as points $\mathbf{a}_{T1}$ and $\mathbf{a}_{T2}$ move further apart, we can expect the waves to become progressively different from one another. To assess the orthogonality of the waves, we can evaluate their overlap integral in $\mathcal{C}$, which is given by

$$\kappa(\mathbf{a}_{T2}, \mathbf{a}_{T1}, \lambda) = \int_{\mathcal{C}} G^H(\mathbf{a}_{T2}, \mathbf{a}_C, \lambda) G(\mathbf{a}_{T1}, \mathbf{a}_C, \lambda) d\mathbf{a}_C \quad (7.26)$$

We now use a simplistic model to obtain some approximate results. As mentioned above, the overlap is expected to be large initially and decrease as the two source points are separated. It is presumed that for any point $\mathbf{a}_{T1}$, there is a finite length Δl

for which $\kappa(\mathbf{a}_{T2}, \mathbf{a}_{T1}, \lambda)$ is finite and approximately constant. In addition, $\kappa(\mathbf{a}_{T2}, \mathbf{a}_{T1}, \lambda)$ is approximately zero for $\mathbf{a}_{T2}$ outside this interval. To estimate the dimension of Δl , only $\mathcal{A}_T$ and $\mathcal{C}$ that are far apart compared with their linear dimensions are considered, such that paraxial approximations can be made. Subsequently we have

$$\begin{aligned}
 & G^H(\mathbf{a}_{T2}, \mathbf{a}_C, \lambda) G(\mathbf{a}_{T1}, \mathbf{a}_C, \lambda) \\
 &= \frac{1}{(4\pi d_T)^2} \exp \left[-i \frac{2\pi}{\lambda} (|\mathbf{a}_C - \mathbf{a}_{T1}| - |\mathbf{a}_C - \mathbf{a}_{T2}|) \right] \\
 &\approx \frac{1}{(4\pi d_T)^2} \exp \left(-i \frac{2\pi}{\lambda} \left\{ (y_C - y_T) + \frac{(x_C - x_{T1})^2}{2d_T} - (y_C - y_T) - \frac{(x_C - x_{T2})^2}{2d_T} \right\} \right) \quad (7.27) \\
 &= \frac{1}{(4\pi d_T)^2} \exp \left(-i \frac{2\pi}{\lambda} \left\{ -\frac{x_{T2}^2}{2d_T} + \frac{x_{T1}^2}{2d_T} + \frac{1}{d_T} [x_C (x_{T2} - x_{T1})] \right\} \right) \\
 &= \frac{1}{(4\pi d_T)^2} F_T(\mathbf{a}_{T2}, \lambda) F_T^H(\mathbf{a}_{T1}, \lambda) \exp \left[-i \frac{2\pi x_C (x_{T2} - x_{T1})}{\lambda d_T} \right]
 \end{aligned}$$

where $F_T(\mathbf{a}_T, \lambda)$ is the focusing function corresponding to a spherical wave which is centered on the receive area and is separated to extract the underlying spherical focusing of the source [MIL00]. Also refer to Fig. 7.6 for definitions of all the other variables. The remaining exponential expression is unity for small arguments. However after $(x_{T2} - x_{T1})$ becomes sufficiently large, the function overall becomes oscillatory as we move through $\mathcal{C}$, for which the integral will tend to average to zero. Essentially, we are unable to avoid the effects of interference in integrating over $\mathcal{C}$, no matter how well the relative phases of the sources are chosen. A characteristic size of $(x_{T2} - x_{T1})$ for which this interference becomes negligible occurs when

$$\left| \frac{2\pi \cdot 2R(x_{T2} - x_{T1})}{\lambda d_T} \right| \approx 2\pi$$

That is, when

$$\left| (x_{T2} - x_{T1}) \right| \approx \frac{\lambda d_T}{2R} \quad (7.28)$$

In this case, as we integrate from one end of $\mathcal{C}$ to another extreme in the x -direction, the phase of the integrand will change by 2π .

Next we select a basis which is essentially uniform within Δl near the point $\mathbf{a}_{T1}$ and is zero elsewhere, i.e., within Δl , we have the eigenfunction

$$\psi_1(\mathbf{a}_{T1}, \lambda) = \frac{1}{\sqrt{\Delta l}} F_T(\mathbf{a}_{T1}, \lambda) \quad (7.29)$$

To form another function orthogonal to that of (7.29), we can simply move sideways to an adjacent array segment of the same length and construct a similar second eigenfunction in that segment. The second eigenfunction would be orthogonal to the first since no overlap occurs between the functions. This process continues until all the aperture is used, following which we have a number $O(\lambda)$ of orthogonal functions

$$O(\lambda) = \frac{2RL}{\lambda d_T} \quad (7.30)$$

all having the same eigenvalues, where L is the transmit array length.

The simplistic approximate eigenfunctions are not similar to the exact prolate spheroidal functions derived from rigorous mathematical approaches, since they fill the entire aperture extent rather than our solutions that are localized each within a sub-division Δl . However, all those approximate eigenfunctions are degenerate and correspond to the same eigenvalue, thus we can take orthogonal linear combinations of such functions instead. And it is easy to construct sets which fill the whole aperture with the number of orthogonal functions remaining unchanged by such linear combinations. Though the functions should not have the abrupt cutoff at the

Fig. 7.7. A microscopic view of DOF

edges that our simplistic model has, smoothing them out nevertheless would not alter the basic argument.

By applying the same arguments presented in Section 7.2.2, there is a need to take into account of the system-dependency of R for DOF computation. Following from (7.30), the differential DOF on $\delta\mu$ is essentially the area of each rectangle scaled by $\delta y/y(0) \times \delta f/W$ in Fig. 7.7, which can be obtained as

$$\delta \text{DOF} = \frac{4R(2a_T(y), f)a_T(y)}{\lambda d_T} \times \frac{\delta y \cdot \delta f}{y(0) \cdot W} \quad (7.31)$$

and the total available communication modes are simply

$$\text{DOF} = \sum \delta \text{DOF} = \iint_{\mathcal{S}} \frac{4fR(2a_T(y), f)a_T(y)}{v_c d_T y(0)W} dy df \quad (7.32)$$

If the scattering is highly correlated, $\mathbf{H}(\cdot, \cdot)$ will be rank-deficient. In such a scenario, the effective DOF are limited by the keyhole population (7.23) with the overall DOF estimate

$$\overline{\text{DOF}} = \min \{\text{DOF}, \text{KP}\} \quad (7.33)$$

It is worth mentioning here that (7.32) is a more optimistic upper bound for the DOF measure as compared to the upper bound for the DIV estimate in (7.21). The reason is because it has been assumed that a *complete* picture of the scattering environment can be obtained in order to fully explore the DOF offered by the channel, whereas this requirement is not enforced for the calculation of diversity gains in Section 7.2.2.

7.3 DIV and DOF in General Scattering Environments with Imperfect Receiving Conditions

7.3.1 Diversity Gain Calculation

The indoor channel measurements have shown that physical paths are clustered both in the temporal and spatial domains [PH03, COR01, SJJ++00]. To apply the methodology presented in Section 7.2 to more general situations, we first identify three classes of clusters as depicted in Fig. 7.8(a):

- Class-One (C-I) Cluster being the scattering cluster illuminated by the Tx array only, has the subtended angular interval $\Omega_{T,q}^I$, cluster radius R_q^I , and keyhole radius r_q^I where $q = 1, 2, 3, \dots, Q$.

Fig. 7.8. (a) Classification of scattering clusters and (b) the propagation process for DIV calculation

- Class-Two (C-II) Cluster being the scattering cluster observed from the Rx array only, has the subtended angular interval $\Omega_{R,s}^{II}$, delay interval T_s^{II} , cluster radius R_s^{II} , and keyhole radius r_s^{II} where $s = 1, 2, 3, \dots, S$.

- Class-Three (C-III) Cluster has a LoS connection to both the Tx and Rx arrays.

Therefore, each transmit interval $\Omega_{T,l}^{\text{III}}$ ($l = 1, 2, 3, \dots, L$) is one-to-one mapped to a corresponding receive interval $\Omega_{R,l}^{\text{III}}$ such that the joint spectra cannot be described by a product of separate AoA and AoD spectra (i.e., a single bounce channel). It has a delay interval of T_l^{III} , cluster radius R_l^{III} , and keyhole radius r_l^{III} .

We can thus obtain the general expression for channel outputs as (see also Fig. 7.8(b))

$$z(t, \mathcal{G}_R) = \frac{1}{\sqrt{2E}} \left\{ \underbrace{\sum_s \sum_{\tilde{j}} \sum_{\tilde{k}} \tilde{a}_{s,\tilde{j},\tilde{k}} u(t - \tau_{s,\tilde{j}}, \mathcal{G}_T \mapsto \mathcal{G}_{R,s,\tilde{j},\tilde{k}}) \exp(-i2\pi\tau_{s,\tilde{j}})}_{\text{C-II cluster}} + \underbrace{\sum_l \sum_j \sum_k a_{l,j,k} u(t - \tau_{l,j}, \mathcal{G}_T \mapsto \mathcal{G}_{R,l,j,k}) \exp(-i2\pi\tau_{l,j})}_{\text{C-III cluster}} \right\} \quad (7.34)$$

$\tau_{l,j}$ is the propagation delay caused by the j th scattering strip of the l th C-III cluster.

$\mathcal{G}_T \mapsto \mathcal{G}_{R,l,j,k}$ denotes the mapping of waveforms from $\mathcal{A}_T$ to $\mathcal{A}_R$ in the ACA domain

by the k th subdivision of the j th scattering strip in the l th C-III cluster. The variables

for the C-II clusters are defined in a similar manner. The approximation in (7.4) still

applies to both the C-II and C-III clusters and is given as follows

$$\begin{cases} u(t - \tau_{s,\tilde{j}}) \approx u\left(t - \frac{\tilde{n}}{W}\right) & \text{provided that } \left| \tau_{s,\tilde{j}} - \frac{\tilde{n}}{W} \right| \leq \frac{1}{2W} \\ u(t - \tau_{l,j}) \approx u\left(t - \frac{n}{W}\right) & \text{provided that } \left| \tau_{l,j} - \frac{n}{W} \right| \leq \frac{1}{2W} \end{cases} \quad (7.35)$$

where $\tilde{n}$ and n are both integers. While (7.5) remains valid for C-III cluster, for C-II

cluster it has to be reformulated as

$$u\left(\mathcal{G}_T \mapsto \mathcal{G}_{R,s,\tilde{j},\tilde{k}}\right) \stackrel{(1)}{\equiv} \sum_{q,\kappa} u\left(\mathcal{G}_T \mapsto \mathcal{G}_{T,q,\kappa}\right) \approx \sum_{\tilde{m}} u\left(\mathcal{G}_T \mapsto \frac{\tilde{m}}{L/\lambda}\right) \quad (7.36)$$

provided that $\left|\mathcal{G}_{T,q,\kappa} - \frac{\tilde{m}}{L/\lambda}\right| \leq \frac{1}{2L/\lambda}$

for any s , $\tilde{j}$, and $\tilde{k}$. $\mathcal{G}_{T,q,\kappa}$ is the AoD from the Tx array that impinges upon the κ th angular subdivision of the q th C-I cluster and $\tilde{m}$ is an integer. The equality (1) is arrived by assuming that the intermediate scattering process (ISP) is DIV-conservative when the waveforms are transformed from C-I clusters to C-II clusters, and no signal is transformed from C-I clusters to C-III clusters.

Following the same procedures as presented in Section 7.2.2 and assuming an infinite-rank channel transfer matrix, (7.34) can be rewritten as a combination of many known spatial-temporal functions multiplied by statistically independent random variables $\tilde{b}_{\tilde{n},\tilde{m}}$ and $b_{n,m}$, viz.,

$$z(t, \mathcal{G}_R) \approx \sum_{\tilde{n}} \sum_{\tilde{m}} \tilde{b}_{\tilde{n},\tilde{m}} u\left(t - \frac{\tilde{n}}{W}, \mathcal{G}_T \mapsto \frac{\tilde{m}}{L/\lambda}\right) + \sum_n \sum_m b_{n,m} u\left(t - \frac{n}{W}, \mathcal{G}_T \mapsto \frac{m}{L/\lambda}\right) \quad (7.37)$$

It can be easily shown that the number of independent channels through which replicas of the same information signal can be transmitted is $\mathcal{J} \times W \times \left|\bigcup_s \mathcal{T}_s^{\text{II}}\right| \times (L/\lambda) \times \left|\bigcup_q \Omega_{T,q}^{\text{I}}\right|$ for C-II clusters and $W \times \left|\bigcup_l \mathcal{T}_l^{\text{III}}\right| \times (L/\lambda) \times \left|\bigcup_l \Omega_{T,l}^{\text{III}}\right|$ for C-III clusters, where $\mathcal{J}$ is the average number of distinctive AoAs from C-II clusters to the Rx array.

Along the lines of derivation for (7.20) and noting that the delay interval, T , and angular interval, Ω_T , are both system-dependent parameters, the overall DIV is thus obtained as

$$\begin{aligned} \text{DIV} = \iint_{\mathcal{S}} \mathcal{J} \cdot \frac{2a_T(y)f \left| \bigcup_q \Omega_{T,q}^I(2a_T(y), f) \right| \left| \bigcup_s T_s^{\text{II}}(2a_T(y), f) \right|}{y(0)v_c} dydf \\ + \iint_{\mathcal{S}} \frac{2a_T(y)f \left| \bigcup_l \Omega_{T,l}^{\text{III}}(2a_T(y), f) \right| \left| \bigcup_l T_l^{\text{III}}(2a_T(y), f) \right|}{y(0)v_c} dydf \end{aligned} \quad (7.38)$$

Equation (7.38) calculates the DIV for receivers with perfect beamforming capabilities and infinite bandwidth. In practical situations, however, receivers have finite aperture window L' (henceforth finite array resolution) and filtering bandwidth W' . The overall DIV can be straightforwardly deduced as

$$\text{DIV} = \text{DIV}^{\text{II}} + \text{DIV}^{\text{III}} \quad (7.39)$$

with

$$\begin{aligned} \text{DIV}^{\text{II}} = \iint_{\mathcal{S}'} \frac{2a_T(y)f \left| \bigcup_q \Omega_{T,q}^I(2a_T(y), f) \right|}{v_c} \times \frac{L'f \left| \bigcup_s \Omega_{R,s}^{\text{II}}(2a_T(y), f) \right|}{v_c} \\ \times \frac{\left| \bigcup_s T_s^{\text{II}}(2a_T(y), f) \right|}{y(0)} dydf \end{aligned} \quad (7.40)$$

$$\begin{aligned} \text{DIV}^{\text{III}} = \iint_{\mathcal{S}'} \min \left\{ \frac{2a_T(y)f \left| \bigcup_l \Omega_{T,l}^{\text{III}}(2a_T(y), f) \right|}{v_c}, \frac{L'f \left| \bigcup_l \Omega_{R,l}^{\text{III}}(2a_T(y), f) \right|}{v_c} \right\} \\ \times \frac{\left| \bigcup_l T_l^{\text{III}}(2a_T(y), f) \right|}{y(0)} dydf \end{aligned} \quad (7.41)$$

where $\mathcal{S}' = \mathcal{A}_T \times (\mathcal{F} \cap \mathcal{F}')$ with $\mathcal{F}'$ being the frequency space spanned by the filter at the Rx. Equations (7.40) and (7.41) provide the DIV estimate for C-II and C-III clusters, respectively.

To derive the general formulation of DIV for scattering clusters with non-vanishing keyholes, we simply compare the DIV density of *each* cluster to its KP density function and substitute the smaller one to either (7.40) or (7.41). For example, consider the ℓ th C-III cluster, if

$$\min \left\{ \frac{2a_T(y)f|\Omega_{T,\ell}^{\text{III}}(2a_T(y),f)|}{v_c}, \frac{L'f|\Omega_{R,\ell}^{\text{III}}(2a_T(y),f)|}{v_c} \right\} \\ \times \frac{|T_\ell^{\text{III}}(2a_T(y),f)|}{y(0)} > \left[\frac{R_\ell^{\text{III}}(2a_T(y),f)}{r_\ell^{\text{III}}(2a_T(y),f)} \right]^2 \frac{1}{|\hat{S}'|} \quad \forall (y,f) \in \hat{S}'$$

(7.41) is modified as

$$\text{DIV}^{\text{III}} = \iint_{\hat{S}} \left\{ \min \left\{ \frac{2a_T(y)f|\bigcup_{\ell \neq t} \Omega_{T,\ell}^{\text{III}}(2a_T(y),f)|}{v_c}, \frac{L'f|\bigcup_{\ell \neq t} \Omega_{R,\ell}^{\text{III}}(2a_T(y),f)|}{v_c} \right\} \right. \\ \left. \times \frac{|\bigcup_{\ell \neq t} T_\ell^{\text{III}}(2a_T(y),f)|}{y(0)} \right\} + \frac{1}{|\hat{S}'|} \left[\frac{R_\ell^{\text{III}}(2a_T(y),f)}{r_\ell^{\text{III}}(2a_T(y),f)} \right]^2 dydf$$

7.3.2 Degrees-of-freedom Calculation

For C-I and C-II clusters, the spatial channels involve the communicating with waves between two scattering areas in the ISP. Consider a multi-hop propagation path $\mathfrak{P}$ from the Tx to Rx via the q th C-I cluster and the s th C-II cluster as depicted in Fig. 7.9. We denote the largest free-run-length of waves as $d_{\mathfrak{p}}$, which is the maximum distance traveled by a plane wave before it intercedes with the next scattering cluster. We further define the cluster radius of the two clusters connected by $d_{\mathfrak{p}}$ as $\bar{R}_{\mathfrak{p}}$ and $\hat{R}_{\mathfrak{p}}$, while their keyhole radius as $\bar{r}_{\mathfrak{p}}$ and $\hat{r}_{\mathfrak{p}}$. If the clusters and arrays are far apart from one another compared with their linear dimensions, we can safely infer that the spatial DOF associated with $\mathfrak{P}$ is constrained by the free path of waves. Following from (7.32), the overall DOF for path $\mathfrak{P}$ (see also Fig. 7.9)

$$\text{DOF}_{\mathfrak{p}} = \min \{ \text{DOF}_{\mathfrak{p}1}, \text{DOF}_{\mathfrak{p}2} \} \quad (7.42)$$

where

$$\text{DOF}_{p1} = \iint_{\mathcal{S}} \min \left\{ \frac{2fR_s^{\text{II}}(2a_T(y), f)L'}{v_c d_s^{\text{II}}}, \frac{4fR_q^{\text{I}}(2a_T(y), f)a_T(y)}{v_c d_q^{\text{I}}} \right\} \frac{1}{y(0)W} dydf \quad (7.43)$$

$$\text{DOF}_{p2} = \iint_{\mathcal{S}} \frac{4f\bar{R}_p(2a_T(y), f)\hat{R}_p(2a_T(y), f)}{v_c d_p y(0)W} dydf \quad (7.44)$$

d_q^{I} and d_s^{II} represent the distances between the Tx and the q th C-I cluster, the Rx and the s th C-II cluster, respectively. Equations (7.43) and (7.44) provide the DOF estimate for the signal flow at two ends of the path link and the ISP, respectively.

The general expression of $\overline{\text{DOF}}_p$ for realistic environments with non-vanishing keyholes can be obtained by selecting the smaller value of DOF_p and the minimum KP among all the clusters associated with this path link, i.e.,

Fig. 7.9. DOF evaluation for two typical propagation paths: $\mathcal{P}$ and $\mathcal{B}$. The former involves a multiple-scattering process whereas the latter signifies a single-bounce channel.

$$KP_b = \min_{\eta \in \Xi_{c,b}} \left\{ \iint_{\hat{S}'} \frac{1}{|\hat{S}'|} \frac{R_\eta^2(2a_T(y), f)}{r_\eta^2(2a_T(y), f)} dydf \right\} \quad (7.45)$$

where $\Xi_{c,b}$ is the set of all clusters along β .

Now we consider the orthogonal communication modes related to the wave propagation via the l th C-III cluster. We denote this path link as β as depicted in Fig. 7.9. Similarly, by assuming a free-run-length-limited capacity gain, the DOF estimate can be easily obtained as

$$\begin{aligned} \overline{DOF}_\beta &= \min \{ \overline{DOF}_\beta, KP_\beta \} \\ &= \min \left\{ \iint_{\hat{S}} \min \left\{ \frac{2fL'R_l^{\text{III}}(2a_T(y), f)}{v_c d_{R,l}^{\text{III}}}, \frac{4fa_T(y)R_l^{\text{III}}(2a_T(y), f)}{v_c d_{T,l}^{\text{III}}} \right\} \frac{1}{y(0)W} dydf \right. \\ &\quad \left. , \min_{\gamma \in \Xi_{c,\beta}} \left\{ \iint_{\hat{S}'} \frac{1}{|\hat{S}'|} \frac{R_\gamma^2(2a_T(y), f)}{r_\gamma^2(2a_T(y), f)} dydf \right\} \right\} \end{aligned} \quad (7.46)$$

$d_{T,l}^{\text{III}}$ and $d_{R,l}^{\text{III}}$ correspond to the distances of separation between the l th C-III cluster and the Tx and Rx arrays, respectively. $\Xi_{c,\beta}$ stands for the set of all clusters associated with β . Subsequently, the overall DOF is simply the summation of all parallel channels in the wireless link, and reads

$$\overline{DOF} = \sum_b \overline{DOF}_b + \sum_\beta \overline{DOF}_\beta \quad (7.47)$$

7.4 Numerical Examples

7.4.1 SCSB Channel with Perfect Receiver Conditions

In the first study case, we will focus on a fully scattered environment where each keyhole is regarded as an ideal point reradiator. For the sake of simplicity, it is

assumed that the dependencies of T , Ω and R are separable in the aperture and frequency domains [cf. (7.15)~(7.18)],

$$\begin{cases} \Omega(2a, f) = \sqrt{\Omega_a(2a)\Omega_f(f)} = \sqrt{2 \sin\left(\beta^{(\Theta, 2a)} e^{2\gamma^{(\Theta, 2a)} a/\lambda}\right) \times 2 \sin\left(\beta^{(\Theta, f)} e^{\gamma^{(\Theta, f)} f}\right)} \\ T(2a, f) = \sqrt{T_a(2a)T_f(f)} = \sqrt{\left(\beta^{(T, 2a)} e^{2\gamma^{(T, 2a)} a/\lambda}\right) \times \left(\beta^{(T, f)} e^{\gamma^{(T, f)} f}\right)} \\ R(2a, f) = \sqrt{R_a(2a)R_f(f)} = \sqrt{\left(\beta^{(R, 2a)} e^{2\gamma^{(R, 2a)} a/\lambda}\right) \times \left(\beta^{(R, f)} e^{\gamma^{(R, f)} f}\right)} \end{cases} \quad (7.48)$$

It is worth mentioning here that the relationship for the cluster radius versus the array length is assumed to follow an exponential function as well.

Subsequently, we study four types of typical PSD functions, where the separability of signal spectrum is assumed.

$$f_p(\mathbf{a}_T, f) = \begin{cases} C_3 \exp(-2|a|/\sigma_a) \exp(-|f - f_0|/\sigma_f) & |a| \leq L_0/2, |f - f_0| \leq W_0/2 & (7.49.1) \\ 1/(L_0 W_0) & |a| \leq L_0/2, |f - f_0| \leq W_0/2 & (7.49.2) \\ 1/(L_0 W_0) & |a| \leq L_0/2, |f - f_1| \leq W_0/2, f_1 > f_0 & (7.49.3) \\ 1/(L_1 W_0) & |a| \leq L_1/2, |f - f_0| \leq W_0/2, L_1 > L_0 & (7.49.4) \\ 1/(L_0 W_1) & |a| \leq L_0/2, |f - f_0| \leq W_1/2, W_1 > W_0 & (7.49.5) \end{cases}$$

Equation (7.49.1) is a multiplication of two truncated Laplacian distributions with C_3 being a normalization factor. It has a sharp peak at $(0, f_0)$ and characterizes signals of narrow power spectrum while (7.49.2) represents signals of flat spectrum. For (7.49.3), the system operates at a higher frequency as compared to (7.49.2). Finally, the transmit power for (7.49.4) and (7.49.5) is distributed over an array of larger aperture length and wider frequency interval, respectively, both of which are compared to (7.49.2). The parameters used in the examples are listed in Table 7.1. After substituting (7.48) and (7.49) into (7.21), the numbers of DIV for the five PSDs are calculated as 23, 31, 33, 29, and 60. Following from (7.32) yields the DOF: 1.96,

Table 7.1. Parameters used in the numerical examples

Parameters in (7.48)		Parameters in (7.49)	Other parameters
$\beta^{(\Theta,2a)}$ 20	$\beta^{(\Theta,f)}$ 20	L_0 30 cm	L' 25 cm
$\gamma^{(\Theta,2a)}$ -0.14	$\gamma^{(\Theta,f)}$ -0.13	W_0 2 GHz	W' 2 GHz
$\beta^{(T,2a)}$ 18	$\beta^{(T,f)}$ 7.5	L_1 60 cm	f' 3 GHz
$\gamma^{(T,2a)}$ -0.25	$\gamma^{(T,f)}$ 0.16	W_1 4 GHz	d_T^{III} 50 m
$\beta^{(R,2a)}$ 16	$\beta^{(R,f)}$ 25	f_0 4 GHz	d_R^{III} 50 m
$\gamma^{(R,2a)}$ -0.15	$\gamma^{(R,f)}$ -0.2	f_1 6 GHz	
		σ_a 5 cm	
		σ_f 1 GHz	

3.16, 3.36, 4.68, and 6.13. Therefore, the following observations can be made for this numerical example: (i) both DIV and DOF will increase if the transmission scheme has a flat power spectrum, operates at a higher frequency, or occupies a broader bandwidth; (ii) both DIV and DOF will not be scaled linearly with respect to the array length as usually believed; (iii) if the multipath channel is defined system-independently, the diversity and capacity gains would be no different for the transmission schemes depicted in (7.49.1), (7.49.2), and (7.49.3).

7.4.2 SCSB Channel with Imperfect Receiver Conditions

If the receiver operates at imperfect conditions (i.e., with a finite receive aperture L' and frequency band $[f'-W'/2, f'+W'/2]$, see Table 7.1), the numbers of DIV and DOF can be obtained as 13 and 1.27 by substituting (7.48) and (7.49.2) into (7.41) and (7.42), assuming the same channel conditions as in Section 7.4.1 and the transmit

signals in the form of (7.49.2). Evidently, both DIV and DOF are greatly reduced by more than half in this case, which certainly holds an intuitive appeal.

7.4.3 MCMB Channel with Imperfect Receiver Conditions

In this section, we will briefly investigate the effect of multiple scattering and multiple clusters on the system diversity estimate. Consider an abstract channel with one C-I and one C-II clusters. In order to make a fair comparison, we further assume that the system operating conditions and cluster properties are similar with the examples in Section 7.4.2 with the only exception that the original AoA-AoD characteristics of the single C-III cluster are now decoupled into two clusters. Following from (7.40), the number of DIV is calculated as 19 which is larger than the figure obtained in the previous case. Thus, in terms of the maximum diversity gain, MCMB channels are better than SCSB channels. However, the attainable DOF would remain the same as might be inspected by intuition. Therefore, characterizing the bouncing frequency in different physical environments helps system designers to deploy more efficient space-time diversity and multiplexing schemes.

As a final remark of this section, our arguments have resulted from an oversimplification in the model, but they do correctly count the available DIV and DOF and yield a substantially correct intuitive picture. Nonetheless, the implementation of the general framework to represent the real-world situation would have to be justified by further measurements.

7.5 Conclusions

The DIV and DOF calculation for system-dependent wideband directional channels has been presented in this chapter. It has been shown that the possible diversity and multiplexing gains are greatly dependent on the power spectrum density of the transmitted waveforms, which is based on the fact that the wireless channel could be perturbed to a certain extent when signals are filtered either in the azimuth or frequency domain, thereby allowing for a more flexible signaling strategy optimization. This leaves a very interesting open-ended question for future investigation.

CHAPTER 8 UWB SOURCE LOCALIZATION USING A PARTICLE-SWARM-OPTIMIZED CAPON ESTIMATOR: A FREQUENCY-DEPENDENT CHANNEL MODELING PERSPECTIVE

8.1 Introduction

Ultrawideband (UWB) channel features three distinct characteristics: presence of large angle spreads for multiple uncorrelated scattering clusters [MFP03], wideband array signals [HUS02], and frequency-dependent dispersive effects [QIU02, PH03]. Therefore, fundamental differences exist between the conventional and UWB array signal processing techniques [HUS02, GCB02, CRA00]. In this chapter, we extend the simple yet accurate generalized-Capon-estimator for narrowband stationary signals [HSG04] to UWB scenarios based on a realistic signal/channel model in the angular-frequency domain. The particle swarm optimization (PSO) [KE95, KES01], which is a robust stochastic evolutionary computation technique and is new to the signal processing community, is then implemented to perform the multidimensional search for the parametric setting of the underlying channel model. The chapter is organized as follows. In Section 8.2, we present the analytical model in angular-frequency domain of UWB impulse waveforms, and provide the formulation of the generalized broadband Capon estimator. Based on this estimation method, a detailed explanation of the PSO algorithm for our problem of interest is presented in Section

8.3. Computer simulation results are presented in Section 8.4, and conclusions are drawn in Section 8.5.

8.2 Signal and Channel Models for UWB Transmission Schemes

For simplicity, we assume that the distributed signals are on the same plane as the array (i.e., a two-dimensional propagation) and the total fields are dominated by the scattered fields. The angular-frequency domain expression of the channel response can be formulated as

$$h(f, \theta) = \sum_{k=1}^K \sum_{l \in S_k} \beta_{kl} e^{j\phi_{kl}} e^{-j2\pi f \tau_{kl}} A_{kl}(f, \theta_{kl}) \delta(\theta - \theta_{kl}) \quad (8.1)$$

where K is the total number of the scattering centers (clusters) and S_k denotes the set of propagation paths (point source components) in the k th cluster. $\delta(\cdot)$ is the delta function. It is further assumed that the K diffuse sources and the point scattering components within each source are independent of each other. This assumption is often used to simplify calculations in many applications [AP00] and has been experimentally shown to be relevant for the indoor channel [CRA00]. Moreover, all the scattering centers have well-defined frequency dependencies as given by their leading-order geometry theory of diffraction (GTD) scattering coefficients.

For each point source, we model it using four parameters $\{\beta_{kl} e^{j\phi_{kl}}; A_{kl}; \theta_{kl}; \tau_{kl}\}$ corresponding to {complex amplitude; frequency dependence; angle-of-arrival; path delay time}. A special form of $A_{kl}(f, \theta_{kl}) \propto f^{\alpha_{kl}}$ is used to model the frequency dependence of common structures [QIU02, QL99, QIU96, QIU04a, QIU04b]. For

example, a ray coming from a line-of-sight (LoS) path has $\alpha_{kl} = 0$ and a ray coming from a wall edge has $\alpha_{kl} = -0.5$. If the plane wave is diffracted at an endpoint or tip such as the desk corner, $\alpha_{kl} = -1$.

If these point source components are densely clustered together, we replace each of the point sources by a density function spreading over a small angular volume $\Delta\theta$ for any frequency component f . Equation (8.1) can thus be reformulated as¹

$$h(f, \theta) = \lim_{\Delta\theta \rightarrow 0} \sum_{k=1}^K \sum_{l \in S_k} \beta_k e^{j\phi_k} e^{-j2\pi f \tau_k} f^{\alpha_k} \sqrt{\rho_k(f, \theta_k, \gamma_k) \Delta\theta} \cdot \delta(\theta - \theta_k) \quad (8.2)$$

where $\rho_k(f, \theta, \gamma_k)$ is the power azimuthal spectrum (PAS) characterizing the distribution of the k th cluster at each f . $\theta \in \Theta_k(f)$ where Θ_k is the region for the k th scattering center within which the diffuse signal spreads and is frequency-dependent [POO04]. γ_k is a parameter vector used to define the location of the k th cluster. As mentioned in Chapter 7, Intel Corporation has performed several UWB wireless channel measurements in both non-line-of-sight (NLoS) and LoS scenarios, at both office and residential environments. The frequency was swept from 2 to 8 GHz with resolution of 3.75 MHz. It was also shown that the average cluster angle spread (AS) decreased with increasing frequency as depicted in Fig. 7.4(c), which was arguably because of the fact that an electromagnetic wave of higher frequency attenuated more severely after passing through or bouncing off objects in the environment and channel became less diffuse at higher frequency. In this chapter, instead of an exponential curve, we attempt to describe the AS variation by polynomial regression curves, which offers a greater flexibility to adjust the model order, viz.,

¹ The subscript l has been dropped in (8.2) to indicate that β_k , ϕ_k , A_k , and τ_k correspond to the strongest path in the k th cluster assuming a symmetric density function $\rho_k(f, \theta, \gamma_k)$.

$$\sigma_{\theta,k}(f) = \sum_{m=0}^{\mathcal{O}_k} c_{m,k} f^m \text{ (GHz)} \quad (8.3)$$

where $\mathcal{O}_k$ is the order of the polynomial equation for the k th cluster and $c_{m,k}$'s are the corresponding coefficients. Fig. 8.1 depicts fitting of the empirical data to four 3rd-order polynomial equations, for which reasonable agreement is justified for all the four groups of data. As we further increase the polynomial order, finer curve-fitting can be obtained. Nevertheless, increasing model order will significantly raise the computational load without much improvement in the underlying model structure. Following from the above arguments, the parameter vector γ_k is described as $[\underline{\theta}_k, c_{0,k}, \dots, c_{\mathcal{O}_k,k}]$ with $\underline{\theta}_k$ being the mean angle-of-arrival (AoA) of the k th cluster. The most common shapes suggested for the PAS function $\rho_k(f, \theta, \gamma_k)$ include uniform [ECS++98], Gaussian [ECS++98], and Laplacian [CRA00, POO04, MFP03], depending on the propagation environments.

Fig. 8.1. Average cluster angle spreads in the townhouse and office environments [POO04] and their polynomial regression curves

Consider that the signals of K scattering centers impinge on an array of q sensors.

The complex envelope of the array output can be expressed as

$$\begin{aligned} \mathbf{y}(t) &= \sum_{k=1}^K \mathbf{s}_k(t) + \mathbf{n}(t) \\ &= \sum_{k=1}^K \int_{f_{\min}}^{f_{\max}} \int_{\theta \in \Theta_k(f)} \beta_k e^{j\phi_k} e^{-i2\pi f \tau_k} f^{\alpha_k} \sqrt{\rho_k(f, \theta, \gamma_k)} d\theta e^{i2\pi f(t + \tau(\theta))} dv(f) + \mathbf{n}(t) \end{aligned} \quad (8.4)$$

where $\mathbf{y}(t)$ is the $q \times 1$ array snapshot vector, $\mathbf{s}_k(t)$ is the $q \times 1$ vector that describes the contribution of the k th scattering center to the array output, and $\mathbf{n}(t)$ is the vector of sensor noise. $[f_{\min}, f_{\max}]$ is the spectral support of the signal, and $dv(f)$ denotes a measure of the signal spectrum at f . $\tau(\theta)$ is a $q \times 1$ vector representing the propagation delay corresponding to an AoA θ with respect to a given reference sensor. For the case of a uniform linear array, $\tau(\theta) = [0 \ \Delta \sin \theta / v_c \ 2\Delta \sin \theta / v_c \ \cdots \ (q-1)\Delta \sin \theta / v_c]^T$ where Δ is the inter-element spacing and v_c is the speed of electromagnetic waves. Assuming uncorrelated AoAs, the (q_1, q_2) th element of the spatial covariance matrix of the received signal from the k th cluster can be expressed as

$$\begin{aligned} [\mathbf{R}_k(\alpha_k, \gamma_k)]_{q_1, q_2} &= \int_{f_{\min}}^{f_{\max}} \int_{f'_{\min}}^{f'_{\max}} \int_{\theta \in \Theta_k(f)} \beta_k^2 e^{-i2\pi(f-f')\tau_k} (ff')^{\alpha_k} \sqrt{\rho_k(f, \theta, \gamma_k) \rho_k(f', \theta, \gamma_k)} \\ &\quad \times e^{i2\pi f \tau_{q_1}(\theta) - i2\pi f' \tau_{q_2}(\theta)} d\theta E[dv(f) dv^*(f')] \end{aligned} \quad (8.5)$$

We further make the common assumption that the spectral components at different frequencies are mutually orthogonal and the typical UWB pulse is the Gaussian monocycle [SM04, CRA00], which has been widely used for analytical evaluation of UWB systems. Hence,

$$E[dv(f) dv^*(f')] = S(f) = Af^4 T^6 \exp(-\pi f^2 T^2) \delta(f - f') \quad (8.6)$$

where $S(f)$ denotes the power spectral density of the source [CRA00], A is a normalization constant, and T is a parameter determining the temporal width of the pulse. Casting (8.6) into (8.5) yields

$$\begin{aligned} & [\mathbf{R}_k(\alpha_k, \gamma_k)]_{q_1, q_2} \\ &= \int_{f_{\min}}^{f_{\max}} \int_{\theta \in \Theta_k(f)} A \beta_k^2 T^6 f^{2\alpha_k+4} \exp \left\{ i 2\pi f \frac{(q_1 - q_2) \Delta}{v_c} \sin \theta - \pi f^2 T^2 \right\} \rho_k(f, \theta, \gamma_k) d\theta df \end{aligned} \quad (8.7)$$

In the cases of uniform, Gaussian, and Laplacian distributions, $\rho_k(f, \theta, \gamma_k)$ takes the forms of

- Uniform distribution

$$\rho_k(f, \theta, \gamma_k) = Q_{U,k} \cdot \left\{ \varepsilon \left[\theta - (\underline{\theta}_k - \sigma_{\theta,k}(f)/2) \right] - \varepsilon \left[\theta - (\underline{\theta}_k + \sigma_{\theta,k}(f)/2) \right] \right\} \quad (8.8.1)$$

- Gaussian distribution

$$\rho_k(f, \theta, \gamma_k) = Q_{G,k} \cdot \frac{2}{\sigma_{\theta,k}(f) \sqrt{2\pi}} \exp \left[-\frac{2(\theta - \underline{\theta}_k)^2}{\sigma_{\theta,k}^2(f)} \right] \quad (8.8.2)$$

- Laplacian distribution

$$\rho_k(f, \theta, \gamma_k) = Q_{L,k} \cdot \frac{\sqrt{2}}{\sigma_{\theta,k}(f)} \exp \left[-\frac{2\sqrt{2}|\theta - \underline{\theta}_k|}{\sigma_{\theta,k}(f)} \right] \quad (8.8.3)$$

respectively, where $\varepsilon[\cdot]$ is the step function and $Q_{U,k}$, $Q_{G,k}$, and $Q_{L,k}$ are normalization constants. Assuming the additive noise at each sensor to be white Gaussian, the overall covariance matrix is given by

$$\mathbf{R}_y = \mathbf{R}_x(\mathbf{a}, \gamma) + \sigma^2 \mathbf{I} = \sum_{k=1}^K \mathbf{R}_k(\alpha_k, \gamma_k) + \sigma^2 \mathbf{I} \quad (8.9)$$

where $\mathbf{R}_y = E\{\mathbf{y}(t)\mathbf{y}^H(t)\}$ is the array covariance matrix and $\mathbf{R}_s(\mathbf{a}, \boldsymbol{\gamma}) = E\{\mathbf{s}(t)\mathbf{s}^H(t)\}$ is the overall covariance matrix of the signals from the multiple incoherently distributed scattering centers with $\mathbf{a} = [\alpha_1, \alpha_2, \dots, \alpha_K]$ and $\boldsymbol{\gamma} = [\gamma_1, \gamma_2, \dots, \gamma_k]$.

A generalized Capon AoA estimator for narrowband sources has been proposed in [HSG04] and shows a substantially improved performance relative to several popular spread source localization techniques such as the DISPARE algorithm [MSW96] and the root-MUSIC based estimator [BO00]. We will further extend this method to UWB scenarios as presented below. The minimum variance distortionless response problem is formulated as

$$\min_{\mathbf{w}} \mathbf{w}^H \mathbf{R}_y \mathbf{w} \quad \text{subject to} \quad \mathbf{w}^H \mathbf{R}_s(\boldsymbol{\Lambda}) \mathbf{w} = 1 \quad (8.10)$$

Following from (8.10), the proposed spatial filter maintains distortionless spatial response to multiple hypothetical sources with the parameter vector $\boldsymbol{\Lambda} = [\mathbf{a}, \boldsymbol{\gamma}]$ while rejecting the contribution of any other sources. Such a response is now represented by means of the covariance matrix $\mathbf{R}_s(\boldsymbol{\Lambda})$ and thus the distortionless response is maintained in (8.10) in the mean power sense rather than in the deterministic sense. According to the derivations in [HSG04], the minimum value of the objective function $\mathbf{w}^H \mathbf{R}_y \mathbf{w}$ is equal to the smallest generalized eigenvalue of the matrix pencil $\{\mathbf{R}_y, \mathbf{R}_s(\boldsymbol{\Lambda})\}$, and the generalized Capon pseudo-spectrum is defined as the beamformer output power when the beamformer is steered to sources with the parameter $\boldsymbol{\Lambda}$. The generalized Capon estimator is thus given by

$$P(\boldsymbol{\Lambda}) = \lambda_{\min} \{\mathbf{R}_y, \mathbf{R}_s(\boldsymbol{\Lambda})\} = \frac{1}{\sigma_{\max} \{\mathbf{R}_y^{-1} \mathbf{R}_s(\boldsymbol{\Lambda})\}} \quad (8.11)$$

where $\lambda_{\min} \{\cdot\}$ stands for the minimum generalized eigenvalue of a matrix pencil and $\sigma_{\max} \{\cdot\}$ is the maximum eigenvalue of a matrix. The parametric vector $\boldsymbol{\Lambda}$ can then be

estimated from the global maximum of (8.11). Apparently, it involves a multidimensional search, which can hardly be solved by traditional nonlinear programming techniques. This motivates us to resort to more powerful evolutionary algorithms. In the discussions that follow, instead of (8.11), we will reformulate the objective function in a different manner and thus change it into a constrained minimization problem

$$\begin{aligned} & \text{minimise } o(\Lambda) = \frac{1}{P(\Lambda)} = \sigma_{\max} \{ \mathbf{R}_y^{-1} \mathbf{R}_s(\Lambda) \} \quad \Lambda \in \Omega \\ & \text{subject to } \sum_{m=0}^{Q_k} c_{m,k} f_{\min}^m \leq \sigma_{\theta \max}, \quad \sum_{m=0}^{Q_k} c_{m,k} f_{\max}^m \geq \sigma_{\theta \min}, \quad \sum_{m=1}^{Q_k} m c_{m,k} f_{\max}^{m-1} \leq 0 \end{aligned} \quad (8.12)$$

where Ω is usually defined by the upper and lower bounds of the decision variable Λ . The constraints are imposed to ensure that the angle spread is limited within the range $[\sigma_{\theta \min}, \sigma_{\theta \max}]$ over a predefined frequency range, $\zeta f \triangleq [f_{\min}, f_{\max}]$, and it decreases monotonically as the frequency increases.

8.3 Particle Swarm Optimization for Locating the Global Minimum

In 1995, Eberhart and Kennedy proposed the PSO algorithm inspired from the simulation of social behavior of flocking organisms, such as swarms of birds and fish schools [KE95, KES01]. Since its invention, PSO has gained increasing popularity because it is simple in concept and easy to implement. Previous study has shown the PSO to be an effective alternative to more established evolutionary algorithms for certain kinds of problems [RR04, BW04, KS98]. Furthermore, the particle swarm paradigm retains the conceptual simplicity of genetic algorithm (GA) while being

subject to only one operator greatly reduces the complexity of applying a generic optimization routine to an arbitrary problem. This technique has been successfully applied to many optimization applications such as tackling multi-objective problems [CPL04], minimax problems [PV02], integer programming problems [PV02], and many engineering applications [RR04, BW04].

PSO can be understood by imagining a swarm of birds that search for foods in an open field. Without any *a priori* knowledge of the field, the individual birds spread out and begin their search in random locations. Each bird can remember the location that it found the most foods, and share this information with its neighbors. As time goes on, the birds profit from the discoveries and previous experience of all other companions during the search for foods by exploring the field in this manner: continuously updating their speed and direction-of-travel depending on how successful at finding foods they have been in comparison to the rest of the swarm. Constantly, they are checking the field they fly over against previously encountered locations of highest food concentration. Eventually, the whole field is probed and the birds end up swarming around the location with the globally highest concentration of foods.

In a mathematical formulation, each companion, called particle or agent, is treated as a point in a D -dimensional solution space which adjusts its own flying according to its flying experience as well as the experience of other particles. Let us represent the i th particle by the D -dimensional vector $\mathbf{x}_i = [x_{i1}, x_{i2}, \dots, x_{iD}]$ and denote the globally best particle in the swarm by the position vector $\mathbf{gbest} = [g_1, g_2, \dots, g_D]$. The best previous position of the i th particle is recorded and represented as $\mathbf{pbest}_i =$

$[p_{i1}, p_{i2}, \dots, p_{iD}]$, and the velocity of the i th particle is $\mathbf{v}_i = [v_{i1}, v_{i2}, \dots, v_{iD}]$. The particles evolve according to the equations

$$v_{id} = wv_{id} + c_1r_1(p_{id} - x_{id}) + c_2r_2(g_d - x_{id}) \quad (8.13)$$

$$x_{id} = x_{id} + v_{id} \quad (8.14)$$

where $d = 1, 2, \dots, D$; $i = 1, 2, \dots, I$, and I is the population size. w is the inertial weight. c_1 and c_2 are two positive constants. r_1 and r_2 are two random values in the range $[0, 1]$. The first equation is used to calculate the i th particle's new velocity by taking into account three terms: the particle's previous velocity, the distance between its best previous and current positions, and the distance between the position of the best particle in the swarm and the i th particle's current position. In general, the performance of each particle is measured according to a problem-dependent fitness function. The inertia weight w determines to what extent the particle remains along its original course unaffected by the pull of **pbest** or **gbest**. It regulates the tradeoff between the exploration and exploitation abilities of the swarm. A large w tends to encourage global exploration as a result of the particle being less moved by the pull of **pbest** and **gbest**, whereas a small one tends to facilitate local exploitation in that particles are rapidly pulled toward **pbest** and **gbest**. A general rule of thumb suggests that it is better to initially set w to a large value, in order to make a better global exploration of the search space, and gradually decrease it to get finer solutions. c_1 and c_2 are scaling factors that determine the relative pull of **pbest** and **gbest**. These are also referred to as the cognitive and social rates, respectively. c_1 determines how much the particle is influenced by the memory of its best location, and c_2 is a factor determining how much the particle is influenced by the rest of the swarm. As default values, $c_1 = c_2 = 2$ were proposed in [KES01], although experimental results

Fig. 8.2. A pictorial illustration of how individual particles are pulled toward the location of the best solution **gbest**, and the location of their own personal best solutions, **pbest**.

indicated that alternative combinations sometimes led to superior performance [SUG99]. A pictorial representation of the above process is shown in Fig. 8.2.

This algorithm limits the resulting velocity to a maximum allowed velocity $v_{\max}$. Kennedy *et al.* [KES01] noted that the PSO performed better if $v_{\max}$ in each dimension was set equal to the dynamic range of that dimension. To further ensure that the particles are always confined within the solution space, the rejecting strategy is implemented here: the particles are allowed to fly without any physical constriction. However, particles that travel outside the allowed solution space are not evaluated for fitness (i.e., assigned with a sufficiently large fitness value for our minimization problem) so that the swarm is repelled away from those infeasible regions. This technique helps to save computation time by evaluating what is in the feasible region, while not interfering with the natural motion of the swarm.

To define a reasonable solution space (i.e., specification of maximum and minimum values for each dimension) for the AoA estimation problem, we propose the following algorithm to determine the range of polynomial regression coefficients:

Step 0: Define a reasonable range for angle spread, $\sigma_{\theta 1}$, such that $\sigma_{\theta 1} \in \Omega_{\theta} \triangleq [\sigma_{\theta \min}, \sigma_{\theta \max}]$ over the spectral support, ζf , for the concerned environment type.

Step 1: Discretize the spectral support ζf into a set of grid points with equal separation, denoted as the vector $\zeta \mathbf{f} = [f_1, f_2, \dots, f_N]$.

Step 2: Randomly generate the first angle spread $\sigma_{\theta 1}$ corresponding to f_1 . Note that $\sigma_{\theta 1}$ is uniformly distributed in Ω_{θ} .

Step 3: Randomly generate the second angle spread $\sigma_{\theta 2}$ corresponding to f_2 . Note that $\sigma_{\theta 2}$ is uniformly distributed in $[\sigma_{\theta \min}, \sigma_{\theta 1}]$, which is to ensure that the angle spread is always non-increasing as the frequency increases.

Step 4: Repeat step 3 for $\sigma_{\theta 3}, \sigma_{\theta 4}, \dots, \sigma_{\theta N}$ and denote the vector $\sigma_{\theta} = [\sigma_{\theta 1}, \sigma_{\theta 2}, \dots, \sigma_{\theta N}]$.

Step 5: Perform the $\mathcal{O}^{\text{th}}$ -order polynomial regression for the two sets of data, $\zeta \mathbf{f}$ and σ_{θ} .

Step 6: Repeat step 2 to 5 for several runs.

Step 7: Based on the simulation results, choose the upper and lower bounds of the search space for each dimension. Sometimes the obtained range may be further divided into several subdivisions to assist the searching process.

In addition, the range for the central angle $\underline{\theta}$ and frequency dependence α can be easily determined.

To alleviate the problem that PSO may end up finding local minima and fail to locate the global minimum in the landscape, we can apply the “deflection”, “stretching”, and “repulsion” techniques [PV04] to transform the original objective

function $o(\Lambda)$ immediately after a local minimum Λ_i has been detected. For this approach, the most fundamental question is how to identify a local minimum. Parsopoulos *et al.* [PV04] used a threshold value κ such that when $o(\Lambda_i) \leq \kappa$ for the i th particle, the particle is isolated from the swarm and marked as a potential global solution. However, κ depends strongly on the *a priori* knowledge of the global minimum as well as the objective function's landscape and dimensionality, which cannot be availed for our AoA estimation problem. Therefore, we could apply another approach by monitoring changes in the fitness of a particle. We track the standard deviation in particle i 's fitness, ς_i , over several iterations. When ς_i is smaller than a threshold value δ , a minimum is identified. δ is a more intuitive and tunable parameter than κ and is less problem-dependent. Another difficulty is that most stochastic optimization algorithms including PSO and GA suffer from the "curse of dimensionality", which simply implies that the performance deteriorates as the dimensionality of the search space increases. This phenomenon may arise in UWB source localization problems when there is a large number of scattering centers, each of them exhibiting a high-order AS variation pattern. A possible solution to overcome this problem is to employ the corporative PSO (CPSO) through using multiple swarms to optimize different components of the solution vector cooperatively [VE04]. Each component would correspond to a single scattering center. The entire algorithm described in this section is shown pictorially in Fig. 8.3.

8.4 Experimental Results

In the following examples, we consider a uniform linear array of 10 sensors with a sensor-spacing of half-wavelength corresponding to the center frequency of the

Fig. 8.3. Flowchart illustrating (a) a plain PSO and (b) a modified PSO

band. The spectral support of the UWB source is taken to be the interval [3GHz, 7GHz], representing a fractional bandwidth of 80%, and the pulse duration $T = 0.25$ ns. In all the experiments, the signal-to-noise-ratio (SNR) is set to 30 dB. Also note that the *sufficiently large* fitness value applied in the rejecting strategy is 10^5 .

8.4.1 Experiment 1: Uniform PAS

In this experiment, we only consider a single-cluster scenario for the indoor UWB channel and assume that the AoAs from point source components are uniformly distributed over the angular range with a central angle $\underline{\theta} = \pi/6$. We further assume that the underlying model structure of the concerned environment type takes the form

of a 3rd-order polynomial, and the frequency dependence of the propagation mode can be described by $\alpha = -0.5$. Subsequently, the parameter vector to be estimated is assumed to be $[\pi/6, 0.06, -1.2, 2.5, 50, -0.5]$ in a 6-dimensional hypercube and the fitness function is defined by (8.12). A population size of 500 is selected, c_1 and c_2 are both set to 2.0, and the inertial weight w is varied linearly from 0.9 to 0.4 over 500 iterations. The range in which the optimal solution is searched for is obtained by applying the algorithm presented in Section 8.3 and depicted in Fig. 8.4.

It is instructive to examine the calculated fitness value over 500 iterations to graphically show how quickly the PSO algorithm can converge to a solution of the problem. From Fig. 8.5(a), it is clear that the particles start to swarm around the feasible locations at around the 50th iteration. Fig. 8.5(b) shows the central angle estimate error versus the iteration number. As it can be seen from the figure, the estimation performance is satisfactory (about 0.01°) and indeed the result is even better than the original generalized Capon estimator for narrowband array signals with much narrower angle spreads (4° and 5°) presented in [HSG04]. Fig. 8.5(c) shows that the frequency dependence converges to the actual value ($\alpha = -0.5$) at around the 420th iteration and the final result is -0.4953 . Finally, the two curves (the estimated AS variation curve and the actual one) are shown in Fig. 8.5(d) where the estimated curve is obtained at the 500th iteration, demonstrating that PSO can successfully find the polynomial regression coefficients. Nevertheless, several other runs show that the plain PSO sometimes can detect sub-optimal solutions. These results signify a need to implement the modify PSO to provide a way of escape from the local minima when the convergence stalls.

8.4.2 Experiment 2: Gaussian PAS

In this experiment the channel model structure and the PSO parameter setting are identical to the previous example except that the signal is now normally distributed. From the plots of Fig. 8.6, the proposed estimator yields a slightly deteriorated result in such a scenario in terms of the central angle estimate error and frequency

Fig. 8.4. Coefficient range estimate based on the algorithm presented in Section 8.3 ($\sigma_{\theta_{\min}} = 20^\circ$ and $\sigma_{\theta_{\max}} = 70^\circ$ over an extended spectral support)

dependence estimate. Several other runs further testify the above observation. This is plausibly attributable to the fact that larger truncation errors are caused for Gaussian distributed signals when the fitness function (8.12) is evaluated by using numerical methods.

8.4.3 Experiment 3: PAS Shape Mismatch

In this example, we investigate the effect of making a wrong assumption on the distribution of the spread signals. The scenario is identical to Experiment 2, in which a Gaussian distributed signal arrives at a mean AoA of $\underline{\theta} = \pi/6$ to the normal of a uniform linear array. However, we assume that the signal is uniformly distributed, and therefore apply (8.8.1) to (8.7) to calculate the signal covariance matrix.

Fig. 8.7(a) depicts an irreducible fitness floor because of the model mismatch. We can also observe from Fig. 8.7(b)-(d) that the performance of the algorithm has gone down significantly due to the wide signal spread. Another observation is that the PSO's convergence stalls at an earlier stage as compared to the previous two examples, which shows that the original small optimal region (volume of the search space surrounding the global optimum) is replaced by a much larger pseudo-optimal region surrounding many global optima leading to erroneous results.

8.4.4 Experiment 4: AS Variation Order Mismatch

We continue to study the effect of making a wrong assumption on the order of the polynomial model for AS variation over the entire spectral support. This time the signal is uniformly distributed. The scenario is identical to that in Experiment 1, in which the actual AS variation curve has the form of a 3rd-order polynomial. However,

we assume that the AS variation can be best fit by polynomial regressions of order 3, 2, 1, and 0 in the PSO algorithms. As a result of reduction of the dimensions, the population sizes are selected to be 500, 80, 50, and 20, respectively.

Fig. 8.8(a) shows that the probability of generating a sample inside the feasible and optimal region decreases as the dimensionality of the search space increases. It is clear that it becomes much harder to find the global optimum of a high-dimensional problem. In general, by assuming a 3rd-order polynomial model, a large number of iterations are wasted to search for the small isolated feasible island, regardless of the relatively large population size used. Nevertheless, due to the model mismatch as we decrease the polynomial order, the attainable minimum fitness value increases, which is made apparent in Fig. 8.8(b). Therefore, it is important to choose an appropriate model degrees-of-freedom that maintains a good balance between the accuracy of the estimation results and the computational load. Fig. 8.9 depicts that for the experiment presented here, the performance of the proposed estimator deteriorates as the model order decreases. In general, a large estimation error is resulted when the order is reduced to 0, corresponding to the conventional narrowband wireless channel. Similar observations of significant deterioration in performance persist in several other runs.

8.4.5 Experiment 5: Multi-cluster Environments

In this experiment we apply the similar procedure to estimate the angular information and frequency-dependence of two independent UWB sources, both uniformly distributed. It is assumed that the two scattering clusters are separated with $\underline{\theta}_1 = -\pi/6$ and $\underline{\theta}_2 = \pi/6$. Furthermore, the AS variation models take the form of the 1st-

order polynomial. In addition, the frequency dependence of the propagation modes can be described by $\alpha_1 = -0.2$ and $\alpha_2 = -0.5$. As such, it is assumed that the parameter vector to be estimated is $[-\pi/6, -5.7, 66, -0.2, \pi/6, -5.15, 82.3, -0.5]$ in an 8-dimensional hypercube. In the PSO algorithm, a population size of 80 is selected, c_1 and c_2 are both set to 2.0, and the inertial weight w is varied linearly from 0.9 to 0.4 over 500 iterations.

From Fig. 8.10(a) (solid line), it is clear that the particles swarm around the optimal region at around the 300th iteration. Fig. 8.10(b) shows the central angle estimate error versus the iteration number for the two clusters. We observe that the estimation accuracies are still acceptable when compared to the results in [HSG04] for narrowband array signals with much narrower angle spread. Fig. 8.10(c) shows that the frequency dependencies converge to the actual values at around the 400th iteration. Next, the estimated AS variation curves and the actual ones over the entire band are shown in Fig. 8.10(d) where the estimated curves are obtained at the 500th iteration. Note that for this example the estimation is more accurate for cluster 2 when compared to the corresponding results for cluster 1. It is worth mentioning here that even when the actual AS variation curve can be better fit by higher-order polynomial equations, there would be moderate deterioration in the accuracy by assuming a 1st-order model.

Finally, we probe into the effect of making a wrong assumption on the number of scattering clusters. Fig. 8.10(a) (dash-dot line) illustrates that if the cluster number is mistakenly set to 1, there will be an irreducible fitness floor, which is much larger than the expected minimum fitness value. This information is very useful when the

number of clusters is unknown or inaccurate. By observing the minimum attainable fitness, it is possible to identify the exact number of clusters.

8.5 Conclusions

This chapter introduced a new method for estimating the angular information of UWB sources and frequency-dependence of the underlying propagation modes. The proposed method is based on a realistic description of UWB channel models: presence of local scattering, wideband array signals, and frequency-dependent dispersive effects. All of these facts have greatly complicated the estimation process and have not been jointly considered in the existing literature to the best of our knowledge. Armed with the powerful and easy-to-implement PSO algorithms, the generalized Capon estimator has been extended to the UWB scenarios and applied for both the source localization and the structure frequency-dependence identification. Simulation results showed a satisfactory performance within a reasonable computation time for different environment types. The future work may include a more comprehensive investigation on the applicability of the modified PSO (e.g., objective-function-transformed PSO and CPSO) on NP-hard UWB source localization problems (e.g., curse of dimensionality occurring in more complicated multi-cluster environments).

Fig. 8.5. Convergence curves of (a) calculated fitness of the PSO algorithm, (b) central angle estimate error, (c) frequency dependence estimate, and (d) the actual AS variation curve over the entire spectral support versus the estimated curve at the 500th iteration for Experiment 1 (uniform PAS)

Fig. 8.6. Convergence curves of (a) calculated fitness of the PSO algorithm, (b) central angle estimate error, (c) frequency dependence estimate, and (d) the actual AS variation curve over the entire spectral support versus the estimated curve at the 500th iteration for Experiment 2 (Gaussian PAS)

Fig. 8.7. Convergence curves of (a) calculated fitness of the PSO algorithm, (b) central angle estimate error, (c) frequency dependence estimate, and (d) the AS

variation curve over the entire spectral support versus the estimated curve at the 500th iteration for Experiment 3 (PAS shape mismatch)

Fig. 8.8. Convergence curves of calculated fitness of the PSO algorithm for Experiment 4 (AS variation order mismatch): (a) the original plot (b) the zoomed-in plot

Fig. 8.9. (a) Central angle and frequency dependence estimate error ($\delta\alpha$ is the difference between the actual and estimated α ; R_α is the searching range) and (b) the actual AS variation curve over the entire spectral support versus the estimated curve at the 500th iteration for Experiment 4 (AS variation order mismatch)

Fig. 8.10. Convergence curves of (a) calculated fitness of the PSO algorithm, (b) central angle estimate error, (c) frequency dependence estimate error, and (d) the actual AS variation curves over the entire spectral support versus the estimated curves at the 500th iteration for Experiment 5 (Multi-cluster environments)

CHAPTER 9 CONCLUSIONS AND OUTLOOK

9.1 Conclusions

This dissertation has presented a unified framework for system-dependent space-time channel modeling. This work was composed of two parts. The first part proposed various geometric channel models (GCMs) for emerging wireless systems from a system-dependent viewpoint.

Chapter 2 presented an overview of the state-of-the-art channel modeling approaches. Starting with a comprehensive discussion of the system and correlation functions, some of the established and emerging channel models have been reviewed with a special emphasis on GCMs. We have also presented our outlook of system-independent and system-dependent channel modeling methodologies.

Chapter 3 introduced a semi-geometrically based statistical (SGBS) model for wireless channels. Three parameters, including the effective path length (EPL), geometry-dependent parameter (GDP), and normalized space-dependent intensity measure (NSDIM), were introduced to establish the relationship between the canonical model and the physical wave-propagation process. The concept of system-dependent channel characterization has been implied in a set of heuristic rules proposed. These rules were then applied to reexamine several widely used GCMs. It was shown that both the circular and elliptical scattering models could be inspected using the proposed analytical tool. The investigation also unveiled that the model assumptions in some of the GCMs might be implausible.

Chapter 4 proposed a parabolic scattering model (PSM) to complement the widely used circular scattering model (CSM) for street-dominated environments. Based on the SGBS model, we have derived expressions for certain important channel properties. The recently introduced angular constriction (AC) was used to provide a quantitative measure on the constriction of received signals in the azimuth domain. The model results were then compared with empirical data obtained in the relevant environments. It was concluded that the PSM for longer-delayed waves would serve to be a more useful solution in the simulation studies of street-guided environments as compared to the typical CSM.

Chapter 5 probed into the influence of antenna directivity on the spatial pattern of scattering points. It has been made clear that the model structure for systems implementing directional antennas is not merely a subset of the model structure for systems with omnidirectional antennas. The versatile von Mises function including a variety of nonisotropic scenarios has been proposed to describe the beam shapes and the exponential power distribution has been used to model the distribution of GDP. The utility of the proposed model was demonstrated by applying it to study the channel characteristics in several numerical examples. It was shown that employing the directional antenna at the mobile terminal is more effective than employing it at the base station for mitigation of multipath interference. However, the former introduced a much larger forbidden zone of correlation, which was undesirable for multiple-input-multiple-output (MIMO) systems.

Chapter 6 introduced an azimuth-frequency-domain geometric microcell (AFDGM) model for ultrawideband (UWB) transmission. Distribution of the scatterers was defined in a hypothetical azimuth-frequency space. Subsequently,

three heuristic rules have been introduced to exploit the underlying assumptions made by any geometric model concerning the actual wave propagation process in an UWB channel. It has been shown that channel should not be defined as a separate functional block from the transmit and receive blocks. We have also derived the important channel parameters including the power azimuthal spectrum (PAS) and power delay spectrum (PDS). Theoretical derivations have testified that the pulse duration does not have a major impact on these two parameters for practical sounding pulses. It was also shown that the channel properties derived using the proposed model exhibited a good agreement with the empirical results reported in other works.

The second part of the thesis discussed the influence of system-dependency of multipath channels on wireless systems.

Chapter 7 calculated the diversity (DIV) and degrees-of-freedom (DOF) associated with a system-dependent wideband directional channel. The paradigms of DIV and DOF densities were proposed, which gave rise to more accurate estimation of the two measures. It has been shown that the possible DIV and DOF gains are greatly dependent on the power spectrum density of the transmitted waveforms, which is because the wireless channel could be perturbed to a certain extent when signals are filtered either in the azimuth or frequency domain, thereby allowing for a more flexible signaling strategy optimization.

Chapter 8 introduced a new method for UWB sources localization in a frequency-dependent propagation channel. The proposed method has incorporated three distinct features of UWB channels: presence of local scattering, wideband array signals, and frequency-dependent dispersive effects. Armed with the robust particle-swarm-optimization (PSO) algorithms, all the model parameters can be estimated within a

reasonable computational time. Simulation results showed a satisfactory performance for different environment types.

There is still a large body of research work to be done in the abovementioned fields. Let us close with 4 possible extensions of the work in this thesis that we are currently considering.

9.2 Future Directions

9.2.1 Dynamic Space-Time Channel Models

The geometry-based models proposed in this dissertation deal with stationary wireless channels. As the future systems will be operating in a shrinking coverage area with dramatic human traffic close to the transmitter and receiver, there is an obvious need to consider the nonstationarity of propagation environments. However, few references can be found in the literature that seek to find dynamic space-time models, which are capable of reflecting the directional nature of the wireless link in terms of angular spectrum at two ends of the connection, as well as the nonstationarity of the wireless applications. Some of the existing models that describe the dynamic nature of indoor propagation channels include the ray-tracing model [ZFW02], the temporal-domain model [NAA01], and the measurement-based multi-step, 4-state Markov model [CLM03].

Currently, we are developing a spatial-temporal dynamic channel (STDC) simulator for nonstationary channels [CD05c]. It has incorporated both the directional properties of wireless channels as well as the evolutionary behavior of scattering clusters in an indoor office environment. In the former case, the model identifies several classes of clusters and establishes a number of typical environment

types. We also give double-directional formulation for the channel impulse responses for which the typical values of model parameters are suggested based on the existing channel measurement campaigns [CHÖ++04, YLC++04, SJJ++00, POO04, CRA00, CTL++03]. In the latter case, the concepts of dynamic local area (DLA), transit area (TA), and dynamic global area (DGA) have been used to elucidate the different dynamic activities of clusters within different service boundary regions (see also Fig. 9.1). A semi-Markov model is introduced to take into account the births and deaths of clusters as the mobile enters into a TA. The model also provides an improvement over the conventional Markov model which considers the state duration to follow an exponential distribution and is not necessarily realistic. The power and spatial-temporal variation of each cluster within its lifespan are also investigated based on the existing dynamic channel measurements [ZFW02, NAA01, CLM03]. It will be of much value to explore the possibility of enhancing the previous geometry-based models with the ability of capturing the nonstationary property of wireless channels.

Fig. 9.1. Definition of “dynamic local area”, “transit area” and “dynamic global area” in the STDC model

9.2.2 Other Physically Based Channel Models

We are also in the process of developing a new physically based model, which contrives a computationally efficient method for a sufficiently accurate prediction of indoor radio propagation [CD05d]. Firstly, the channel is simulated through a set of key parameters which are able to replicate a wide range of situations, from sufficiently specific to sufficiently random. The former is useful when an accurate assessment of a particular environment is required whereas the latter is desired when the system designer is more interested at the average behavior of a wide class of statistical ensemble. Secondly, the channel can be treated as a *continuous multimodal fitness function* of the scatterer locations. Each local optimum and its surrounding area are the result of smaller, closely associated objects that are part of a very similar group of paths. Subsequently, an analogy between the ray tracing process and the identification of multiple optima using the niching particle swarm optimization (NichePSO) [BRI02] is developed.

The results of channel output prediction using the proposed model are compared to a double-directional UWB measurement in a Japanese wooden house [HTK04]. The frequency range was from 3.1 to 10.6 GHz. In the measurement, channel transfer functions were extracted by a vector network analyzer along with the synthetic arrays at both ends of the radio link. With the channel sounder, up to 100 rays were extracted and the paths were identified with the real environment, in which the various scattering clusters were detected based on the double-directional concept. The original floor plan is redrawn in Fig. 9.2(a). We focus on the identification of scattering clusters contributing to the 1st-order reflection. Fig. 9.2(b) shows the spatial distribution of scatterers at the last iteration using the proposed algorithm (i.e.,

Fig. 9.2. (a) Floor plan of the home environment [HTK04] and (b) identified clusters contributing to the 1st-order reflection using the proposed model

the “+” in the figure). Four clusters are clearly identified. Cluster 1 (referred to as Cluster B1 in [HTK04]) has a wide range of azimuthal angle of arrival and corresponds to the displays near the wall. Cluster 2 (referred to as Cluster A1 in [HTK04]) is the reflection from the window behind Rx. Cluster 3 (referred to as Cluster E1 in [HTK04]) consists of scattering from the glass in the wall. Cluster 4

(referred to as Cluster D1 in [HTK04]) corresponds to the paths due to reflection from the wooden door behind Tx. In summary, there is a one-to-one correspondence between the simulated clusters and the detected 1st-order scattering clusters reported in [HTK04]. In future, we will study the influence of various model parameters on the channel outputs.

9.2.3 Indoor Geolocation Using AoA-Based Techniques

Previous indoor directional measurements showed that physical paths were clustered around the receive directions and each cluster corresponded to a major path to the receiver. Using the Bartlett and SAGE estimates overlaid, Czink *et al.* found a mean number of 7.6 clusters in a cluttered office environment [CHÖ++04]. Using a spatial filter on SAGE estimates, 4 to 5 clusters were obtained in [YLC++04]. The authors of [SJJ++00] observed an average number of 3 to 4 clusters with limited angular resolution. For UWB channel measurements [POO04, CRA00], the number of multipath clusters ranges from 1 to 5 as summarized in [POO04]. Hence, at the receiver, several dominant AoAs can be identified as depicted in Fig. 9.3 (multiple AoAs at the two anchors), each corresponding to the spectral peak of a dominant cluster. Note that in [CHÖ++04], a cluster was defined to be “dominant” if its power was not more than 10 dB below the power of the strongest cluster. We will assume that one of the identifiable dominant AoAs is the result of the direct path, which is not necessary to be the strongest ray.

In the proposed algorithm [CD05e], two passive anchors A and B in known positions and with known orientation are measuring the directions to the mobile device (Device C). As indicated in Fig. 9.3, the position of Device C can be

determined as the intersection of two lines of position. Due to the clustering effect, multiple pseudo devices are located all together. To identify the actual device, the algorithm relies on tracking of the movement of the two paths AC and BC by allowing the mobile to move along a predefined route (e.g., $C_0 \rightarrow C_1 \rightarrow C_2 \rightarrow C_0$ as illustrated in Fig. 9.3) in a local area. We assume that a simple dead-reckoning system is implemented at Device C, which is able to provide a rough estimate of the distance traveled by the mobile (i.e., d_0 , d_1 , and d_2 in Fig. 9.3). Subsequently, the method of least square can be applied to estimate the location of the mobile. Due to the vector-ranging nature of the AoA-based technique, there will be attractive possibilities to achieve accurate location estimation when it is combined with other feasible ranging techniques.

Fig. 9.3. The Anchor A, Anchor B, and Device C geometry and location updates using the concept of dead reckoning

9.2.4 Spatial-temporal UWB Measurements in Indoor Open Space Environments

There is little work reported in the literature on spatial-temporal UWB measurements in indoor open space environments, which are supposed to be one of the major service areas for future UWB-based wireless applications. A measurement campaign has been planned for UWB channel characterization in Research TechnoPlaza, Nanyang Technological University [YOG04]. Two typical open-space environments will be considered during the campaign using the peer-to-peer network configuration: the foyer of the TechnoPlaza and the carpark (see also Fig. 9.4). The frequency will be swept from 3 to 10 GHz using a vector network analyzer. Two virtual arrays (linear positioners with omnidirectional UWB antennas) will be implemented at both ends of the radio link to provide a full characterization of the space-time radio channel. The empirical results should prove useful to UWB system designers and future measurement campaigns.

Fig. 9.4. Spatial-temporal channel measurements will be made in indoor open space environments including the foyer (upper figure) and parking lot (lower figure).

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2. Y. Chen and V. K. Dubey, "A dynamic simulation model of indoor wideband directional channels," accepted for publication in *IEEE Trans. Veh. Technol.*, Aug. 2005.
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APPENDIX INFLUENCE OF ARRAY SYSTEM
SENSITIVITY ON MEASURED AOA STATISTICS: A
STUDY CASE

Consider the situation depicted in Fig. A.1(a), where a large number of multipaths emanating from a finite region are impinging on the N -element uniform linear array (ULA) $\mathcal{A}$, which has an array factor of

$$F_{\mathcal{A}}(\theta) = \frac{\sin(Nkd \sin \theta/2)}{N \sin(kd \sin \theta/2)} \tag{A.1}$$

where k is the free space wavenumber, d is the interelement distance, and θ is the AoA measured from broadside. In this case study, a general family of probability density functions (pdfs) that encompass the Gaussian, Laplacian, and uniform distributions are proposed to provide a wide variety of situations for the true

Fig. A.1. (a) Multiple rays impinging on a uniform linear array and (b) illustration of various parameters

distribution of angles:

$$f_{\theta}(\theta) = \frac{u_1(\alpha)}{\sigma} \exp\left(-u_2(\alpha) \left|\frac{\theta - \theta_p}{\sigma}\right|^{\frac{2}{1+\alpha}}\right) \quad (\text{A.2})$$

with

$$u_1(\alpha) = \frac{\{\Gamma[3(1+\alpha)/2]\}^{1/2}}{(1+\alpha)\{\Gamma[(1+\alpha)/2]\}^{3/2}}$$

$$u_2(\alpha) = \left\{ \frac{\Gamma[3(1+\alpha)/2]}{\Gamma[(1+\alpha)/2]} \right\}^{\frac{1}{1+\alpha}}$$

where $\theta_p \in (-\pi, \pi]$ accounts for the mean AoA. Γ is the gamma function, the scale parameter $\alpha > -1$ and the variance of the distribution is σ^2 . For $\alpha = 0$, the generic pdf corresponds to a normal distribution, while for $\alpha = 1$ the pdf is Laplacian. As α tends to -1 , the pdf tends toward a rectangular distribution. These AoA models are typical for mobile radio environments and will be sufficient for the curve fitting of empirical data.

In [SJJ++00], the scanning of the scattering region was done mechanically with a 6° horn at 7 GHz over a 360° range. Whereas in [PMF00], the directional properties of the channel were resolved by post-processing of measurement data, i.e., electronic beam steering. In either case, it is equivalent to pointing the array in a range of directions and noting the value of the average power at each orientation as illustrated in Fig. A.2(b). At any lookup angle Θ , the received total signal will be a weighted sum of all the incident rays. The sum will vary randomly due to the random amplitude and phase of each incident ray. For a narrowband case the transmission from the scattering region to the array is equal to

$$H(\Theta) = C_1 \int A(\theta) F_A(\theta - \Theta) d\theta \quad (\text{A.3})$$

where $A(\theta)$ is the complex scattering strength in direction θ and C_1 is a normalization constant. Assuming that the magnitude of the received complex envelope $|H(\Theta)|$ has a Nakagami fading, $|H(\Theta)|^2$ will be a Gamma variate. The expectation of $|H(\Theta)|^2$ is

$$\begin{aligned} E\{|H(\Theta)|^2\} &= C_1^2 E\left\{\int A(\theta) F_A(\theta - \Theta) d\theta \int A^*(\theta) F_A(\theta - \Theta) d\theta\right\} \\ &= C_1^2 E\left\{\iint A(\theta_1) A^*(\theta_2) F_A(\theta_1 - \Theta) F_A(\theta_2 - \Theta) d\theta_1 d\theta_2\right\} \\ &= C_1^2 \int E\{|A(\theta)|^2\} [F_A(\theta - \Theta)]^2 d\theta \\ &= C_1^2 \int f_r(\theta) [F_A(\theta - \Theta)]^2 d\theta \end{aligned} \quad (\text{A.4})$$

by assuming uncorrelated scattering for rays from various directions. We can interpret (A.4) as representing the average power azimuthal profile (APAP). Since the usual way to eliminate the effect of instrumental noise is truncating the multipath components whose power lies below the noise threshold [ROS99], the measured APAP is thus determined by

$$\begin{aligned} p_{\text{me}}(\Theta) &= \int_{\mathcal{N}} p \cdot f_{p|\Theta}(p|\Theta) dp \\ &= C_2 \int_{\mathcal{N}} p \cdot \left\{ \frac{m}{E\{|H(\Theta)|^2\}} \right\}^m \frac{p^{m-1}}{\Gamma(m)} \exp\left\{ -\frac{mp}{E\{|H(\Theta)|^2\}} \right\} dp \end{aligned} \quad (\text{A.5})$$

where $f_{p|\Theta}$ is the probability of power conditioned on the angle, C_2 is a normalization constant to ensure that $f_{p|\Theta}$ is a pdf, $\mathcal{N}$ is the noise threshold and m is the Nakagami shape factor. Note that as $\mathcal{N}$ approaches to 0, $p_{\text{me}}(\Theta)$ approaches to $E\{|H(\Theta)|^2\}^2$. As a first inspection, when $m = 1$, the Nakagami variate reduces to the Rayleigh variate and $p_{\text{me}}(\Theta) = \mathcal{N} + E\{|H(\Theta)|^2\}$, which shows that the general influence of noise threshold is to elevate the measured APAP.

It is also useful to find the antenna-deconvolved power azimuthal profile (PAP), $p_{ad}(\theta)$, which is the power spectrum after the antenna effects are deconvolved from the measurement data. It can be obtained from the equation given below:

$$p_{me}(\Theta) = C_1^2 \int p_{ad}(\theta) [F(\theta - \Theta)]^2 d\theta \quad \forall \Theta \quad (\text{A.6})$$

We provide a numerical solution through discretization of (A.6). By sampling at a rate of L samples per full AoA range, (A.6) can be rewritten as

$$\frac{1}{C_1^2} p_{me}[l] = \sum_{s=-L/2}^{L/2} p_{ad}[s] F_A^2[l-s], \quad l = -\frac{L}{2} \dots \frac{L}{2} \quad (\text{A.7})$$

where l and s are both integers. After expressing p_{me} in vector form, the convolution in (A.6) becomes a matrix multiplication of the Toeplitz matrix $\mathbf{F}$ with the sample vector $\mathbf{p}_{ad}$

$$\frac{1}{C_1^2} \underbrace{\begin{pmatrix} p_{me}[-L/2] \\ p_{me}[-L/2+1] \\ \vdots \\ p_{me}[L/2] \end{pmatrix}}_{\mathbf{p}_{me}} = \underbrace{\begin{pmatrix} F_A^2[0] & F_A^2[-1] & \cdots & F_A^2[-L] \\ F_A^2[1] & F_A^2[0] & \cdots & F_A^2[-L+1] \\ \vdots & \vdots & \cdots & \vdots \\ F_A^2[L] & F_A^2[L-1] & \cdots & F_A^2[0] \end{pmatrix}}_{\mathbf{F}} \underbrace{\begin{pmatrix} p_{ad}[-L/2] \\ p_{ad}[-L/2+1] \\ \vdots \\ p_{ad}[L/2] \end{pmatrix}}_{\mathbf{p}_{ad}} \quad (\text{A.8})$$

Assuming that the inverse of $\mathbf{F}$ exists, $\mathbf{p}_{ad}$ can be readily obtained as $\mathbf{p}_{ad} = C_1^{-2} \mathbf{F}^{-1} \mathbf{p}_{me}$.

Numerical examples are now provided in order to study the influence of various measurement conditions on the AoA statistics. In all the examples, $d = \lambda/4$ and $m = 1$. We first study the sensitivity of the measured APAP to the noise threshold, which is quantified through the system dynamic range ζ , defined as the number of dB below the maximum APAP. The number of antenna elements $N = 31$ corresponding to an array length of 7.5λ and half power beamwidth (HPBW) of 6.5° , the shape parameter of the true distribution of angles $\alpha = 0$ representing a Gaussian pdf, and the standard deviation $\sigma = 10^\circ$. As depicted in Fig. A.2(a), $p_{me}(\theta)$ becomes more deviated from

$f_{tr}(\theta)$ as ζ decreases from 30 dB to 10 dB. It is also clear that the general influence of the noise level is to raise the tails of the spectrum above the cutoff. Therefore, setting the dynamic range would make the angular spread dependent on the measurement device. Furthermore, the relatively long tails of $p_{me}(\theta)$ as compared to $f_{tr}(\theta)$ are commonly observed in indoor [SJJ++00, CTL++03, PH03], outdoor [PMF00], and outdoor-to-indoor field measurements [KP02], which is due to the smoothing effect of the antenna and the dynamic range of the sounder. Second, the effect of the number of array elements is investigated with the parameter setting of $\zeta = 30$ dB, $\alpha = 0$, and $\sigma = 10^\circ$. Fig. A.2(b) shows the result for $N = 3, 11$, and 31 (corresponding to array length $0.5\lambda, 2.5\lambda, 7.5\lambda$, and HPBW $72.4^\circ, 18.5^\circ, 6.5^\circ$). Note that for a low resolution array with $N = 3$, its HPBW is much wider than the source distribution and thus the measured APAP significantly deviates from the true distribution. In the limit when the array size approaches to 0 (omnidirectional), $p_{me}(\theta)$ would be unchanged over the entire angular range. Nevertheless, if a *single burst* rather than *averaging* is used to find the APAP at each scanning angle, the mechanism of fluctuating AoA is found to be one of the usual phase mixing of unresolved components [AP02].

The plots of Fig. A.2(c) and A.2(d) illustrate the influence of the shape of environment response on the measured APAP. It is shown in Fig. A.2(c) that the inaccuracy in parameter extraction is heavily dependent on the nongaussianity of $f_{tr}(\theta)$ for $N = 31$. It can be observed that the measured APAP exhibits a better fit to the true distribution as α decreases from 1 to -0.99. With a smaller number of antenna elements, the measured APAP significantly deviates from the true distribution for all the different values of α as illustrated in Fig. A.3. Lastly, Fig. A.2(d) illustrates that for $\zeta = 30$ dB, $N = 31$, and $\alpha = 0$, the measured APAP exhibits a closer match to the

true distribution as the angular spread increases from 10° to 30° , which is reasonable because a larger scattering radius enables the array to obtain finer details of the incoming power spectrum.

The plots of antenna-deconvolved PAP are shown in Fig. A.4. It is found that $p_{\text{ad}}(\theta)$ has a better fit to the true distribution if we increase the dynamic range of the system, reduce the number of antenna elements, or scan a scattering cluster with a wider spread. In general, the effect of nongaussianity is negligible as can be seen in Fig. A.4(c) and Fig. A.5, regardless of the array length. Finally, it is worth emphasizing here that the distortion of AoA statistics due to the finite sensitivity of measurement device is less severe for an array of larger HPBW, since the power content is spread over a wider angular range and it would be less affected by truncation of the tails that are below the noise cutoff. Nonetheless, in reality we still prefer to use an array with a narrow angular resolution due to the following reasons: first, the measurement data have not been deconvolved with the radiation pattern of the antenna in most of the existing experiments (see e.g. [SJJ++00, CTL++03, PH03, PMF00] and [KP02]). In such a case, employment of a narrow beam antenna would obtain finer details of the scatterer distribution as depicted in Fig. A.2(b); and second, the difference in the resultant angular profiles with and without deconvolution is marginal for a high-resolution array (comparing Fig. A.2(b) and Fig. A.4(b)). Note that other practical concerns like instrumental impulsive noise or limited scanning interval may also disfavor the use of broad-beam antennas.

In a nutshell, the above examples clearly demonstrate that the interpretation of the channel is jointly determined by the system property (i.e., system dynamic range in this case) and the underlying multipath structure.

Fig. A.2. Measured APAP for different (a) noise cutoff levels, (b) numbers of antenna elements, (c) scatterer clustering conditions, and (d) angular spreads of true distribution. The maximum value of each plot is normalized to 0 dB and the y-axis is in dB. For the sake of clarity, in (c), the plots for $\alpha = 0$ and $\alpha = 1$ have been shifted upward by 10 and 20 dB, respectively.

Fig. A.3. Measured APAP for different values of α (a smaller array with $N = 11$ is used in this example.) The maximum value of each plot is normalized to 0 dB and the y-axis is in dB. For the sake of clarity, the plots for $\alpha = 0$ and $\alpha = 1$ have been shifted upward by 10 and 20 dB, respectively.

Fig. A.4. Antenna-deconvolved PAP for different (a) noise cutoff levels, (b) numbers of antenna elements, (c) scatterer clustering conditions, and (d) angular spreads of true distributions. The maximum value of each plot is normalized to 0 dB and the y-axis is in dB. For the sake of clarity, in (c), the plots for $\alpha = 0$ and $\alpha = 1$ have been shifted upward by 10 and 20 dB, respectively.

Fig. A.5. Antenna-deconvolved PAP for different values of α (a smaller array with $N = 11$ is used in this example.) The maximum value of each plot is normalized to 0 dB and the y-axis is in dB. For the sake of clarity, the plots for $\alpha = 0$ and $\alpha = 1$ have been shifted upward by 10 and 20 dB, respectively.