Propagation channel models for wireless communication systems

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Propagation Channel Models for Wireless Communication Systems

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
Doctor of Philosophy

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Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original re-
search and has not been submitted for a higher degree to any other University or
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Jiang Lei

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Summary

With the development of wireless communication systems, the design, optimization and evaluation of the system become more and more important. Channel modeling provides a deep insight into the characteristics of the mobile propagation channels, which can be of great help to the system design, optimization and evaluation. This thesis mainly focuses on deriving propagation channel models for wireless communication systems.

An uniform theory of diffraction (UTD) based path loss model is proposed for the urban cellular propagation environment. This model takes into account the geometrical shape of the buildings. Rectangular shaped blocks are used to model the rows of buildings, and equal building heights are assumed. The dominant propagation mechanism is multiple diffraction over rooftops. To simplify the derivation, the incident field is assumed to be a vertically polarized plane wave. Closed form expressions are derived to calculate the total received field. When applying average values to the building width and center-to-center spacing of two neighboring buildings, a more explicit path loss model is thus obtained. Good agreement between our theoretical results and measurements shows that our UTD channel model is accurate and outperforms the existing UTD-based over rooftop

model.

To evaluate the spatial and delay properties of the propagation channel, a geometrically based uniform distribution single bounce scatterer model is proposed, which assumes that the scatterers are uniformly distributed around the base station within a circular region. The probability density functions (PDF) for the angle of arrival (AOA) and the time of arrival (TOA) are derived for both omnidirectional and directional transmitting antenna systems. The effect of the beamwidth of the transmitting antenna on Doppler spread is also analyzed.

Furthermore, the uniform distribution single bounce scatterer model is extended to a more general case, where the distribution of the scatterers can be arbitrary. The joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are derived. By employing different distributions, this model can be applied to different propagation environments. Comparisons between our theoretical calculations and the empirical results as well as the measurement data reported in the literature show that the Rayleigh distribution scatterer model is accurate for the outdoor propagation environment, while the exponential distribution scatterer model is applicable for the indoor propagation environment. The uniform distribution scatterer model is found to be a special case of the Rayleigh distribution scatterer model.

The power azimuth spectrums (PAS) for these geometrically based models are also analyzed. The uniform and Rayleigh distribution scatterer models are found to be applicable to outdoor environments, since the PAS curves for them are approximately Laplacian shaped, which is reported to be the best fit distribution for

the PAS of outdoor environment. The power azimuth spectrum of the exponential distribution scatterer model is compared with the measured profile, the good agreement between them again demonstrates that this model is accurate for the indoor propagation environment.

List of Abbreviations and Symbols

Abbreviations

AOA	angle of arrival
BER	bit error rate
BS	base station
CDF	cumulative distribution function
CMR	cellular mobile radio
FDTD	finite-difference time-domain
GBSBCM	geometrically based single bounce circular model
GBSBEM	geometrically based single bounce elliptical model
GO	geometrical optics
GTD	geometrical theory of diffraction
ISB	incident shadow boundary
LOS	line of sight
NLOS	non line of sight
OLOS	obstructed line of sight
PAS	power azimuth spectrum

PDF	probability density function
PDS	power delay spectrum
PL	path loss
PO	physical optics
PSD	power spectrum density
RF	radio frequency
RSB	reflection shadow boundary
TOA	time of arrival
UTD	uniform theory of diffraction

Symbols

$A(\cdot)$	spreading factor
A_{mu}	median attenuation relative to free space
c	speed of the light
D	separation distance between the base station and the mobile receiver
$D_{(\cdot)}$	diffraction coefficients
d	average center-to-center spacing of two neighboring buildings
d_1	average building width
d_m	width of building m
$F_{\theta}(\theta)$	CDF of the AOA
$F_{\tau}(\tau)$	CDF of the TOA
$f_{\theta}(\theta)$	PDF of the AOA

$f_{\tau}(\tau)$	PDF of the TOA
f_d	Doppler spread
f_m	maximum Doppler shift
G_r	gain of the receiving antenna
G_t	gain of the transmitting antenna
$G(h_{re})$	height gain factor of the mobile antenna
$G(h_{te})$	height gain factor of the base station antenna
H	antenna height above the building rooftop
h	average building height
h_{re}	effective mobile antenna height
h_{te}	effective base station antenna height
k	wave number
L_F	free space path loss
L_{ex}	excess loss
L_{mean}	median value of the propagation path loss
P_r	received power
P_t	transmitted power
$P(\theta)$	power azimuth spectrum
R	radius of the scatterer region
$R_{\perp, \parallel}$	parallel and perpendicular reflection coefficients
$S(f)$	power spectrum density
v	velocity of the mobile receiver
W_m	distance between the left edges of building m and building $m + 1$

Z_0	free space impedance
2β	beamwidth of the directional transmitting antenna
β_0	half angle of the diffraction cone
β'_0	incident angle relative to the edge in diffraction
λ	wave length
σ_r	conductivity
θ	angle of arrival at the mobile receiver
τ	time delay
ϵ_r	relative permittivity

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Chapter 1

Introduction

1.1 Background

The ability to communicate with people on the move has evolved remarkably since Guglielmo Marconi first demonstrated radio's ability to provide continuous contact with a ship sailing in the English Channel [1]. That was in 1897, and since then new wireless communications methods and services have been enthusiastically adopted by people throughout the world. Particularly during the past ten years, the mobile radio communications industry has grown by an order of magnitude, fueled by digital and radio frequency (RF) circuit fabrication improvements, new large-scale circuit integration, and other miniaturization technologies which make portable radio equipment smaller, cheaper, and more reliable.

With the development of the communication systems, the design, optimization and evaluation of the system has become more and more important. When signals are transmitted from the base station (BS) to the mobile receiver, they undergo numerous possible paths and suffer losses because of the reflection, diffraction or scattering as shown in Figure 1.1. The received signal at the mobile receiver is

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sensitive to the path loss between the transmitter and the receiver. Therefore, path loss information is vital for the system design and optimization at the base station.

Figure 1.1: The multipath propagation channel

Reflecting objects and scatterers in the channel dissipate the signal energy in amplitude, phase and time. These effects result in multiple versions of the transmitted signal that arrive at the receiving antenna, displaced with respect to one another in time and spatial orientation. The random phase and amplitudes of the different multipath components cause fluctuations in signal strength, thereby inducing the fading, signal distortion, or both [2].

The ill-effects of the multipath fading in wireless systems can be reduced by employing spatial diversity, where a multi-element antenna at the receiver is employed [3]. In an array antenna, the signals received by the various elements can be

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weighted appropriately to result in a combined signal that fluctuates less rapidly than the individual signals. This array will then require less power to achieve a given bit error rate (BER) than the case where only one element is used.

Figure 1.2: Adaptive antenna beam with radiation minimized in the direction of the interferer

In mobile wireless communications, another limitation is the interference caused by co-channel interferers which occupy the same channel as the desired user. In a high-interference environment, the simple strategy of switching to the strongest signal or extracting maximum signal power from the antennas is not appropriate as it can lead to the signal enhancement of an interferer rather than that of the desired user. However, if the direction of the desired user is known or if a suitable reference signal that correlates more with the desired signal than with any of the interferers can be generated at the receiver, the weights of the array antenna could be chosen to suppress interference and thereby improve the system performance. Hence the smart antennas come into use. The frequency reuse is sensitive to co-channel interferences. Smart antennas enable reduction in co-channel interference, which leads to increase in the frequency reuse factor. Thus smart antennas allow

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more users to use the same frequency spectrum at the same time, and bring about tremendous increase in capacity.

The angle of arrival (AOA) information is required for the antennas such as dynamic phased arrays, which make use of the angle of arrival information from the desired user and steer a beam maximum towards the desired user, thus improving upon the capabilities of a switched beam antenna. The adaptive antenna systems as shown in Figure 1.2 make use of the well defined angle of arrival of the desired user and jammer to determine the weights.

Therefore, with the increasing use of smart antenna systems, in addition to the power level, the spatial and delay properties of the propagation environment such as the angle of arrival, the time of arrival (TOA), angular spread, delay spread, Doppler spread and the power azimuth spectrum are also crucial to the design of the antenna system. Propagation channel models providing such information are in great need.

1.2 Motivation

Although considerable efforts have been put into the channel modeling, a lot of problems still need to be solved. Simple and practical models are needed best in system design and evaluation. The path loss information is one of the fundamental demands of the channel propagation prediction. The most widely used path loss models such as Okumura model [4] and Hata model [5], are based on extensive measurements. The measurements are usually made in a limited region, and the environments are often specified, which lead to the models' low flexibility in

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application. On the other side, the extensive field measurements take time and efforts. The experiment will be very tough when taken in a larger area, for example, the cellular propagation environment. Another way to estimate the path loss of the channel is to simulate the propagation environment and trace all the possible propagation paths. This kind of model can be very detailed and accurate, but correspondingly, the required computing time is incredibly long, especially for complicated environments. Although a lot of work has been done to improve the theory or algorithm, these models are still time consuming because of their essential principles.

Theoretical analysis is an efficient and relatively accurate way to model the propagation channel. By assuming some ideal conditions, they only take into account the major propagation mechanisms and propagation paths, which largely simplifies the analysis and saves computing time. Hence, this model takes advantages of the two types described above.

In urban cellular communication systems, diffraction is usually the major propagation mechanism, especially for long transmitter receiver separation distance. The signals with multiple reflections will suffer great loss due to the large number of reflections in such situation, hence the received power for these paths is negligible. The line of sight (LOS) path is almost impossible in an urban street area, and the path penetrating the buildings is usually unable to reach the receiver. Early theoretical models such as Walfisch and Bertoni model [6], are mostly derived base on this assumption. The physical optics (PO) has been used to calculate the diffracted field, it is accurate but still not efficient enough because of the Kirchhoff-

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Huygens integral, especially for the multiple diffracted field. On the other side, the uniform theory of diffraction (UTD) has been proved to be accurate and efficient in application. It is a ray based theory and closed form expressions can be obtained when calculating the path loss. Therefore, the UTD base theoretical channel model may be a better way to predict the cellular communication system. A simple analytical channel model for urban cellular environment is aimed to be derived in this thesis.

With the advent of the smart antenna systems, the requirements for channel models are no longer restricted to the path loss information, the spatial and the delay properties of the channels become more and more important. The spatial and the delay properties are usually described in a statistical way, hence these models are often called statistical models. The statistical models can also be derived from the measured results, or from theoretical analysis. For the reasons discussed before, we prefer the theoretical analysis to get more flexible models. Among them, the geometrically based single bounce scatterer model is one of the most often used ones. In these models, the basic idea is making assumptions of the scatterers' distribution around the mobile or between the base station and the mobile receiver. However, in a communication system, the location of the mobile changes from time to time, thus the scatterers' distribution around the mobile may also change; on the other hand, the location of the base station is usually fixed, therefore, it is easier to determine the scatterers' distribution around the base station no matter where the mobile receiver is. In this thesis, we aim to study the spatial and delay properties of the channel based on this idea.

1.3 Objectives

Specifically, we aim to

- (i) Propose a simple analytical path loss channel model for urban cellular communication environment. Based on ray reflection and diffraction theory, closed form expressions will be derived to calculate the received field. Analyze and compare the theoretical results with measurements and existing models to verify its veracity and improvement over the path loss prediction for longer distance.
- (ii) Propose a geometrically based channel model to investigate the spatial and delay properties of the communication environments. Use the uniform distribution to model the scatterers for the simplicity of analysis. Derive the probability density functions of the angle of arrival and the time of arrival for both omnidirectional and directional transmitting antenna systems. Study the effect of the directional transmitting antenna on Doppler spread.
- (iii) Further extend the uniform distribution scatterer model to a more general situation. Provide the general formulations of the model and the derivation of the joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA. Apply different scatterer distributions to this model and compare these statistical characteristics with empirical results or measurements, find the suitable models for outdoor and indoor propagation environments.
- (iv) Investigate the power azimuth spectrums for the geometrically based channel

models derived above. Derive the analytic expressions by assuming some simple propagation mechanisms, such as reflection. Compare the theoretical outcomes with empirical results or the measured profile reported in literature to further test their validity.

1.4 Major Contributions of the Thesis

The major contributions of the thesis include the following:

- (i) A simple analytic path loss model is derived for urban cellular communication systems. Rectangular shaped blocks with equal height are used to model the buildings, and multiple diffraction over building rooftops are assumed to be the dominant propagation path. A closed form expression is derived to calculate the total received field. By using average values of the building width and center-to-center spacing of two neighboring buildings, a simpler analytic path loss model is thus obtained. Good agreement between our theoretical results with measurements shows that our UTD channel model is accurate and outperforms the existing UTD-based over rooftop model.
- (ii) A geometrically based single bounce uniform distribution scatterer model is proposed to investigate the spatial and delay properties of the wireless communication systems. The scatterers are assumed to be uniformly distributed around the base station within a circular region. The probability density functions of the arrival angles and the time delays at the mobile receiver are derived for both omnidirectional and directional transmitting antenna sys-

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tems. The Doppler spectrum is analyzed as well to study the effect of the directional transmitting antenna on Doppler spread.

- (iii) A more general situation is investigated for the geometrically based single bounce model proposed above. The scatterers are still assumed to be distributed around the base station within a circular area, but the distribution of the scatterers can be arbitrary. General formulations of the model and the derivation of the joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are outlined. By employing different distributions, this model can be applied to different propagation environments. We focus on the Rayleigh distribution scatterer model and the exponential distribution scatterer model. The former is found to be applicable to the outdoor propagation environment, while the latter predicts the indoor propagation environment very well.
- (iv) The power azimuth spectrum is investigated for all the geometrically based single bounce scatterer models derived above. The single bounce in each model is assumed to be a specular reflection. The power azimuth spectrums are derived and compared with the empirical and measurement results.

1.5 Organization of the Thesis

The thesis is organized as follows.

Chapter 2 provides an overview of the existing propagation channel models. A brief introduction to the background theory for the path loss models is given first.

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Then the path loss models are introduced in three types, namely, empirical models, site-specific models and theoretical models. The advantages and drawbacks of these models are analyzed respectively. In the following section, the wireless multipath channel models are presented. We mainly focus on the empirical models and the geometrically based models.

Chapter 3 presents a UTD based simple analytic channel model for urban cellular propagation environment. Rectangular shaped blocks are used to model the rows of buildings instead of traditional half screens of vanishing thickness. A closed form expression for the received field is derived first for various building widths and inter building spacing. Then, average building width and average center-to-center spacing of any two neighboring buildings are assumed. A more simplified expression is thus obtained. After that, the characteristics of the received field are analyzed, and our path loss model is compared with the existing models and the measurements to show our model's advantage and accuracy.

In Chapter 4, a geometrically based single bounce scatterer model is proposed to provide the spatial and delay properties of the wireless communication channels. The concept of the uniform distribution scatterer model is introduced first. Then, closed form expressions for the probability density functions of the angle of arrival and the time of arrival at the mobile receiver are derived for omnidirectional transmitting antenna system. The main impact parameters for these distribution functions are discussed respectively. Finally, the PDFs of the AOA and TOA for directional transmitting antenna system are derived. The effect of the directional transmitting antenna on Doppler spectrum is analyzed as well.

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In Chapter 5, the channel model proposed in Chapter 4 is further extended to a more general case. The scatterer distribution is no longer limited to the uniform distribution. General formulations of the model and the derivation of the joint probability density function (PDF) of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are outlined first. Then, the Rayleigh distribution scatterer model is discussed. Theoretical PDFs are compared with the empirical results reported in the literature for outdoor environment, and the rms angular spread and delay spread are presented as well. After that, the exponential distribution scatterer model is investigated. All the statistical characteristics listed above are evaluated. The PDF of the AOA is compared with the measured data from different indoor environments to validate the applicability of this model.

In Chapter 6, the power azimuth spectrum is studied for the geometrically based single bounce scatterer models proposed in Chapter 4 and Chapter 5. The received power at the mobile receiver is analyzed first, and then the general formulation of the power azimuth spectrum is derived. Subsequently, we discuss the PASs for uniform, Rayleigh and exponential distribution scatterer models. The important impact parameters are analyzed, and theoretical power spectrums are compared with the empirical results and the measured profile reported in the literature.

Chapter 7 summaries the main issues addressed in this thesis and comments on the future work.

Chapter 2

A Brief Survey of Propagation Channel Models for Wireless Communication Systems

2.1 Introduction

Before implementing designs and confirming planning of wireless communication systems, accurate propagation characteristics of the environment should be known. Propagation analysis provides a good initial estimate of the signal characteristics. The ability to predict radio propagation behavior for wireless personal communication systems, such as cellular mobile radio, is becoming crucial to system design. Since the site measurements are costly, propagation models have been developed as a suitable, low-cost, and convenient alternative [6–23].

[2, 24–26] introduced some of the widely used propagation channel models. Classical propagation prediction usually provides the parameters corresponding to the large-scale path loss. The path loss is vital for the determination of a base station placement and in optimizing it. From this information, we can decide the minimum power needed for a transmitter to radiate to serve a given region.

CHAPTER 2. A BRIEF SURVEY OF PROPAGATION CHANNEL MODELS FOR WIRELESS COMMUNICATION SYSTEMS

However, in mobile communication systems, in order to mitigate the multipath fading, reduce the co-channel interference and increase the capacity of the system, smart antennas become more and more widely used. For smart antenna systems, in addition to signal power level, the spatial and delay properties of the channel have an enormous impact on the performance of the system. Therefore, it is necessary to have channel models that provide the angle of arrival (AOA) of multipath components and the multiple delay profile.

This chapter is organized as follows. In section 2.2, the background theory for path loss propagation models is introduced. The path loss propagation models and the wireless multipath channel models are reviewed in section 2.3 and section 2.4, respectively. Conclusion is given in section 2.5.

2.2 Background Theory for Path Loss Propagation Models

2.2.1 Path Loss

Path loss (PL), which represents the signal attenuation as a positive quantity measured in dB, is defined as the difference (in dB) between the effective transmitted power and the received power, and may or may not include the effect of the antenna gains [2]. The path loss for free space model when antenna gains are included is defined by

$$PL = 10 \log \frac{P_t}{P_r} = -10 \log \left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2} \right] \quad (2.1)$$

where P_t is the transmitted power and P_r is the received power. G_t and G_r are the gains of the transmitting antenna and receiving antenna, respectively, d is the distance between transmitter and receiver, and λ is the wavelength in free space. It is clear that (2.1) does not hold for $d = 0$. For this reason, large-scale propagation models use a close-in distance, d_0 , as a known received power reference point. d_0 is chosen to be in the far field of the antenna, typically 1 m for indoor environments and 100 m or 1 km for outdoor environments [27]. So the expression for free space path loss can be written as

$$PL(d) = PL(d_0) + 20 \log \frac{d}{d_0} \quad (2.2)$$

Path loss is the main ingredient of a propagation model. It is related to the area of the coverage of mobile systems.

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2.2.2 The Uniform Theory of Diffraction

Geometrical optics (GO) is incapable of predicting the field in the shadow regions, and is inaccurate in the vicinity of the shadow boundaries. To supplement the geometrical optics, the diffracted rays are added.

In 1974 Kouyoumjian and Pathak [28] developed a ray-based uniform diffraction theory. By multiplying a transition function, the diffraction coefficients are more general than Keller's geometrical theory of diffraction (GTD) coefficients [29] in the sense that not only do they remain bounded along the shadow boundaries, but they also can be used to calculate the diffraction fields from a wedge with curved surfaces.

The two-dimensional UTD edge diffraction coefficients are given by [30]

$$D_{\perp,\parallel}(L^i, L^{ro}, L^{rn}, \phi, \phi', n) = D_1 + D_2 + R_{\perp,\parallel}(D_3 + D_4) \quad (2.3)$$

where L^i, L^{ro} and L^{rn} are distance parameters, ϕ' is the incident angle, ϕ is the diffraction angle, n is a parameter related to the interior angle of the edge, $R_{\perp,\parallel}$ are the reflection coefficients for parallel and perpendicular polarization, respectively. The component D_1 is associate with the diffracted field that compensates for the discontinuity in the GO field when the o -face is shadowed; D_2 , when the n -face is shadowed; D_3 , when there is reflection from the n -face; D_4 , when there is reflection from the o -face. The concepts of n -face, o -face, and the detailed information about $L^i, L^{ro}, L^{rn}, n, R_{\perp,\parallel}, D_1, D_2, D_3$ and D_4 are given in Appendix A.

2.3 Path Loss Propagation Models

The path loss propagation models can be roughly divided into three types, i.e., the empirical, theoretical, and site-specific models. They will be reviewed respectively in the next three sections .

2.3.1 Empirical Models

Empirical models are usually a set of equations derived from extensive field measurements [5, 8, 13, 23, 31–38]. Empirical models are simple, efficient and accurate for environments with the same characteristics as those where the measurements were made. However, they cannot be used for different environments without modifications, and they are even useless when applied to a quite different environment. For example, the indoor picocells cannot use the empirical model for the outdoor macrocells. The input parameters for the empirical models are usually qualitative and not very specific, while the output parameters are basically range specific, not site specific.

Okumura’s model [2] is one of the most widely used empirical models for signal prediction in urban areas. This model is applicable for frequencies in the range 150 MHz to 1920 MHz (it is also extrapolated up to 3000 MHz) and distances of 1 km to 100 km. It can be used for base station antenna heights ranging from 30 m to 1000 m.

Okumura developed a set of curves giving the median attenuation relative to free space (A_{mu}), in an urban area over a quasi-smooth terrain with a base station (BS) effective antenna height (h_{te}) of 200 m and a mobile antenna height (h_{re}) of

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3 m. The model is expressed as

$$L_{mean}(dB) = L_F + A_{mu}(f, d) - G(h_{te}) - G(h_{re}) - G_{AREA} \quad (2.4)$$

where L_{mean} is the median value of the propagation path loss, L_F is the free-space path loss, A_{mu} is the median attenuation value relative to free space in an urban area, $G(h_{te})$ and $G(h_{re})$ are the height gain factors of base station and mobile antennas, and G_{AREA} is the gain due to the type of the environment.

The Okumura model is wholly based on measured data and does not provide any analytical explanation. It is very practical and has become a standard for system planning in modern land mobile radio systems in Japan. The major disadvantage with the model is its slow response to rapid changes in terrain, therefore the model is fairly good in urban and suburban areas, but not as good in rural areas.

The Hata model [5] is an empirical formulation of the graphical path loss data provided by Okumura, and is valid from 150 MHz to 1500 MHz. Hata presented the urban area propagation loss as a standard formula and supplied correction equations for application to other situations. The standard formula for median path loss in urban areas is given by

$$\begin{aligned} L_{mean}(urban)(dB) = & 69.55 + 26.16 \log f_c - 13.82 \log h_{te} - a(h_{re}) \\ & + (44.9 - 6.55 \log h_{te}) \log d \end{aligned} \quad (2.5)$$

where f_c is the frequency (in MHz) from 150 MHz to 1500 MHz, h_{te} is the effective transmitter antenna height (in meters) ranging from 30 m to 200 m, h_{re} is the effective receiver antenna height (in meters) ranging from 1 m to 10 m, d is the

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distance (in km) between transmitter and receiver, and $a(h_{re})$ is the correction factor for effective mobile antenna height which is a function of the size of coverage area.

The standard Hata formula in (2.5) can be further modified to predict the suburban area and the open rural area. The predictions of the Hata model compare very closely with the original Okumura model, as long as d exceeds 1 km. This model is well suited for large cell mobile systems, but not personal communications systems (PCS) which have cells on the order of 1 km radius.

The European Co-operative for Scientific and Technical research formed the COST-231 working committee to develop an extended version of the Hata model. COST-231 proposed the following formula to extend Hata's model to 2 GHz. The proposed model for path loss is

$$L_{mean}(urban)(dB) = 46.3 + 33.9 \log f_c - 13.82 \log h_{te} - a(h_{re}) + (44.9 - 6.55 \log h_{te}) \log d + C_M \quad (2.6)$$

where f_c , h_{te} , h_{re} , d and $a(h_{re})$ are defined in (2.5), and

$$C_M = \begin{array}{ll} 0 \text{ dB} & \text{for medium size city and suburban areas} \\ 3 \text{ dB} & \text{for metropolitan centers} \end{array}$$

The COST-231 extension of the Hata model is restricted to the following range of parameters:

$$\begin{array}{ll} f & : 1500 \text{ MHz to } 2000 \text{ MHz} \\ h_{te} & : 30 \text{ m to } 200 \text{ m} \\ h_{re} & : 1 \text{ m to } 10 \text{ m} \\ d & : 1 \text{ km to } 20 \text{ km} \end{array}$$

2.3.2 Site-Specific Models

Site-specific models are based on numerical methods such as the ray-tracing method [39–61] and the finite-difference time-domain (FDTD) method. The input parameters can be very detailed and accurate. The disadvantages of the site-specific methods are the large computational overhead that may be prohibitive for some complex environments.

Generally there are two approaches to ray-tracing: the image method [39, 43, 58] and the ray launching method [39, 43–45, 52, 56, 57]. Image method is a simple and accurate method for determining the ray trajectory. In the image method, the reflection points are determined geometrically by considering the transmitter's equivalent image. Figure 2.1(a) shows the basic idea of the image method. The image of T_X due to $W1$ is first determined (T_{X1}), then the image of T_{X1} due to $W2$ is calculated (T_{X2}). Connecting R_X and T_{X2} , a reflection point $P2$ is found on $W2$. Another reflection point $P1$ is the intersection point of the line connecting $P2$ and T_X with $W1$. For the image method, the main drawback is that the effects of randomly distributed scatterers cannot be easily taken into considerations and it also suffers from inefficiency when the number of objects is large. In the ray launching method [39, 43–45, 52, 56, 57], rays are launched at a constant angular separation $\Delta\theta$ from the transmitter as shown in Figure 2.1(b). At the receiving point, a reception sphere is constructed with radius equal to $\Delta\theta d/\sqrt{3}$, where d is the total distance of the propagation from transmitter to receiver, including paths due to all multiple reflections. The rays that reach the reception sphere

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Figure 2.1: Ray-tracing methods

are considered to have arrived at the receiver. The major drawback of the ray launching technique is that at each receiver position, intersection tests must be performed on every scatterer for all the test rays, and a reception test is needed after each intersection to determine whether a particular ray will be received. When the number of objects is large, the testing can be very time consuming and inefficient.

Tan and Tan [46] proposed a hybrid method combining the multiple image and ray tracing launching method. The ray launching method is used to quickly identify a possible ray trajectory from T_X to R_X . The exact reflection positions can then be accurately found by the multiple image method. The method has the advantages of the ray launching and image method. After determining all the

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trajectories from T_X to R_X , the field at the receiver can now be calculated and then the path loss.

2.3.3 Theoretical Models

Theoretical models are derived physically assuming some ideal conditions. They are more efficient than the site-specific models and more site-specific than the empirical models. A typical propagation model for urban cellular environments is

Figure 2.2: Propagation over-rooftops

the over-rooftop model. Walfisch and Bertoni [6] presented a physical model of the propagation process that takes place in urban environments outside the high-rise urban core. The rows of buildings are replaced by opaque absorbing half-screens of vanishing thickness. The heights of the buildings are assumed to be equal and the spacing between the buildings is identical. The transmitted fields propagate over the roof tops by a process of multiple diffraction past rows of buildings that act as cylindrical obstacles, as suggested by path 1 in Figure 2.2.

Walfisch and Bertoni [6], using physical optics (PO) theory, calculated the field by direct numerical evaluation of the Kirchhoff-Huygens integral as shown in Figure 2.3. The field $H_{n+1}(y)$ incident on screen $(n + 1)$ due to the field $H_n(y')$ above the n th half-screen is given by (2.7). The symbol y' and y correspond to a

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Figure 2.3: Geometry for Kirchhoff-Huygens evaluation of plane wave diffraction by series of half-screens

position above the current and the next building respectively.

$$H_{n+1}(y) = \frac{e^{j\pi/4}}{2\sqrt{\lambda}} \int_0^{\infty} H_n(y') \frac{e^{-jkr}}{\sqrt{r}} (\cos \delta + \cos \alpha) dy' \quad (2.7)$$

where $r = \sqrt{d^2 + (y - y')^2}$ and $\cos \delta = d/r$. The evaluation of the Kirchhoff-Huygens integral adopted here is time consuming and not efficient.

Later, some improved solutions based on physical optics for diffractions by multiple absorbing half-planes were developed [17–20]. Xia and Bertoni [17] developed a closed-form expression for the field via Boersma’s functions [62] for cylindrical and plane waves when the source is placed at a distance from the edges equal to the separation between them. The solution can be applied for incidence either from above or below the plane containing the edges. Maciel *et al.*, [18] presented theoretical results for the dependence on base station antenna height of the average received signal for mobiles at street level. The results can be applied to residential and commercial sections of cities and to all ranges of antenna height

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from well above to below that of the surrounding buildings. An explicit solution to the multiple edge diffraction over arbitrary numbers of equal height parallel half planes is presented in [19]. In [20], two models of mobile radio wave propagation over buildings are presented. The Flat Edge model provides a simple yet accurate representation when buildings are assumed to be of constant height and spacing. By combining the Flat Edge model with a rapid method of calculating multiple edge diffraction, the deterministic predictions for arbitrary building height and spacing become possible.

The uniform theory of diffraction (UTD) is a ray-based theory and is more efficient than physical optics in calculation. Therefore, the UTD-based theoretical models will be more applicable in quick estimation of the multiple diffraction [21, 63–67]. Neve and Rowe [21] investigated a UTD-based model solving multiple diffractions over rooftops. The major assumptions are same as in previous cases, the buildings are replaced by opaque absorbing half-screens of vanishing thickness, equal building heights and spacing are assumed. The total received field at screen $(n + 1)$ is given by the summation of the multiple diffraction fields from screen $1, \dots, n$ plus the direct field contribution at screen $(n + 1)$. Zhang [65] simplified Neve and Rowe's formula and developed it including wide-band characteristics of the environment. In [66], he further extended this formulation for multiple building forward diffraction to predict the propagation in urban microcellular environments. The total received signal in the urban side and parallel street is focused on. Balis and Hinton [67] proposed a UTD-based model for prediction of propagation path loss and shadowing variability in urban mobile environments. Buildings are rep-

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resented as conducting half planes or screens, and the model includes the effect of building height variation along all intervening screens between base station and mobile. This model has low computational complexity and could be applied to small urban cells over regular terrain.

Figure 2.4: Geometry of plane wave incidence over a series of perfectly conducting wedges with the same separation w

Juan-Llácer and Rodríguez [68] have investigated multiple diffractions based on UTD and PO. Perfectly conducting wedges are used instead of half screens as shown in Figure 2.4. This model allows for numerical evaluation of a large number of wedges. When the angle of incidence α is positive, the field attenuation tends to be independent of the number of wedges.

In over-rooftop models, the path loss in decibels is the sum of free-space loss and the so-called excess loss (L_{ex}). The excess loss is further divided into two parts $L_{ex} = L_{ex1} + L_{ex2}$, i.e., the diffraction of the fields at the rooftop before the mobile down to the street level, and the reduction of the field at this rooftop as a result of propagation over the previous rows of buildings.

2.4 Wireless Multipath Channel Models

With the advent of smart antenna systems for both interference cancellation and position location applications, the spatial and delay properties of the wireless communication channels will have an enormous impact on the performance of smart antenna systems; hence, an understanding of these properties is paramount to effective system design and evaluation [69]. Therefore, the information regarding the signal multipath delay and the angle of arrival becomes an important part of channel modelling [70–92].

Deriving empirical models from measurement data is a typical and common way of investigating channel models. It is simple and efficient. However, it has the drawbacks we have mentioned before, the flexibility is low, the models are limited to use in the environments with the same characteristics as those where the measurements were made. In contrast, the geometrically based theoretical models are more general than the empirical models. They idealize the spatial relationships among the transmitter, the scatterers, and the receiver for a wide class of propagation settings, the results are applicable generally within that class of channels. In this type of models, the propagation environments are usually modeled by some geometric shapes with scatterers of certain distributions. By assuming some simple propagation mechanisms, the spacial and delay properties are estimated in a statistical way. We will introduce these two types of channel models respectively in the following sections.

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2.4.1 Empirical Models

[70, 71, 73–79] are the empirical models derived from measurements for outdoor and indoor environments. Pedersen *et al.*, [70, 71] reported an empirical stochastic model of the delay and azimuthal dispersion for the outdoor propagation environments. The statistical properties are extracted from macrocellular measurements conducted in a variety of urban environments. It is found that in typical urban environments the power azimuth spectrum is accurately described by a Laplacian function, while a Gaussian PDF matched the azimuth PDF. Moreover, the power delay spectrum and the delay PDF are accurately modeled by an exponential decaying function. Couto de Miranda and Cole [73] made measurements on a 4.1 km path across central London. Analysis of the probability density function estimates of angle of arrival shows virtually no departure from a Gaussian distribution.

[76–79] investigated the indoor multipath channels based on the indoor propagation measurements. In all these models, the received signals are observed to arrive in clusters, both in time and azimuthal angle. The angle of arrival within one cluster for indoor environment is found to follow Laplacian distribution [76, 78, 79]. Different standard deviations for the zero-mean Laplacian PDFs were reported in these three models. In [78], data was collected at 7 GHz in delay-angle domain with impulse response measurements carried out using narrow beam dish antenna. The experiment was performed within the office buildings on the Brigham Young University (BYU) campus. The distribution of the arrival angles were observed to fit a zero-mean Laplacian PDF with standard deviation of 25.5° for Clyde Build-

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ings, while 21.5° for Crabtree Building. Cramer *et al.*, [79] took the measurement in an office/laboratory environment, analysis shows that a value of 38° is the best fit. Chong *et al.*, [76] made the experiment in the University of Bristol, three different values of standard deviation were reported for three different kinds of environments, 3.9° for LOS scenario, 9.0° for obstructed-LOS (OLOS) scenario, and 7.3° for non-LOS (NLOS) scenario. In these papers, no theoretical model is presented, only empirical parameters for the distributions are derived from the measured data, and they are obviously not suitable for the environment that is different from where the measurements were made. In [77], the PDF of the AOA was found to be different from what had been observed in [76, 78, 79]. It turns out that Gaussian distribution provides a better fit. In despite of this, most of the existing evidence shows that a zero-mean Laplacian distribution will be the best fit for the PDF of the arrival angles. The PDF of the time of arrival is modeled by an exponential decaying function. For the power azimuth spectrum, it is found that a Laplacian distribution fits the PAS best for indoor environments [76, 77], which is same as what has been observed for outdoor environments.

2.4.2 Geometrically Based Models

[83, 84, 86–92] are analytic channel models, derived by assuming some ideal conditions. Among them, the most popular radio propagation models are the single bounce scattering geometrical models. Liberti and Rappaport [83] proposed a geometrically based single bounce elliptical model (GBSBEM) as shown in Figure 2.5. It assumes that the scatterers are uniformly distributed within an ellipse with foci

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at the base station and mobile receiver. The major axis of the ellipse is given by $c\tau_m$, where τ_m is the maximum delay associated with scatterers within the ellipse, and c is the speed of light. θ is the angle of arrival at the base station. This model was proposed for microcell environments where antenna heights are relatively low, and therefore multipath scattering near the base station is just as likely as multipath scattering near the mobile. This model only accounts for the multi-

Figure 2.5: The elliptical scatterer model

path signals which arrive with an absolute delay $\leq \tau_m$. Ignoring components with larger delays is possible since signals with longer delay will experience greater path loss, and hence have relatively low power compared to those with shorter delays.

Later, Petrus *et al.*, [84] proposed a geometrically based single bounce circular model (GBSBCM) for macrocell mobile environments. It assumes that the scatterers lie within a circle of radius R about the mobile as shown in Figure 2.6. The separation distance from the base station to the mobile receiver is D . Often the requirement $R < D$ is imposed. θ is the angle of arrival at the base station. Since the scatterers are confined to a circle around the mobile, the AOA of the multipath components is restricted to an angular region of $2\theta_{max}$. This model is based on the fact that the antenna heights are typically higher than surrounding scatterers in macrocell environments. Hence it assumes that multipath reflection is caused

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Figure 2.6: The circular scatterer model

by scatterers located close to the mobile receiver, the effects of the scatterers near the base station are ignored. The PDF of AOA is found to be a function of D/R , the smaller the value of D/R , the more concentrated the arrival angles at the base station. The effects of beamforming on the Doppler spectrum for narrowband signals are investigated in [84, 93]. It is observed that the rate and the depth of the envelope fades are significantly reduced when a beamformer with narrow beam width is used.

Ertel and Reed [86] proposed a more general approach that the spatial and delay properties for both elliptical [83] and circular [84] scattering models can be evaluated using a common approach. In addition to the marginal PDFs of the AOA and TOA, the joint AOA/TOA PDFs are derived as well. The distribution of the arrival angles has been analyzed for both at the base station and at the mobile receiver.

Olenko *et al.*, [87] reported a geometrically based single bounce hollow disc scatterer model. The scatterers are assumed to be distributed uniformly on a two-dimensional hollow-disc (i.e., a thick ring) as shown in Figure 2.7, which indicates that there will be more reflections close to but not immediately adjacent to the

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Figure 2.7: The hollow-disc scatterer model

mobile. By varying the hollow-disc's thickness, this model can degenerate to the well-known circular ring model (when $r = R$) proposed by Lee [94], or the circular disc model (when $r = 0$) we discussed above.

For a realistic communication environment, the scatterers usually will not be uniformly distributed. The geometrical-based models for non-uniformly distributed scatterers are investigated in [88, 92]. [88] presented a Gaussian scatter density model, which assumed that the mobile station is surrounded by scatterers of Gaussian distribution. The closer the distance to the mobile, the more the effective scatterers. Expressions for the probability density function of the angle of arrival, the power azimuth spectrum, the time of arrival and the time delay spectrum are provided. Laurila *et al.*, [92] discussed the influences of different scatterer distributions on the power delay profiles and the azimuthal power spectrum. Three distributions have been considered, i.e., uniform, Rayleigh and one-side Gaussian distribution (all defined with respect to the scalar distance between the scatterers and the mobile receiver). The one-sided Gaussian distribution is found to give the best agreement with measurements.

2.5 Conclusion

In this chapter, a brief survey of the propagation channel models is presented. Two categories of models are reviewed in detail. The classical one is the path loss model. It provides the path loss information of the propagation channel, which is important to the communication system design, such as the base station placement determination and optimization. The path loss model can be further divided into three types according to their investigating methods, i.e., the empirical, site-specific and theoretical models. The empirical models are a set of equations derived from the measured data, the site-specific models are based on numerical methods such as ray-tracing, and the theoretical models are derived physically assuming some ideal conditions for the propagation environments. With the advent of smart antenna systems comes the need to better understand the spatial and delay properties of the wireless communications channel. This kind of model can be derived empirically from measurements as well, however, the geometrically based model is more flexible in application. It idealizes the spatial relationships among the transmitter, the scatterers, and the receiver for a wide class of propagation settings, the results are applicable generally within that class of channels.

Chapter 3

Path Loss Model for Urban Cellular Communication Systems

3.1 Introduction

As has been mentioned in Chapter 2, the path loss is an important and fundamental parameter for system planning of wireless communications. A typical scenario in cellular mobile radio (CMR) systems is the over rooftop multiple diffraction caused by rows of buildings in the city outside the high-rise urban core. The over rooftop propagation models are usually derived by assuming the buildings to be rows of half screens with vanishing thickness. In this chapter, a UTD-based theoretical channel model for urban cellular environment is proposed, different from previous models. The geometrical shape of the buildings is taken into account and rectangular shaped blocks are used to model the buildings. Closed form expressions will be derived to calculate the total received field. Our theoretical model will be compared with measurements and the existing model to show its accuracy and better performance.

This chapter is organized as follows. In section 3.2, a brief introduction to an

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existing UTD-based model is given, and the limitation of this model is outlined as well. Section 3.3 introduces our improved UTD-based model. Closed form expressions are derived for two cases: irregular building widths and spacing, uniform building width and spacing. The numerical results and discussion are given in section 3.4. Section 3.5 concludes the work of this chapter.

3.2 UTD-Based Over Rooftop Model

Before we go to our UTD-based model, a brief review of Neve and Rowe's model [21] is given here. Figure 3.1 shows the two dimensional geometry of this model,

Figure 3.1: Two dimensional geometry of the over rooftop model

the buildings are replaced by opaque absorbing screens of vanishing thickness. The field incident on the screen m is diffracted by the screen m first, then this field will propagate horizontally across the remaining screens $(m+1), \dots, (n+1)$. The total field at screen $(n+1)$ is the summation of the contribution of the diffracted field from screen $1, \dots, n$ plus the direct field incident upon screen $(n+1)$. A closed

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form expression has been derived as follows:

$$E_{tot}(n+1) = E_0 \left[e^{-jkn d \cos \alpha} + \frac{D_\alpha D_{\pi/2}^n}{\sqrt{d^{n+1}}} e^{-jkd(1+n+\cos \alpha)} \times \sum_{m=1}^n \left[\frac{D_{\pi/2}}{\sqrt{d}} e^{jkd(\cos \alpha - 1)} \right]^{-m} \right] \quad (3.1)$$

where D_α is the diffraction coefficient with incident angle α , and $D_{\pi/2}$ is the diffraction coefficient for horizontal diffractions over the tops of the half screens. Equation (3.1) is a closed form expression and can be evaluated rapidly on a desktop workstation. Compared with measurement results, Neve and Rowe UTD model works well only in a relative small range. Figure 3.2 and Figure 3.3 show the compar-

Figure 3.2: Comparisons between theoretical models and the measurements, Location 4 (Saratoga)

isons of Neve and Rowe model with measurements [95] and Walfisch and Bertoni model [6]. It is observed that the Walfisch and Bertoni model is in general more

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Figure 3.3: Comparisons between theoretical models and the measurements, Location 5 (Ivy Ridge)

accurate than Neve and Rowe model. As α becomes less than about 0.5 degree, Neve and Rowe model begins to depart from the measurement data, which means, when the receiver is far from the transmitter, this model is no longer accurate.

3.3 Improved Channel Model Based on UTD

To remove the limitation of Neve and Rowe model, an improved UTD-based channel model is proposed. Rows of buildings are no longer replaced by opaque absorbing screens of vanishing thickness as in previous models. The building width is taken into account. Rectangular shaped blocks are used to model the rows of buildings (see Figure 3.4). The incident field propagating over the rooftops will be diffracted by both the left and right edges of the buildings instead of diffracted only once by each half-screen or wedge.

The major assumptions of this model are given below:

- (i) The incident field is assumed to be a vertically polarized plane wave.
- (ii) All the buildings are assumed to be of equal height with perfectly conducting surfaces.
- (iii) The dominant propagation path is the multiple diffraction over rooftops.
- (iv) Back diffracted rays are not considered for the reasons outlined in [6].

3.3.1 Irregular Building Widths and Spacing

Our theoretical model is illustrated in Figure 3.4, $d_1, d_2, \dots, d_{n+1}$ represent the $(n + 1)$ buildings' widths, respectively. W_m is the distance from the left edge of building m to the left edge of building $(m + 1)$. α is the angle of incident at the edges of the buildings for direct incident fields.

Unlike Neve and Rowe model, the incident field will now be incident upon both the left and right edges of the building as indicated in Figure 3.4. The direct fields

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Figure 3.4: Geometry of the propagation model

incident upon building m are given by

$$E_{mL}^i = E_0 e^{-jkW \cos \alpha} \quad (3.2)$$

$$E_{mR}^i = E_0 e^{-jk(W+d_m) \cos \alpha} \quad (3.3)$$

where

$$W = \begin{cases} \sum_{l=1}^{m-1} W_l & m > 1 \\ 0 & m = 1 \end{cases}$$

The superscript i or d refers to an incident or diffracted field component, respectively, and the subscripts L , R identify the left and right edges of the particular building in question.

The direct incident fields are first diffracted by the left and right edges of building m , and then, these diffracted fields will propagate horizontally over $(n - m)$ buildings until they reach the building $(n + 1)$. Diffraction coefficients are symbolized by $D_{\alpha Lm}$, $D_{\alpha Rm}$, D_{Lm} and D_{Rm} , where $D_{\alpha Lm}$ and $D_{\alpha Rm}$ refer to the left and

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right edge diffraction coefficients with an angle α that the incident field component makes with building m , respectively. D_{Lm} and D_{Rm} refer to the horizontal left and right edge diffraction coefficients of building m .

The field diffracted by building m and incident upon left edge of building $(m+1)$ can now be calculated as

$$\begin{aligned} E_m^d &= E_{(m+1)L}^i \\ &= E_{mL}^i D_{\alpha Lm} A(d_m) e^{-jk d_m} D_{Rm} A(W_m - d_m) e^{-jk(W_m - d_m)} \\ &\quad + E_{mR}^i D_{\alpha Rm} A(W_m - d_m) e^{-jk(W_m - d_m)} \end{aligned} \quad (3.4)$$

where $A(s) = 1/\sqrt{s}$ is the spreading factor. The first term of equation (3.4) is the diffracted field caused by the direct incident field on the left edge of building m , the second term is the diffracted field caused by the direct incident field on the right edge of building m . E_m^d will now propagate horizontally across the remaining buildings $(m+1), \dots, n$ to building $(n+1)$. The field at the left edge of building $(n+1)$ due to the field incident upon building m can be calculated as

$$E_{(n+1)L}^i = E_{(m+1)L}^i \times \prod_{t=m+1}^n \left[D_{Lt} D_{Rt} \frac{e^{-jk W_t}}{\sqrt{d_t(W_t - d_t)}} \right] \quad (3.5)$$

Since the electric field is assumed to be vertically polarized, our calculation shows that the diffraction coefficient $D_{\alpha Rm}$ is very small compared with the other diffraction coefficients and may be neglected. The contribution of the direct incident field upon building m to the total received field is given by

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$$E^d(m) = E_{mL}^i D_{\alpha Lm} D_{Rm} \frac{e^{-jkW_m}}{\sqrt{d_m(W_m - d_m)}} \times \prod_{t=m+1}^n \left[D_{Lt} D_{Rt} \frac{e^{-jkW_t}}{\sqrt{d_t(W_t - d_t)}} \right] \quad (3.6)$$

The diffraction coefficients are given as

$$D_{\alpha Lm} = D_{\parallel}(L, \phi = \frac{3}{2}\pi, \phi' = \frac{\pi}{2} + \alpha) \quad (3.7)$$

$$D_{Lm} = D_{\perp}(L, \phi = \frac{3}{2}\pi + \delta\phi, \phi' = \frac{\pi}{2}) \quad (3.8)$$

$$D_{Rm} = D_{\parallel}(L, \phi = \pi + \delta\phi, \phi' = 0) \quad (3.9)$$

The distance parameters for $D_{\alpha Lm}$, D_{Lm} and D_{Rm} are given as below

$$L = \begin{cases} d_m & \text{for } D_{\alpha Lm} \\ \frac{(W_{m-1} - d_{m-1})d_m}{(W_{m-1} - d_{m-1}) + d_m} & \text{for } D_{Lm} \\ \frac{(W_m - d_m)d_m}{W_m} & \text{for } D_{Rm} \end{cases} \quad (3.10)$$

In D_{Lm} and D_{Rm} , a small, positive angular perturbation $\delta\phi$ is introduced to ensure that the field point is always in the shadow zone of the incident field. As long as $\delta\phi$ is reasonably small, no convergence problems will arise since diffraction fields in the UTD do not behave asymptotically at the shadow boundaries, unlike the GTD. The total field at building $(n + 1)$ is given by the summation of (3.6) for buildings $1, \dots, n$ plus the direct field contribution at building $(n + 1)$, namely

$$E_{tot}(n+1)_L = \sum_{m=1}^n \left[E_0 e^{-jkW \cos \alpha} \frac{D_{\alpha Lm} D_{Rm} e^{-jkW_m}}{\sqrt{d_m(W_m - d_m)}} \times \mathbb{D} \right] + E_0 e^{-jk \sum_{l=1}^n W_l \cos \alpha} \quad (3.11)$$

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where

$$\mathbb{D} = \begin{cases} \prod_{t=m+1}^n \left[D_{Lt} D_{Rt} \frac{e^{-jkW_t}}{\sqrt{d_t(W_t - d_t)}} \right] & m < n \\ 1 & m = n \end{cases} \quad (3.12)$$

(3.11) is a closed form expression and can be evaluated rapidly on a desktop PC.

3.3.2 Uniform Building Width and Spacing

Although expression (3.11) can be calculated quickly, it is still a little complicated.

In order to get an explicit expression, we assume that the center-to-center spacing of any two neighboring buildings is of the same length d , and all the buildings are of equal width d_1 as shown in Figure 3.5. As in section 3.3.1, we first find out the

Figure 3.5: Uniform building width and spacing

direct incident field upon building m , given as below

$$E_{mL}^i = E_0 e^{-jk(m-1)d \cos \alpha} \quad (3.13)$$

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$$E_{mR}^i = E_0 e^{-jk((m-1)d+d_1)\cos\alpha} \quad (3.14)$$

The field diffracted by building m is given by

$$\begin{aligned} E_m^d &= E_{(m+1)L}^i \\ &= E_{mL}^i D_{\alpha L} A(d_1) e^{-jkd_1} D_R A(d-d_1) e^{-jk(d-d_1)} \\ &\quad + E_{mR}^i D_{\alpha R} A(d-d_1) e^{-jk(d-d_1)} \end{aligned} \quad (3.15)$$

The first term of equation (3.15) is the diffracted field caused by the direct incident field on the left edge of building m , the second term is the diffracted field caused by the direct incident field on the right edge of building m . Since we assume that the buildings are of uniform width and center-to-center spacing, the diffraction coefficients will now be equal for all the buildings. The denotations $D_{\alpha Lm}$, $D_{\alpha Rm}$, D_{Lm} and D_{Rm} in section 3.3.1 are simplified as $D_{\alpha L}$, $D_{\alpha R}$, D_L and D_R , where

$$D_{\alpha L} = D_{\parallel}(L = d_1, \phi = \frac{3}{2}\pi, \phi' = \frac{\pi}{2} + \alpha) \quad (3.16)$$

$$D_{\alpha R} = D_{\parallel}(L = d - d_1, \phi = \pi, \phi' = \alpha) \quad (3.17)$$

$$D_L = D_{\perp}(L = \frac{d_1(d-d_1)}{d}, \phi = \frac{3}{2}\pi + \delta\phi, \phi' = \frac{\pi}{2}) \quad (3.18)$$

$$D_R = D_{\parallel}(L = \frac{d_1(d-d_1)}{d}, \phi = \pi + \delta\phi, \phi' = 0) \quad (3.19)$$

The diffraction coefficient $D_{\alpha R}$ will also be neglected in the following analysis for the same reason explained above. The contribution of the direct field incident upon building m to the total received field at building $(n+1)$ is

$$E^d(m) = E_0 \frac{D_{\alpha L} D_R e^{-jkd(1+(m-1)\cos\alpha)} / \sqrt{d_1(d-d_1)}}{d_1^{(n-m)/2} (d-d_1)^{(n-m)/2}} \times (D_L D_R)^{n-m} e^{-jk(n-m)d} \quad (3.20)$$

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So the total received field at the left edge of building $(n + 1)$ is the summation of the contribution of the previous n buildings plus the directly incident field, given by

$$E_{tot}(n + 1)_L = E_0 e^{-j k n d \cos \alpha} + E_0 \frac{D_{\alpha L} D_R e^{-j k d (1 + (n-1) \cos \alpha)}}{\sqrt{d_1 (d - d_1)}} \times \frac{1 - \left[D_L D_R e^{j k (\cos \alpha - 1) d} / \sqrt{d_1 (d - d_1)} \right]^n}{1 - D_L D_R e^{j k (\cos \alpha - 1) d} / \sqrt{d_1 (d - d_1)}} \quad (3.21)$$

(3.21) is a closed form expression and much simpler than (3.11).

3.4 Results and Discussion

The variation in field amplitude $|E_{tot}(n+1)_L|$ given by (3.11) and (3.21) are plotted in Figure 3.6 and Figure 3.7 as a function of the building number n , respectively. In Figure 3.6, the intervening spaces $W_1 \dots W_n$ between the left edges of two neighboring buildings are uniformly distributed in the range of (35 m -10 m, 35 m +10 m), building width d_m is a quarter of W_m . The received fields are fluctuating because W_m and d_m are not constants. In Figure 3.7, $d = 35$ m, $d_1 = d/4$, $|E_{tot}(n + 1)_L|$ is observed to settle to a nominally constant value once the number of buildings n becomes large enough. Although there seems no settled field in Figure 3.6, the total received field is actually oscillating around a certain value. To get a clear view of this, we combine Figure 3.6 and Figure 3.7 in Figure 3.8. It is seen that the fields in Figure 3.6 are fluctuating according to the fields in Figure 3.7, that is to say, the settled values in Figure 3.7 are the mean values of the fluctuated fields in Figure 3.6. This can also be explained as, the parameter $d = 35$ m is the mean of the parameters $W_1 \dots W_n$, which are uniformly distributed

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Figure 3.6: $|E_{tot}(n+1)_L|$ /building number n for equation(3.11)

in the range of (35 m -10 m, 35 m +10 m). In the following parts, only (3.21) is used to simplify the analysis.

In Figure 3.7, the reason that received field settles to a constant is due to the

Figure 3.7: $|E_{tot}(n+1)_L|$ /building number n for equation(3.21)

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Figure 3.8: Combination of Figure 3.6 and Figure 3.7

reduced perturbative effect of the buildings with small building number n upon the field of buildings with large n . The settled value is seen to increase with the increasing of α , while for a fixed α , the settled field amplitude is seen to increase as the center-to-center spacing increases.

As we mentioned before, in equation (3.15), the diffraction coefficient $D_{\alpha R}$ is very small compared with others, thus the latter part of the equation can be ignored to make the derivation simpler. To validate this, Figure 3.9 plots the theoretical results from equation (3.21) and those considering $D_{\alpha R}$. For different values of α , 0.2° , 1.0° and 3.2° , these two results are practically the same, that is to say, the neglect of the field caused by diffraction coefficient $D_{\alpha R}$ is reasonable, and the simplified equation (3.21) can be used directly to calculate the total received field.

As in previous models [6, 21], the received field is found to be a function of $\alpha\sqrt{d/\lambda}$. The received field versus $\alpha\sqrt{d/\lambda}$ is plotted in Figure 3.10, the building

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Figure 3.9: $|E_{tot}(n+1)_L|$ versus building number n

width d_1 is chosen as $0.25d$. It is observed that for the same value $\alpha\sqrt{d/\lambda}$, the received field amplitudes are nearly the same when d equals 50λ and 200λ , which means the settled field is a function of $\alpha\sqrt{d/\lambda}$, not the particular variable d or

Figure 3.10: $|E_{tot}(n+1)_L|$ versus $\alpha\sqrt{d/\lambda}$

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α . Hence an empirical expression can also be derived when the settled field versus $\alpha\sqrt{d/\lambda}$ is plotted using logarithmic scales.

Figure 3.11: $|E_{tot}(n+1)_L|$ versus $\alpha\sqrt{d/\lambda}$. $d = 200\lambda$, $d_1 = 0.25d$

Figure 3.11 is a plot of $|E_{tot}(n+1)_L|$ versus $\alpha\sqrt{d/\lambda}$ in logarithmic scales. The slope of the curve is sensitive to the value of d_1/d . As d_1/d varies, the slope of the curve will change. Furthermore, since our closed form expression (3.21) can be calculated rapidly on a desktop PC, the empirical expression is not very necessary to reduce the simulation time as in [6].

The total path loss is expressed as

$$L_p = L_F + L_{ex} \quad (3.22)$$

L_F is the free space loss given by

$$L_F(dB) = -20 \log \left(\frac{\lambda}{4\pi d_{(T-R)}} \right) \quad (3.23)$$

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where $d_{(T-R)}$ is the separation distance between the transmitter and the receiver. L_{ex} is the excess loss caused by multiple diffractions of the buildings. The received signal power can be found by subtracting L_p from the radiated power plus antenna gains, expressed in dBm.

Figures 3.12 to 3.17 show all the measurements of average received signal in Philadelphia by Ott and Plitkins, reported in [95], at 820 MHz with radiated power and antenna gains totaling 53.3 dBm. Six fixed antenna locations were used with antenna heights ranging between 45 ft and 255 ft above the curb. The average building height h is assumed to be 7.5 m. The mobile antenna height h_m is about 1.5 m. We use $d = 35$ m as the center-to-center spacing. All the parameters above follow those in [6], except the additional parameter building width d_1 . Our simulation results show that good agreement between theoretical calculation and measurements are obtained for d_1 from 8 m to 12 m. Thus the average value of $d_1 = 10$ m is chosen. Figure 3.12 to 3.17 reproduce the measurements for all the six locations. The dots represent sector averages of the received signal level in dBm plotted as a function of range d_t in miles from the transmitter. The transmitting antenna height above the assumed building rooftop level H is indicated for each site.

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Figure 3.12: Location 1 Arch

Figure 3.14: Location 3 Dewey

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Figure 3.13: Location 2 Pennypacker-High

Figure 3.15: Location 4 Saratoga

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Figure 3.16: Location 5 Ivy Ridge

Figure 3.17: Location 6 Jefferson

Comparisons between results from this theoretical model and measurements

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have shown good agreement for all the six measured sites reported in [95]. The reason for the not very good agreement in the first two figures has been explained in [6], i.e., these two sites are located near downtown Philadelphia, whose tall buildings may be responsible for the reduction in the signal below the predicted values. The remaining four sites are away from the downtown region. For these sites the measured received signal strength is in good agreement with the theoretical result. Generally speaking, our improved UTD model performs better than Neve and Rowe model especially for $\alpha < 0.5^\circ$. It is comparable with Walfisch and Bertoni model and matches the measurements very well. Furthermore, since our model is ray-based, it saves computing time compared with Walfisch and Bertoni model.

3.5 Conclusion

In this chapter, we developed a simple analytical UTD based channel model for predicting the average path loss in urban cellular communication systems. The buildings are modeled by rectangular shaped blocks to take into account the buildings' width. Vertically polarized plane wave incident field, equal building height and perfectly conducting building surfaces are assumed. Closed form expressions are derived to calculate the total received field. When applying the average values to the building width and center-to-center spacing of two neighboring buildings, a simpler analytic path loss model is thus obtained. The received field is observed to settle to a constant value once the number of buildings becomes large enough, and the settled field is found to be a function of $\alpha^2 d / \lambda$ provided that we fix the ratio of d_1 / d as shown in Figure 3.5. The performance of our proposed model has been contrasted with measurements and two reported models [6, 21], i.e., the physical optics based Walfisch and Bertoni model and the UTD based Neve and Rowe model. Good agreement between our theoretical results with measurements and Walfisch and Bertoni model shows that our UTD model is accurate and outperforms Neve and Rowe model.

Chapter 4

Analytical Propagation Channel Model

4.1 Introduction

In Chapter 3, a path loss model for urban cellular environment was proposed. However, this information is only one of the fundamental demands of the channel modeling, with the advent of smart antenna systems comes the need to better understand the spatial and delay properties of the wireless communications system, the traditional path loss model cannot solve such problems. Therefore, the geometrically based channel model becomes a more practical way in evaluating the spatial and delay properties. The basic idea of the existing geometrically based single bounce channel models is making assumptions of the scatterers' distribution around the mobile or between the base station and the mobile receiver. Different from these previous models, in this chapter, a simple analytical propagation channel model is proposed assuming that the scatterers are uniformly distributed around the base station inside a circle, of which the radius is determined by the coverage area of the transmitting antenna. The mobile receiver is located within

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the scatterer region. The PDF of the AOA and the PDF of the TOA for multiple paths arriving at the mobile receiver are discussed for both omnidirectional and directional transmitting antenna systems. The effect of the beamwidth of the directional transmitting antenna on the Doppler spread will be analyzed as well.

This chapter is organized as follows. In section 4.2, the PDFs of the AOA and the TOA are derived for omnidirectional transmitting antenna, the angular spread and delay spread are also investigated. Section 4.3 analyzes the PDF of AOA, the PDF of TOA and the Doppler spectrum for directional transmitting antenna system. The conclusion is given in section 4.4.

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4.2 Model for Omnidirectional Transmitting Antenna Systems

4.2.1 The Angle of Arrival

In mobile radio system design, hexagons are often used to model coverage areas and base stations are usually depicted as being at the center of the cell. Hence it is reasonable to approximate the cell as a circle with the base station in the center as shown in Figure 4.1. The base station and mobile receiver are marked as B and M , respectively.

Figure 4.1: AOA at the mobile receiver when the base station uses an omnidirectional antenna

M , respectively. R is the radius of the cell, and D denotes the distance between the base station and the mobile receiver. The receiver is located within the cell.

To simplify the formulation, the following assumptions are made:

- (i) The scatterers are assumed to be uniformly distributed inside the circle around the base station.
- (ii) Each scatterer is an omnidirectional reradiating element.
- (iii) Only one bounce is assumed for each propagation path.

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(iv) Scattering/re-radiation loss is neglected.

Let the length of line segment ME be r , and this line makes an angle θ with line MB . By using the law of Cosines, the following expression can be obtained

$$R^2 = D^2 + r^2 - 2Dr \cos \theta \quad (4.1)$$

From (4.1), r is easily derived as

$$r = D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta} \quad (4.2)$$

where θ is the angle of arrival at the mobile receiver.

Since the scatterers are uniformly distributed and each multipath component from the transmitter to receiver undergoes only one bounce, the area within the shadowed strip as shown in Figure 4.1 is proportional to the probability of the AOA at the mobile receiver. The cumulative distribution function (CDF) of the AOA can be derived as below

$$F_\theta(\theta) = \frac{A_{strip}}{A_{circle}} = \frac{\frac{1}{2} \int_0^\theta r^2 dx}{\pi R^2} \quad (4.3)$$

where A_{strip} is the area within the strip and A_{circle} is the area of the circle. Substituting (4.2) into (4.3) gives

$$F_\theta(\theta) = \frac{1}{2\pi R^2} \int_0^\theta \left(D \cos x + \sqrt{R^2 - D^2 \sin^2 x} \right)^2 dx \quad (4.4)$$

The PDF of the AOA is the derivative of the CDF with respect to θ and given by

$$f_\theta(\theta) = \frac{dF_\theta(\theta)}{d\theta} = \frac{1}{2\pi} \left(\frac{D}{R} \cos \theta + \sqrt{1 - \frac{D^2}{R^2} \sin^2 \theta} \right)^2 \quad -\pi \leq \theta \leq \pi \quad (4.5)$$

The PDF of the AOA at the mobile receiver depends only on the ratio of D to R .

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Figure 4.2: PDF of the AOA at the mobile receiver

The comparisons in Figure 4.2 show that our theoretical PDFs match the normalized simulated histograms very well. The simulated histograms are generated by creating uniformly distributed scatterers within the cell, and for each scatterer, the angle of arrival at the mobile receiver is calculated. By carrying out 100 000 trials, the normalized histograms for the AOA can be easily obtained. Figure 4.2 plots the PDF curves for different values of D/R when the base station uses an omnidirectional antenna. The PDF value at $\theta = 0^\circ$ accounts for reflection from scatterers located along the same line as the transmitter. From this figure we can see that the AOA at the receiver ranges mostly from about -100° to 100° for larger value of D/R , i.e., the mobile receiver is far from the base station. The abscissas of the intersections of the three curves are around 60° . This can be explained as in Figure 4.3, the area of the shadowed strip, which is proportional to

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Figure 4.3: The variations of the area of the shadowed strips in Figure 4.1 for different mobile receiver locations

the probability of the AOA, is almost the same for different values of D/R (that is, the different locations of the mobile receiver M_1 , M_2 , M_3 once the scatterer region is determined) when the AOA is around 60° . For the AOA that is smaller than 60° , the area of the shadowed strip will increase as the value of D/R increases, and when AOA becomes larger than 60° , the area of the shadowed strip will decrease as the value of D/R increases. These are the reasons for the variations of the PDF curves in Figure 4.2.

Due to the scattering, multipath components arrive at angles different from the direct component, and the angular spread is a measure used to determine the angular dispersion of the channel. The angular spread is defined based on the second central moment as below

$$\sigma_\theta = \sqrt{\overline{\theta^2} - \bar{\theta}^2} \quad (4.6)$$

where

$$\bar{\theta} = \int_{-\pi}^{\pi} \theta f_\theta(\theta) d\theta \quad (4.7)$$

$$\overline{\theta^2} = \int_{-\pi}^{\pi} \theta^2 f_\theta(\theta) d\theta \quad (4.8)$$

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Figure 4.4: Plot of angular spread versus the value of D/R

Using the model derived above, we can easily obtain the angular spread. The plot of angular spread as a function of the ratio of the distance between the base station and the mobile receiver to the radius of the scattering region is shown in Figure 4.4. The value of D/R ranges from 0 to 1. From the figure we can see that the angular spread decreases almost linearly from about 104° to 32.5° as D/R increases. Also the angular spread is independent of the particular value of R or D , because the PDF of the AOA depends only on D/R . In Figure 4.4, the angular spreads for $R = 1000$ m and $R = 5000$ m are plotted, and the two lines are observed to overlap each other.

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4.2.2 Time of Arrival

The scatterers contributing the same time delay τ lie on an ellipse with foci at the base station and the mobile receiver as shown in Figure 4.5. The major axis of

Figure 4.5: TOA at the mobile receiver when the base station uses an omnidirectional antenna

the ellipse is given by $c\tau$, where c is the speed of light. The area of intersection of the ellipse with the circular scatterer region can be easily derived using the polar coordinated equations of the circle and ellipse referred to the polar coordinate system defined at the base station, (ρ, ϕ) . ϕ is assumed to increase in the clockwise direction for convenience. With respect to this coordinate system, the circle and the ellipse are described by the following two equations, respectively

$$\rho = R \quad (4.9)$$

$$\rho = \frac{c^2\tau^2 - D^2}{2c\tau - 2D\cos\phi} \quad (4.10)$$

$(2R - D)/c$ is the maximum time delay that keeps the ellipse inside the circle, thus the area of the intersection is the area of the ellipse when $\tau \leq (2R - D)/c$. As the time delay increases, the intersecting area becomes part of the ellipse, which can

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be viewed as the summation of two parts shown in Figure 4.5, one is a sector of the circle with angle 2α , and the other is a part of the ellipse which area can be obtained by integrating in polar coordinate system. α is determined by substituting α for ϕ and solving (4.9) and (4.10) simultaneously

$$\alpha = \cos^{-1} \left(\frac{D^2 + 2Rc\tau - c^2\tau^2}{2RD} \right) \quad (4.11)$$

The ratio of the area of the intersection of ellipse with circle to the area of circle leads to an expression for the cumulative distribution function of the time delay. The area of intersection of ellipse with the circular region is given by

$$A = \begin{cases} \frac{\pi c\tau}{4} \sqrt{c^2\tau^2 - D^2} & \frac{D}{c} < \tau \leq \frac{2R-D}{c} \\ R^2\alpha + \int_{\alpha}^{\pi} \left(\frac{c^2\tau^2 - D^2}{2c\tau - 2D\cos x} \right)^2 dx & \tau > \frac{2R-D}{c} \end{cases} \quad (4.12)$$

where D/c is the minimum time delay.

The CDF of the multipath delay is given by $F_{\tau}(\tau) = \frac{A}{\pi R^2}$. The PDF of the time of arrival is the derivative of the CDF with respect to τ and is given as below

$$f_{\tau}(\tau) = \begin{cases} \frac{c(c^2\tau^2 + m_1^2)}{4R^2m_1} & \frac{D}{c} < \tau \leq \frac{2R-D}{c} \\ \frac{m_5}{\pi R D \sqrt{1-m^2}} + \frac{c(c^2\tau^2 + m_1^2)}{4R^2m_1} - \frac{\mathbb{Z}_1 + \mathbb{Z}_2}{\pi R^2} & \tau > \frac{2R-D}{c} \end{cases} \quad (4.13)$$

where

$$m = \frac{D^2 + 2Rc\tau - c^2\tau^2}{2RD} = \cos \alpha \quad (4.14)$$

$$m_1 = \sqrt{c^2\tau^2 - D^2} \quad (4.15)$$

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$$m_2(x) = \tan^{-1} \left(\sqrt{\frac{c\tau + D}{c\tau - D}} \tan \frac{x}{2} \right) \quad (4.16)$$

When $x = \alpha$, $m_2(\alpha)$ is obtained.

$$m_3 = c\tau - D + (c\tau + D) \tan^2 \frac{\alpha}{2} \quad (4.17)$$

$$m_4 = c\tau - Dm \quad (4.18)$$

$$m_5 = c^2\tau - Rc \quad (4.19)$$

$$\mathbb{Z}_1 = \frac{c^2\tau D\sqrt{1-m^2}}{2m_4} + \frac{m_1^2 m m_5}{4Rm_4\sqrt{1-m^2}} - \frac{Dc^2\tau m_1^2\sqrt{1-m^2}}{4Rm_4^2} \quad (4.20)$$

$$\mathbb{Z}_2 = \frac{c(c^2\tau^2 + m_1^2)}{2m_1} m_2(\alpha) + \frac{c\tau}{2m_3} \left[\frac{m_1^2 m_5}{RD(1+m)\sqrt{1-m^2}} - cD \tan \frac{\alpha}{2} \right] \quad (4.21)$$

The time delay is restricted to $D/c < \tau \leq (2R + D)/c$. The minimum delay is determined by the distance between the base station and the mobile receiver D , while the maximum time delay that is valid for the formula depends on both D and the radius of the cell R .

The PDF curves of the path delay are plotted in Figure 4.6. The distance between the base station and the mobile receiver is fixed at 600 m, and the radius of the scatterers is selected as 1000 m, 1200 m and 1400 m for test cases. The minimum time delays for these three curves are the same since D is fixed, but the maximum time delays are different because of different R . As R increases, the length of the increasing part of the curves increases, that is to say, there will be larger probability of having longer time delay path. When τ becomes larger than $(2R - D)/c$, as the time delay increases, the PDF will decrease, which means that

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Figure 4.6: PDF of the TOA at the mobile receiver (Omnidirectional transmitting antenna, D is fixed at 600 m, R is various as shown in the figure)

the probability of receiving signals with extreme long time delay is small. From the analysis above we can see that, most of the rays that reach the mobile receiver with time delay about $(2R - D)/c$. When $\tau > (2R + D)/c$, the PDF is zero, i.e., no ray will arrive at the receiver after time delay $(2R + D)/c$. This is for the reason that we assume the effective scatterers are located within a circle of radius R , the effects of the scatterers outside this circle are neglected.

Delay spread is a measure of time dispersion of the channel. It is defined based on the second central moment as below

$$\sigma_\tau = \sqrt{\overline{\tau^2} - \bar{\tau}^2} \quad (4.22)$$

where

$$\bar{\tau} = \int_{\tau_{min}}^{\tau_{max}} \tau f_\tau(\tau) d\tau \quad (4.23)$$

$$\overline{\tau^2} = \int_{\tau_{min}}^{\tau_{max}} \tau^2 f_\tau(\tau) d\tau \quad (4.24)$$

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Figure 4.7: Plot of delay spread versus the value of R/D (D is fixed at 600 m and R is varied)

τ_{min} and τ_{max} are the minimum and maximum time delay discussed above. The delay spread for this model is plotted in Figure 4.7. D is maintained at 600 m and R is varied, the ratio R/D is ranged from 1 to 10. It can be observed that the delay spread increases almost linearly as the value of R/D increase. That is to say, once the distance between the base station and the mobile receiver is fixed, the delay spread will increase linearly as the radius of the scatterer region increases.

4.3 Model for Directional Transmitting Antenna Systems

4.3.1 Angle of Arrival

Figure 4.8: AOA at the mobile receiver when the base station uses a directional antenna of beamwidth 2β

For directional antenna communication system, each antenna only radiates a specified sector. In Figure 4.8, a sector antenna of beamwidth 2β is used at the base station, and scatterers are assumed to be uniformly distributed within the sector. Since the mobile receiver can lie anywhere within the sector, we assume that the line connecting the base station and the mobile receiver makes angle ϑ with the angle bisector of $\angle EBF$. θ_{m1} is the angle between line EM and BM , θ_{m2} is the angle between line BM and FM . θ is the angle of arrival at the mobile receiver. The mobile receiver is in the upper part of the sector when ϑ is positive, and lower part when ϑ is negative. θ_{m1} and θ_{m2} are both assumed to be positive for the convenience of analysis.

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The cumulative distribution function of the AOA is expressed as

$$F_{\theta}(\theta) = \frac{A_{strip}}{A_{sector}} = \frac{\frac{1}{2} \int_0^{\theta} r^2 dx}{\beta R^2} \quad (4.25)$$

where A_{strip} is the area within the shadowed strip, A_{sector} is the area of sector with angle 2β , and r can be determined as below:

(i) $0 \leq \theta < \pi$

When $\theta \leq \theta_{m1}$, by using the Law of Sines, we get

$$\frac{r}{\sin(\beta - \vartheta)} = \frac{D}{\sin(\pi - (\beta - \vartheta) - \theta)} \quad (4.26)$$

When $\theta > \theta_{m1}$, by using the Law of Cosines, we get

$$R^2 = D^2 + r^2 - 2Dr \cos \theta \quad (4.27)$$

Thus r can be obtained as

$$r = \begin{cases} \frac{D \sin(\beta - \vartheta)}{\sin(\beta - \vartheta + \theta)} & \theta \leq \theta_{m1} \\ D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta} & \theta > \theta_{m1} \end{cases} \quad (4.28)$$

For θ_{m1} , it can be determined from the Law of Sines,

$$\frac{R}{\sin \theta_{m1}} = \frac{\overline{ME}}{\sin(\beta - \vartheta)} = \frac{\sqrt{R^2 + D^2 - 2RD \cos(\beta - \vartheta)}}{\sin(\beta - \vartheta)} \quad (4.29)$$

θ_{m1} is an acute angle when $D/R > \cos(\beta - \vartheta)$, and an obtuse angle when $D/R < \cos(\beta - \vartheta)$. Therefore,

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$$\theta_{m1} = \begin{cases} \sin^{-1} \left(\frac{R \sin(\beta - \vartheta)}{\sqrt{R^2 + D^2 - 2RD \cos(\beta - \vartheta)}} \right) & \frac{D}{R} \geq \cos(\beta - \vartheta) \\ \pi - \sin^{-1} \left(\frac{R \sin(\beta - \vartheta)}{\sqrt{R^2 + D^2 - 2RD \cos(\beta - \vartheta)}} \right) & \frac{D}{R} < \cos(\beta - \vartheta) \end{cases} \quad (4.30)$$

(ii) $-\pi \leq \theta < 0$

r and θ_{m2} are obtained following the same way that r and θ_{m1} have been derived in case (i). The expressions for them are given by

$$r = \begin{cases} \frac{D \sin(\beta + \vartheta)}{\sin(\beta + \vartheta + |\theta|)} & |\theta| \leq \theta_{m2} \\ D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta} & |\theta| > \theta_{m2} \end{cases} \quad (4.31)$$

$$\theta_{m2} = \begin{cases} \sin^{-1} \left(\frac{R \sin(\beta + \vartheta)}{\sqrt{R^2 + D^2 - 2RD \cos(\beta + \vartheta)}} \right) & \frac{D}{R} \geq \cos(\beta + \vartheta) \\ \pi - \sin^{-1} \left(\frac{R \sin(\beta + \vartheta)}{\sqrt{R^2 + D^2 - 2RD \cos(\beta + \vartheta)}} \right) & \frac{D}{R} < \cos(\beta + \vartheta) \end{cases} \quad (4.32)$$

The main difference between these two cases is that the $\angle MBE$ equals $(\beta - \vartheta)$ in (i) but $\angle MBF$ equals $(\beta + \vartheta)$ in (ii).

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The PDF of the AOA is the derivative of the CDF with respect to θ as below

$$f_{\theta}(\theta) = \begin{cases} \frac{\left(D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}\right)^2}{2R^2 \beta} & \theta_{m1} \leq \theta \leq \pi \text{ or } -\pi \leq \theta \leq -\theta_{m2} \\ \frac{D^2 \sin^2(\beta - \vartheta)}{2R^2 \beta \sin^2(\beta - \vartheta + \theta)} & 0 < \theta < \theta_{m1} \\ \frac{D^2 \sin^2(\beta + \vartheta)}{2R^2 \beta \sin^2(\beta + \vartheta + |\theta|)} & -\theta_{m2} < \theta < 0 \end{cases} \quad (4.33)$$

It can be seen that, the PDF of AOA depends on not only the ratio of D/R , as in the omnidirectional transmitting antenna system, but also the beamwidth 2β and the angle ϑ .

For a simpler case when $\vartheta = 0$, the mobile receiver lies on the bisector of $\angle EBF$ as shown in Figure 4.9. By substituting 0 for ϑ , (4.33) can now be simplified as

Figure 4.9: $\vartheta = 0$

$$f_{\theta}(\theta) = \frac{dF_{\theta}(\theta)}{d\theta} = \begin{cases} \frac{D^2 \sin^2 \beta}{2R^2 \beta \sin^2(\beta + |\theta|)} & |\theta| < \theta_m \\ \frac{1}{2R^2 \beta} \left(D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}\right)^2 & |\theta| \geq \theta_m \end{cases} \quad (4.34)$$

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where $\theta_{m1} = \theta_{m2} = \theta_m$ is given by

$$\theta_m = \begin{cases} \sin^{-1} \left(\frac{R \sin \beta}{\sqrt{R^2 + D^2 - 2RD \cos \beta}} \right) & \frac{D}{R} \geq \cos \beta \\ \pi - \sin^{-1} \left(\frac{R \sin \beta}{\sqrt{R^2 + D^2 - 2RD \cos \beta}} \right) & \frac{D}{R} < \cos \beta \end{cases} \quad (4.35)$$

In this case, the PDF of the AOA depends both on the value of D/R and the beamwidth 2β .

Figure 4.10: AOA at the mobile receiver when the base station uses a directional antenna of beamwidth 120° , $\vartheta = 10^\circ$, $D/R = 0.5$

Figure 4.10 shows the PDF of the AOA for a 120° sector transmitting antenna system. The distance between the base station and the mobile receiver is $0.5R$. The PDF of the AOA for this directional antenna system is not symmetrical with respect to the line $\theta = 0^\circ$ as in Figure 4.2. It is obvious that the probability of receiving signals with $\theta < 0^\circ$ is larger than the probability of receiving signals with

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$\theta > 0^\circ$. This is for the reason that, in the simulation, ϑ is selected for a positive value of 10° .

Figure 4.11: AOA at the mobile receiver when the base station uses a directional antenna of beamwidth 120° , $D/R = 0.5$

In order to get a clear view of the effect of angle ϑ , we plot a 3D figure in Figure 4.11. The X, Y and Z axes are ϑ , the angle of arrival and the probability density function, respectively. When $\vartheta < 0$, the receiver is in the lower part of the sector EBF. Since the scatterers are uniformly distributed, the total number of scatterers within the region GBF is smaller than that in region EBG, which leads to the fact that the probability of the AOA having negative values is smaller than the probability of the AOA having positive values. As ϑ increases, the number of negative values increases. When ϑ reaches zero, the probabilities of having positive AOAs and negative AOAs are equal. This can also be explained from equation (4.34), which is an even function and is symmetrical with respect to the line $\theta = 0^\circ$.

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For the case that $\vartheta > 0$, the AOAs are mostly negative. From the analysis above, we can see that as the angle ϑ increases, the probability of having negative AOA increases while the probability of having positive AOA decreases.

Figure 4.12: PDF of the AOA at the mobile receiver (120° sector antenna)

For communication system using a 120° sector antenna, the comparisons between the PDF of the AOA for different values of D/R are shown in Figure 4.12. Same as before, the AOA mainly lies in the range of $(-100^\circ, 100^\circ)$ for larger value of D/R , furthermore, there are peaks when θ equals to 0° or $\pm\theta_m$ for the reason that the area of the shadowed strip reaches maximum at these values.

Figure 4.13 plots the PDF curves of the AOA when base station uses 60° , 90° and 120° sector antennas. The ratio of D to R is maintained at 0.8. It can be seen that the effect of the beamwidth is not that obvious as the effect of D/R when the AOA is larger than θ_m . The dotted lines in the figure show where the

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Figure 4.13: PDF of the AOA at the mobile receiver ($D/R = 0.8$)

θ_m is for these three PDF curves. The PDFs of the AOA change dramatically when $\theta \leq |\theta_m|$, but almost remain the same when $\theta > |\theta_m|$. This is for the reason

Figure 4.14: Illustration of the AOA at the mobile receiver when using transmitting antennas of different beamwidths

that the area of the shadowed strip, which is proportional to the probability of the AOA, is sensitive to the beamwidth of the sector antenna when $\theta \leq |\theta_m|$, as shown in Figure 4.14. For AOAs that are larger than θ_m , the beamwidth of the

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antenna will not affect the area of the strip since we assume that all the scatterers are located within the sector.

4.3.2 Time of Arrival

Without loss of generality, we assume that the angle between the line connecting the base station and the mobile receiver and the angle bisector of $\angle EBF$ is zero. Again, we calculate the intersection area of the ellipse with the sector scatterer region in Figure 4.15. The ellipse is described in (4.10), and the intersection area

Figure 4.15: TOA for directional antenna system

can be calculated as

$$\begin{aligned}
 A &= \begin{cases} \int_0^\beta \left(\frac{c^2\tau^2 - D^2}{2c\tau - 2D\cos x} \right)^2 dx & \frac{D}{c} < \tau \leq \frac{2R-D}{c} \\ R^2\alpha + \int_\alpha^\beta \left(\frac{c^2\tau^2 - D^2}{2c\tau - 2D\cos x} \right)^2 dx & \tau > \frac{2R-D}{c} \end{cases} \\
 &= \begin{cases} \mathbb{Z}_3(\beta) & \frac{D}{c} < \tau \leq \frac{2R-D}{c} \\ R^2\alpha + \mathbb{Z}_3(\beta) - \mathbb{Z}_3(\alpha) & \tau > \frac{2R-D}{c} \end{cases}
 \end{aligned} \tag{4.36}$$

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where $\mathbb{Z}_3(\beta)$ and $\mathbb{Z}_3(\alpha)$ are obtained by substituting β and α for x in the following expression, respectively

$$\mathbb{Z}_3(x) = \frac{D(c^2\tau^2 - D^2) \sin x}{4(c\tau - D \cos x)} + \frac{c\tau\sqrt{c^2\tau^2 - D^2}}{2} \tan^{-1} \left(\sqrt{\frac{c\tau + D}{c\tau - D}} \tan \frac{x}{2} \right) \quad (4.37)$$

So the CDF of the time delay is given by $F_\tau(\tau) = \frac{A}{R^2\beta}$, and the PDF can be obtained as

$$f_\tau(\tau) = \frac{dF_\tau(\tau)}{d\tau} = \begin{cases} \frac{D(cD^2 - c^2\tau D \cos \beta) \sin \beta}{4(c\tau - D \cos \beta)^2 R^2 \beta} + \frac{c(c^2\tau^2 + m_1^2)m_2(\beta)}{2m_1 R^2 \beta} & \frac{D}{c} < \tau \leq \frac{2R - D}{c} \\ \frac{m_5}{\beta R D \sqrt{1 - m^2}} + \frac{D \sin \beta (cD^2 - c^2\tau D \cos \beta)}{4(c\tau - D \cos \beta)^2 R^2 \beta} + \frac{c(c^2\tau^2 + m_1^2)m_2(\beta)}{2m_1 R^2 \beta} - \frac{(\mathbb{Z}_1 + \mathbb{Z}_2)}{R^2 \beta} & \tau > \frac{2R - D}{c} \end{cases} \quad (4.38)$$

where m_1 , $m_2(\beta)$, m_4 , m_5 , $\mathbb{Z}_1$ and $\mathbb{Z}_2$ are defined in (4.15), (4.16), (4.18), (4.19), (4.20) and (4.21), respectively. The maximum time delay is restricted by the condition

$$R = \frac{c^2\tau^2 - D^2}{2c\tau - 2D \cos \beta} \quad (4.39)$$

Solving (4.39), we get the maximum τ that is valid for this directional antenna system model

$$\tau_{direc} = \frac{R + \sqrt{R^2 + D^2 - 2DR \cos \beta}}{c} \quad (4.40)$$

The path delay for directional antenna system is plotted in Figure 4.16, the beamwidth is 120° , the distance between the base station and the mobile receiver is also fixed at 600 m, and the radius of the scatterers is selected as 1000 m,

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Figure 4.16: PDF of the TOA at the mobile receiver (120° sector antenna, D is fixed at 600 m, R is various as shown in the figure)

1200 m and 1400 m for test cases. Same as before, the minimum time delays for these three curves are the same, and the maximum time delays vary with different values of R . Compared with Figure 4.6, it can be seen that, for the same parameters D and R , the maximum time delay for sector antenna system is smaller than that for omnidirectional transmitting antenna system. This is for the reason that the time delay for the sector antenna system is restricted to $\tau_{direc} = (R + \sqrt{R^2 + D^2 - 2DR \cos \beta})/c$, as long as $\beta < \pi$ ($\beta = \pi$ becomes omnidirectional antenna), $\tau_{direc} < \tau_{omni} = (2R + D)/c$. The peak value of the curves happens when time delay equals to $(2R - D)/c$.

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Figure 4.17: PDF of the TOA at the mobile receiver for different beamwidth antenna systems, D is fixed at 600 m, $R = 1000$ m.

The effect of the beamwidth on the TOA is not very obvious once the values of D and R are determined. But the maximum time delay is sensitive to the beamwidth of the base station antenna, which can be figured out from (4.40).

τ_{direc} increases as the beamwidth of the transmitting antenna increases.

4.3.3 Doppler Spectrum

Now we will analyze the spectral shape due to the Doppler spread for this model. The received signal at the mobile receiver experiences Doppler spread due to the motion of the receiver. Assume that the mobile moves at a constant velocity v , and an angle ϕ with respect to the direct line-of-sight component. θ is the angle of arrival at the mobile receiver. An omnidirectional antenna is used at the receiver. The difference in path lengths of the received multipath signals lead to the phase

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change, and hence the apparent change in frequency, or Doppler shift, is given by

$$f_d = f_m \cos(\phi - \theta) = \frac{v}{\lambda} \cdot \cos(\phi - \theta) \quad (4.41)$$

where f_m is the maximum Doppler shift obtained when $\phi = \theta$, λ is the wavelength of the carrier signal. It can be seen that the Doppler shift is related to the mobile velocity and the spacial angle between the direction of motion of the mobile and the direction of arrival of the wave.

The power spectral density (PSD) is given by Jakes in [96]

$$S(f) = \frac{P_0}{\sqrt{f_m^2 - f^2}} [f_\theta(\phi + \cos^{-1}(f/f_m)) + f_\theta(\phi - \cos^{-1}(f/f_m))] , \quad |f| < f_m \quad (4.42)$$

where P_0 is the average power that would be received by an isotropic antenna, $f_\theta(\theta)$ is the PDF of the AOA at the mobile receiver. In order to see the effect of directional antenna, let us consider a test case where the mobile is travelling at a velocity of 54 km/h and a carrier frequency of 2 GHz, follow those in [84]. The ratio of D/R is selected as 0.5 and $\vartheta = 0$. The beamwidth of the sector antenna is assumed to be 30° , 60° and 120° . Figure 4.18 and Figure 4.20 show the Doppler spectrum for $\phi = 0^\circ$ and $\phi = 90^\circ$, respectively.

For $\phi = 0^\circ$, there are more negative Doppler frequency components than positive Doppler frequency components. This is explained by dividing the scatterer region into two parts by a line through the mobile receiver perpendicular to the LOS component, since the mobile is moving towards the base station in the direction of LOS component, the scatterers in the right part contribute to the positive Doppler shift and the scatterers in the left part contribute to the negative Doppler

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Figure 4.18: Doppler spectrum when the mobile is moving at the direction of 0° with respect to the direct component. $D/R = 0.5$ and $\vartheta = 0$

Figure 4.19: Illustration of the scattering region that causes positive and negative frequency components when the mobile is moving toward the base station

shift as shown in Figure 4.19. Larger area of the left part results in more negative Doppler frequency components. Meanwhile, there exists a small peak for each curve, which happens when the angle of arrival at the mobile equals to $\pm\theta_m$ (see Figure 4.13). This is for the reason that the Doppler spectrum is proportional to the PDF of the AOA, which reaches its maximum at $\pm\theta_m$.

For $\phi = 90^\circ$, the Doppler spectrum is symmetrical with respect to the zero

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Figure 4.20: Doppler spectrum when the mobile is moving at the direction of 90° with respect to the direct component. $D/R = 0.5$ and $\vartheta = 0$

Figure 4.21: Illustration of the scattering region that causes positive and negative frequency components when the mobile is moving perpendicular to the base station

Doppler shift. Again, we divide the scatterer region into two parts by the line joining the base station and the mobile as shown in Figure 4.21. Equal areas of the two parts lead to the equal positive and negative Doppler frequency components. The Doppler spectrum of Clarke's model [96] is shown in these two figures for comparison, it is observed that by using directional transmitting antenna, the Doppler spread at the mobile receiver can be reduced. The narrower the beamwidth

of the directional antenna, the less the Doppler spread.

4.4 Conclusion

A simple analytical propagation channel model for mobile communication system is derived in this chapter. The scatterers are assumed to be uniformly distributed around the base station inside a circle, which radius is determined by the coverage area of the base station. The mobile receiver is located within this area, and each multipath component is assumed to undergo only one bounce traveling from the transmitter to the receiver. The probability density functions for the angle of arrival and the time of arrival at the mobile receiver are derived for both omnidirectional and directional transmitting antenna systems. Good agreement has been shown between our theoretical results and the normalized simulated histograms. This model also shows that the angular spread decreases almost linearly as the value of D/R increases, while the delay spread increases almost linearly as the radius of the scatterer region increases. By analyzing the effect of the directional transmitting antenna on the Doppler spectrum, it is found that by reducing the beamwidth of the antenna, the Doppler spread at the mobile receiver will be greatly reduced.

Chapter 5

Channel Models for Outdoor and Indoor Propagation Environments

5.1 Introduction

In Chapter 4, the geometrically based single bounce scatterer model was proposed to evaluate the spatial and delay properties of the wireless propagation channel. However, simple uniform distribution is not good enough to model the various propagation environments. In this chapter, a more general situation is discussed, where the distribution of the scatterers can be arbitrary. The joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are derived. By employing different distributions, e.g. Rayleigh distribution, exponential distribution, etc., this model can be applied to different propagation environments.

This chapter is organized as follows. In section 5.2, general formulations of the model and the derivation of the joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are outlined. The Rayleigh distribution scatterer model is discussed in section 5.3, and the exponential distribution scatterer model will be described in section 5.4. A brief conclusion is given at the end.

5.2 General Formulation

As mentioned in Chapter 4, the scatterers are assumed to be distributed around the base station within a circular region of radius R . The mobile receiver is located within the scatterer region. To simplify the formulation, the following assumptions are made:

- (i) The angle of arrival at the receiver includes only the azimuthal coordinates.
- (ii) Each scatterer is an omnidirectional reradiating element.
- (iii) Only one bounce is assumed for each propagation path.

5.2.1 Joint TOA/AOA PDF

Figure 5.1: Scatterer geometry

Figure 5.1 shows the geometry of the scatterers around the base station B . The mobile receiver is marked as M (without loss of generality, the mobile receiver is located on the x axis), and the radius of the cell is R . D denotes the distance between the base station and the mobile receiver. A point S in this circle is denoted

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by the polar coordinates (ρ, ϕ) or the rectangular coordinates (x, y) with the origin at the base station. In this chapter, the polar coordinate (ρ, ϕ) is different from what has been defined in Chapter 4. ϕ is conventionally orientated and increases anti-clockwise. (x', y') denotes the coordinate system with the mobile receiver at the origin as shown in Figure 5.1. The correspondence polar coordinates with respect to the rectangular coordinates (x', y') is (r', θ) , where θ is the angle of arrival at the mobile receiver. The PDF of the scatterer around the base station is given by $f_{\rho\phi}(\rho, \phi)$, $f_{xy}(x, y)$, $f_{x'y'}(x', y')$ or $f_{r'\theta}(r', \theta)$ according to the different coordinates described above. With a given $f_{\rho\phi}(\rho, \phi)$, the joint PDF of AOA and TOA can be derived as below.

It is difficult to do the transformation directly from polar coordinates (ρ, ϕ) to (r', θ) since the radial coordinate and the angular coordinate will both change. But if we transform from (x, y) to (x', y') , it will be much easier. The joint PDF of $f_{xy}(x, y)$ is found using [97]

$$f_{xy}(x, y) = |J(x, y)| f_{\rho\phi}(\rho, \phi) \Big|_{\substack{\rho=\sqrt{x^2+y^2} \\ \phi=\tan^{-1}(y/x)}} \quad (5.1)$$

where $|J(x, y)|$ is Jacobian of the inverse transformation given by

$$J(x, y) = \begin{vmatrix} \frac{\partial \rho}{\partial x} & \frac{\partial \rho}{\partial y} \\ \frac{\partial \phi}{\partial x} & \frac{\partial \phi}{\partial y} \end{vmatrix} = \frac{1}{\sqrt{x^2 + y^2}} \quad (5.2)$$

The scatterer density $f_{x'y'}(x', y')$ expressed in (x', y') system is given by

$$f_{x'y'}(x', y') = f_{xy}(x, y) \Big|_{\substack{x=x'-D \\ y=y'}} \quad (5.3)$$

The polar coordinates expression for (5.3) is written as

$$f_{r'\theta}(r', \theta) = |J(r', \theta)| f_{x'y'}(x', y') \Big|_{\substack{x'=r' \cos \theta \\ y'=r' \sin \theta}} \quad (5.4)$$

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where $|J(r', \theta)|$ is the Jacobian of the inverse transformation given by

$$J(r', \theta) = \begin{vmatrix} \frac{\partial x'}{\partial y'} & \frac{\partial x'}{\partial \theta} \\ \frac{\partial r'}{\partial y'} & \frac{\partial r'}{\partial \theta} \end{vmatrix} = r' \quad (5.5)$$

Since the path propagation delay is calculated as

$$\begin{aligned} \tau &= \frac{r' + \rho}{c} \\ &= \frac{1}{c} \left(r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta} \right) \end{aligned} \quad (5.6)$$

then, r' can be solved as

$$r' = \frac{c^2 \tau^2 - D^2}{2(c\tau - D \cos \theta)} \quad (5.7)$$

The joint PDF of TOA/AOA is given by

$$f_{\tau\theta}(\tau, \theta) = |J(\tau, \theta)| f_{r'\theta}(r', \theta) \Big|_{\substack{r'=(c^2\tau^2-D^2)/(2c\tau-2D\cos\theta) \\ \theta=\theta}} \quad (5.8)$$

where $|J(\tau, \theta)|$ is the Jacobian of the inverse transformation given by

$$J(\tau, \theta) = \begin{vmatrix} \frac{\partial r'}{\partial \tau} & \frac{\partial r'}{\partial \theta} \\ \frac{\partial \theta}{\partial \tau} & \frac{\partial \theta}{\partial \theta} \end{vmatrix} = \frac{c^3 \tau^2 + cD^2 - 2c^2 \tau D \cos \theta}{2(c\tau - D \cos \theta)^2} \quad (5.9)$$

5.2.2 Marginal PDF of the AOA and the TOA

Figure 5.2 illustrates the geometry of the angle of arrival for this channel model. θ is the angle of arrival at the mobile receiver and lies on $[-\pi, \pi]$ (with the assumed position of the mobile receiver). It is obvious that the distribution of the arrival angles are symmetrical with respect to 0° . To simplify the process, we start with the distribution for positive angles, by replacing $|\theta|$ with θ , the PDF for the whole AOA can be obtained.

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Figure 5.2: Angle of arrival for the single bounce scatterer model

The line ME expressed in polar coordinates (ρ, ϕ) is as below

$$\rho = \frac{D \tan \theta}{\sin \phi - \tan \theta \cos \phi} \quad (5.10)$$

Let $\angle MBE$ be γ , then

$$r^2 = D^2 + R^2 - 2DR \cos \gamma \quad (5.11)$$

and γ can be obtained as

$$\gamma = \cos^{-1} \left(\frac{D^2 + R^2 - r^2}{2DR} \right) \quad (5.12)$$

where r is the length of line ME and equals to $r = D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}$.

The cumulative distribution function (CDF) of the angle of arrival can now be calculated as the ratio of the weighted area of EMF to the weighted area of the whole circle.

$$F_{\theta}(\theta) = \frac{\int_0^{\pi-\gamma} \int_0^R f_{\rho\phi}(\rho, \phi) \, d\rho d\phi + \int_{\pi-\gamma}^{\pi} \int_0^{\frac{D \tan \theta}{\sin \phi - \tan \theta \cos \phi}} f_{\rho\phi}(\rho, \phi) \, d\rho d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) \, d\rho d\phi} \quad (5.13)$$

The area of EMF is viewed as two parts, one is the sector EBF with angle $\pi - \gamma$, the other is the triangle EBM , which are represented by the two terms

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in the numerator of the expression (5.13). The PDF of the angle of arrival is the derivative of the CDF with respect to θ . To calculate the derivative of (5.13), the following theory [98] is needed:

$$\frac{d}{dx} \int_{u(x)}^{v(x)} f(x, t) dt = v'(x)f(x, v(x)) - u'(x)f(x, u(x)) + \int_{u(x)}^{v(x)} \frac{\partial}{\partial x} f(x, t) dt \quad (5.14)$$

By employing (5.14), the PDF of the AOA is given as

$$\begin{aligned} f_{\theta}(\theta) = & \frac{1}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} \times \\ & \left[-\gamma'_{\theta} \int_0^R f_{\rho\phi}(\rho, \pi - \gamma) d\rho + \gamma'_{\theta} \int_0^{\frac{D \sin |\theta|}{\sin(\gamma + |\theta|)}} f_{\rho\phi}(\rho, \pi - \gamma) d\rho \right. \\ & \left. + \int_{\pi - \gamma}^{\pi} \frac{D \sin \phi}{\sin^2(\phi - |\theta|)} f_{\rho\phi} \left(\frac{D \sin |\theta|}{\sin(\phi - |\theta|)}, \phi \right) d\phi \right] \end{aligned} \quad (5.15)$$

where γ'_{θ} is given by

$$\gamma'_{\theta} = \frac{2rr'_{\theta}}{\sqrt{4D^2R^2 - (D^2 + R^2 - r^2)^2}} \quad (5.16)$$

and

$$r'_{\theta} = -D \sin |\theta| - \frac{D^2 \sin 2|\theta|}{2\sqrt{R^2 - D^2 \sin^2 \theta}} \quad (5.17)$$

Substitute (5.12), (5.16) and (5.17) into (5.15), and after some tedious work, the PDF of the AOA can be simplified as

$$f_{\theta}(\theta) = \frac{\int_{\pi - \gamma}^{\pi} \frac{D \sin \phi}{\sin^2(\phi - |\theta|)} f_{\rho\phi} \left(\frac{D \sin |\theta|}{\sin(\phi - |\theta|)}, \phi \right) d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} \quad (5.18)$$

For the time of arrival, the scatterers contributing the same time delay lie on an ellipse with foci at the base station and the mobile receiver as shown in

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Figure 5.3: Time of arrival for the single bounce scatterer model

Figure 5.3. The major axis of the ellipse is given by $c\tau$, where c is the speed of light. The cumulative distribution function is calculated as the ratio of the weighted intersection area of the ellipse with the scatterer region to the weighted area of the scatterer region as below

$$F_{\tau}(\tau) = \begin{cases} \frac{2 \int_0^{\pi} \int_0^{\frac{c^2\tau^2 - D^2}{2c\tau + 2D \cos \phi}} f_{\rho\phi}(\rho, \phi) d\rho d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} & \tau \leq \frac{2R - D}{c} \\ \frac{2 \int_{\alpha}^{\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi + 2 \int_0^{\alpha} \int_0^{\frac{c^2\tau^2 - D^2}{2c\tau + 2D \cos \phi}} f_{\rho\phi}(\rho, \phi) d\rho d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} & \tau > \frac{2R - D}{c} \end{cases} \quad (5.19)$$

where

$$\rho = \frac{c^2\tau^2 - D^2}{2c\tau + 2D \cos \phi} \quad (5.20)$$

is the expression for the ellipse in Figure 5.3 in polar coordinates system (ρ, ϕ) .

The scatterer region is expressed as below

$$\rho = R \quad (5.21)$$

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The intersection angle α is determined by substituting α for ϕ and solving (5.20) and (5.21) simultaneously

$$\alpha = \cos^{-1} \left(\frac{c^2 \tau^2 - D^2 - 2Rc\tau}{2RD} \right) \quad (5.22)$$

$(2R - D)/c$ is the maximum time delay that keeps the ellipse inside the circle, thus the area of the intersection is the area of the ellipse. When $\tau > (2R - D)/c$, the intersection area becomes part of the ellipse. This area can be viewed as the summation of two parts as shown in Figure 5.3, one is the sector of the circle with angle $2(\pi - \alpha)$, and the other is a part of the ellipse which area can be obtained by integrating in polar coordinate system. The probability density function (PDF) of the time of arrival is the derivative of the CDF with respect to τ given as

$$f_\tau(\tau) = \begin{cases} \frac{2 \int_0^\pi f_{\rho\phi} \left(\frac{c^2 \tau^2 - D^2}{2c\tau + 2D \cos \phi}, \phi \right) \left(\frac{c^2 \tau}{c\tau + D \cos \phi} - \frac{c(c^2 \tau^2 - D^2)}{2(c\tau + D \cos \phi)^2} \right) d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} & \tau \leq \frac{2R - D}{c} \\ \frac{2}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} \left[-\alpha'_\tau \int_0^R f_{\rho\phi}(\rho, \alpha) d\rho + \alpha'_\tau \int_0^{\frac{c^2 \tau^2 - D^2}{2c\tau + 2D \cos \alpha}} f_{\rho\phi}(\rho, \alpha) d\rho \right. \\ \left. + \int_0^\alpha f_{\rho\phi} \left(\frac{c^2 \tau^2 - D^2}{2c\tau + 2D \cos \phi}, \phi \right) \left(\frac{c^2 \tau}{c\tau + D \cos \phi} - \frac{c(c^2 \tau^2 - D^2)}{2(c\tau + D \cos \phi)^2} \right) d\phi \right] & \tau > \frac{2R - D}{c} \end{cases} \quad (5.23)$$

where α'_τ is given by

$$\alpha'_\tau = \frac{Rc - c^2 \tau}{RD\sqrt{1 - \cos^2 \alpha}} \quad (5.24)$$

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By substituting (5.22) and (5.24) into (5.23), the PDF of the TOA can be simplified

as

$$f_{\tau}(\tau) = \begin{cases} \frac{2 \int_0^{\pi} f_{\rho\phi} \left(\frac{c^2\tau^2 - D^2}{2c\tau + 2D \cos \phi}, \phi \right) \left(\frac{c^2\tau}{c\tau + D \cos \phi} - \frac{c(c^2\tau^2 - D^2)}{2(c\tau + D \cos \phi)^2} \right) d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} & \tau \leq \frac{2R - D}{c} \\ \frac{2 \int_0^{\alpha} f_{\rho\phi} \left(\frac{c^2\tau^2 - D^2}{2c\tau + 2D \cos \phi}, \phi \right) \left(\frac{c^2\tau}{c\tau + D \cos \phi} - \frac{c(c^2\tau^2 - D^2)}{2(c\tau + D \cos \phi)^2} \right) d\phi}{\int_0^{2\pi} \int_0^R f_{\rho\phi}(\rho, \phi) d\rho d\phi} & \tau > \frac{2R - D}{c} \end{cases} \quad (5.25)$$

The analysis above is applicable to any distribution of scatterers around the base station. In this chapter, we focus on two non-uniform distributions, i.e., the Rayleigh distribution and the exponential distribution.

5.3 Rayleigh Distribution Scatterer Model

In a typical wireless communication environment, it is reasonable to assume that the scatterer density gradually tapers off with distance from the base station. A zero mean Gaussian distribution with standard deviation σ of the scatterers around the base station is the most often used assumption [88, 92]. This leads to the Rayleigh distribution in the polar coordinate system for the distance between the base station and the scatterers

$$f_\rho(\rho) = \frac{\rho}{N_R \sigma^2} e^{-\frac{\rho^2}{2\sigma^2}} \quad 0 \leq \rho \leq R \quad (5.26)$$

where $N_R = 1 - e^{-\frac{R^2}{2\sigma^2}}$ is the normalization constant since we restrict ρ to $[0, R]$. Strictly speaking, (5.26) is a truncated Rayleigh distribution, but for simplicity, we will call it Rayleigh distribution. Furthermore, since the scatterer distribution is independent of the azimuthal angle, the joint PDF $f_{\rho\phi}(\rho, \phi)$ can be easily obtained as

$$f_{\rho\phi}(\rho, \phi) = \frac{\rho}{2\pi N_R \sigma^2} e^{-\frac{\rho^2}{2\sigma^2}} \quad (5.27)$$

Following the steps in section 5.2.1,

$$f_{xy}(x, y) = \frac{1}{2\pi N_R \sigma^2} e^{-\frac{x^2+y^2}{2\sigma^2}} \quad (5.28)$$

$$f_{x'y'}(x', y') = \frac{1}{2\pi N_R \sigma^2} e^{-\frac{(x'-D)^2+y'^2}{2\sigma^2}} \quad (5.29)$$

$$f_{r'\theta}(r', \theta) = \frac{r'}{2\pi N_R \sigma^2} e^{-\frac{(r' \cos \theta - D)^2 + (r' \sin \theta)^2}{2\sigma^2}} \quad (5.30)$$

Then the joint TOA/AOA PDF is given by

$$f_{\tau\theta}(\tau, \theta) = \frac{(c^3\tau^2 + cD^2 - 2c^2\tau D \cos \theta)(c^2\tau^2 - D^2)}{8\pi N_R \sigma^2 (c\tau - D \cos \theta)^3} e^{-\frac{\left(\frac{(c^2\tau^2 - D^2) \cos \theta}{2c\tau - 2D \cos \theta} - D\right)^2 + \left(\frac{(c^2\tau^2 - D^2) \sin \theta}{2c\tau - 2D \cos \theta}\right)^2}{2\sigma^2}} \quad (5.31)$$

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The joint PDF of TOA/AOA at the mobile receiver for Rayleigh distribution scatterer model is plotted in Figure 5.4. The radius of the cell R is chosen as 1000 m and the distance between the base station and the mobile receiver D is assumed to be $0.5R$. The parameter σ is determined by the environment. We will first discuss the case that $\sigma < R$, a value of $0.5R$ is chosen for a test case. It can be observed from Figure 5.4 that the receiver tends to receive signals with small angle of arrival or short time delay. The probability of receiving signals with large angle of arrival (about 180°) and long time delay is relative low.

Figure 5.4: Joint TOA/AOA PDF for Rayleigh distribution scatterer model

Substitute (5.27) into (5.18), the PDF of the AOA can be obtained as

$$f_{\theta}(\theta) = \frac{\int_{|\theta|}^{\gamma+|\theta|} \frac{\mathbb{C}'_{\theta}}{\sin^2 \phi} e^{\frac{\mathbb{C}}{\sin^2 \phi}} d\phi - e^{-\frac{D^2}{2\sigma^2}} + e^{-\frac{R^2}{2\sigma^2}}}{2\pi \left(e^{-\frac{R^2}{2\sigma^2}} - 1 \right)} \quad (5.32)$$

where $\mathbb{C} = -D^2 \sin^2 \theta / 2\sigma^2$, $\mathbb{C}'_{\theta} = -D^2 \sin 2|\theta| / 2\sigma^2$. The integral in (5.32) can be calculated easily using numerical integration.

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Figure 5.5: PDF for the AOA of multipath components at the mobile receiver for Rayleigh distribution scatterer model, $D/R = 0.5$. The PDF curve for $R = 2000$ m is shifted by 0.002

Figure 5.5 shows a comparison between the theoretical PDF of the AOA at the mobile receiver and the normalized histograms for Rayleigh distribution scatterer model. The radius of the cell is assumed as 1000 m and 2000 m, the distance between the base station and the mobile receiver is $0.5R$. σ is still assumed to be $0.5R$ as above. The simulated histograms are generated by creating scatterers following the Rayleigh distribution (5.27) within the circle, and for each scatterer, the angle of arrival at the mobile receiver is calculated. By carrying out 30 000 trials, the normalized histograms for the AOA can be easily obtained. The simulation takes approximately 20 secs on a desktop PC. From the figure we can see that the theoretical PDF matches the simulated histograms closely. Since σ has been determined as $0.5R$, the PDF of the AOA only depends on the ratio of D/R .

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This can be figured out directly from (5.32),

$$\mathbb{C} = -2 \sin^2 \theta \times \frac{D^2}{R^2} \quad (5.33)$$

$$\mathbb{C}'_{\theta} = -2 \sin 2|\theta| \times \frac{D^2}{R^2} \quad (5.34)$$

(5.32) becomes a equation with the prime parameter D/R . In Figure 5.5, the PDFs are plotted when $R = 1000$ m and $R = 2000$ m, the ratio of the separation distance between the base station and the mobile receiver to the radius of the scatterer region is assumed to be 0.5 for both cases. The two curves actually overlap each other, so in order to get a clear view, we shift the one for $R = 2000$ m by 0.002. From the analysis above, we conclude that, for a given σ , the PDF of the AOA at the mobile receiver depends only on the ratio of D/R , which is the same as what has been observed in the uniform distribution scatterer model.

Figure 5.6: The Gaussian approximation for the PDF of the AOA.

For the AOA, [70, 73] had indicated that the Gaussian distribution fitted the

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PDF of the AOA for outdoor environment best. Figure 5.6 shows that our theoretical PDF is also approximately a Gaussian distribution. The approximation for the theoretical PDF is obtained using least squares fitting. The best fit for our theoretical PDFs when D/R equals to 0.8 and 0.4 are the zero mean Gaussian distributions with standard deviation 32.5° and 61° . It is observed that as the value of D/R increases, the theoretical PDF will approach more closely to the Gaussian distribution. Therefore, the Rayleigh distribution scatterer model can be applied to the outdoor environment.

Figure 5.7: Rms angular spread versus D/R for Rayleigh distribution scatterer model

Figure 5.7 plots the rms angular spread versus D/R . The angular spread decreases monotonically from about 102° to 22° as the value of D/R increases from about 0 to 1. This is because we assume that the scatterer density tapers off with the distance from the base station, as the mobile moves away from the

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base station, the number of the scatterers contributing the larger angle of arrival decreases, hence the angular spread decreases.

The influence of the standard deviation σ upon the PDF of the AOA is displayed in Figure 5.8. The scatterer region R is 1000 m and D/R is selected as 0.5. From

Figure 5.8: The influence of the standard deviation σ upon the PDF of the AOA

the figure we can see that, as σ becomes smaller, i.e., the scatterers are denser near the base station, the arrive angles will be more concentrated in a small range around 0, since the number of the scatterers contributing to the large AOA is relatively small. As σ increases, the distribution of the scatterers becomes less concentrated, and approaches to the uniform distribution eventually. There will be more and more signals arriving at large angles, until finally the PDF of the AOA is equal to that of the uniform distribution scatterer model. In the following paragraph, we will show how the Gaussian distribution scatterer model can be approximated when σ is larger than R .

When $\sigma > R$, (5.27) can be simplified by employing the Maclaurin series [99]

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as below

$$\begin{aligned} f_{\rho\phi}(\rho, \phi) &= \frac{\rho}{2\pi N_R \sigma^2} \left(1 - \frac{\rho^2}{2\sigma^2} + \frac{\rho^4}{8\sigma^4} + \dots \right) \\ &\approx \frac{R^2}{2N_R \sigma^2} \times \frac{\rho}{\pi R^2} \end{aligned} \quad (5.35)$$

where the higher order of ρ/σ is ignored because the value of ρ/σ is smaller than 1. The second term of (5.35) represents the uniform distribution in a rectangular coordinate. The first term of (5.35) is a constant that will not affect the calculation of the PDF of the AOA and the PDF of the TOA. Thus the uniform distribution scatterer model derived in Chapter 4 can be viewed as a special case of the Rayleigh distribution scatterer model when $\sigma > R$. Since the higher order components of ρ/σ are ignored, the larger the σ , the closer the approximation in (5.35). The

Figure 5.9: Comparison between the uniform distribution scatterer model and the Rayleigh distribution scatterer model with large σ . The value of D/R for both models is set to 0.5.

comparison between the uniform distribution scatterer model and the Rayleigh

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distribution scatterer model with larger σ is shown in Figure 5.9. The value of D/R is set to 0.5 for both models. From the figure we can see that when $\sigma = R$, the Rayleigh distribution scatterer model matches the uniform distribution scatterer model closely; as σ increases to $5R$, they almost overlap each other. This validates the approximation in (5.35), the Rayleigh distribution scatterer model can be approximated to the uniform distribution scatterer model when $\sigma > R$, the larger the σ , the closer the approximation.

The PDF of the TOA can be easily obtained by substituting (5.27) into (5.25).

$$f_{\tau}(\tau) = \begin{cases} \frac{\int_0^{\pi} e^{\frac{\mathbb{C}1}{(c\tau + D \cos \phi)^2}} \left(\frac{\mathbb{C}1'_{\tau}}{(c\tau + D \cos \phi)^2} - \frac{2c\mathbb{C}1}{(c\tau + D \cos \phi)^3} \right) d\phi}{\pi \left(e^{-\frac{R^2}{2\sigma^2}} - 1 \right)} & \tau \leq \frac{2R - D}{c} \\ \frac{\int_0^{\alpha} e^{\frac{\mathbb{C}1}{(c\tau + D \cos \phi)^2}} \left(\frac{\mathbb{C}1'_{\tau}}{(c\tau + D \cos \phi)^2} - \frac{2c\mathbb{C}1}{(c\tau + D \cos \phi)^3} \right) d\phi}{\pi \left(e^{-\frac{R^2}{2\sigma^2}} - 1 \right)} & \tau > \frac{2R - D}{c} \end{cases} \quad (5.36)$$

where $\mathbb{C}1 = -(c^2\tau^2 - D^2)^2/8\sigma^2$, $\mathbb{C}1'_{\tau} = -(c^2\tau^2 - D^2)c^2\tau/2\sigma^2$.

The PDF of TOA for Rayleigh distribution scatterer model is plotted in Figure 5.10. The minimum time delay is the time propagating from the base station to the mobile receiver via the LOS path, that is D/c , the maximum time delay for this model is $(2R + D)/c$. The signal transmits from base station B to F , then to the mobile M (see in Figure 5.2). The PDF of TOA at the mobile receiver decreases to zero as the time increases from the minimum path delay to the maximum path delay, no ray will arrive at the receiver after time delay $(2R + D)/c$. The reason for this is that we assume all the effective scatterers are located within a circular region of radius R , the effects of the scatterers outside this circle are

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Figure 5.10: PDF of the TOA for Rayleigh distribution scatterer model, $R = 1000$ m, $D/R = 0.5$, $\sigma = 0.5R$

Figure 5.11: Rms delay spread versus R/D for Rayleigh distribution scatterer model

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neglected.

The delay spread versus R/D for Rayleigh distribution scatterer model is plotted in Figure 5.11, the separation distance between base station and the mobile receiver is fixed at 600 m, and R varies. It is seen that the rms delay spread is approximately proportional to the scatterer region. This is for the reason that, when the scatterer region enlarges, the number of the scatterers contributing to the longer time delay increases as well, hence the delay spread. Meanwhile, if R is given, the rms delay spread will also increase monotonically as the mobile receiver moves away from the base station. For this case, the delay spread increases just because the time of arrival becomes longer as the separation distance between the base station and the mobile receiver increases.

5.4 Exponential Distribution Scatterer Model

In this section, we introduce an (truncated) exponential distribution to model the scatterers around the base station. In this case, the distribution for the distance between the base station and the scatterer in the polar coordinate is assumed to be

$$f_{\rho}(\rho) = \frac{a}{N_E} e^{-a\rho} \quad 0 \leq \rho \leq R \quad (5.37)$$

where $N_E = 1 - e^{-aR}$ is the normalization constant since we restrict ρ to $[0, R]$. a is a constant. The scatterer density $f_{\rho\phi}(\rho, \phi)$ is given by

$$f_{\rho\phi}(\rho, \phi) = \frac{a}{2\pi N_E} e^{-a\rho} \quad (5.38)$$

Following the steps in section 5.2.1, we have

$$f_{xy}(x, y) = \frac{a}{2\pi N_E \sqrt{x^2 + y^2}} e^{-a\sqrt{x^2 + y^2}} \quad (5.39)$$

$$f_{x'y'}(x', y') = \frac{a}{2\pi N_E \sqrt{(x' - D)^2 + y'^2}} e^{-a\sqrt{(x' - D)^2 + y'^2}} \quad (5.40)$$

$$f_{r'\theta}(r', \theta) = \frac{ar'}{2\pi N_E \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} e^{-a\sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \quad (5.41)$$

The joint TOA/AOA PDF is given by

$$f_{\tau\theta}(\tau, \theta) = \frac{c^3\tau^2 + cD^2 - 2c^2\tau D \cos \theta}{2(c\tau - D \cos \theta)^2} \times \frac{ar'}{2\pi N_E \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} e^{-a\sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \quad (5.42)$$

where r' is given by (5.7).

Figure 5.12 shows the joint TOA/AOA PDF for exponential distribution scatterer model. The radius of the cell and the distance between the base station and

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Figure 5.12: Joint TOA/AOA PDF for exponential distribution scatterer model the mobile receiver are chosen as above, i.e., 1000 m and $0.5R$. The parameter a is assumed to be $3/R$. Compared with Figure 5.4, the received signals are mostly with small angle of arrival and short time delay. The probability of receiving signals with large angle of arrival or long time delay is relative low.

By substituting (5.38) into (5.18), the PDF of the AOA can be obtained as

$$f_{\theta}(\theta) = \frac{aD \cos \theta \int_{|\theta|}^{\gamma+|\theta|} \frac{e^{-\frac{aD \sin|\theta|}{\sin \phi}}}{\sin \phi} d\phi + e^{-aD} - e^{-aR}}{2\pi(1 - e^{-aR})} \quad (5.43)$$

The PDF of the AOA for the exponential distribution scatterer model also depends on the ratio of D/R . Since a is a constant and R is known, a can be expressed in the form of $\mathbb{C}_a/R$, where $\mathbb{C}_a = Ra$. Replace a with $\mathbb{C}_a/R$ in (5.43), the whole expression becomes a function of D/R now.

Figure 5.13 shows a comparison between the theoretical PDF of the AOA at the mobile receiver and the normalized histograms for the exponential distribution

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Figure 5.13: PDF for the AOA of multipath components at the mobile receiver for Exponential distribution scatterer model, $D/R = 0.5$. The PDF curve for $R = 2000$ m is shifted by 0.005

scatterer model. The radius of the cell is assumed as 1000 m and 2000 m, the distance between the base station and the mobile receiver is $0.5R$. a is assumed to be $3/R$ as above. Since the separation distance between the base station and the mobile receiver is assumed to be $0.5R$ for both cases, the two curves actually overlap each other. In order to get a clear view, we shift the one for $R = 2000$ m by 0.005. The simulated histograms are generated by creating scatterers following the exponential distribution (5.38) within the cell, and for each scatterer, the angle of arrival at the mobile receiver is calculated. By carrying out 30 000 trials, the normalized histograms for the AOA can be easily obtained. It can be seen that the theoretical PDF matches the simulated histograms very well.

For the exponential distribution scatterer model, the PDF of the AOA plotted in Figure 5.13 is approximately Laplacian shaped. In [78, 79, 100], indoor multipath

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channels has been proposed, and the measurement results show that the zero-mean Laplace distribution is the best fit for the PDF of the AOA. Consequently, we will compare our theoretical results with the measurements in [78, 79, 100] to show that the exponential distribution scatterer model can be applied to the indoor environments. Cramer *et al.*, [79, 100] collected data in an office/laboratory building. In the experiment, the location of the transmitting antenna is fixed and rectangular arrays of measurements are made by moving the receiving antenna to 49 points in a 7×7 array with 6-inch spacing. [79] reported that the relative azimuth arrival angles of the recovered UWB signals were best fit to a Laplacian density, with a standard deviation of 38° . In order to compare with the measurements, the parameters for our theoretical model should be determined first. A value of $3/R$ is found to be the best choice of a for the indoor environment. Since the PDF of the AOA depends on D/R , we only need to determine the value of D/R , not the particular value of D and R . Figure 5.14 is the rms angular spread versus D/R for the exponential distribution scatterer model.

A value of 0.42 is required for D/R to produce the standard deviation 38° in [79]. Substituting a and D/R into (5.43), the theoretical PDF of the AOA at the mobile receiver is obtained. Figure 5.15 is the comparison for the PDF of the AOA between our theoretical result and the measurements reported in [79]. The figure indicates that our theoretical PDF matches the measurements very well. Figure 5.16 compares the theoretical result with the measurements of the ray arrival angles at 5° of resolution reported in [100]. A Laplacian density with $\sigma = 37^\circ$ is observed to fit the measurement result best. In order to get the standard deviation

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Figure 5.14: Rms angular spread versus D/R for the exponential distribution scatterer model

Figure 5.15: Comparison of the theoretical PDF of the AOA for the exponential distribution scatterer model with measurements in Cramer’s paper (measured at 1° of resolution)

Figure 5.16: Comparison between the PDF of the AOA for the exponential distribution scatterer model and the angle of arrival measured at 5° of resolution

37° , D/R is selected as 0.44. These two values are not indicated in Figure 5.14, just to avoid the crowded display of the ticks on the scales. Good agreement has been shown between our theoretical result and the measured data.

Another comparison is made with the measurements in [78]. Spencer *et al.*, [78] collected data at 7 GHz in delay-angle domain with impulse response measurements carried out using narrow beam dish antenna. The experiment was performed within the office buildings on the Brigham Young University (BYU) campus. Clusters of arrivals were observed to exist in angle, we make comparison with the angular data within one cluster with respect to the cluster mean. The angle of arrival in [78] is measured in Clyde Building (CB), and found to fit a Laplace distribution with standard deviation 25.5° . From Figure 5.14 we know that a value of 0.6 for

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Figure 5.17: Comparison of the theoretical PDF of the AOA for the exponential distribution scatterer model with measurements in Spencer's paper

D/R is required to produce a standard deviation of 25.5° . The comparison for the histogram of relative ray arrivals with respect to the cluster mean for CB with the theoretical result is shown in Figure 5.17. The theoretical result also matches the measurement perfectly.

The good agreement between our theoretical results and the measurements made in the office/laboratory environments in Figure 5.15, Figure 5.16 and Figure 5.17 shows that the exponential distribution scatterer model is applicable for the office/laboratory indoor propagation environments. Furthermore, this model performs better than the Laplacian distributions in [78, 79, 100]. The tails of the Laplacian distributions infinitely approach zero, which underestimates the arrival angles with large value. However, in our exponential distribution scatterer model, no such problem exists, the AOA can be predicted very well on $[-\pi, \pi]$. To get a

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clear view, the enlarged plot of the comparisons is shown in Figure 5.18.

Figure 5.18: Enlarged plot of comparisons between theoretical results and measurements

The PDF of the TOA is

$$f_{\tau}(\tau) = \begin{cases} \frac{\int_0^{\pi} e^{\frac{\mathbb{C}2}{c\tau + D \cos \phi}} \left(\frac{\mathbb{C}2'_{\tau}}{c\tau + D \cos \phi} - \frac{c\mathbb{C}2}{(c\tau + D \cos \phi)^2} \right) d\phi}{\pi(e^{-aR} - 1)} & \tau \leq \frac{2R - D}{c} \\ \frac{\int_0^{\alpha} e^{\frac{\mathbb{C}2}{c\tau + D \cos \phi}} \left(\frac{\mathbb{C}2'_{\tau}}{c\tau + D \cos \phi} - \frac{c\mathbb{C}2}{(c\tau + D \cos \phi)^2} \right) d\phi}{\pi(e^{-aR} - 1)} & \tau > \frac{2R - D}{c} \end{cases} \quad (5.44)$$

where $\mathbb{C}2 = -a(c^2\tau^2 - D^2)/2$, $\mathbb{C}2'_{\tau} = -ac^2\tau$. Figure 5.19 is the PDF of the TOA for the exponential distribution scatterer model. The radius of the cell is chosen as 10 m, the ratio D/R and the parameter a are assumed to be 0.5 and $3/R$, respectively. The PDF of the TOA at the mobile receiver decreases monotonously to zero as the time increases from the minimum path delay to the maximum path delay.

[77] had found that the arrival delay PDF fits the exponential decay function. The PDF curve in Figure 5.19 is also observed to be nearly exponentially decayed. Figure 5.20 shows the exponential approximations for the PDF curves. Since the

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Figure 5.19: PDF of the TOA for the exponential distribution scatterer model, $R = 10$ m, $D/R = 0.5$, $a = 3/R$

Figure 5.20: The Exponential approximation for the PDF of the TOA, $R = 10$ m, $a = 3/R$.

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PDF of the TOA approaches infinite near the minimum time delay D/c , we ignore those values. By using the least squares fitting, the exponential approximations are obtained as the dashed lines in the figure. The PDFs of the TOA for $D/R = 0.4$ and $D/R = 0.6$ have been shifted horizontally to make them more recognizable. It can be seen that the PDF of the TOA decays exponentially as a whole, the smaller the D/R , the closer the approximation.

Figure 5.21: Rms delay spread versus R/D for the exponential distribution scatterer model, $D = 6$ m

The rms delay spread for the exponential distribution scatterer model is plotted in Figure 5.21. The delay spread is found to be approximately proportional to the radius of the coverage area R once the separation distance between the base station and the mobile receiver has been given. Meanwhile, if R has been decided, the rms delay spread will increase monotonously as the mobile receiver moves away from the base station. This is the same as what has been observed in Rayleigh distribution

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scatterer model. The reasons are the same as what have been explained in section 5.3.

When the decay constant a is very small, i.e., $a \ll 1/R$, (5.37) can be simplified as

$$\begin{aligned} f_\rho(\rho) &= \frac{a}{N_E} e^{-a\rho} \\ &= \frac{a}{1 - e^{-aR}} \left(1 - a\rho + \frac{a^2 \rho^2}{2} - \frac{a^3 \rho^3}{6} + \dots\right) \approx \frac{a}{1 - e^{-aR}} \end{aligned} \quad (5.45)$$

Assume $a = b/R$, where $0 < b \ll 1$, the above expression can be further rewritten as

$$f_\rho(\rho) \approx \frac{b/R}{1 - e^{-b}} \approx \frac{b/R}{1 - (1 - b)} = \frac{1}{R} \quad (5.46)$$

Thus the $f_{xy}(x, y)$ is obtained as

$$f_{x,y}(x, y) = \frac{1}{2\pi R \sqrt{x^2 + y^2}} \quad (5.47)$$

In such condition, the scatterer density is inversely proportional to the distance from the base station, which is another simple assumption for the scatterer density around the base station.

5.5 Conclusion

In this chapter, geometrically based single bounce scatterer models are derived by assuming that the scatterers are located within a circle around the base station. The radius of the circle is determined by the coverage area of the base station. General formulations of the model and the derivation of the joint PDF TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are outlined. By employing different scatterer distributions, this model can be applied to different propagation environments. Analysis shows that the Rayleigh distribution scatterer model can predict the outdoor environment. The uniform distribution scatterer model derived in Chapter 4 is found to be a special case of the Rayleigh distribution scatterer model. Good agreement between our theoretical results and the measurements indicates that the exponential distribution scatterer model is applicable for the indoor propagation environment, and performs better than the Laplacian distribution fit in predicting the probability density of the arrival angles with large value.

Chapter 6

Power Azimuth Spectrum for Geometrically Based Channel Models

6.1 Introduction

In Chapter 4 and Chapter 5, geometrically based statistical channel models for outdoor and indoor environments were proposed. We focus on the probability density functions of the angle of arrival and the time of arrival. However, in addition to these parameters, power azimuth spectrum is also an important measure, which helps design and optimize the antenna power pattern. In Chapter 4 and Chapter 5, the PDF of the AOA and the PDF of the TOA are derived without considering the transmitted/received power. In this chapter, these geometrically based single bounce scatterer models are further extended to estimate the power azimuth spectrum. The single bounce in each model is assumed to be a specular reflection in order to simplify the analysis. In addition to the usual geometrical parameters, i.e., the radius of the scatterer region, the separation distance between the base station and the mobile receiver, the distribution of the scatterers, etc., the

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carrier frequency of the transmitting signal, the relative permittivity and conductivity of the reflecting surface can be taken into account. They are not considered in Chapter 4 and Chapter 5 since no power is involved during the analysis of the AOA and the TOA

This chapter is organized as follows. Section 6.2 gives the general assumptions of the model and the derivation of the power azimuth spectrum. The PASs for uniform, Rayleigh and exponential distribution scatterer models will be analyzed in section 6.3. Section 6.4 concludes the work of this chapter.

6.2 Power Azimuth Spectrum for Geometrically Based Channel Models

To simplify the derivation of the power azimuth spectrum, following assumptions are made :

- (i) The transmitting and mobile antennas are both omnidirectional and vertically polarized.
- (ii) All the reflecting surfaces are assumed to be smooth.
- (iii) Only one reflection is assumed for each propagation path, multiple reflections may be neglected.
- (iv) No line of sight path between transmitting and mobile antennas are included.

6.2.1 The Power Arriving at the Mobile Receiver

The total received power at the mobile receiver can be calculated as [101]

$$P = \frac{\lambda^2}{8\pi Z_0} |E_r|^2 \quad (6.1)$$

where $Z_0 = 120\pi \Omega$ is the free space impedance, E_r is the electric field received at the mobile, and λ is the wavelength, given by

$$\lambda = \frac{c}{f} \quad (6.2)$$

where c is the speed of the light, and f is the carrier frequency of the transmitted signal. Assume that there are N multipath components arriving at the mobile receiver, E_r is then given by

$$E_r = \sum_{m=1}^N E_0 \frac{e^{-jks_m}}{s_m} \mathbb{R}_m \quad (6.3)$$

where E_0 is the incident electric field intensity at 1 m from the transmitting antenna, s_m is the propagation path length, and $\mathbb{R}_m$ is the reflection coefficient of the m th multipath component, respectively. k is the wave number obtained as

$$k = \frac{2\pi}{\lambda} \quad (6.4)$$

To calculate the received power with arriving angle θ , (6.1) can be rewrite as

$$P(\theta) = \frac{\lambda^2}{8\pi Z_0} |E_r(\theta)|^2 \quad (6.5)$$

$$= \frac{\lambda^2}{8\pi Z_0} \left| \sum_{m=1}^{N_\theta} E_0 \frac{e^{-jks_m}}{s_m} \mathbb{R}_m \right|^2 \quad (6.6)$$

where θ is the angle of arrival at the mobile receiver, $P(\theta)$ describes the arriving power in azimuth angle, $E_r(\theta)$ is the total received field at angle θ , and N_θ is the number of multipath components arriving at the mobile with angle θ .

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6.2.2 The Power Azimuth Spectrum at the Mobile Receiver

Figure 6.1: Power received at the mobile with angle of arrival θ

Figure 6.1 illustrates the geometrically based single bounce scattering model in Chapter 4 and Chapter 5. The base station and the mobile receiver are marked as B and M respectively. θ is the angle of arrival at the mobile receiver. The scatterers are assumed to be distributed around the base station within a circle of radius R . The distance between the base station and the mobile receiver is D . The scatterer distribution around the base station is given by $f_{\rho\phi}(\rho, \phi)$ (refer to the polar coordinate system defined at the base station), or $f_{r'\theta}(r', \theta)$ (refer to the polar coordinate system defined at the mobile receiver).

In Figure 6.1, the transmitted field is departure from B , reflected at an arbitrary point S (denoted by (r', θ) in $r' - \theta$ polar coordinate system), and then reaches the mobile receiver, the received field can be calculated as

$$E_S = E_0 \frac{e^{-jk(r'+\rho)}}{r' + \rho} \mathbb{R}(r', \theta) \quad (6.7)$$

where r' is the distance from the reflection point S to the mobile receiver, ρ is the

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distance from the base station to the reflection point S , and $\mathbb{R}(r', \theta)$ is the reflection coefficient at S . The scatterer distribution at the point S can be expressed by $f_{r'\theta}(r', \theta)$, therefore, the received electric field at the mobile receiver from the scatterer element dr' is

$$\begin{aligned} dE_r &= E_S \times f_{r'\theta}(r', \theta) dr' \\ &= E_0 \frac{e^{-jk(r'+\rho)}}{r' + \rho} \mathbb{R}(r', \theta) f_{r'\theta}(r', \theta) dr' \end{aligned} \quad (6.8)$$

Thus the total received field at the mobile receiver with respect to the arriving angle θ can be obtained by integrating dE_r over line MG , and is given as below

$$E_r^\theta = \int_0^r E_0 \frac{e^{-jk(r'+\rho)}}{r' + \rho} \mathbb{R}(r', \theta) f_{r'\theta}(r', \theta) dr' \quad (6.9)$$

where r is the length of line MG given by

$$r = D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta} \quad (6.10)$$

Substituting (6.9) into (6.5), the power azimuth spectrum is obtained as

$$P(\theta) = \frac{\lambda^2 E_0}{8\pi Z_0} \left| \int_0^r \frac{e^{-jk(r'+\rho)}}{r' + \rho} \mathbb{R}(r', \theta) f_{r'\theta}(r', \theta) dr' \right|^2 \quad (6.11)$$

In (6.11), ρ can be expressed in terms of r' and θ as given below

$$\rho = \sqrt{r'^2 + D^2 - 2r'D \cos \theta} \quad (6.12)$$

Since the transmitted field is assumed to be vertically polarized, the reflection coefficient $\mathbb{R}(r', \theta)$ can be calculated as

$$\mathbb{R}(r', \theta) = R_\perp = \frac{\cos \varphi - \sqrt{\varepsilon - \sin^2 \varphi}}{\cos \varphi + \sqrt{\varepsilon - \sin^2 \varphi}} \quad (6.13)$$

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where φ is the angle of incidence. Since the refraction is specular, the incident angle φ equals half of $\angle BSM$, that is, $\varphi = \psi/2$, and ψ is given by

$$\psi = \cos^{-1} \left(\frac{r'^2 + \rho^2 - D^2}{2r'\rho} \right) \quad (6.14)$$

where ρ is given in (6.12). ε is given by

$$\varepsilon = \varepsilon_r - j60\sigma_r\lambda \quad (6.15)$$

ε_r and σ_r are the relative permittivity and conductivity of the reflecting surfaces.

6.3 Results and Discussion

6.3.1 PAS for Uniform Distribution Scatterer Model

By substituting different scatterer distribution functions for $f_{r'\theta}(r', \theta)$, (6.11) can be applied to different propagation environments. In Chapter 4, the uniform distribution scatterer model is discussed, which assumes that the scatterers are uniformly distributed around the base station. The scatterer distribution is expressed as

$$f_{r'\theta}^{(Uni)}(r', \theta) = \frac{r'}{\pi R^2} \quad (6.16)$$

Substitute (6.16) into (6.11), the power azimuth spectrum for the uniform distribution scatterer model is obtained as

$$P(\theta) = \frac{\lambda^2 E_0}{8\pi Z_0} \left| \int_0^{D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}} \frac{e^{-jk(r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta})} \mathbb{R}(r', \theta)}{r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \times \frac{r'}{\pi R^2} dr' \right|^2 \quad (6.17)$$

Figure 6.2 shows the power azimuth spectrum of the uniform distribution scatterer model. The radius of the scatterer region is assumed to be $R = 1000$ m, the separation distance between the base station and the mobile receiver is $D = 800$ m.

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Figure 6.2: PAS at the mobile receiver for uniform distribution scatterer model, $R = 1000$ m, $D/R = 0.8$, $f = 5$ GHz

Furthermore, the carrier frequency of the transmitted signal is selected as 5 GHz, the relative permittivity ϵ_r and conductivity σ_r are assumed to be 7 and 0.2 S/m following the previous authors [51]. It can be seen that, the power azimuth spectrum is approximately Laplacian shaped. Our calculation shows that, this result is not sensitive to the relative permittivity and conductivity of the reflecting surfaces; a change of ϵ_r from 5 to 15 and σ_r from 0.1 to 1 S/m does not produce significant difference. By using the least squares fitting, the best fit Laplacian distribution is obtained as the superimposed dashed curve in Fig. 6.2. The standard deviation of the Laplacian distribution is 26.5° . The fluctuation of the PAS curve in Fig. 6.2 is caused by the constructive and destructive interference of the phase factor in (6.17), which is usually ignored in other geometrically based theoretical channel models [88].

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Figure 6.3: PAS at the mobile receiver for uniform distribution scatterer model, $R = 1000$ m and 2000 m, $D/R = 0.8$, $f = 5$ GHz

Figure 6.3 compares the PAS of the uniform distribution scatterer model for different scatterer regions. The radius R is chosen as 1000 m and 2000 m, other parameters are same as those in Figure 6.2. It is found that, the power azimuth spectrum depends on the ratio of D/R , not the particular value of D or R . The PAS curves for $R = 1000$ m and $R = 2000$ m are overlapped as a whole, the difference in fluctuation between them is caused by the variation of the phases of the multipath components. Since D/R is fixed, the change of R leads to the change of D , hence the variation in phase factor in (6.17), and ultimately, the amplitude difference between these two PAS curves.

In (6.17), although the reflection coefficient is included, it is observed that this coefficient does not affect the power azimuth spectrum much. Since the received power plotted in the figure is normalized, if we assume the reflection coefficient

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Figure 6.4: PAS at the mobile receiver for uniform distribution scatterer model, $R = 1000$ m, $D/R = 0.8$, $f = 5$ GHz

is a constant, the PAS will have nothing to do with the reflection coefficient. The comparison between the PAS curves with and without reflection coefficient is plotted in Figure 6.4. The two curves almost overlap each other. As a result, we can assume $\mathbb{R}$ is a constant. A more explicit expression is obtained as

$$P(\theta) = \frac{\lambda^2 E_0 \mathbb{R}}{8\pi Z_0} \left| \int_0^{D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}} \frac{e^{-jk(r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta})}}{r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \times \frac{r'}{\pi R^2} dr' \right|^2 \quad (6.18)$$

6.3.2 PAS for Rayleigh Distribution Scatterer Model

In Chapter 5, the Rayleigh distribution scatterer model is discussed, which assumes that the scatterers around the base station follow Rayleigh distribution within a circle of radius R . The scatterer distribution is expressed as

$$f_{r'\theta}^{(Ray)}(r', \theta) = \frac{r'}{2\pi N_R \sigma^2} e^{-\frac{(r' \cos \theta - D)^2 + (r' \sin \theta)^2}{2\sigma^2}} \quad (6.19)$$

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where $N_R = 1 - e^{-\frac{R^2}{2\sigma^2}}$ is the normalization constant since the radius of the scatterer region is restricted to $[0, R]$. Substitute (6.19) into (6.11), the power azimuth spectrum for the Rayleigh distribution scatterer model is obtained as

$$P(\theta) = \frac{\lambda^2 E_0}{8\pi Z_0} \left| \int_0^{D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}} \frac{e^{-jk(r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta})} \mathbb{R}(r', \theta)}{r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \times \frac{r'}{2\pi N_R \sigma^2} e^{-\frac{(r' \cos \theta - D)^2 + (r' \sin \theta)^2}{2\sigma^2}} dr' \right|^2 \quad (6.20)$$

Figure 6.5: PAS at the mobile receiver for Rayleigh distribution scatterer model, $R = 1000$ m, $\sigma = R/2$, $D/R = 0.8$, $f = 5$ GHz

Figure 6.5 shows the PAS of the Rayleigh distribution scatterer model. The radius of the scatterer region is chosen as 1000 m, $\sigma = R/2$, the separation distance between the base station and the mobile receiver is 800 m, and the carrier frequency of the transmitted signal is 5 GHz, furthermore, $\varepsilon_r = 7$ and $\sigma_r = 0.02$ S/m. The power azimuth spectrum for Rayleigh distribution scatterer model is also observed to be approximately Laplacian shaped. The best fit Laplacian distribution is obtained by using the least squares fitting, and has the standard deviation

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of 18.8° .

In [70, 102, 103], the power azimuth spectrum for outdoor urban environments is reported to match the Laplacian distribution best. Therefore, our uniform distribution and Rayleigh distribution scatterer models are applicable to the outdoor environments. Various environments with different scatterer distributions can be modeled by changing the value of the standard deviation σ in (6.19). The received power is highly concentrated around 0° . This indicates that a significant fraction of the received power propagates from the base station to the mobile receiver via rooftop diffractions, which we have discussed in Chapter 3.

Figure 6.6: PAS at the mobile receiver for Rayleigh distribution scatterer model, $R = 1000$ m, $\sigma = R/2$, $f = 5$ GHz

For Rayleigh distribution scatterer model, the power azimuth spectrum also depends on the ratio of D/R . Figure 6.6 shows the comparison between the PAS curves when D/R equals to 0.5 and 0.8, other parameters are the same for these

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two cases. It can be seen that, the received power attenuates more quickly with the increase of the arrival angles for larger value of D/R . This can be explained as, the effective scatterers are restricted to be inside a certain range, as the mobile receiver moves away from the base station, the scatterers contributing large arrival angles will decrease, hence the received power.

Same as in section 6.3.1, the reflection coefficient is not an important factor when calculating the received power spectrum, we can just consider a constant loss for each reflection for the simplicity of the calculation. The rewritten expression will not be displayed here, we can refer to (6.18).

6.3.3 PAS for Exponential Distribution Scatterer Model

In Chapter 5, the exponential distribution scatterer model is discussed as well, which assumes that the scatterers around the base station follow the exponential distribution within a circle of radius R . The scatterer distribution is expressed as

$$f_{r'\theta}^{(Exp)}(r', \theta) = \frac{ar'}{2\pi N_E \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} e^{-a\sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \quad (6.21)$$

where $N_E = 1 - e^{-aR}$ is the normalization constant since the radius of the scatterer region is restricted to $[0, R]$. Substitute (6.21) into (6.11), the power azimuth spectrum for the Rayleigh distribution scatterer model is obtained as

$$P(\theta) = \frac{\lambda^2 E_0}{8\pi Z_0} \left| \int_0^{D \cos \theta + \sqrt{R^2 - D^2 \sin^2 \theta}} \frac{e^{-jk(r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta})} \mathbb{R}(r', \theta)}{r' + \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} \times \frac{ar'}{2\pi N_E \sqrt{r'^2 + D^2 - 2r'D \cos \theta}} e^{-a\sqrt{r'^2 + D^2 - 2r'D \cos \theta}} dr' \right|^2 \quad (6.22)$$

Same as in uniform and Rayleigh distribution scatterer models, the PAS of the exponential distribution scatterer model depends on the ratio of D/R , and

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is not sensitive to the reflection coefficient. Since they have been discussed in section 6.3.1 and 6.3.2, detailed analysis will not be shown here. In Chapter 5, we have mentioned that exponential distribution scatterer model is suitable for indoor environment, in this section, the theoretical PAS will be compared with the measured data.

Figure 6.7: PAS at the mobile receiver for exponential distribution scatterer model compared with the measurement, $R = 50$ m, $D = 15$ m, $f = 5.375$ GHz, $a = 3/R$

Poon [77] took the measurement in a two-floor townhouse and an office of INTEL for LOS and NLOS scenarios. The frequency of measurement is from 2 to 8 GHz with resolution of 3.75 MHz. The average angle power profile in NLOS was measured at 5-5.75 GHz. To compare with the measurement, a mean value of 5.375 GHz is selected. Furthermore, since the separation distance between the transmitter and the receiver varies from 3 m to 25 m, we choose 15 m as an average. The scatterer region is chosen as twice the maximum separation distance, i.e., 50

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m. The value of a is still selected as $3/R$ as in Chapter 5. Since the relative permittivity and the conductivity of the reflecting surfaces are unknown, and the reflection coefficient does not affect the PAS much, we ignore the effect of the reflection coefficient when compared with the measurement. If detailed information is provided, the calculated theoretical result may be more precise. Figure 6.7 shows the comparison between the PAS at the mobile receiver for exponential distribution scatterer model and the measurement taken by Poon [77]. It can be seen that, our theoretical result matches the measured data in the office/laboratory reasonably well, which again demonstrates that the exponential distribution scatterer model is applicable to the office/laboratory indoor propagation environment.

6.4 Conclusion

In this chapter, by assuming each single bounce to be a specular reflection, the power azimuth spectrums for the geometrically based single bounce scatterer models proposed in Chapter 4 and Chapter 5 are derived. The PAS is found to depend on the ratio of D to R ; as the value of D/R increases, the received power attenuates more quickly with the increase of the arrival angles. The PAS curves of the uniform distribution scatterer model and the Rayleigh distribution scatterer model fit the Laplacian distribution best, which has been observed to match the PAS for outdoor environment. The exponential distribution scatterer model is compared with the measurement, the theoretical PAS matches the measured data very well, which further validates the applicability of the exponential distribution scatterer model to the office/laboratory indoor environment.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

Channel modeling plays an important role in wireless communication system design, evaluation and optimization, the objective of this thesis is to derive propagation channel models for mobile communication environments. Our research has been focused on the analytic channel models, which are fast in evaluating, and more flexible in application than the empirical models derived from extensive field measurements.

Path loss is one of the fundamental demands of the channel propagation prediction. We start our thesis with the path loss model for urban cellular communication system. Most of the existing over rooftop multiple diffraction models assume the buildings to be of vanishing thickness [6, 17–21]. In our thesis, the building width is taken into account. Rectangular shaped blocks are used to model the rows of building instead of the traditional half screens. The building width is assumed to be d_m and the inter building spacing is assumed to be $(W_m - d_m)$, where W_m is the

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distance from the left edge of building m to the left edge of building $(m + 1)$. By using the uniform theory of diffraction, closed form expression is derived to calculate the total received field. It is observed that the received field waves up and down with the increase of the distance from the transmitter. In order to get a more explicit expression, average center-to-center spacing of two neighboring buildings d , and average building width d_1 are assumed. The received field is then observed to settle to a nominally constant value once the number of buildings becomes large enough. If we choose W_m and d_m randomly distributed around d and d_1 respectively, the received field of the former case is actually waving around the settled field of the latter one. The settled field is found to be a function of $\alpha\sqrt{d/\lambda}$, where α is the incident angle, λ is the wavelength. The performance of our model has been contrasted with measurements and two reported models, i.e., the physical optics based Walfisch and Bertoni [6] model and the UTD based Neve and Rowe [21] model. Good agreement between our theoretical results with measurements and Walfisch and Bertoni model shows that our UTD model outperforms Neve and Rowe model. Furthermore, since our model is base on UTD, it saves computing time in derivation compared with Walfisch and Bertoni model.

With the advent of smart antenna systems for both interference cancellation and position location applications, in addition to signal power level, the spatial and delay properties of the channel have an enormous impact on the performance of the system. Hence the propagation channel models providing the information about the spatial and delay properties are needed. In Chapter 4, a geometrically based single bounce scatterer model is proposed, which assumes that the scatterers are

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uniformly distributed around the base station within a circle of radius R , where R is determined by the coverage area of the base station. The mobile receiver is located inside the scatterer region, and separated from the base station by a distance of value D . The probability density functions for the angle of arrival and the time of arrival have been derived for both omnidirectional and directional transmitting antenna systems. It is observed that the PDF of the angle of arrival depends only on the ratio of D/R in the omnidirectional transmitting antenna system, the rms angular spread decreases almost linearly as the value of D/R increases. The PDF of the time of arrival depends on both D and R , where D determines the minimum time delay, and together with R , the maximum time delay is settled. The delay spread increases linearly with the radius of the scatterer region once the separation distance D has been fixed. For directional transmitting antenna system, the beamwidth of the transmitting antenna is also a prominent factor for the PDFs of the AOA and TOA. The analysis of the Doppler spectrum shows that, by reducing the beamwidth of the transmitting antenna, the Doppler spread at the mobile receiver will be greatly reduced.

However, in realistic communication systems, simple uniform distribution scatterer model discussed in Chapter 4 is not enough to model the various propagation environments. In Chapter 5, a more general situation is discussed, where the distribution of the scatterers can be arbitrary. The joint PDF of TOA/AOA, the marginal PDF of AOA and the marginal PDF of TOA are derived. By employing different distributions, e.g. Rayleigh distribution, exponential distribution, etc., this model can be applied to different propagation environments. The marginal

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PDF of AOA for Rayleigh distribution scatterer model is approximately a Gaussian function, which has been reported as the distribution of the arrival angles for outdoor environment. The uniform distribution scatterer model proposed in Chapter 4 can be viewed as a special case of the Rayleigh distribution scatterer model when $\sigma > R$. For exponential distribution scatterer model, the PDF of the AOA is approximately Laplacian shaped. Good agreement between the theoretical results and the measurements shows that the exponential distribution scatterer model is applicable to the office/laboratory indoor environment. The PDF of the TOA decays almost exponentially with the increase of time delay.

In Chapter 6, the power azimuth spectrums are analyzed for the geometrically based single bounce scatterer models proposed in Chapter 4 and Chapter 5. The single bounce in each model is assumed to be a specular reflection on a smooth surface for the simplicity of the analysis, multiple reflections are neglected. The transmitting and mobile antennas are both assumed to be omnidirectional and vertically polarized. The uniform and Rayleigh distribution scatterer models are found to be applicable to outdoor environments, since the PAS curves for them are Laplacian shaped, which is reported to be the best fit distribution for the PAS of outdoor environment. The power azimuth spectrum of the exponential distribution scatterer model is compared with the measured profile, the good agreement between them again demonstrates that this model is accurate for the office/laboratory indoor propagation environment.

For a wireless communication system, the path loss helps to determine the transmitting power. Then, the AOA, TOA, angular spread, delay spread, Doppler

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spectrum and PAS help design and optimize antenna parameters such as radiation pattern, array spacing, horizontal beamwidth, gain, etc. Consequently, the system level channel model is obtained.

7.2 Recommendations for Future Work

Although an intensive study on the channel modeling has been carried out in this thesis, there still exists some issues that would be interesting to consider. In this section, we present a number of possible extensions of current work.

- (i) Even though the UTD based path loss model is expected to be generally valid for non-perfectly conducting wedge due to the assumption of application to high frequency cases, it would be interesting if a more detailed research is done to investigate the accuracy of the UTD path loss expressions when diffraction occurs at a low conductivity material wedge.

Figure 7.1: Geometry of the scatterers

- (ii) For the geometrically based channel models proposed in this thesis, all the propagation paths between the transmitting and mobile antennas are as-

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sumed to undergo only one bounce, which may lead to fact that, the time delay of the theoretical results is too short compared with the measurements in urban area. Therefore, it is reasonable to take multiple scattering into account. To include a second bounce, we can further restrict a small range around the mobile receiver as shown in Figure 7.1. All the scatterers inside the circle with radius R may contribute to the first bounce, but only the scatterers inside the circle with radius R' will contribute to the second one, thus a two bounces scatterer model is obtained. The statistical characteristics of the propagation channel may be derived using this model. Furthermore, We may extend this model to include multiple bounce.

- (iii) The geometrically based channel models can be further extended to model the three-dimensional environments. The transmitting and receiving antenna heights, the ground reflections may also be included in these models.
- (iv) In our theoretical models, specified simplifications are inevitable, which may bring in the limitation and inaccuracy when employing these models. In the future work, a computerized tomography may be used as a basis for radio system planning, especially for outdoor downtown environments. Detailed information of the environments is stored, and more accurate results can be obtained for each individual environment.
- (v) There is also a great need for more published measurements with well documented parameters for the geometry of the propagation environments to enable comparisons of theory with measurements.

Author's Publications

Journal Paper

1. L. Jiang and S. Y. Tan, "A simple analytic path loss model for urban cellular communication systems," *J. Electromag. Waves and Appl.*, vol. 18, pp. 1017–1032, Aug. 2004.
2. L. Jiang and S. Y. Tan, "Simple geometrical-based AOA model for mobile communication systems," *IEE Electron. Lett.*, vol. 40, pp. 1203–1205, Sept. 2004.
3. L. Jiang and S. Y. Tan, "Geometrically-based channel model for mobile-communication systems," *Microwave Opt. Technol. Lett.*, vol. 45, pp. 522–528, June 2005.
4. L. Jiang and S. Y. Tan, "Statistical channel models for outdoor and indoor propagation environments," *IEEE Trans. Veh. Technol.*, revised in Jan. 2006.
5. L. Jiang and S. Y. Tan, "Geometrically based power azimuth spectrum model for mobile communication systems," submitted to *IEEE Trans. Wireless Commun.*, 2005.

AUTHOR'S PUBLICATIONS

Conference Paper

1. L. Jiang and S. Y. Tan, "An improved UTD propagation model for urban cellular communication system," in *Proc. IEEE ICICS-PCM*, Singapore, pp. 376–380, Dec. 15-18 2003.
2. L. Jiang and S. Y. Tan, "Geometrical-based propagation model for mobile communication systems," in *International Conference on Microwave and MillimeterWave Technology, ICMMT 2004*, (Beijing, P.R.C.), pp. 834–837, Aug.18-21 2004.

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Appendix A

The Uniform Theory of Diffraction

A.1 The Law of Diffraction

The generalized Fermat’s principle states that an edge diffracted ray between a point S and another point P is a curve that has a stationary optical length among all curves between S and P with one point on the edge. Keller noted that the

Figure A.1: Cone of diffracted rays

diffracted rays propagate along parallel cones with the edge as their common axis. He therefore postulated the law of diffraction as follows [29]: A diffracted ray and the corresponding incident ray make equal angles with the edge at the point of diffraction, provided they are in the same medium. They lie opposite sides of the

APPENDIX A. THE UNIFORM THEORY OF DIFFRACTION

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plane normal to the edge at the point of diffraction.

Figure A.1 illustrates the concept of the diffraction cone. A ray is incident at an angle β'_0 relative to an edge and consequently the diffracted rays form a cone with half-angle β_0 as shown. From the law of diffraction we know that $\beta_0 = \beta'_0$. When the angle $\beta_0 = 90^\circ$, the incident ray is perpendicular to the edge and the diffraction cone degenerates to a disk. This is the two-dimensional case, all rays are by definition perpendicular to the edge; that is, $\sin \beta_0 = 1$.

The diffracted field will, in general, have the following form:

$$E^d(s) = E^i(Q_e) \cdot D \cdot A(s) e^{-jks} \quad (\text{A.1})$$

where $E^i(Q_e)$ is the field incident at the point of diffraction (Q_e) on an edge, D is a dyadic diffraction coefficient, $A(s)$ is a spreading factor, and s is the distance from diffraction point to the field point. Since the entire edge acts as a virtual line source for the diffracted rays, the wave fronts of the diffracted fields are cylindrical and hence the spread factor becomes

$$A(s) = \frac{1}{\sqrt{s}} \quad (\text{A.2})$$

A.2 Shadow Boundaries

Before discussing the diffraction theories, we first take a look at the shadow boundaries. The boundaries controlling as they do the fate of the incident rays are usually called incident shadow boundaries (ISB). Those delineating the regions of space in which reflected rays do or do not exist are called reflection shadow boundaries (RSB).

It is customary to label the two faces of the wedge the *o*-face and *n*-face, respectively. Although the designation of which face is called the *o*-face and which is the *n*-face is in fact arbitrary, as a matter of convention, we will consider the incident angle ϕ' and diffraction angle ϕ to be measured from the *o*-face. The

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wedge is assumed to have an interior angle α , which is related to the parameter n by

$$\alpha = (2 - n)\pi \quad (\text{A.3})$$

or

$$n = \frac{2\pi - \alpha}{\pi} \quad (\text{A.4})$$

with α in radius; the exterior wedge angle is equal to $n\pi$. Figure A.2 shows the

Figure A.2: Incident shadow boundary

cases where the n -face and o -face are shadowed. In general, the incident shadow boundaries are located at

$$|\phi_{ISB} - \phi'| = \pi \quad (\text{A.5})$$

Figure A.3: Reflection shadow boundary

Figure A.3 shows the cases where there are reflections from the o -face and n -face. When the reflection is from o -face, the RSB is located at

$$\phi_{RSB} + \phi' = \pi \quad (\text{A.6})$$

when the reflection is from the n -face, the RSB is located at

$$\phi_{RSB} + \phi' = (2n - 1)\pi \quad (\text{A.7})$$

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A.3 The Uniform Theory of Diffraction

The diffraction coefficients developed by Kouyoumjian and Pathak [28] are more general than Keller's coefficients in the sense that not only do they remain bounded along the shadow boundaries, but they also can be used to calculate the diffraction fields from a wedge with curved surfaces.

A.3.1 Two-Dimensional UTD Diffraction Coefficients

The two-dimensional UTD edge diffraction coefficients are given by [30]

$$D_{\perp,\parallel}(L^i, L^{ro}, L^{rn}, \phi, \phi', n) = D_1 + D_2 + R_{\perp,\parallel}(D_3 + D_4) \quad (\text{A.8})$$

For a conducting wedge, the reflection coefficients are given as

$$R_{\perp,\parallel} = \mp 1 \quad (\text{A.9})$$

L^i , L^{rn} and L^{ro} are the so-called distance parameters. The components of the diffraction coefficients are given by

$$D_1 = \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cot \left[\frac{\pi + (\phi - \phi')}{2n} \right] F \left[kL^i a^+(\phi - \phi') \right] \quad (\text{A.10})$$

$$D_2 = \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cot \left[\frac{\pi - (\phi - \phi')}{2n} \right] F \left[kL^i a^-(\phi - \phi') \right] \quad (\text{A.11})$$

$$D_3 = \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cot \left[\frac{\pi + (\phi + \phi')}{2n} \right] F \left[kL^{rn} a^+(\phi + \phi') \right] \quad (\text{A.12})$$

$$D_4 = \frac{-e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cot \left[\frac{\pi - (\phi + \phi')}{2n} \right] F \left[kL^{ro} a^-(\phi + \phi') \right] \quad (\text{A.13})$$

In the two-dimensional case if the n -face and o -face are both planes, and incident fields are assumed to have cylindrical wavefronts, then the distance parameters are given by

$$L^{rn} = L^{ro} = L^i \quad (\text{A.14})$$

$$L^i = \frac{ss'}{s + s'} \quad (\text{A.15})$$

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where s' is the distance from the line source to the diffraction point and s is the distance from the diffraction point to the field point. For $a^\pm$ in equation (A.10), (A.11) and (A.12).

$$a(\phi \pm \phi') = 2 \cos^2 \left(\frac{\phi \pm \phi'}{2} \right) \quad (\text{A.16})$$

for a^+ in (A.13)

$$a^+(\phi + \phi') = 2 \cos^2 \left[\frac{2\pi n - (\phi + \phi')}{2} \right] \quad (\text{A.17})$$

There are several distinct differences between the uniform diffraction coefficients and Keller's original diffraction coefficients. The most obvious differences are the transition function ($F(x)$) and the fact that now there are four terms instead of Keller's original two [30]. Four terms are needed, as both faces can create incident shadow boundaries and reflection shadow boundaries. The component D_1 is associated with the diffracted field that compensates for the discontinuity in the GO field when the o -face is shadowed; D_2 , when the n -face is shadowed; D_3 , when there is reflection from the n -face; D_4 , when there is reflection from the o -face.

A.3.2 Transition Function

The transition function in the diffraction coefficients is given by [30]

$$F(x) = 2j\sqrt{x}e^{jx} \int_{\sqrt{x}}^{\infty} e^{-ju^2} du \quad (\text{A.18})$$

in which we take the principal (positive) branch of the square root. It is approximated [104] by

$$F(x) \simeq 1 \quad (\text{A.19})$$

for $x > 10$. Also it is appropriate to write

$$F(x) \simeq \sqrt{\pi x} e^{j(\pi/4+x)} \quad (\text{A.20})$$

for $x < 10^{-3}$. In other cases, it is advantageous to write

$$F(x) \simeq \sqrt{\pi x} e^{j(\pi/4+x)} - 2j\sqrt{x}e^{jx} \int_0^{\sqrt{x}} e^{-ju^2} du \quad (\text{A.21})$$

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When $x < 0$, the transition function is evaluated as

$$F(x) = F^*(|x|) \tag{A.22}$$

Figure A.4: Transition function

Figure A.5: UTD diffraction coefficients

The magnitude and the phase of the transition function are plotted in Fig-

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ure A.4. Because the transition function approached zero at the same rate as that at which the cotangent terms in the diffraction coefficients approach infinity at the shadow boundaries, the diffraction coefficients shown in Figure A.5 are finite at the shadow boundaries and hence the diffracted fields remain bounded in the immediate vicinity of the shadow boundaries. This feature makes the UTD uniform.