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Microcellular Mobile Radio Channel Simulation: A Geometrical Optics Approach

By

Syed A. Abbas

**A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirement for the degree of
Master of Engineering**

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
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Microcellular Mobile Radio Channel Simulation: A Geometrical Optics Approach

Submitted By Syed A. Abbas B.Sc(Eng)

in partial fulfillment of the requirements for the degree of
Master of Engineering

Thesis supervisor

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Ottawa-Carleton Institute for Electrical Engineering
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ABSTRACT

This thesis investigates the issues involved in microcellular mobile radio propagation through computer simulation of the propagation mechanism in an urban environment. Propagation, which is understood to be of multipath nature, is caused by a possible line of sight, reflected and diffracted components. A simulation model is developed using geometrical optics. The ray tracing technique which is extensively used in geometrical optics for the design of optical systems is also found suitable to simulate radio propagation environment. Conventional ray tracing, which uses laws of reflection and refraction is extended to incorporate diffracted rays into ray tracing borrowing the concept from geometrical theory of diffraction. The behavior and approximations needed for reflection and diffraction coefficient are discussed. A simulation model is developed and then used to simulate and obtain path loss, fast and slow fading data for a microcellular environment. The simulation results for path loss, slow and fast fading are compared to actual measurements reported in literature. The fast fading component of the path loss is curvefitted for Rayleigh, Rice and Nakagami distribution to compare the three and find the best fit distribution. The nature of near field radio propagation is also investigated from a theoretical standpoint and certain observations are made.

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LIST OF ACRONYMS

AWGN	Additive White Gaussian Noise
BPSK	Binary Phase Shift Keying
<i>cdf</i>	Cumulative Distribution Function
<i>D</i>	Diffraction Coefficient
dB	Decibel
GHz	Giga Hertz
<i>L</i>	Path Loss
LOS	Line of Sight
LTI	Linear Time-Invariant
LTV	Linear Time-Variant
<i>m</i>	Nakagami Distribution Parameter
MHz	Mega Hertz
<i>pdf</i>	Probability Density Function
<i>R</i>	Reflection Coefficient
<i>s</i>	Rice Distribution Parameter

The growing interest in Personal Communication Services has resulted in the reinvestigation of certain issues that are of concern in Mobile Cellular Systems. One issue of particular interest is that of the number of users on the existing resources. Increasing the number of users requires efficient use of allocated bandwidth so that channel capacity is optimally used. The understanding of signal propagation mechanism and consequently the resulting channel model and its characterization plays an important part in determining how many users can be accommodated. In mobile radio systems the channel is not stationary as the propagation mechanism changes from one geographical location to another. This has brought about the classification of the mobile radio propagation environment as urban, suburban and rural. Among these three, the urban environment is the more complex and versatile and is therefore further subdivided into so-called microcells and macro cells (medium and large cells). The microcells are different from the commonly used cells firstly in that they cover relatively small geographical area and secondly that the base station antenna height is much lower (around hundred meter lowered to the lamp post heights). The downtown of the cities are the target environment. In North America, this environment is characterized by the grid structured layout and high-rise buildings (also known as the Manhattan environment). The transmitted signal reaches the receiver through multi-

path. These multipath are composed of a line of sight (LOS) path, if a line of sight exists, ground reflected path, single or multiple reflected paths and diffracted and many combinations of these paths.

This thesis aims to look into different aspects of Microcellular Mobile Radio Propagation Channel. These aspects include understanding the propagation mechanism, a simulation based on that propagation mechanism, path loss characteristics and statistical characterization of the variability in received signal strength which is known to have a rapid variation superimposed on a slowly varying component.

The Ottawa downtown environment is selected for this simulation whose main features are the grid structure, combination of low and high-rise buildings with open parking spaces between them. The results are compared with path loss measurements reported in literature. The simulation tool is derived from geometrical optics which is ray tracing and is extended to incorporate the diffracted multipath from the edges of the buildings. The geometrical theory of diffraction helps determine the diffracted rays.

1.1 Thesis Contributions

A geometrical optics simulation model is developed for microcellular mobile radio propagation. This model includes both reflected and the diffracted multipath. The model is primarily an extension of the existing reflection models. The behavior and approximations involved in determination of reflection and diffraction coefficients are discussed. Effects of transmitter antenna location, height and frequency of operation are investigated. An attempt is made to provide a comparison of Nakagami and Rice/Rayleigh models for small area statistics. The probability of error of a BPSK signal over a Nakagami channel is derived independently. Finally an observation is made about the statistical characterization of the signal in the near field region.

1.2 Thesis Organization

The organization of the thesis is described below.

Chapter Two provides a literature survey on issues under investigation, i.e., path loss, slow and fast fading and radio propagation simulation methods. It is noted in the first section that most popular methods for path loss calculations are not very suitable for micro-

cellular propagation environment. The second section observes that there are two basic approaches for statistical characterization of fast fading, that is, Nakagami and Rice/Rayleigh models. These were independently developed, tested and found to be in close conformity between theory and practice. In this thesis, we test and compare the two theories for microcellular environment, among others. The propagation simulation techniques are surveyed in the last section.

Chapter Three introduces the theory behind communication channel modelling and relates it to geometrical optics. In the first section a general physio-geometric approach of channel modelling is explained. In the second section, ray tracing technique and the geometrical theory of diffraction are reviewed. The behavior of reflection and diffraction coefficients are explained. A geometrical optics simulation model is developed in the last section to simulate the propagation mechanism of the microcellular environment.

Chapter Four presents the simulation results. In the first section simulated path loss results are presented and compared with the measured results. The second section is devoted to the presentation of results on fast fading. Nakagami and Rayleigh/Rice model are compared. In the third section slow fading results and in the last section a plausible interpretation of near field propagation is presented.

Chapter Five discusses the result of this work and forms the conclusions. Applications and recommendations for further work are made. This is followed by a few appendices and references.

State of the Art in Mobile Radio Propagation

The radio propagation phenomenon between a source and a sink is in general highly dependent on the physical environment. In most recent communication systems two scenarios arise. First is that of fixed transmitter and receiver locations which consequently determines the propagation environment. The typical example of such a system is a line of sight microwaves radio relay systems. The second situation is that of mobile communication where either source or sink is mobile. Therefore, this leads into a problem where the environment is not fixed and understanding the propagation mechanism in this case becomes more cumbersome and is generally subdivided into a set of certain typical environments. In this thesis we are primarily concerned with the latter case. Mobile radio propagation is broadly characterized into urban, suburban and rural categories. In all the three categories propagation is understood to have multipath. The number of significant multipath, however, change in the above mentioned three categories. The rural environment, devoid of any major obstruction, typically gives rise to two paths. One path is direct or line of sight (LOS) while the other is a ground reflected one. In the case of the presence of hilly terrain we expect more path depending upon the number of reflectors in proximity to the source or sink. In a suburban environment, we again generally have a LOS and a number of reflected paths. The reflected paths, in general, carry much less energy. This

thesis, however does not delve into the above categories. The focus of this work is on the third type of mobile radio propagation which is urban multipath propagation. The urban multipath propagation has been considered as one of the worst environment with, most often, an absence of a direct path and a very large number of multipath with different energy content as a result of multiple reflections and diffractions from buildings. In modern mobile cellular systems, the urban areas have been subdivided into smaller cells known as microcells for increased user density and spectrum efficiency. This results into a simpler and perhaps tractable environment. This thesis is concerned with the study of microcellular mobile radio channel modelling and simulations.

Parameters of interest while investigating multipath radio propagation include the nature of path loss, statistical characterization of path loss and multipath delay spread. The typical received signal strength curve variations are composed of a slowly varying mean and a more rapidly varying component superimposed on that slowly varying mean. The former is known as slow fading and depends on the large area geography. The latter is termed as fast fading and depends upon the small area geography.

In physical measurements, an unmodulated carrier is generally transmitted at the frequencies of interest. The received signal then contains information about propagation loss and fading statistics. This narrow-band channel sounding technique does not provide information on delay spread in actual experiments. Although, with ray tracing, one can get delay spread with much fine resolution. In contrast, a wide band pulse is transmitted to observe the frequency selectivity and delay spread. Here, we focus our attention on path loss and envelope fading statistics.

2.1 Propagation Loss

Future Personal Communication Systems will rely on the microcellular concept to make efficient use of scarce frequency spectrum and to provide inexpensive infrastructure and small size subscriber's units, [1]. Newly proposed microcellular systems operate over a relatively short radio path (on the order of 200 to 1000 meters) using low antennas (about the height of a lamp post) and low transmitter power (on the order of 10mW). This concept utilizes some of the special characteristics of the radio propagation for small cells to reduce the severe multipath fading and shadowing in a land mobile communication envi-

ronment. Over a relatively shorter propagation path, radio link is arranged to be a clear LOS. This scenario is quite different from the conventional cellular mobile radio where base station antenna is very high and cell size is comparatively large.

Approach in the field strength prediction for a typical mobile cellular system is to provide some basic median field strength curves for one type of service area and then correction factors are given for other types of terrains. The typical terrain types are identified as, urban, suburban, open and isolated mountain areas, rolling hills, sloping terrain and land-sea mixed paths. A list of different propagation models is provided in [2]. Here we are summarizing some of the most widely used.

2.1.1 Free Space

This fundamental theoretical model predicts the losses due to radial spreading for propagation in an ideal region without boundaries. Propagation loss is given by, [2],

$$L = 87 + 20\log D \text{ (dB)} , \quad (2.1)$$

where D is distance between base and mobile antenna in kilometers. The L is median propagation loss in dB's, between half wave dipoles for 900 MHz operating frequency. This model is used as a reference for comparing propagation loss given by other models.

2.1.2 Plane Earth

The plane earth model has been derived theoretically taking into account the presence of an idealized plane earth; with finite conductivity, [3]. For 900 MHz, median propagation loss for a 1.5 m high mobile antenna is given by, [2],

$$L = 111.6 - 20\log h + 40\log D \text{ (dB)} . \quad (2.2)$$

This expression is equal to free space loss when $D = h/17$ (km). The h is the height of the base station antenna in meters. This model is also useful as a reference for comparing other models. Calculated results for land mobile service at 900 MHz are not sensitive to the values chosen for the ground constants. This model is an oversimplification in that it does not include important factors such as terrain profiles, vegetation and buildings.

2.1.3 Bullington's Nomograms

In addition to free space and plane earth effects, the Bullington's nomograms give field strength and median propagation loss as a function of antenna heights for certain idealized

type of intervening terrain, specifically smooth curved earth with: 1) no obstruction; 2) a single sharp ridge; and 3) hills, [4]. The smooth earth and single ridge results come directly from physical theory. The treatment of hills is based on an approximate solution to the theoretical problem of diffraction by two knife edges located at the respective horizons of the transmitting and receiving antennas. The model is an oversimplification since vegetation, and other terrain factors are not accounted for.

2.1.4 Egli

This model consists of empirical formulas that provide a “terrain factor” to be applied to the theoretical plane-earth field strength, [5]. The terrain factor for 900 MHz has a median value of 27.5 dB. The variation of field strength with antenna heights and distance is that of the underlying plane earth model. Median propagation loss for a 1.5 m high mobile antenna is equal to, [2],

$$L = 139.1 - 20\log h + 40\log D \text{ (db)} . \quad (2.3)$$

The Egli’s model is a systematic interpretation of measurements covering a wide span of frequencies (90 to 1000 MHz) with some sacrifice of the accuracy.

2.1.5 Okumura

Okumura gives a method for predicting field strength and service area for a given terrain for frequencies in the range 150-2000 MHz, distance of 1 to 100 km and base station effective antenna heights typical of land mobile applications, [6]. The basic median field strength curve for 900 MHz applies to an urban area. Correction factors are given for suburban, open and isolated mountain areas, rolling hills, sloping terrain and mixed land-sea paths. The method is based on measurements made in Tokyo and surrounding suburbs at 240, 453, 922, 1310, 1430 and 1920 MHz. Statistical analysis of measurements was used to determine distance and frequency dependence of median field strength, location variability and antenna height gain factors. The urban curves with suburban correction factors seem to be most suited for cities in the United States.

2.1.6 Discussion

All of the above mentioned propagation models can not be outrightly applied for micro-cellular radio propagation. There are two basic reasons for this. Firstly, all measurements on which these models are based were taken for conventional mobile cellular systems with

radii greater than 1 km. Secondly, base station antenna heights are much greater than those being proposed for microcellular systems. This results in all of the above being generally inapplicable except free space and plane earth models which again serve as a basis for comparison.

In the past few years, extensive measurements have been made in United States [7], [8], [9], Europe, [11], and Japan [12], for microcellular environment. Most of the experiments performed in downtown environments have shown some general gross feature. The experiments used the base antenna heights of 5 to 8.7 meters and mobile antenna heights of 1.6 to 3.65 meters, substantially lower than the surrounding buildings. At distances close to the transmitter and line of sight propagation loss is proportional to d^{-2} as in free space. At far off distances (distances greater than the break-point discussed next) with line of sight, loss is approximately proportional from d^{-2} to d^{-4} . With no line of sight, a distance dependence of d^{-4} to d^{-8} is observed, respectively, for near and far off distances, [8], [10]. The approximate break point at which the propagation law changes in LOS case is defined as the distance between antennas for which the ground just begins to obstruct the first fresnel zone. The distance in which the first fresnel zone just touches the ground is given by, [7],

$$d = \frac{1}{\lambda} \sqrt{(\Sigma^2 - \Delta^2)^2 - 2(\Sigma^2 + \Delta^2) \left(\frac{\lambda}{2}\right)^2 + \left(\frac{\lambda}{2}\right)^4} \quad (2.4)$$

where $\Sigma = h_b + h_m$ and $\Delta = h_b - h_m$. The h_b and h_m are the base and mobile antenna heights respectively. For higher frequencies, this expression is approximated as a simple function of wavelength and antenna heights as follows, [7], [8],

$$d = \frac{4h_b h_m}{\lambda} \quad (2.5)$$

The observation of break point is manifest only in those measurements which are taken in big downtowns such as New York where streets are several kilometer long. In small downtowns like Ottawa with streets one kilometer or slightly larger this effect is hard to observe and LOS coverage gives rise to free space propagation or d^{-2} propagation law, as observed in [10].

2.2 Fast Fading

Different attempts have been made to comprehend the radio propagation in urban environment. A matter of interest is to obtain knowledge about the random fluctuations in mobile radio reception in small areas or distances. Initial attempts include some experiments by Young, [13], in as early as 1951. Young had found that the Rayleigh distribution gives a good fit to the observed amplitude fluctuations in mobile radio reception at 150, 450, 900 and 3700 MHz in New York city provided that the sample area is less than about 1000 square feet. The other experiments performed by the Jakes, [14], and Rustoko, [15], have verified the above results. Small areas and presence of a direct wave resulted in the Rician distribution as a better fit than the Rayleigh distribution. Ossana, [16], appears to be the first to come forth with a reflection model for mobile radio propagation. However, he did not specifically look into the statistics. Gilbert, [17], using the Ossana's reflection model for multipath fading of mobile radio signal looks into the statistical properties. The Ossana's reflection model assumes that multipath are due to nearby, randomly placed, vertical, plane and good reflectors. Vertical polarization is assumed and vehicle is supposed to move through the standing wave pattern caused by the reflections. Clarke, [18], presents a model which he terms as scattering model. Clark's model assumes that at any point the received field is made up of a number of generally horizontally travelling free-space plane waves, whose azimuthal angles of arrivals occur at random for different positions of receiver and whose phases are completely random such that the phase is rectangularly distributed through out 0 to 2π . The phase and angle of arrival of each component wave is assumed to be statistically independent. In addition, it is assumed that at every point there are exactly N component waves and that these N waves have the same amplitude. The transmitted field is vertically polarized and that polarization is unchanged on scattering so that the received field is also vertically polarized. He also terms this scattered field as "incoherent field", because it's phase is completely random.

Under above assumptions the field components at a point are composed of superposition of N waves, the n th wave arriving at an angle α_n (Figure 2.1),

$$E_z = E_0 \sum_{n=1}^N e^{j\theta_n}. \quad (2.6)$$

FIGURE 2.1

Clarke model's geometry; n th component wave make angle α_n with x-axis

$$H_x = \frac{-E_0}{\eta} \sum_{n=1}^N \sin \alpha_n e^{j\varphi_n}, \quad (2.7)$$

$$H_y = \frac{E_0}{\eta} \sum_{n=1}^N \cos \alpha_n e^{j\varphi_n}. \quad (2.8)$$

In these equations, E_0 is the common(real) amplitude of the N waves, η is the intrinsic impedance of the free space. The time variations are considered to be of the type of $e^{j\omega t}$. Using reference [19], Clarke shows that the envelope of all three field components will be Rayleigh distributed and their phase shall be rectangularly distributed through out $0 \rightarrow 2\pi$.

This approach is based on work by Rice, [20], who has shown that the nature of the envelope of the random noise is Rayleigh distributed. We pursue this concept further because it is of fundamental importance and needs to be clearly understood. We may write noise current (or voltage) $I(t)$ as a fourier series in the limit when $M \rightarrow \infty$, such that:

$$I(t) = \sum_{m=1}^M [a_m \cos \omega_m t + b_m \sin \omega_m t], \quad (2.9)$$

where a_m and b_m are assumed to be independent and normally distributed random variables. Another way to write noise current is:

$$I(t) = \sum_{m=1}^M c_m \cos (\omega_m t - \varphi_m), \quad (2.10)$$

where it is being assumed that c_m is constant and φ_m is a random variable and $\varphi_1, \varphi_2, \varphi_3, \dots$ are distributed at random between $0 \rightarrow 2\pi$.

Considering the noise current flowing in the output of a relatively narrow band filter which has the character of a sine wave of, roughly, the midband frequency whose amplitude fluctuates irregularly, the rapidity of fluctuations being of the order of bandwidth. To study the fluctuations of the envelope of such a wave, we first define the envelope. Let f_n be a representative midband frequency. Then if $\omega_n = 2\pi f_n$, the noise current may be represented by:

$$I(t) = \sum_{m=1}^M c_m \cos (\omega_m t - \omega_n t - \varphi_m + \omega_n t) = I_c \cos \omega_n t + I_s \sin \omega_n t. \quad (2.11)$$

Where,

$$I_c = \sum_{m=1}^M c_m \cos(\omega_m t - \omega_n t - \phi_m), \quad (2.12)$$

$$I_s = \sum_{m=1}^M c_m \sin(\omega_m t - \omega_n t - \phi_m). \quad (2.13)$$

Envelope R is a function of both time and random phase and defined by,

$$R = [I_c^2 + I_s^2]^{1/2}, \quad (2.14)$$

Both I_c and I_s are random variables. The central limit theorem implies that I_c and I_s should be normally distributed with zero mean and same variance, a consequence of the fact that $I(t)$ is a zero mean random process. They are independent since $\overline{I_c I_s} = 0$ (orthogonality). Consequently, the probability that point (I_c, I_s) lies within the elementary triangle $dI_c dI_s$ is,

$$\frac{dI_c dI_s}{2\pi\Omega^2} \exp\left[-\frac{(I_c^2 + I_s^2)}{2\Omega^2}\right], \quad (2.15)$$

where Ω^2 is the variance. It is convenient to introduce another variable θ , where,

$$I_c = R \cos \theta, \quad (2.16)$$

$$I_s = R \sin \theta. \quad (2.17)$$

Since I_c and I_s are random variables so are R and θ . The differentials are related by $dI_c dI_s = R d\theta dR$, and the distribution functions of R and θ are obtainable from (2.15) when the change of variables is made. i.e.,

$$\frac{d\theta}{2\pi} \frac{R dR}{\Omega^2} e^{-\frac{R^2}{2\Omega^2}}. \quad (2.18)$$

Since this may be expressed as a product of terms involving R and θ only, R and θ are independent random variables, θ being uniformly distributed between $0 \rightarrow 2\pi$ and R , the value of envelope, having the probability density,

$$p(R) = \frac{R}{\Omega^2} e^{-\frac{R^2}{2\Omega^2}}. \quad (2.19)$$

The (2.6)-(2.8) may in fact be written as the functional form of (2.10). For example multiplying (2.6) by $e^{j\omega t}$, collecting constant term under one constant c_m and taking the real

part shall yield the right hand side of (2.10). Consider transmitting an unmodulated narrow-band carrier. The multipath of different lengths result into different amplitudes and phases of the arriving multipath at the receiver. Thus E_r becomes a random variable and envelope probability density function then may be shown similarly to be as Rayleigh distributed.

Although, Clarke's theoretical model is one of the most popular models. However, when looked at from a physically motivated channel modelling approach, it is found to be based on quite a few unrealistic assumptions. For example, I_c and I_s are assumed to have zero mean which may be hardly true in case of an actual channel. The other fact is that to show I_c and I_s are normally distributed we assume to have infinite or very large component waves or multipath. The significant multipath are generally very few but we may reasonably assume large N multipath if we allow the multipath to have unequal amplitudes. Physically these multipath are due to multiple reflection, diffraction or combinations of both. This would, however, result into non-zero mean for both random variables I_c and I_s . In this case the envelope R shall be Rician distributed which is written as,

$$p(R) = \frac{R}{\Omega^2} e^{-\frac{R^2 + s^2}{2\sigma^2}} I_0\left(\frac{Rs}{\Omega^2}\right) \quad ; (R > 0). \quad (2.20)$$

If we perform the transformation as shown in [21], the same results are to be obtained if there is a strong LOS component present between transmitter and receiver. This has been verified while analyzing experimental data, [22], [23], for those environment where a strong specular component was present.

Summarizing at this point, we observe that while analyzing the scattered field characteristics the following approximations are supposed to be valid, [24];

- large number of partial waves
- identical component waves amplitudes
- no correlation between different component waves
- no correlation between phase and amplitude of any component waves
- phase distribution uniform between $0 \rightarrow 2\pi$

Physically, these approximations are equivalent to homogeneous diffused scattering field resulting from randomly distributed point scatterers and lead to application of central

limit theorem giving a complex Gaussian process. This picture, however, is not representative of the actual communication channels for component waves have different amplitudes and may not be sufficiently large to make central limit theorem applicable, as mentioned earlier. Moreover, correlation between amplitude and phase of same or different component waves may not be ruled out. Thus need is to investigate the fading or random phase and amplitude problem under the most general conditions.

Nakagami, [25], has considered this problem in the above mentioned more general conditions. He gives an approximate solution to the above problem. This solution now known as Nakagami distribution was originally developed for fast fading processes in ionospheric and tropospheric propagation. It has also been reported as a fitting model for mobile radio channel, [24], [26]. In the following we develop the necessary mathematics required to arrive at Nakagami distribution while considering fast fading on the general basis as reported in [24], [25].

It is supposed that the signal intensity ξ at an observing point is composed of some component fields $E_n = E_{0,n} \exp [j\phi_n]$; ($n = 1, 2 \dots n$) which have travelled on different paths and whose amplitudes and phases vary according to certain unknown statistical law. Under these conditions, ξ may generally be written as;

$$\xi = \left| \sum_{n=1}^N E_{0,n} \exp [j\phi_n] \right| = |x + jy|. \quad (2.21)$$

Let $\xi(E_n)$ be a positive definitive function of random variables E_n where ($n = 1, 2 \dots n$). Then distribution of envelope R , equal to ξ , can be formally expressed as,

$$p(R) = \overline{\delta(R - \xi)}. \quad (2.22)$$

where $\delta(R)$ denotes the delta function after Dirac, and bar means the average with respect to random variables. The delta function in (2.22) can be written as a Hankel Transformation,

$$\delta(R - \xi) = R^{\nu+1} \int_0^\infty d\lambda \lambda \frac{J_\nu(\lambda R) J_\nu(\lambda \xi)}{\xi^\nu}. \quad (2.23)$$

where J_ν is the Bessel function of arbitrary complex order ν with $Re \nu \geq -1/2$. We may re-write equation (2.23) as,

$$p(R) = \frac{R^{\nu+1}}{2^{\nu}\Gamma(\nu+1)} \int_0^{\infty} d\lambda \lambda^{\nu+1} J_{\nu}(\lambda R) F_{\nu}(\lambda), \quad (2.24)$$

with,

$$F_{\nu}(\lambda) = 2^{\nu}\Gamma(\nu+1) \overline{\left(\frac{J_{\nu}(\lambda \xi)}{\lambda^{\nu} \xi^{\nu}} \right)}, \quad (2.25)$$

which may be expanded in powers of $\lambda \xi$ leading to,

$$F_{\nu}(\lambda) = \Gamma(\nu+1) \sum_{k=0}^{\infty} \frac{(-1)^k \left(\frac{\lambda}{2}\right)^{2k} \overline{\xi}^{2k}}{k! \Gamma(\nu+k+1)}. \quad (2.26)$$

Nakagami approximates $F_{\nu}(\lambda)$ as a function of two parameters dependent upon the moments $\overline{\xi^2}$ and $\overline{\xi^4}$, i.e.,

$$m = \frac{\overline{\xi^2}^2}{(\overline{\xi^2} - \overline{\xi^4})^2}, \quad (2.27)$$

and,

$$\Omega = \overline{\xi^2}, \quad (2.28)$$

by invoking a decoupling approximation for moments of order higher than four by replacing the Lyapunov inequality $\overline{\xi^{2k+2}} \geq \overline{\xi^{2k}} \overline{\xi^2}$ by an equality and expressing these moments in terms of $\overline{\xi^2}$ and $\overline{\xi^4}$, i.e., Ω and m . Choosing ν , which is an arbitrary complex number in (2.23) ($\text{Re } \nu \geq -1/2$), as satisfying $\nu = m - 1$ yields,

$$F_{\nu}(\lambda) \approx e^{(-\Omega \lambda^2)/(4(\nu+1))}, \quad (2.29)$$

and substituting this into (2.24),

$$p(R) \approx R^m \int_0^{\infty} d\lambda \lambda^m J_{m-1}(\lambda R) e^{-(\Omega/(4m))\lambda^2}. \quad (2.30)$$

$$p(R) = \frac{2m^m R^{2m-1}}{\Gamma(m) \Omega^m} e^{-(m/\Omega)R^2} \quad ; (R > 0, m \geq \frac{1}{2}). \quad (2.31)$$

the Nakagami m -distribution with the two parameters m and Ω .

In scattering processes generating merely diffuse wavefield, $m = 1$ and the Nakagami distribution is identical to Rayleigh distribution. In case of a direct component, the Nakagami m -distribution approximates the Rice distribution with $m > 1$. The simple algebraic form of m and Ω allows a straightforward extraction of these parameters from measured data. In narrow band measurements usually many scattered components are superimposed and m is slightly greater than one even if the direct component is missing.

Having said this, the natural question to be asked is that "how Nakagami, Rayleigh and Rice compare to each other for the same propagation environment". This is one of the questions we seek to answer in this thesis based on microcellular mobile radio propagation environment.

2.3 Slow Fading

Slow fading statistics are needed to characterize slowly varying component or large area statistics of the received signal envelope. As opposed to fast fading statistics, these statistics are reported to be lognormally distributed nearly unanimously, [25], [26]. Suzuki, [26], showed both Nakagami and log-normal distributions as a good fit for envelope fading statistics even for small area. But those areas were yet quite big to make lognormal distribution as a good fit. When median path loss is in decibels, then functional form reduces to that of a normal distribution as given below,

$$p(R) = \frac{1}{\sqrt{2\pi}\sigma_R} e^{-\frac{(R - m_R)^2}{2\sigma_R^2}}, \quad (2.32)$$

where R denote path loss in terms of dB's. The m_R and σ_R^2 are mean and variance in dB's.

2.4 Mobile Radio Propagation Simulation

It is highly desirable to develop tools for simulation of mobile radio propagation. It is due to the fact that propagation is not stationary and a reasonable simulation can help by giving extensive information about path loss in the area to be covered, shadow regions, and statistical character of the received signal envelope. To simulate the propagation, we need to know how this propagation channel may be modelled. Based on that channel model, one may simulate the propagation environment and then obtain the desired informa-

tion. In this section we review the state of the art in mobile radio channel modelling. We again emphasize here that unique channel models are unrealistic due to changing environment in mobile communication.

2.4.1 Non-Urban Environment

Non-urban environment here includes both rural and suburban environments. In rural environment, a two ray model is considered sufficient for simulation purposes. These two rays are due to a LOS ray and a ground reflected ray as shown in Figure 2.2. Simplicity of environment results into this representation. In a mountainous region we may need more than two rays. The number of extra rays depend on position of receiver and presence of ridges account for extra reflected and/or diffracted rays.

In a suburban environment comprising of residential area with wide streets, the exact number of rays is generally difficult to determine. Whatever the number of rays, those are understood to be due to a possible LOS, reflected, edge diffracted and diffracted-reflected rays.

2.4.2 Urban Environment

In urban environment, for a classical mobile radio system, the number of rays increase tremendously. A tractable channel model may result if conventional cellular systems are broken down into smaller microcellular systems. A few of these are mentioned here. For a non-microcellular environment one of the models for theoretical predictions of mean field strength, [27], suggests three modes of propagation. They are ray-theoretical modes (D and R, diffracted and reflected by buildings), scattered mode (S), and guided mode (G) as shown in Figure 2.3. In, [9], a six ray model as shown in Figure 2.4 has been presented for LOS microcellular and personal communications. Six rays are due to a LOS (ray #1), ground reflection (ray #2) and single and double reflections (ray #3-6) on both sides of the streets. It may be noted that diffraction due to edges of the buildings is neglected from this model. In fact the inclusion of diffracted rays depend upon the scope of the simulation. It is observed that for determination of path loss exclusion of diffracted paths may not effect the result much. In general a model is desired to be simple with minimum number of rays. We look into this aspect of microcellular channel in this thesis.

FIGURE 2.2

A two ray Multipath propagation model for a non-urban environment

FIGURE 2.3

Multipath propagation model for a non-microcellular urban environment, after reference [27]

FIGURE 2.4

Multipath propagation model for a microcellular environment, after reference [9]

2.4.3 Ray Tracing for Multipath Channel Simulation

The use of ray tracing for multipath radio channel simulation is becoming very popular nowadays. It has been extensively used for portable indoor radio channel simulation, [28], [29]. It is also reported to have been used for the outdoor portable communication systems [30]. In indoor environment the conventional ray tracing is used to its full capability to account for the reflection and refraction through walls. However, it is not possible to account for the diffraction and extension of conventional ray tracing is required to account for diffraction. In outdoor environment the refraction of rays into the building is of not much significance. To apply ray tracing to outdoor environment, therefore, one needs some means to account for diffraction to yield reasonable results. One approach is presented in [30].

Next Chapter is devoted to the development of a general channel model and its applicability for geometrical optics. The simulation model is then developed and possible trade off are discussed.

Communication Channel Modelling and Geometrical Optic

In designing a reliable communication system, precise knowledge of the propagation medium is highly desirable. Different propagation media have been used for information transmission depending upon physical constraints, required reliability, economy and above all, user requirements. We may broadly divide communication channels into wireless and wire channels. Wire channels have a physical connection between transmitter and receiver, for example, open wire, coaxial cable, wave guides and optical fibre transmission systems. On the other hand, wireless channels being more flexible are invariably radio channel using free space as propagation medium, with an exception of sonars, which use water as propagation medium.

The transmission characteristics of some of the above systems may have enormous differences. But still one may find certain characteristics which may be considered common to all. For example, a certain input power is injected at the transmitting station and an output power is received at the receiving station. It has also been observed that for some practical purposes, majority of channels may be considered linear. They do have nonlinearities but are usually quite low to play a significant role and are generally ignored. The received power has some undesirable power component called noise. This additive interference is present even if no input is applied, which shows the independence of this

highly undesirable feature of the channel to that of the applied signal. The transmitted signal while passing through the channel experiences attenuation and requires finite time to reach the destination. This time delay and loss may be fixed or time-variant. On the contrary, both or one of them may be dependent on certain physical parameters of channel and time-variant in nature. These characteristics of channel suggest that it may be represented by a linear time-variant network or filter with two or more terminal pairs. This representation is very general in nature and is applicable on any type of channel. Therefore, it is representative of radio channels in general and mobile radio channel in particular.

The need to characterize a channel by a generalized mathematical model emanates from the fact that it is of high interest to determine the effect of propagation medium on the transmitted signal. This knowledge is ultimately used to design signals which are less affected by channel impairments and an optimum receiver which consequently improves the degree of reliability and that how much power must be delivered.

3.1 A General Model

The simplest type of channel may be represented as shown in Figure 3.1. Signal $v_i(t)$ is applied at the input of the channel, which is modified by the channel and $v_o(t)$ is received at the output. Mathematically, $v_o(t)$ may be written as:

$$v_o(t) = A(t) v_i[t - \tau(t)] + v_N(t), \quad (3.1)$$

which means that channel operates on the input signal in such a fashion that a delayed and multiplied version is received with some undesirable signal $v_N(t)$ added to it. Multiplicative constant $A(t)$ and delay $\tau(t)$ may be constant or time-variant over an observation period.

This simple representation may be useful in certain cases but for lots of other a simple model may not be representative. For example, in a diversity reception system, more than one signal may be transmitted or received. Therefore this simple representation of the channel may be extended by a multi-input/multi-output linear time-variant network, as shown in Figure 3.2.

A system of equations describing the above channel may be as follows:

FIGURE 3.1

A simple channel model

FIGURE 3.2

Multi-input multi-output channel model

nally observable effect of whatever causal propagation mechanism are at work is analyzed. Although these effects are supposedly computable from the knowledge of the mechanism established from the physio-geometric approach, in practice, this is not always possible because of the requirement of a detailed knowledge of the causal forces and inability to predict the detailed interaction. The problem is simplified by resorting to the concept of random processes and channel is modelled as a time-variant filter. This channel is represented by randomly time-variant system function. This approach provides a parametric representation of the channel generally required for signal design and performance analysis purposes. Then analysis of transmission performance is then aimed to determine the channel parameters. These parameters may be identified from equation (3.1) as the multiplicative disturbances $K_{nm}(t)$, delays $\tau_{nm}(t)$ and the additive disturbances $v_N(t)$. Because all of the above three parameters are the consequence of a very complex physical process, they are characterized as a random process.

This thesis uses the physio-geometric approach of the theory of communication channel to investigate the channel parameters for microcellular mobile radio channel through simulation using geometrical optics. Review of the systems-function approach is presented in Appendix A.

It was mentioned earlier that in physio-geometric approach, the main attention is focussed on the propagation mechanism. From this standpoint, one of the system of equations in (3.2) may be modelled as shown in Figure 3.3.

In the next section we review the geometrical optics. It would be shown later that channel models, depicted in Figures 3.3 can be transformed into the geometrical optics term.

3.2 Geometrical Optics

Geometrical Optics, [32], [33], is the study of the behavior and properties of light based on an idealized nature of light. A ray of light is a physical abstraction in the sense that it represents only a part or aspect of physical reality. Geometrical optics is an approximation to more detailed and complete theories of light and, in fact, is the chief tool used in the design of optical systems and instruments. Geometrical optics arise as a result of a series of approximations and simplifications of the most rigorous theory.

FIGURE 3.3**A tapped delay line model**

According to modern views, light is an electromagnetic wave of the same kind as wireless waves but with shorter wavelengths. In fact a complete range of spectrum of electromagnetic radiations is known of having wavelengths several thousands of meters down to 10^{-13} meters. Geometrical optics is in principle applicable to all of these radiations but in practice lenses, mirrors and prisms are used mainly between the range from the infra-red to ultra-violet, commonly known as light. The propagation of the electromagnetic waves is known to form surfaces of constant phases. The direction of propagation of wave at a point in space around the source of electromagnetic wave is considered to be normal to surface of constant phase. We can arrive at the concept of ray by considering the behavior of waves as wavelength is made smaller and smaller. The other concept in geometrical optics is that of geometrical wavefronts. The geometrical wavefronts are surfaces in the limiting process of the surfaces of constant phase in the wave motion as the wavelength tends to zero and for convenience are simply called wavefronts, though geometrical optics is strictly not considered with waves. To recapitulate, rays are normal to wavefronts and wavefronts are surfaces orthogonal to rays. In the next section we look into ray tracing from the perspective of optical as well as radio systems.

3.2.1 Ray Tracing

Ray tracing is the single most important tool used in optical design and Snell's law is the main ingredient in formulating procedures for ray tracing. However, it should be noted that Snell's law is a special case of Fermat's principle.

To trace a ray in an optical system an object point is selected and a ray direction defined. Next, the point where the ray intersect the first surface of the optical system is found. At that point normal to the surface is calculated. This normal and the incident ray determine the plane of incidence. The angle of incidence " i " is the angle between incident ray and the surface normal. Knowing the refractive index and using Snell's law one determines the direction of the refracted ray. The reflected ray is determined in the medium containing incident ray making angle " i " with the surface normal on the other side of the incident ray. Starting with the point of intersection of the ray with the first surface and knowing the direction of reflection and refraction from that point one finds the point of intersection of these two daughter rays with next surface. One repeats the process until one runs out of surfaces.

From above, one can observe that in ray tracing two operations are involved. First determines the direction of reflected and refracted rays, while second involves finding the point of intersection of a ray with the surface. The input to the first operation are: (i) direction of surface normal; (ii) two refractive indices; (iii) the direction of incident ray. While input to the second operations are: (i) the shape of optical surface; (ii) the position of the starting point of the ray relative to this surface; (iii) the ray's direction.

Ray tracing, [28], approximates the scattering of electromagnetic waves by simple reflection and refractions and can be much less demanding of computations but provides much more details such as path loss, delay spread and fast and slow fading. Application of ray tracing in radio propagation systems requires certain approximations. The accuracy of ray tracing depends mainly on the ratio of the wavelength to the dimension of scatterer and the volume of interest. Ray tracing is most accurate when the point of observation is many wavelengths away from the nearest scatterer and all scatterers are large compared to a wavelength and smooth i.e., have surface features that are small compared to a wavelength. If ray tracing is to be used to describe the microcellular radio propagation mechanism, ray tracing must account for the diffraction of rays at the edges and corners of the buildings. We find that it can be done through borrowing ideas from Keller's Geometrical Theory of Diffraction which is to be introduced in Section 3.2.3.

The basic idea can be understood with the help of Figure 3.4. It shows a source T, a sink R, two reflecting surfaces A & B and various images labelled by appending the sequence of reflecting surfaces that generated them. The straight line distance between an image and receiver is equal in length to the corresponding actually reflected ray paths. For example, the line TxyR has the same length as straight line from image I_{AB} to R. It is desirable to generate the images of a source object in all the planes which help computing 2nd round of reflections and so forth. For N reflecting planes, there would be N first order, $N(N-1)$ second order or two reflection images, $N(N-1)(N-1)$ three reflection images etc., which shows that incorporation of multiple reflections of rays increase the computation time nearly exponentially. It is, therefore, desirable to incorporate only as many multiple reflections as may be essential and others be rejected. One priority may be to use receiver noise threshold as criteria and any ray weaker than this threshold may be discarded. Moreover, all images of the source may not be visible at the destination. To determine whether an im-

FIGURE 3.4 Illustration of ray tracing, ray source T , sink R and two reflectors A and B

age of the source is visible at the destination is to determine whether the corresponding reflected ray intersects all the necessary surfaces and thus, reaches destination and contributes in the received power. Using above procedure all possible multipath and their lengths are enumerated. The phasor summation of these multipath yield the received signal. To determine the strength of the reflected paths, the knowledge about the Reflectivity is important. It is described in the next subsection.

3.2.2 Reflection Coefficient

The reflection coefficient R is complex in general and given by the following equation, [34],

$$R_v = \frac{\tilde{\epsilon}\tilde{\mu}\cos\theta - \sqrt{\tilde{\epsilon}\tilde{\mu} - (\sin\theta)^2}}{\tilde{\epsilon}\tilde{\mu}\cos\theta + \sqrt{\tilde{\epsilon}\tilde{\mu} - (\sin\theta)^2}}, \quad (3.4)$$

if incident E field lie in the plane of incidence (vertical polarization), and for incident E field perpendicular to the plane of incidence (horizontal polarization) by,

$$R_h = \frac{\cos\theta - \sqrt{\tilde{\epsilon}\tilde{\mu} - (\sin\theta)^2}}{\cos\theta + \sqrt{\tilde{\epsilon}\tilde{\mu} - (\sin\theta)^2}}, \quad (3.5)$$

here θ is the angle of incidence and,

$$\tilde{\epsilon}\tilde{\mu} = \frac{\epsilon_2\mu_2}{\epsilon_1\mu_1}, \quad (3.6)$$

where $\epsilon = \epsilon_0\epsilon_r$, and $\mu = \mu_0\mu_r$.

In most applications the two adjoining medias are non-magnetic with $\mu_1 = \mu_2 = \mu_0$; moreover, in most applications media one is free space or air with $\epsilon_1 = \epsilon_0$, so that (3.4) and (3.5) reduce, respectively, to

$$R_v = \frac{\tilde{\epsilon}\cos\theta - \sqrt{\tilde{\epsilon} - (\sin\theta)^2}}{\tilde{\epsilon}\cos\theta + \sqrt{\tilde{\epsilon} - (\sin\theta)^2}}, \quad (3.7)$$

$$R_h = \frac{\cos\theta - \sqrt{\tilde{\epsilon} - (\sin\theta)^2}}{\cos\theta + \sqrt{\tilde{\epsilon} - (\sin\theta)^2}}, \quad (3.8)$$

where,

$$\tilde{\epsilon} = \frac{\epsilon}{\epsilon_0} = k + j60\lambda\sigma, \quad (3.9)$$

is the relative complex permittivity (dielectric constant) of the medium two where k is its real part. If the two media are non-magnetic, but medium one is not vacuum, then (4.6) and (4.7) will hold if $\tilde{\epsilon}$ is replaced by ϵ/ϵ_1 .

For a perfect conductor, $\sigma \rightarrow \infty$, hence $\tilde{\epsilon} \rightarrow \infty$, so that (3.7) and (3.8) will yield;

$$R_v = 1 \quad R_h = -1. \quad (3.10)$$

The reflection coefficients in (3.7) and (3.8) are generally complex. From the definition, it follows that the absolute value of R will give the ratio of the amplitudes of the reflected to incident waves, whilst the argument of the R equals the phase shift caused by the reflection at the interface (not considering the natural phase shift due to propagation to and from the surface).

It can be verified from (3.8) that R_h cannot vanish (except for the trivial case when $\tilde{\epsilon}/\epsilon_1$ equals unity, i.e. when the interface is fictitious). On the other hand it follows from (3.7) that R_v will vanish for;

$$\tan \theta_c = \frac{1}{\sqrt{\tilde{\epsilon}}}. \quad (3.11)$$

or more generally, for an arbitrary (non-magnetic) medium of incidence one,

$$\tan \theta_c = \sqrt{\frac{\tilde{\epsilon}}{\epsilon_1}}. \quad (3.12)$$

The angle θ_c satisfying (3.12) is called the Brewster angle or polarizing angle, as under above conditions the polarization plane is perpendicular to the plane of incidence. The component of electric field normal to the plane of incidence is completely transmitted leaving only that component to be reflected which lie in the plane of incidence. Since θ_c must be real, it follows from (3.12) that only a dielectric has a Brewster angle for incidence from free space. If the medium is conducting R_v cannot be made to vanish; however, $|R_v|$ will attain a minimum for some value of the θ , and this value is sometimes referred to as the *quasi-Brewster* or *pseudo-Brewster angle*.

Apart from these general properties, the reflection coefficient depend on the angle of incidence, the permittivities and conductivities of the adjoining media and if one of the media is conducting then also on the incident wavelength. Variations of ϵ_r , except for values very close to one effect R_i 's magnitude slightly for frequencies (consequently wavelengths) above 100 MHz [16], [49]. The R may vary over its full range as the incident angle is varied from $0 \rightarrow \pi/2$. Figures 3.5 and 3.6 show the variations of the magnitude of the complex reflection coefficients R_v and R_h as a function of angle of incidence θ for different relative permittivities ϵ_r of reflecting material assuming same conduction properties. We may observe the angle of incidence θ_c at which R_v tends to approach zero or the Brewster angle. At small angles of incidence, both R_v and R_h are less than one. For increasing values of ϵ_r , however, they approach one. As ϵ_r increases, θ_c approaches $\pi/2$ radians.

For those cases where the incident polarization is neither perpendicular nor parallel to the plane of incidence, the determination of reflection coefficient is quite involved mathematically. This may, however, be observed that the value of reflection coefficient would be bounded by the two curves of the R_v and R_h . In that situation one possible approximation is suggested in Figure 3.7, where reflection coefficient assumes a constant value till Brewster angle and then increases linearly.

This behavior of the magnitude of the complex reflection coefficient is subject to certain interpretations which results in possible approximations and simplifications in ray tracing. It may be appreciated that for a given receiver position the angle of incidence of a multiple reflected path shall always be smaller than the angle of incidence of a single reflection path. Furthermore, the more the number of multiple reflections, the smallest the angle of incidence. Therefore, even though the multiple reflected paths would be strongly reflected, they would be extremely weak as they arrive at the receiver when compared to the single reflection path as long as the reflecting surface is not an infinitely conducting surface. Thus, the multiple reflected paths may be discarded as they contribute little in received signal. Another criterion to discard multiple reflected paths is to neglect those paths whose magnitude is smaller than the receiver noise level.

The ray tracing described above considers the reflection and refraction but does not include the diffraction phenomenon. The extension of the basic idea is needed to incorporate

FIGURE 3.5

Reflection coefficient magnitude as a function of angle of incidence for horizontal polarization

FIGURE 3.6

Reflection coefficient magnitude as a function of angle of incidence for vertical polarization

FIGURE 3.7

A possible approximation for reflection coefficient magnitude for neither vertical nor horizontal polarization

the diffraction process. This may be achieved through the use of Geometrical Theory of Diffraction which is described in next section.

3.2.3 Geometrical Theory of Diffraction

Keller, in his classical paper, [35], describes an extension of the geometrical optics which accounts for the diffraction of light. Before Keller's theory geometrical optics only accounted for the reflection and refraction phenomena. Similar to geometrical optics he assumes that light travels along certain straight and curved lines called rays but he also introduces certain new rays called diffraction rays, in addition to usual rays. Diffracted rays are produced by incident rays when they hit on edges, corners and vertices of the boundary surfaces. They do so by means of several laws of diffractions which are similar to reflection and refraction and are deduced from Fermat's principle, appropriately modified. The diffracted rays when propagating away from the diffractor behave like ordinary rays and constructed by use of Hygen's wavelet construction. In short, all fundamental principles of ordinary geometrical optics can be extended to the geometrical theory of diffraction. The only difficulty occurs in obtaining the initial value of the field at the point of diffraction. In the case of ordinary ray, the field of a ray emerging from a source is specified at the source. But on a reflected or transmitted ray, the initial value is obtained by multiplying the field on the incident ray by a reflection or transmission coefficients. By analogy, the initial value of the field on a diffracted ray is obtained by multiplying the field on the incident ray by diffraction coefficients which are different for edge, corner and vertex diffractions.

In microcellular mobile radio propagation, only edges of the buildings gives rise to significant diffracted rays to be accounted due to low antenna heights. Therefore, we only look into the edge diffraction in the following section.

3.2.4 Edge Diffraction

Keller, [35], hypothesizes that when a ray hits an edge or vertex or grazes a boundary surface such ray must give rise to diffracted rays. This hypothesis was tested mathematically in those cases where diffraction problems have been solved by other means. One such case is the diffraction of a plane wave by a semi-infinite screen with a straight edge. The solution to this problem was obtained by Sommerfeld, [36], and consisted of incident

and reflected wave and a third wave called diffracted wave. When the incident wave is propagating in a direction normal to the edge of screen, the ray is cylindrical with the edge as its axis. The straight lines orthogonal to the cylindrical wavefronts of the diffracted wave appear to come from the edge and can be identified as diffracted rays. This suggests that an incident ray normal to the edge produces diffracted rays which are also normal to the edge and which leave it in all directions, as shown in Figure 3.8. On the other hand, when the incident rays hit the edge of the screen obliquely, the diffracted field in Sommerfeld's solution is conical i.e. diffracted wavefronts are parallel cones with the edge on their common axis. The straight lines orthogonal to wavefronts also appear to come from edge and are identified as diffracted rays, Figure 3.9 on this Keller suggests "law of edge diffraction". "A diffracted and the corresponding incident ray make equal angles with the edge of the point of diffraction, provided that they are both in the same medium. They lie on the other side of the plane normal to the edge at the point of intersection. When the two rays lie in the different medium, the ratio of the sines of the angles between incident and diffracted rays and the normal plane is reciprocal to the indices of the refraction of two media". The second part of this law pertaining to diffraction in two different mediums is suggested by Snell's law instead. While both parts of the law are consequences of the following modified form of Fermat's principle which is known as Fermat's principle for edge diffraction. "An edge diffracted ray between two points is a curve which has stationary optical path length among all curves between the two points with one point on the edge".

Let us now consider the field E_d on a diffracted ray from an edge assuming ray is in an homogenous medium with E_i , the complex incident field, on the edge with phase ϕ_i and amplitude A_i . It can be shown that,

$$E_d = DA_i r^{-1/2} e^{j(kr + \phi_i)} = DE_i r^{-1/2} e^{jk r}, \quad (3.13)$$

where k is the propagation constant, r is the distance from the edge, D is the diffraction coefficient and $E_i = A_i e^{j\phi_i}$. Comparing this with Sommerfeld's, [36], result when asymptotically expanded for large values of r , it agrees perfectly with above provided that, [35],

$$D = \left[-\frac{e^{j\frac{\pi}{4}}}{2(2\pi k)^{1/2} \sin \beta} \right] \left[\cos \left(\frac{\theta - \alpha}{2} \right)^{-1} \pm \sin \left(\frac{\theta + \alpha}{2} \right)^{-1} \right], \quad (3.14)$$

FIGURE 3.8

 Diffracted rays of an edge for normal incidence

FIGURE 3.9

 Diffracted rays of an edge for oblique incidence

here β is the angle between incident ray and the edge. The angle between the incident and diffracted ray and normal to the screen are θ and α respectively. The upper sign applies when the boundary conditions on the half plane is $E = 0$ and lower one if $\partial E / (\partial n) = 0$.

Similarly, in case the half plane is replaced by a wedge of angle $(2 - n)\pi$, with $\beta \neq \pi/2$, the Sommerfeld's, [36], exact solution for wedge yields agreement when, [35],

$$D = \left[\frac{e^{\frac{j\pi}{4}} \sin \frac{\pi}{4}}{n(2\pi k)^{1/2} \sin \beta} \right] \left[\left(\cos \frac{\pi}{n} - \cos \frac{\theta - \alpha}{n} \right)^{-1} \mp \left(\cos \frac{\pi}{n} - \cos \frac{\theta + \alpha + \pi}{n} \right)^{-1} \right]. \quad (3.15)$$

The diffraction coefficient of a wedge (actually the edge of a wedge) are more complicated than the reflection coefficient of a plane. They not only are algebraically more involved but also depend on more variables; which are, angle of incidence and scattering and the vertex angle of the wedge. Moreover, two angles are needed to determine the direction of incidence uniquely. The geometry is illustrated in Figures 3.10 and 3.11.

In actual environment, certain parameters are fixed. For example, the shape of the buildings (consequently vertex angle) and base station positions are fixed, thus are the angle α and β . Only the position of the receiver changes as a result only angle ϕ will change. The plot of the absolute value of D with respect to ϕ is shown in Figure 3.12 for ϕ varying from zero to 2π . The value of n used is $3/2$ for a vertex angle of $\pi/2$. Each curve corresponds to a different value of β with $\alpha = 0$. In the Figure 3.13 solid curve is for $\beta = \pi/2$, dotted for $\beta = \pi/3$, dash-dot for $\beta = \pi/4$ and dashed for $\beta = \pi/6$. We may observe that as β is increased the value of $|D|$ is decreased. However, we also observe that value of $|D|$ shows a singularity at $\phi = 0$, $\phi = \pi$ and $\phi = 2\pi$, the reason and effect of which is explained in the following paragraph.

Now consider the Figure 3.14 where each curve corresponds to a different value of α (the choice of $\alpha = 0$ is for illustration). Certain observations may be made. At reflection and shadow boundaries (Figure 3.13), we obtain illegal values for $|D|$ (i.e greater than one). It has been mentioned that equations (3.14) and (3.15) are not valid at and in the vicinity of reflection and shadow boundaries. Sommerfeld, [36], explains that the infinitely large value which this expression assumes on the shadow boundary is illusory and is due to the fact that the asymptotic approximation to get the diffracted ray solution is not

FIGURE 3.10

Illustration of geometry for equation (3.15)

(a)

(b)

FIGURE 3.11

(a) Projection of incident and diffracted rays in x - y plane (b) incident and diffracted rays from edge which is along z -axis

FIGURE 3.12

Plot of equation (3.15) as a function of φ for different values of incident angle β , here $\alpha = 0$

FIGURE 3.13

Illustration of reflection and shadow boundaries

FIGURE 3.14

Plot of equation (3.15) as a function of ϕ for same incident angle β but different α . At reflection and shadow boundaries magnitude of diffraction coefficient has singularities elsewhere it is nearly constant

valid here. Instead exact solution involving a Fresnel integral is to be used which forms the basis of Uniform Theory of Diffraction, [37]. However, it will make D too involved and requires huge computation as the number of factors on which D depends increase. Observing that except at shadow and reflection boundaries the value of D does not change very much and it is a small number, we may assume a single value for D based on the non-shadow boundary region to be also valid for all ϕ 's. This numerical value is considered to be 0.2 in this simulation. The effect of this choice will be this that at those positions of the receiver which are in the vicinity of these boundaries, the contribution of diffracted paths would be lesser than it would otherwise have. This assumption, however, shall reduce the computation very much in the simulation, particularly, for the fast fading computations where all diffraction contributions are included.

A heuristic extension which allows the approximate treatment of finitely conducting locally rough wedge diffraction has been made in, [38], and, [39]. We do not consider these issues in this thesis.

3.3 Geometrical Optics Simulation Model

The geometrical optics simulation model uses the properties of the wave incident on a plane surface. The angle of reflection equals the angle of incidence. This results in an image of a fixed transmitter appearing as a fixed point on the other side of the reflective surface in a position such that both the transmitter and the image lie on the line normal to the reflecting surface and are equidistant from the surface (Figure 3.4). This helps in determination of images of a fixed transmitter site once the environment is selected. An image may or may not be visible at certain receiver positions as the reflective surface may not extend far enough to intersect the image to receiver ray. The diffracted rays from the edges of the reflecting surface are determined from geometrical theory of diffraction, [35].

Following assumptions are made so that ray tracing technique could be applicable for radio propagation simulation, [28]. It is assumed that:

- the point of observation is many wavelengths from the nearest scatterer
- all scatterer are large compared to a wavelength
- all scatterer are smooth, i.e. have surface features that are small compared to a wavelength.

One of the proposed microcells structures is known as Manhattan environment, [40], (after downtown Manhattan New York City) with transmitter at about lamp-post height, and which provides coverage to those receivers which are in LOS streets. The cells are either half square cell, full square cell or rectangular cells as shown in Figures 3.15-3.17, respectively. In these figures different shades are used to represent the possible frequency reuse pattern where a circle represents the base station antenna location. We may observe that antennas are placed in such a way that proposed cell coverage area is within line of sight. Another observation may be that full square and rectangular cells utilize lower number of base stations than half square cells making them more suitable practically.

A possible propagation mechanism for a rectangular cell using [7], [9], and geometrical theory of diffraction (GTD), [35], is shown in Figure 3.18. The three dimensional environment is represented in two dimensions by taking a section at $z = z_1$, assuming transmitter and receiver at the same height i.e. $z = z_1$ where ground is represented by x - y plane at $z = 0$. An edge is therefore represented by a point and the building walls by lines. The above assumption will result into normal incidence of the transmitted wave at the edges of the buildings. The resultant diffracted waves are cylindrical in nature as shown at edges. The transmitter is shown by a circle and receiver positions by a square. The receiver position corresponds to the situation of a LOS street where besides a LOS and ground reflected paths, there may be building reflected paths and edge diffracted paths. Diffracted paths are shown by shaded rays in Figure 3.18. In a real environment there would be LOS, ground reflected, single or multiple reflected, single or multiple diffracted, single or multiple wall-reflected/edge-diffracted and edge-diffracted/wall-reflected rays and their combination in infinite possible ways. However, for simulation purposes, the number of rays may be reduced to a very few depending upon the aim of the simulation.

We assume, in addition to a direct path, L reflected (L includes ground reflected path), M diffracted and N multiple reflected-diffracted/diffracted-reflected paths are received by the dipole, therefore, the received field strength, E_r , as the phasor summation of the all constituent multipath is written as,

$$E_r = C \left(\frac{e^{j\phi}}{r} + \sum_{i=1}^L \frac{R_i e^{j\Delta\phi_i}}{r_i} + \sum_{j=1}^M \frac{D_j e^{j\Delta\phi_j}}{r_j} + \sum_{k=1}^N \frac{S_k e^{j\Delta\phi_k}}{r_k} \right) + z, \quad (3.16)$$

FIGURE 3.15

Microcellular cell structure, a half square cell layout

FIGURE 3.16

Microcellular cell structure, a full square cell layout

FIGURE 3.17

Microcellular cell structure, a rectangular cell layout

FIGURE 3.18

A possible dominant propagation mechanism. Multiple wall reflections are neglected

where C is a constant which takes into account the effects of frequency, transmitted field strength and antenna gains. The terms R_i , D_j , and S_k are reflection, diffraction and reflection-diffraction/diffraction-reflection coefficients, respectively. The S_k is obtained by product of R_i 's or D_j 's or R_i 's and D_j 's. The r , r_i , r_j and r_k are, respectively, path lengths of the direct, reflected, diffracted and reflection-diffraction/diffraction-reflection rays. z is complex noise which is assumed to be normally distributed. The differential phase is given by

$$\Delta\phi = (2\pi\Delta l)/\lambda, \quad (3.17)$$

where Δl is differential path length between the direct and the reflected or diffracted or reflected-diffracted/diffracted-reflected ray giving $\Delta\phi_i$, or $\Delta\phi_j$, or $\Delta\phi_k$, respectively, and λ is wavelength. The external surfaces of the high rises are predominantly covered by glass, therefore we assume that in the simulation environment R_i is caused by glass with $\epsilon_r = 6$. For ground reflections, the dielectric constant for average ground $\epsilon_r = 15$ is used. The D_j 's in (3.16) are the diffraction coefficients and assumed to be constant as discussed in last section (p. 42).

As we argued earlier (p. 32) that multiple reflected paths may be discarded, multiple diffracted and multiple reflected-diffracted and reflected-diffracted-reflected paths may also be discarded using the same argument. All of above mentioned multipath may be neglected to help simplify (3.16) in the following form,

$$E_r = C \left(\frac{e^{j\phi}}{r} + \sum_{i=1}^L \frac{R_i e^{j\Delta\phi_i}}{r_i} + \sum_{j=1}^M \frac{D_j e^{j\Delta\phi_j}}{r_j} + \sum_{n=1}^N \frac{R_n^2 e^{j\Delta\phi_n}}{r_n} \right) + z', \quad (3.18)$$

where fourth term in the brackets accounts for double surface reflections and random variable z' may be thought to have absorbed the omitted terms as well as the affects arising due to basic assumptions involved for application of ray tracing, such as, flat surfaces. Heuristically, third, fourth and fifth terms in (3.16) may be considered to again give rise to a normally distributed random variable z' in (3.18) due to application of central limit theorem.

The channel model depicted in Figures 3.3 may now be seen through the perspective of the geometrical optics. The main issue is the translation of parameters τ and K in geometrical optics term. The tapped delay line model is applicable due to multipath reception of the transmitted signal. In this case, τ will be the delay associated with the different paths

and K would be the path strengths. In geometrical optics terms, τ corresponds to the different phases of the multipath signals and K would be dependent on path length, reflection and diffraction coefficients, among others.

Using the simulation model developed in the last Chapter we investigate the path loss, slow and fast fading statistics based on Ottawa downtown. The simulated path loss is compared with [10]. However, from [10] certain informations, such as receiver path and exact transmitter location, needed for a detailed comparison can not be extracted precisely. Therefore, we will only be able to compare the gross features of the simulated path loss with the measured one.

The simulation included transmitting a 910 MHz continuous wave (CW) signal from a fixed base station to a mobile receiver and recording the received signal envelope as a function of receiver location. The simulated base station antenna was located at Slater Street approximately as in [10] at a height of 8.5 m while mobile receiver was 3.65m above the ground. The value of reflection coefficient R is approximated as in Figure 3.7. The value of diffraction coefficients D is chosen approximately to be equal to 0.2 as discussed in the Chapter 3. Simulated recordings are made on LOS along the street for a rectangular cell structure in a straight line in the middle of the street assuming receiver is moved from initial to final position at a uniform speed. The simulation records 10 samples/meter to determine median path loss over an interval of 2 meters and 1000 samples/meter or approximately 330 samples/wavelength of the received signal envelope to inves-

tigate fast fading statistic. The frequency used in most of the results is 910 MHz. However, some simulations were performed at 1.6 GHz and 11 GHz to compare the path loss results with 910 MHz. The above parameters are assumed in all results unless otherwise specified.

4.1 Path Loss

Simulation results for path loss are presented in this subsection. A varying signal strength is experienced by a mobile due to its passage through an interference pattern developed by multipath propagation. In this section we will investigate the effect of antenna heights, carrier frequency and physical location of base station antenna on the median path loss characteristics of our selected environment.

Slater and Queens Streets of the Ottawa downtown are selected for the simulations. The building layout of the two streets is shown in the Figure 4.1 on an approximate scale. Transmitter positions are also identified. The length of the streets being considered is about 900 meters. Whereas, the width of the street is determined by the buildings on both sides of the street which varies from 18 to 22 meters for Slater and between 20 to 23 for Queens Street.

We noted earlier that the constituent multipath are due to LOS, ground reflected, buildings reflected and edge diffracted paths. To begin with we simulate a two ray interference pattern and then add buildings reflected paths and the edge diffracted paths. Figure 4.2 shows free space and two ray path loss curves. In the propagation loss curves ordinate is median path loss in dB's and abscissa is distance in meters. The minimum path loss point in the graph is the closest point to the transmitter. In Figure 4.3, building surface reflected paths and diffracted paths are included. Certain observations may be made. The effect of adding these multipath is to distort a rather definable interference pattern of the LOS and ground reflected paths. Surface reflected and diffracted paths sometimes add constructively and at other times destructively.

Figure 4.4 is taken from reference [10]. When we compare it with Figure 4.3, we note certain similarities and dissimilarities. Close to the transmitter the measured and simulated results appear to be quite similar but after a distance of few hundred meters the simulation gives lesser path loss than the measured one. A number of factors may be responsible for

FIGURE 4.1

Building lay-out of the Slater and Queens Street in Ottawa Downtown on an approximate scale for Figures 4.3 and 4.5

FIGURE 4.2

Propagation loss versus distance for plane earth reflection model

FIGURE 4.3

Propagation loss including reflected and diffracted multipath for Slater Street

FIGURE 4.4

Measured propagation loss for Slater Street from reference [10]

FIGURE 4.5

Propagation loss for Queens Street

this disagreement. First of all, the transmitter location used in [10] is not known exactly. Based on available informations we can only reduce the uncertainty within an area of approximately four square meter. We will show later that the transmit antenna location is important in determination of received signal power for this environment. Secondly, the mobile receiver's path along which the measurements in [10] have been taken for Figure 4.4 is not known which also is an important factor to be known to have a better comparison. Moreover simulation does not take into account the traffic on the street. Also the number of samples of the median is taken over an interval of two meters vary from one interval to another due to changing speed of mobile. A number of approximations has also been made to apply ray tracing. All of these factors combined make it difficult to form a candid detailed comparison. Only gross features may possibly be verified, which are: i) median path loss is close to that of free space propagation loss; ii) occasional fades of up to 10 dB's; as reported in, [10]. These features are clearly visible in Figure 4.3. A similar plot is shown for Queens Street in Figure 4.5.

We mentioned earlier that received median path loss curve depends on the path chosen for the simulated mobile receiver. This is verified by plotting the median path loss for Slater and Queens Streets both along (900 m) and across (17 m) the street covering an area of 900 x 17 meters and shown in Figures 4.6 and 4.7 respectively. Each point is a median of 1.5 meter square area taken 3 meter along and 0.5 meter across the street. Transmitter is 12 meters from zero along width axis and 235 meters from zero along length axis. Certain observations may be made. In the east of transmitter, path loss is distributed somewhat symmetrically along the street. Two dome like pattern are identified for the Slater Street of about the same radius. The same is true about Queens Street but there are three domes with centre dome having larger radius than the ones on the sides, apparently a consequence of the difference in buildings layouts. This emphasizes the fact that a mobile which is not travelling in a straight path will certainly experience a different median path loss than the one doing so.

To investigate the effect of transmitter location on received median path loss and propagation pattern, we changed the transmitter location in two steps, firstly along and then across the street. In first step we moved the transmitter towards Lyon Street in the west, about 40 meters while keeping the lateral position same as that for the transmitter location

FIGURE 4.6

Median path loss distribution pattern of Slater Street

FIGURE 4.7

Median path loss distribution pattern of Queens Street

of Figure 4.6. The result is shown in Figure 4.8. No significant change is observed besides a slight change in the radius of domes, which may still be identified. In the second step, we moved the transmitter 9 meters across the street in northerly direction. The resulting path loss distribution is shown in the Figure 4.9. Interestingly, we observe a significant change in the path loss distribution which is now more evenly distributed. It has some interesting implications. Firstly, the distribution of path loss is function of transmitter position. Secondly, moving the transmitter along the street does not affect the loss distribution pattern significantly for smaller changes thus preserving the main features. Whereas a lateral change affects the loss distribution pattern more rapidly than along the street. This appears to be due to the fact that a lateral change results into disruption of building lay out symmetry faster than along the street. This observation has to have important consequence on selection of a suitable fixed base station antenna location.

Next we investigated the effect of increasing transmitter antenna heights on the received path loss while keeping it below average building heights. To this end, we vary the transmit antenna height in 0.5 meter steps starting with both transmitter and receiver having same height of 3.5 meters and ending with the transmitter height of 20 meters for the same mobile receiver path along the Slater Street. The result is plotted in Figure 4.10, where x-axis is the path length and y-axis is the change in transmit antenna height in 0.5 meter steps. Two observations are in order. Firstly close to the transmitter as antenna height is increased, the same receiver point goes into multiple fades which are shown in Figure 4.11. It is expected because of small distance between the transmitter and receiver. A change in transmitter height shall effect the length of ground reflected path significantly and result into rapid multiple fades as a function of transmitter height. Secondly, as receiver moves away from the transmitter, the number of deep fades start reducing as a function of distance between transmitter and receiver. At far off distances this effect becomes negligible and also show an increase in received signal power. It is expected because in this case the LOS and ground reflected components start becoming nearly equal.

Now we look into the effect of change in carrier frequencies. Uptill now a 910 MHz carrier was used for all the results presented. Considering now the Slater Street and all parameters, same as for Figure 4.6 (transmitter location and receiver locations), except the frequency which is now changed to 1.6 GHz. Result is shown in Figure 4.12. The effect is

FIGURE 4.8

Median path loss distribution pattern of Slater Street; transmitter is moved along the street

FIGURE 4.9

Median path loss distribution pattern of Slater Street; transmitter is moved across the street

FIGURE 4.10

Median path loss distribution pattern as a function of transmitter antenna height for the same straight line path on Slater Street

FIGURE 4.11

Path loss for a fixed receiver point on Slater Street as a function of transmitter antenna height. The distance is mentioned with the curves

that of some distortion in the path loss distribution pattern. When the frequency is changed to 11 GHz, about an order of magnitude higher, resultant path loss distribution is shown in Figure 4.13. The choice of 11 GHz is to observe the effect of change in frequency on propagation characteristics. For a candid comparison, all parameters such as transmitter and receiver locations, reflection and diffraction coefficients are same as in Figure 4.10. The resultant path loss, now, has large number of propagation holes whereas a 20 dB's path loss offset is also evidenced due to a change in frequency of an order of magnitude.

Lastly, before we turn to fading statistics, we look into the path loss results from the coverage viewpoint. For this purpose some contour plots are drawn. Figures 4.14-4.17 are the contour plot corresponding to Figures 4.6-4.9 and their parameter (transmitter locations, receiver locations and frequency) for a -80 dB threshold. Transmitter locations are shown by shaded ovals. The coverage area comprises of closed and line contours. Between two broken line contours the signal strength is either above or below the threshold value. This is identified in graphs by letters A and B for above and below respectively. Area inside a closed contour in region A corresponds to signal strength below threshold and vice versa. The -80 dB contour plot in Figure 4.15 corresponds to the Figure 4.8 in which transmitter is moved along the street 40 meters in westward direction. When comparing Figures 4.14 and 4.15 one observes changes in the contour patterns in most of the coverage area. An important change is the disappearance of below threshold signal strength corridor in Figure 4.14 in the middle of the street as observed in Figure 4.15. In Figure 4.16, transmitter is moved across the street as was done in Figure 4.9. It results into further change in contour pattern. Observations include a large number of small closed contour and only one big area having signal strength below threshold. Figure 4.17 is -100 dB contour plot corresponding to the Figure 4.13 of Slater Street path loss for 11 GHz frequency. Number of contours are large and everywhere showing frequent deep fade region or propagation holes, each of which, however, cover relatively small area. All of the above observations verify the fact that service quality and reliability in microcellular systems depends upon where a base station antenna is located as well as the frequency of operation. Because the frequency of operation is a regulatory matter and finally fixed, therefore, one needs to select the transmitter site in such a way to optimize service quality and reliability

FIGURE 4.12

Median path loss distribution pattern of Slater Street; $f = 1.6$ GHz

FIGURE 4.13

Median path loss distribution pattern of Slater Street; $f = 11$ GHz

FIGURE 4.14

The -80 dB contour plot for Slater Street corresponding to Figure 4.6

FIGURE 4.15

The -80 dB contour plot for Slater Street corresponding to Figure 4.7.
Transmitter moved along the street

FIGURE 4.16

The -80 dB contour plot for Slater Street corresponding to Figure 4.8.
Transmitter moved across the street

FIGURE 4.17

The -100 dB contour plot for Slater Street corresponding to Figure 4.13;
frequency=11GHz

for those services which have existing frequency standards. However, for emerging frequency standards, this may help choose the more appropriate frequency band.

4.2 Fast Fading

While investigating small area/fast fading statistics, the type of best fit distribution and typical and worst case values of its parameters are of major concern. Here, the problem include, firstly, to find a criteria for “curvefitting”, secondly, finding what distribution function is best fit for the given criterion and thirdly, answering the question “which of the simulated or experimentally determined parameters may be considered as typical to represent the statistics”. In this simulation criterion for curvefitting is Non-Linear Least Square error and goodness of fit is determined through relative comparison of different candidates distribution functions based on the least square curvefitting error.

In non-linear least squares curvefitting, [41], the model which depends non-linearly on the set of M numbers of parameters a_k , $k = 1, 2, \dots, M$, is fitted based on a merit function χ^2 . The model to be fitted is generally written as,

$$y = y(x; a_k) \quad k = 1, 2, \dots, M, \quad (4.1)$$

and the merit function χ^2 is given by,

$$\chi^2(a_k) = \sum_{i=1}^N \left[\frac{y_i - y(x_i; a_k)}{\sigma_i} \right]^2 \quad k = 1, 2, \dots, M. \quad (4.2)$$

Here σ_i is the standard deviation of the measurement error of the i th data point. If the measurement error is not known then it is set to a constant value $\sigma = 1$, which is the case in this simulation. The best fit parameters are obtained by the minimization of the merit function. The minimization process proceeds iteratively and for a given trial value of parameters, the curvefitting procedure is repeated until χ^2 stops (or effectively stops) decreasing.

Rayleigh, Rician and Nakagami distributions were fitted for each interval. The functional form of the distributions are respectively,

$$p(x) = \frac{x}{\Omega^2} e^{-\frac{x^2}{2\Omega^2}} \quad (x \geq 0), \quad (4.3)$$

$$p(x) = \frac{x}{\Omega^2} e^{-\frac{(x^2+s^2)}{2\Omega^2}} I_0\left(\frac{rs}{\Omega^2}\right) \quad (x \geq 0), \quad (4.4)$$

$$p(x) = \frac{2m^m x^{2m-1}}{\Gamma(m) \Omega^m} e^{-\frac{m}{\Omega} x^2} \quad (x \geq 0, m \geq \frac{1}{2}). \quad (4.5)$$

Where in equations (4.3)-(4.5), the Ω , m and s are curvefitting parameters. Here we observe that Rice and Nakagami have two parameters while Rayleigh has only one. Moreover, for $s = 0$ and $m = 1$, both Rice and Nakagami distributions degenerate into Rayleigh.

In actual experiments it is observed that received signal envelope comprises of a rapidly varying component which is superimposed on a slowly varying mean as shown in Figure 4.18. Before determining small area statistics, it is desired to remove the fast fading component from the slowly varying mean. One approach is to find a small interval over which the mean may appear constant and then normalizing the data by that mean value. The dilemma in this approach is to find the smallest area over which fast fading statistics may be found with confidence. In literature, intervals as big as hundreds of wavelengths are used [20], perhaps due to the fact that achievable envelope sampling rate were low and therefore areas were selected to have enough data to make analysis reasonable. In computer simulation of channel one do not has this limitation, thus, permitting a selection of smaller path length.

In another approach, measured or simulated data is passed through a low pass filter with a suitable cutoff frequency, the data is then normalized to this filtered mean. The issue in this approach is "how to select the cut-off period of the low pass filter". The answer to this question as in the last approach is related to the best engineering judgement through trial and error. Figures 4.19 and 4.20 show the approximation of the slowly varying mean using a sixth order Butterworth low pass filter with spatial cutoff at one and two meters, respectively. Figure 4.21 is the normalized envelope obtained by the 2 meter cutoff filtered mean. Here we may appreciate that reducing the spatial cutoff will result into an increased slowly varying component in the normalized envelope data. The choice of spatial cut-off period at 2 meter is arbitrary and obtained by repeated trial. Once mean is removed, Rice model may be fitted by varying the Ω while keeping the value of s to that of normalized

FIGURE 4.18

A typical received envelope of the path loss

FIGURE 4.19

Approximation of a slowly varying component with a one meter cut off filter

FIGURE 4.20

Approximation of a slowly varying component with a two meter cut off filter

FIGURE 4.21

Envelope normalized to two meter cutoff mean to investigate the fast fading

value of unity. However the approach we use is one in which s is allowed to vary around its normalized unity value. This is to account for possible errors in the normalization procedure, due either to the arbitrariness of the selected interval or spatial cut-off selection, whatever be the case.

Which of the above mentioned three distribution functions best represent the fast fading statistics? The answer to this question is sought through fitting the statistical models for a data of about 400 intervals of length 30λ (or approximately 10 meters) and observing that which one of the three fits with lowest least square error. This simulation result when normalized to interval mean is shown in Figure 4.22. Here, increase in the number of intervals corresponds to the increasing distance from the transmitter. Rician distribution seems to be the best fit in majority of the intervals. However, Nakagami distribution fits nearly as well as the Rician. In certain interval either of the Nakagami or Rician fit better than the other, while in majority of the cases both give approximately same least square error. On the other hand Rayleigh turns out to be worst fit in nearly all the intervals except a few ones, where it does as good as the other two. But a trend is observed however, which is this that as intervals get further away from the transmitter, the least square error of the Rayleigh fit decreases and at far off intervals it starts to fit nearly as good as the other two.

Another question which may arise is "what effect the normalization procedure has on the curve fitting of the distribution functions?". The answer may be extracted from the graph in Figure 4.23. All the observations for Figure 4.22 are valid except for this that close to the transmitter the magnitude of the least square error has increased. This may suggest that interval mean normalization should be preferred for curvefitting. Some typical curvefitting results for the three distributions with the simulated pdf's are shown in Figures 4.24 and 4.25.

The other matter of interest is to gain information about the curvefitting parameters. We have two curvefitting parameters for both Nakagami and Rician distributions while one for Rayleigh. Parameter Ω provides information about the spread of the distribution whereas m and s parameters of Nakagami and Rician distributions, respectively, provide information about the shape of the distribution. Rayleigh has no shaping parameter. The shape parameters shall affect the vital fading information contained in the lower tail. It may therefore be said that variation in m and s shall give one an indication of the fading

FIGURE 4.22

Least square error of curvefitting for 400 intervals normalized to interval mean

FIGURE 4.23

Least square error of curvefitting for 400 intervals normalized to filtered mean over each interval

FIGURE 4.24

A typical curvefitting of Nakagami (solid), Rice (dotted) and Rayleigh distribution (dashed) to the simulated *pdf* near transmitter

FIGURE 4.25

A typical curvefitting of Nakagami (solid), Rice (dotted) and Rayleigh distribution (dashed) to the simulated *pdf* away from transmitter

encountered in a specific environment. In 400 intervals over which curvefitting is done here, both m and s change from one interval to another. This change is depicted in Figure 4.26. One observation is the downward trend of the m and s with respect to distance emphasizing a possible direct relationship between the two. Another observation is that for those values of m which are less than one the s parameter assumes a value of zero. This implies that the Rice model in case of severe fading shall reduce to Rayleigh, however, Nakagami shall be able to characterize fading even when fading is worse than Rayleigh. It may also be observed that this variation is to be treated randomly as no simple relation appears to govern this change. The way to characterize the randomly varying parameter is to determine the cumulative distribution function (*cdf*). Figure 4.27 show the *cdf*'s of m and s parameters. Some probable conclusions which may be drawn from Figure 4.27 are that there is a 76% probability that value of m parameters is greater than one over a mobile path. On the other hand, it is 78% probable for s parameter to have a value greater than 0.05. What this may imply physically? Firstly, the channels statistics are non-stationary not only from the point of view of larger areas but also for the smaller areas. Secondly, this channel on the average is better than the Rayleigh channel and we may expect improved performance for a signal transmitted over this channel than a Rayleigh channel. The second conclusion is based on the well founded Rician channel performance results, where s greater than zero implies a line of sight signal component, resulting into an improved performance. Is it so and how, is yet to be confirmed for the Nakagami channel. However, one physical interpretation do exist for the m parameter of the Nakagami distribution.

Nakagami, [25], interprets m parameter as a fading figure. It can be appreciated from Figure 4.28 that as value of m increases the area under the lower tail of the probability density function is lesser. The lower tail corresponds to fades and therefore m is inversely proportional to fading. In a more quantitative manner, we may find the probability of error P_e for a simple BPSK signal over Nakagami fading channel and compare it to Rayleigh and additive white gaussian noise (AWGN) channel. For an AWGN channel, probability of error as a function of received signal to noise ratio (SNR) per bit is given by, [42],

$$P_e(\gamma_b) = \frac{1}{2} \operatorname{erfc}(\sqrt{\gamma_b}). \quad (4.6)$$

FIGURE 4.26

Variation in ' m ' (dotted) and ' s ' parameter (solid)

FIGURE 4.27

The cumulative distribution function of ' m ' and ' s ' parameter

FIGURE 4.28

Nakagami distribution for different values of m parameter with $\Omega = 1$

FIGURE 4.29

Comparison of error probabilities for a BPSK signal over AWGN, Rayleigh and Nakagami channel

where $\gamma_b = \alpha^2 (E_b/N_o)$. Here α is the channel attenuation and considered to be known and fixed. When α is randomly varying, we must average $P_e(\gamma_b)$ over the probability density function of γ_b . That is, we must evaluate the integral,

$$P_{a,e} = \int_0^{\infty} P_e(\gamma_b) p(\gamma_b) d\gamma_b, \quad (4.7)$$

where $p(\gamma_b)$ is the probability density function of γ_b when α is random. In case of α being Rayleigh distributed the result of integration in equation (4.7) for binary PSK is, [42],

$$P_{a,e} = \frac{1}{2} \left[1 - \sqrt{\frac{\gamma_b}{1 + \gamma_b}} \right], \quad (4.8)$$

where $\gamma_b = (E_b/N_o) E\{\alpha^2\}$. The term $E\{\alpha^2\}$ is the average value of α^2 . If α is Nakagami distributed, the integral reduces to the following form, [Appendix B],

$$P_{a,e} = \int_0^{\infty} \frac{A}{2} \operatorname{erfc}(\sqrt{\gamma_b}) \gamma_b^{m-1} e^{-B\gamma_b} d\gamma_b, \quad (4.9)$$

here $A = \frac{m^m}{\Gamma(m) \Omega^m} \left(\frac{N_o}{E_b}\right)^m$ and $B = \left(\frac{m}{\Omega}\right) \left(\frac{N_o}{E_b}\right)$.

The plot of Equation (4.6), (4.8) and (4.9) is shown in Figure 4.29. The fade margin required for Rayleigh fading to achieve a probability of error of 10^{-2} is about 10 dB's. The same is 8 dB's for Nakagami channel with $m=1.3132$, the median value of our curvefitting data while normalized to filtered mean, thus giving a 2 dB's improvement. At 10^{-4} this improvement is around 7 dB's.

4.3 Slow Fading

The path loss data over large intervals are curvefitted to obtain slow fading statistics. We obtain a lognormal fit which is widely reported in literature, [6], [25], [26]. The path loss in decibel is normally distributed as shown in the Figure 4.30. We find the standard deviation to be dependent on the length of interval which has been selected for curvefitting. Measurements report standard deviation of 6.5-8.2 dB's, [6], [43]. Some figures for illustration are about 4-5 dB's for a hundreds meter interval, about 5-6 dB's for two hundred meter intervals and about 7-8 dB's for a three hundred meter interval. We further observed

FIGURE 4.30

A typical normal fit for the slow fading over a large interval, mean = 72.2 dB and standard deviation = 7.2 dB. Interval length is 300 meters

that reducing the interval length increase the least square curvefitting error as opposed to increasing interval lengths when it decreases.

4.4 Comments on Near Field Radio Propagation

The field due to a physical dipole antenna at any distance r is found using Maxwell's equations under appropriate boundary conditions. A number of approximations are made, one of which is that solution is only valid if the point of observation is several wavelengths away from the antenna. This assumption helps writing the far field in a closed form but near field remains ambiguous and no general closed form solution is available in literature. From electromagnetic theory point of view, this is one of the most complex problems. However, this may not be discouraging from the radio engineer's point of view because though one may not be able to tell the exact form of near field solution, but it may be possible to know as much as one do in case of the nature of far field under multipath signal propagation conditions. As invariably all radio propagation environment give rise to multipath which results in the fading of the received signal envelope and faded waveform cannot be written into any closed form and is assumed to vary randomly to be characterized statistically. Here, it is discussed that the received signal envelope in near field may be represented as a fading envelope similar to that of far field and that it is theoretically possible to determine the statistical behavior of near field signal envelope.

The discussion in Appendix C evidence that in the absence of the assumptions made for determination of the field of a physical dipole, the near field is neither a uniform plane wave nor is it sinusoidal for a sinusoidally oscillating current source. We recall that if observation point is very close to dipole then the role of exponential term in (C.8) would be pronounced. In other words, contributions due to each point source would have different phase at the point of observation.

The problem is to look into the field components of a sinusoidal dipole radiator when point of observation is in proximity to radiator. Considering the physical dipole as comprising of N point source current elements, where N is sufficiently large, the field components due to any contributing point source at any observation point are given by (C.1)-(C.3) with equality changed to near equality,

$$H_{\varphi_s} = -\frac{I_0 d \beta_0^2}{4\pi} \sin \theta \left[\frac{1}{j\beta_0 r} + \frac{1}{(j\beta_0 r)^2} \right] e^{-j\beta_0 r}, \quad (4.10)$$

$$E_{rs} = -\frac{I_0 d \beta_0^2}{2\pi} \sqrt{\frac{\mu}{\epsilon}} \cos \theta \left[\frac{1}{(j\beta_0 r)^2} + \frac{1}{(j\beta_0 r)^3} \right] e^{-j\beta_0 r}, \quad (4.11)$$

$$E_{\theta s} = -\frac{I_0 d \beta_0^2}{4\pi} \sqrt{\frac{\mu}{\epsilon}} \sin \theta \left[\frac{1}{j\beta_0 r} + \frac{1}{(j\beta_0 r)^2} + \frac{1}{(j\beta_0 r)^3} \right] e^{-j\beta_0 r}. \quad (4.12)$$

We rewrite the electric field component $E_{\theta s}$ such that for any point source at any observation point:

$$E_{\theta, n} = |E_{\theta, n}| e^{j\varphi_n}, \quad (4.13)$$

where first term on right hand side of the (4.12) is the magnitude of the complex $E_{\theta, n}$, while φ_n is the phase. The resultant E_{θ} field at the observation point is obtained through superposition of N component fields each due to n point sources giving:

$$E_{\theta} = \left| \sum_{n=1}^N E_{\theta, n} e^{j\varphi_n} \right| = |x + jy| \quad (4.14)$$

The nature of $E_{\theta, n}$ and φ_n are of our concern. The $E_{\theta, n}$ depends upon source current, wavelength, physical dimensions of the radiator and at the location of the observation point. The φ_n is a function of wavelength and the radial distance from the point source. Keeping other parameters fixed, from one observation point to another, E_{θ} will be the function of space coordinates only. When observation point is moved around the dipole, both $E_{\theta, n}$ and φ_n shall keep changing and E_{θ} shall vary from one to another point. In other words the field at any observation point shall be as that of the phasor sum of multipath with different strengths and phase due to different path lengths and resulting into fading of the received electric field.

We know that in multipath propagation environment, if the number of multipath could be deterministic then received signal at an observation point shall also be deterministic. This is generally the case when multipath propagation environment is simulated on computers and maximum numbers of rays is known. The addition of more multipath increase the complexity of the received signal which is consequently characterized statistically due to the fact that whole process is no more tractable. Same is the case here for near field radio propagation and we need to characterize the near field statistically.

For statistical characterization, we may treat this problem as the random vector problem. As earlier, it may reasonably be supposed again that the received field intensity, at any observing point is composed of some component fields which have followed different paths and whose amplitude and phase vary according to certain statistical law. Under this condition the equation (4.14) now represent the general problem of random vectors. Nakagami [35] shows that the following distribution,

$$p(E) = \frac{2m^m E^{2m-1}}{\Gamma(m) \Omega^m} e^{-\frac{m}{\Omega} E^2} \quad ; (E > 0) , \quad (4.15)$$

is a general approximate solution for the distribution of magnitudes of resultant random vectors. Here m and Ω are distribution parameters. m controls the skewness and Ω controls the spread of the distribution. The typical values of above parameters are not known and needs to be determined experimentally or through simulations keeping this theoretical development into consideration.

5.1 Discussion of The Results

The following is a summary of this thesis research. Path loss and fading statistics of 900 MHz microcellular mobile radio channel were derived based on a computer simulation of microcellular multipath environment. The interest was threefold: first, to develop an improved understanding of the propagation phenomenon; secondly, to develop a simulation model using ray tracing technique from the geometrical optics and thirdly to investigate path loss characteristics and fading statistics of the channel. The development of simulation model is based on a trade-off between simplicity (to save simulation time) and accuracy because certain essential assumptions (such as flat reflecting surfaces) give rise to inaccuracies, therefore, it seems reasonable to trade it with simplicity in simulation model.

We find the propagation in the microcellular environment to be of a multipath nature. These multipath comprise of reflected and diffracted components besides a line of sight if the cell structure is designed to have one, which is the case for some of the proposed microcells structures. The reflected multipath are due to ground and buildings on each side of the road. The major source of diffracted components are the edges and the corners of the buildings. The multiray modelling approach is used to model this channel. However, the dilemma is to decide how many and which rays must be included in the model. We find

that the number of rays should be related to the purpose of channel modelling. In literature, [7], a four ray model is also considered sufficient for propagation loss estimation. To model fast fading statistics more rays would be needed and both reflected and diffracted paths must be included. However one cannot definitely determine the number of rays because these multipath are a result of a particular physical geometry which will change from one to another environment. Nevertheless, if environment could be broadly categorized, a procedure may be laid down to suggest an appropriate model. For example, in typical grid structured North American downtowns, building reflected paths are essentially considered. Similarly, paths diffracted from the edges of the buildings are recommended for inclusions, whereas, the corner diffracted paths may be neglected because the buildings are very tall when compared to transmitter and receiver antenna heights. The total number of diffracted paths would be dependent upon the number of buildings standing alone along the receiver path.

In the simulation model, it is assumed that received signal is a phasor summation of the multipath. Multipath due to different physical lengths contain different power and phase and when added vectorially give rise to fading in the received signal. The reflected multipath are multiplied by the complex reflection coefficient R which is a function of angle of incidence, reflecting material and the polarization of the incident wave. Because the real propagation environment is too complex to be modelled by neat analytical expressions for vertical and horizontal polarization for reflection coefficient therefore one resorts to approximate values, but, how good are the approximation is only determined through comparing the simulation results to the measured ones. However, if the transmit and the receive antennas height difference is reduced to zero, the expression for reflection coefficient shall be applicable exactly which of course is very unrealistic. The choice of the material constants for the reflecting surface are also to be assumed because the material is hardly same all over the surface. Two materials are predominantly encountered, i.e., a combination of concrete and glass. The choice of dielectric constant and conductivity is that of glass. The selection is based on the observation that generally more than fifty percent of the surface area is covered by glass, while some other buildings have all glass exterior. Another assumption which one is forced to make is that reflecting surfaces are flat to permit the use of ray tracing. It is quite unrealistic for the outdoor environment and there

must be some way to compensate for this. The approach which we use is that the surface roughness effects be included through assuming that the roughness is randomly distributed and further assuming it to be uniform for individual surfaces. But noting that the summation effect of all of multiple reflected paths would be such that it may be considered due to central limit theorem as the normally distributed variables with zero mean and some variance. Where variance is chosen arbitrarily through observation and comparison with the experiments. We neglected this effect in median path loss simulation results but included it in fast fading results.

The diffracted rays are multiplied by the magnitude of the diffraction coefficients D which is required to be between zero and one. We noted that the expression for diffraction coefficient is not valid at and in the vicinity of the reflection and shadow boundaries. We assume a constant value for D for every diffracted path, having observed this that except around the above mentioned boundaries the value of D does not change much. The choice of D to be a constant is also motivated by the fact that it is desirable to reduce the computation and consequently the simulation time. Although geometrical theory of diffraction does not provide the exact expression for the diffraction coefficient, it does provide the basic mechanism and the principle to help determine the diffracted rays, their direction as well as the length which are essential ingredient of this simulation.

The simulated path loss is compared to some experimental results available in literature. We observed that due to line of sight propagation a plane earth propagation model provides a first approximation to path loss. This model is attributed to have a d^2 dependence on distance up to a break-point which is a function of antenna heights and wavelength of the transmitted signal. After this break-point the distance dependence becomes proportional to d^4 . Because the area of interest was small, only d^2 distance dependence was observable. Effect of building reflected and diffracted multipath is found to give rise to microvariations or fast fading in the received signal which distorts the plane earth propagation results. The simulation does not reproduce the experimental results exactly because of the fact that a number of assumptions were unavoidable in the simulation. However, we find simulation to be able provide the gross features of the propagation environment, which are found to be free space propagation with occasional fades of 10 db's. Other aspects of the simulation such as effect of transmitter location, height and carrier frequency

on received path loss were also investigated. We find that received path loss is a function of transmitter location. Moving the transmitter across than along the street had more effect on the path loss distribution pattern in the area of interest. Increasing the transmit antenna height, while keeping it below the average building heights, provides small improvement in received signal power at greater distances. Close to the transmitter the effect was that of changing the interference pattern of the field. Increasing the frequency resulted into a further loss and a more uniform distribution of propagation pattern.

While investigating fading statistics the reference model is the one which separates envelope into a slow and fast fading component. The fast fading component is considered Rayleigh or Rician distributed depending upon the absence or presence of the line of sight component respectively. The slow fading component is assumed to be lognormally distributed. The former is based on a small geographical area and later on a large area. After curvefitting we observed that the small area statistics are Rician distributed as has been reported in literature, whereas large area statistics are lognormally distributed. Certain experiments used a Nakagami model for fast fading and reported a close fit. However, to the best of our knowledge nobody compared the Rice and Nakagami distribution for the same experimental data. Here we compared the two models and using the same simulation data and curvefitting criterion to obtain a reasonable comparison. Results show that the Nakagami model fit as well as the Rician. Another observation worth mentioning is that Rayleigh distribution has comparatively improved fit on that data which is far away from the transmitter than the closer ones. It is reasonable because as receiver moves away from the transmitter, the line of sight component starts to diminish due to several reason (e.g., occasional loss of LOS) and carries just a slightly larger power than the multipath which then amounts to the assumption of equal power of the multipath.

We find it in place to remark on the Rice and Nakagami model for fast fading in the light of the theoretical development of Section 2.2. Fading is due to summation of vectors which are assumed random in their magnitude and phase. Assuming summation of an infinitely large number of random vectors, a gaussian process is assumed which results into Rayleigh or Rician model. On the other hand, considering a sum of finite number of vectors on a general basis, we obtain an approximate solution for the distribution of the magnitude of the resultant vector in terms of the Nakagami model. Interestingly enough, both

models are approximate and fit good on simulated data and both degenerate to Rayleigh under certain conditions. The question arises which one is better and on what basis? There is no unique answer to it as far as we know. Mathematically, Rayleigh is the simplest but inaccurate. Among Rice and Nakagami, which are more accurate, latter may be preferred due to relatively simple mathematical structure (e.g. absence of a Bessel function).

Lastly we observe that near field propagation may be treated in a similar manner as a far field. In other words, the received signal may be treated as a random vector problem and characterized statistically by a Nakagami distribution.

5.2 Applications and Further Work

In design of communication systems propagation characteristics govern the design of the other system components, for example, the modulation techniques, coding scheme, available bandwidth, and power requirement. Therefore, first step towards the system design include the physical measurements of the propagation characteristics which is a very costly and time consuming job. The simulation model of this thesis which is developed for urban multipath environment may be used to acquire the necessary information about the channel. Once an estimate of the channel characteristics is available, the whole system design may be simulated and analyzed before any field trial is planned. The work of this thesis may as well form the basis for a simulation package which could help determine the optimum cell lay-outs and appropriate base station antenna locations.

Further work may include:

- Extending the simulation program to study the propagation mechanism and characteristics of non line of sight propagation in microcellular systems.
- More information can be extracted for Nakagami fading channel model. The relationship of m -parameter with distance has been investigated here, the dependance on other propagation factors, such as frequency, may be investigated.
- Effect of different reflecting material on path loss and fading statistics may be carried out to determine their impact on some quantitative basis.
- The simulation model may also be used with few modifications to help simulate urban propagation for conventional mobile cellular systems.

- Another interesting extension might be to modify the simulation model of this thesis for wide band channel characterization and to investigate the delay spread characteristics of the channel.
- Because, the Rice and Nakagami models provide a very close fit for fast fading statistics of microcellular channel, it shall be very useful to analytically investigate those conditions under which the two distributions are equivalent.
- To look into those criteria which may be used for a more meaningful and physically significant comparison of the Rice and Nakagami fading model.

Linear Time-Invariant/Variant Channels, A System-Function Approach

In the following subsections we discuss linear time-invariant and linear time-variant channels separately. However, it will be shown that once time-variant channel is characterized the time-invariant channel should in fact be a special case of time-variant channels.

A.1 Linear Time-Invariant Channels

Consider the two terminal pair linear network as shown in Figure 3.1. Let us assume it be time-invariant; i.e., having parameters which do not vary in time, commonly called a fixed parameter network. Thus, if an input $x_i(t)$ produces an output $x_o(t)$, then an input $x_i(t - \tau)$ produces a corresponding output $x_o(t - \tau)$. If the system under consideration is linear and the input $x_i(t)$ is of the form $e^{j\omega t}$, then the output also has the same form; i.e., if the input is sinusoidal, then the output is also sinusoidal but may be shifted in phase. Thus, a linear system cannot generate new frequencies. At most only the input frequencies are present at the output.

Now, let $X_i(j\omega)$ and $X_o(j\omega)$ represent the Fourier Transform of the input and output of the linear time-invariant network respectively. Then ratio of the output to the input, here denoted by $H(j\omega)$, is termed the frequency response of the network and may be represented by,

$$H(j\omega) = \frac{X_o(j\omega)}{X_i(j\omega)}. \quad (\text{A.1})$$

Since the frequency response is determined by the ratios of quantities expressed in frequency domain, it is a frequency domain representation of the linear time-invariant network above.

The output obtained when input to the network is a unit impulse is denoted by the weighting function or impulse response of the network. Thus, if $x_i(t) = \delta(t)$, then $X_i(j\omega) = 1$, and $H(j\omega) = X_o(j\omega)$, we obtain,

$$x_o(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(j\omega) e^{j\omega t} d\omega. \quad (\text{A.2})$$

The output obtained under these conditions is denoted by $h(t)$, therefore,

$$h(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(j\omega) e^{j\omega t} d\omega, \quad (\text{A.3})$$

and conversely,

$$H(j\omega) = \int_{-\infty}^{\infty} h(t) e^{-j\omega t} dt. \quad (\text{A.4})$$

It is also possible to obtain output solution by time domain calculations using convolution technique once impulse response is known. In this case,

$$x_o(t) = \int_{-\infty}^t x_i(\tau) h(t - \tau) d\tau. \quad (\text{A.5})$$

The relationship indicates that the input may be considered as an infinite number of impulses, each weighted as a function of $x_i(\tau)$. The output of the network due to each of these impulses is simply the impulse response of the network, $h(t)$. Making use of the superposition principle for linear systems, the output at any given time t may be considered to be the sum of all past impulses. Thus, an impulse $x_i(t) \delta(t - \tau)$ yield an output of $x_i(t) h(t - \tau)$ and we are led to the result in equation (A.5).

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A.2 Linear Time-Variant Channels

Linearity of channels helps to use superposition technique in characterizing linear time-variant (LTV) channels. The most commonly used method of characterizing LTV networks are the impulse response $h(t, \tau)$ and corresponding system function $H(j\omega, t)$ as introduced by Zadeh, [44]. The quantity $h(t, \tau)$ is interpreted as the response of the network at time t to an impulse applied at time τ . The response of the network to any signal $x(t)$ can then be found by superimposing the responses of the system to a weighted series of impulses whose sum is equal to $x(t)$. Mathematically,

$$x(t) = \int_{-\infty}^{\infty} x(\tau) \delta(t - \tau) d\tau. \quad (\text{A.6})$$

Using the linearity property, we can write,

$$y(t) = \int_{-\infty}^{\infty} x(\tau) h(t, \tau) d\tau. \quad (\text{A.7})$$

Assuming causality, i.e., $h(t, \tau) = 0$ for $t < \tau$, we may rewrite,

$$y(t) = \int_{-\infty}^{\infty} x(\tau) h(t, \tau) d\tau. \quad (\text{A.8})$$

The system function is defined by Zadeh [45] as,

$$H(j\omega, t) = \int_{-\infty}^{\infty} h(t, \tau) e^{-j\omega(t-\tau)} d\tau = \int_{-\infty}^{\infty} h(t, \tau) e^{j\omega\tau} d\tau e^{-j\omega t}. \quad (\text{A.9})$$

Therefore,

$$\begin{aligned} y(t) &= \frac{1}{2\pi} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} X(j\omega) e^{j\omega\tau} d\omega h(t, \tau) d\tau \\ &= \frac{1}{2\pi} \int_{-\infty}^{\infty} X(j\omega) \int_{-\infty}^{\infty} h(t, \tau) e^{-j\omega(t-\tau)} d\tau e^{j\omega t} d\omega \end{aligned} \quad (\text{A.10})$$

$$y(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} X(j\omega) H(j\omega, t) e^{j\omega t} d\omega, \quad (\text{A.11})$$

Equation (A.11) completely characterizes the output of the filter if system function can be determined. It may be noted that for non-varying t the system function becomes the transfer function of the time-invariant case.

System function may be given a useful operational definition as the output to an exponential input;

$$H(j\omega, t) = \left. \frac{y(t)}{x(t)} \right|_{x(t) = e^{j\omega t}, (-\infty < t < \infty)} \quad (\text{A.12})$$

Impulse response of the LTI and LTV low pass filters are plotted in Figures A.1 and A.2, respectively. Figure A.1 shows that whenever impulse is applied response remains the same but shifted in time. On the other hand, Figure A.2 shows one of the many possible ways the response of the LTV low pass filter may change with time when an impulse is applied at different instants of time. This in fact shows the sensitivity of the response on the instant of application of impulse.

Equation (A.11) helps characterize the LTV filter output and consequently the output of the channel which has been modelled by the LTV filter. But in typical communication problem, input and channel characteristics, and as a result the output of the channel are not deterministic. What we know are the statistics of the input signal and the channel, from which we require to determine the statistics of the channel output. This is not an easy job, because the statistics may be those of stationary/non-stationary and Gaussian/non-Gaussian random processes. In case of non-Gaussian statistics there is not enough mathematics available yet to find the output statistics. However, the case is different if input and channel statistics are stationary and Gaussian process. Moreover, if the statistics are non-stationary then we are again helpless even if the processes are Gaussian. In case of non-stationary statistics we rely solely on experimental results of the channel characteristics and this varies from case to case.

A stationary Gaussian process is completely specified by its autocorrelation function. Fortunately, there exists a convenient relationship between the input and the output of correlation function of the Gaussian, statistically stationary and time-variant network as shown by Zadeh, [45], [46]. Defining system correlation function as $R_H(j\omega; \tau)$,

FIGURE A.1

Impulse response of a time-invariant low pass filter

FIGURE A.2

One possible impulse response of a time-variant low pass filter

$$R_H(j\omega; t) = \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T H(j\omega; t) H(-j\omega; (t - \tau)) d\tau. \quad (\text{A.13})$$

The autocorrelation function is given by,

$$R_Y(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} R_H(j\omega; \tau) S_X(j\omega) e^{j\omega\tau} d\omega. \quad (\text{A.14})$$

Where,

$$S_X(j\omega) = \int_{-\infty}^{\infty} R_X(\tau) e^{-j\omega\tau} d\tau, \quad (\text{A.15})$$

is the input power spectral density, and,

$$S_Y(j\omega) = \int_{-\infty}^{\infty} R_Y(\tau) e^{-j\omega\tau} d\tau. \quad (\text{A.16})$$

is the output power spectral density.

Summarizing at this point, we recapitulate that channels may be modelled as linear time-variant filters/networks, but this model is not completely satisfactory. It would be necessary in general that characteristics of the channel be measured under operating conditions using physiogeometric approach. While the system correlation function suffices for the Gaussian channels, experimental demonstrations should be used to check whether any given channel can be described by the Gaussian statistics. Due to expense and time factors, it has been always desirable to seek some techniques to be able to estimate channel parameters and characterize it in laboratory. A computer simulation is the solution if the propagation mechanism could be modelled. One popular approach is the use of geometrical optics.

Probability of Error for BPSK Signals over Nakagami Fading Channel

We know that for BPSK signal transmitted over AWGN channel, the probability of error which is a function of required signal to noise ratio (SNR) per bit, is related as,

$$P_e(\gamma_b) = \frac{1}{2} \operatorname{erfc}(\sqrt{\gamma_b}). \quad (\text{B.1})$$

where,

$$\gamma_b = \alpha^2 (E_b/N_o). \quad (\text{B.2})$$

Here E_b is energy per bit, N_o is noise power and α is channel attenuation. If α is randomly varying, consequently, γ_b is also randomly varying and needs to be characterized statistically. Therefore, we must average $P_e(\gamma_b)$ over the probability density function of γ_b . That is, we must evaluate the integral,

$$P_{a,e} = \int_0^{\infty} P_e(\gamma_b) p(\gamma_b) d\gamma_b. \quad (\text{B.3})$$

To determine $p(\gamma_b)$, we first observe that α is Nakagami distributed, therefore, statistics of γ_b are obtained by the transformation of statistics of α through the transformation relationship given in (B.2) as follows,

$$p(\gamma_b) = p(\alpha) \Big|_{\alpha=\gamma_b} \frac{d\alpha}{d\gamma_b}. \quad (\text{B.4})$$

We know that,

$$\frac{d\alpha}{d\gamma_b} = \frac{N_o}{2E_b\alpha}, \quad (\text{B.5})$$

therefore,

$$p(\gamma_b) = \frac{N_o}{2E_b\alpha} p(\alpha) \Big|_{\alpha=\gamma_b}. \quad (\text{B.6})$$

As α is Nakagami distributed,

$$p(\gamma_b) = \frac{N_o}{2E_b\alpha} \cdot \left(\frac{2m^m \alpha^{2m-1}}{\Gamma(m) \Omega^m} \cdot e^{-\left(\frac{m}{\Omega}\right)\alpha^2} \Big|_{\alpha=\gamma_b} \right), \quad (\text{B.7})$$

$$p(\gamma_b) = \frac{m^m}{\Gamma(m) \Omega^m} \cdot \left(\frac{N_o}{E_b}\right)^m \cdot \gamma_b^{m-1} \cdot e^{-\left(\frac{m}{\Omega}\right) \cdot \frac{N_o}{E_b} \cdot \gamma_b}. \quad (\text{B.8})$$

We may rewrite the above equation as,

$$p(\gamma_b) = A \cdot \gamma_b^{m-1} \cdot e^{-B \gamma_b}, \quad (\text{B.9})$$

where $A = \frac{m^m}{\Gamma(m) \Omega^m} \left(\frac{N_o}{E_b}\right)^m$ and $B = \left(\frac{m}{\Omega}\right) \left(\frac{N_o}{E_b}\right)$.

Putting the expression in (B.9) into (B.3), we have,

$$P_{a,e} = \int_0^{\infty} \frac{A}{2} \operatorname{erfc}(\sqrt{\gamma_b}) \gamma_b^{m-1} e^{-B \gamma_b} d\gamma_b. \quad (\text{B.10})$$

This expression can be evaluated numerically for different values of m .

Near Field Radio Propagation

One scenario which may arise in microcellular systems but is rare in conventional cellular systems is the one that receiver may become quite close to the transmitter in the so-called near field region. What is the nature of this near field propagation? The answer to this is not yet clear. The relations which describe the electromagnetic field elsewhere are not valid in the near field region. They are even approximations for the far field. The approximations are due to certain assumption which are made to reach at the analytical results. In this Appendix, we develop these expressions, explicitly mentioning the assumptions and the reasons behind them. A discussion is provided in Chapter Four to dwell upon plausible interpretation of the near field in microcellular environment.

C.1 Electromagnetic Radiation

Electromagnetic theory deals with electric/magnetic fields and their sources. Fields are related to the sources through certain relationships governed by some physical laws. Source of electric and magnetic fields are electrical charges and currents. Both fields are generally classified as time-varying or non-time-varying. Superposition theorem is used to help find field due to some source. Static electric field intensity at a point in space due to a line of electric charge is found by considering the line charge as a continuum of point

charges and superposing the field due to each constituent point charge using Coulomb's law. Similarly, the magnetic field intensity due to a current surface is obtained by approximating the source into infinitesimally small current elements and using superposition. In case of time-varying sources, the resultant fields are found to be propagating and both electric and magnetic fields are coupled through Maxwell's equations. Therefore, determination of one field helps finding the other field through some integration or differentiation of the first one.

Let us assume a hypothetical, infinitesimally small current element of length, d , as a source which is oriented along z -axis where positive sense of current, $I_0 \cos \omega t$, is along positive z -axis. The resultant components of fields, in phasor form, are written in spherical coordinates as follows [47], [48]:

$$H_{\phi s} = -\frac{I_0 d \beta_0^2}{4\pi} \sin \theta \left[\frac{1}{j\beta_0 r} + \frac{1}{(j\beta_0 r)^2} \right] e^{-j\beta_0 r}, \quad (\text{C.1})$$

$$E_{rs} = -\frac{I_0 d \beta_0^2}{2\pi} \sqrt{\frac{\mu}{\epsilon}} \cos \theta \left[\frac{1}{(j\beta_0 r)^2} + \frac{1}{(j\beta_0 r)^3} \right] e^{-j\beta_0 r}, \quad (\text{C.2})$$

$$E_{\theta s} = -\frac{I_0 d \beta_0^2}{4\pi} \sqrt{\frac{\mu}{\epsilon}} \sin \theta \left[\frac{1}{j\beta_0 r} + \frac{1}{(j\beta_0 r)^2} + \frac{1}{(j\beta_0 r)^3} \right] e^{-j\beta_0 r}, \quad (\text{C.3})$$

where $\beta_0 = (2\pi)/\lambda$, r is the distance between source and the observation point, μ and ϵ are permeability and dielectric constants of the propagation medium. In [47], for certain values of parameters H_{ϕ} is plotted against r . For $\lambda \leq r \leq 2\lambda$, it is shown that curve is non-sinusoidal, but for $101\lambda \leq r \leq 102\lambda$ essentially a sinusoid is obtained. It may further be observed that in (C.1)-(C.3) the terms involving $1/r$, $1/r^2$ and $1/r^3$ are present such that at distances ten or more wavelengths from the oscillating current element, all terms except the inverse-distance, $1/r$, may be neglected and distant or far-field becomes:

$$E_{rs} = 0, \quad (\text{C.4})$$

$$H_{\phi s} = j \frac{I_0 d \beta_0}{4\pi r} \sin \theta e^{-j\beta_0 r}, \quad (\text{C.5})$$

$$E_{\theta s} = j \frac{I_0 d \beta_0}{4\pi r} \sqrt{\frac{\mu}{\epsilon}} \sin \theta e^{-j\beta_0 r}, \quad (\text{C.6})$$

$$E_{\theta z} = \sqrt{\frac{\mu}{\epsilon}} H_{\phi z}. \quad (\text{C.7})$$

From (C.7) it may be observed that relationship between $E_{\theta z}$ and $H_{\phi z}$ is that of between electric and magnetic fields of the uniform plane waves.

C.2 A Physical Dipole

The function of the point source electric dipole is to act as an elementary field which can be used to evaluate that of any given current. If we know the current distribution of the source only, the integration of the point-source solution is required to calculate the field. For a very short dipole antenna only current distribution is required for the purpose of calculating the principal features of the radiation.

Consider a conductor forming the arms of physical dipole as shown in Figure C.1. Current on the surface of the cylinder extending from the centres to the end may be denoted by surface current density K_z . We want to find the field at point P distant r from the centre of the dipole using spherical coordinates. To calculate electromagnetic field at this point, it will be sufficient to calculate either electric or magnetic field. The other field may be found by differentiation from Maxwell's equations. We first calculate the effect at point P of the field caused by a small current element as shown in Figure C.1. Using primes to distinguish the position of the current element from that of observation point. The magnetic field produced at point P shall be given as in (C.1):

$$dH_{\phi z} = \frac{K(z') r_0 d\phi' dz' \beta_0^2}{4\pi} \sin\theta'' \left[\frac{1}{j\beta_0 r''} - \frac{1}{(j\beta_0 r'')^2} \right] e^{-j\beta_0 r''}. \quad (\text{C.8})$$

We can note that for a given point P , both θ'' and r'' are functions of z' and ϕ' , the location coordinates of the current element. From Figure C.1, it is clear that

$$|r - r''| \leq \sqrt{(l/2)^2 + r_0^2}. \quad (\text{C.9})$$

Considering point P being far away from the current element in comparison to its dimensions i.e.,

$$r \gg \sqrt{l^2 + 4r_0^2}. \quad (\text{C.10})$$

Or equivalently, we may assume r'' becomes parallel to r . Consequently, it can be validly assumed that $\theta'' \approx \theta$ or $\sin\theta'' \approx \sin\theta$. We further assume that largest dimension of the physical radiator is much smaller than a wavelength or.

$$\sqrt{l^2 + 4r_0^2} \ll \frac{\lambda}{2\pi}. \quad (\text{C.11})$$

The difference between r and r'' , measured in wavelengths determine the relative phase of arrival at point P of the radiations from differential current element on the cylinder. The assumption of r and r'' being parallel is most important with respect to term $\beta_0 r''$ in the exponential. It amounts to assuming there is negligible phase difference at the point of observation between the field contributions from current elements on the cylinder that are separated by the maximum possible distance. Under these assumptions:

$$\beta_0 r'' \approx \frac{2\pi r''}{\lambda} \approx \frac{2\pi r}{\lambda}. \quad (\text{C.12})$$

The (C.8) may now be rewritten to account for $H_{\phi s}$ due to all contributing point source current elements:

$$H_{\phi s} = \frac{\beta_0^2}{4\pi} \sin\theta \left[\frac{1}{j\beta_0 r} - \frac{1}{(j\beta_0 r)^2} \right] e^{-j\beta_0 r} \int_0^{\frac{l}{2}} \int_{-\frac{l}{2}}^{\frac{l}{2}} K(z') r_0 d\phi' dz'. \quad (\text{C.13})$$

The double integral term in above equation amounts to $I_0 d$ and fields due to cylinder ends will tend to cancel each other, [46], therefore, (C.1) shall give r' : entire contribution of the physical dipole to the ϕ component of the magnetic field. Equation (C.1) may now be written as,

$$H_{\phi s} \approx -\frac{I_0 d \beta_0^2}{4\pi} \sin\theta \left[\frac{1}{j\beta_0 r} - \frac{1}{(j\beta_0 r)^2} \right] e^{-j\beta_0 r}. \quad (\text{C.14})$$

We can see that (C.14) is similar to (C.1) except that equality has been changed with near equality due to assumptions which are made here. Thus E_{rs} and $E_{\theta s}$ shall be given by (C.2) and (C.3) with equalities changed to near equalities. As mentioned before, the far-field of the physical dipole antenna can be considered as a uniform plane wave with only $E_{\theta s}$ and $H_{\phi s}$ components present, propagating radially outwards and approximately given

by (C.5) and (C.6). In Chapter Four we use this development when the near field propagation is discussed.

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